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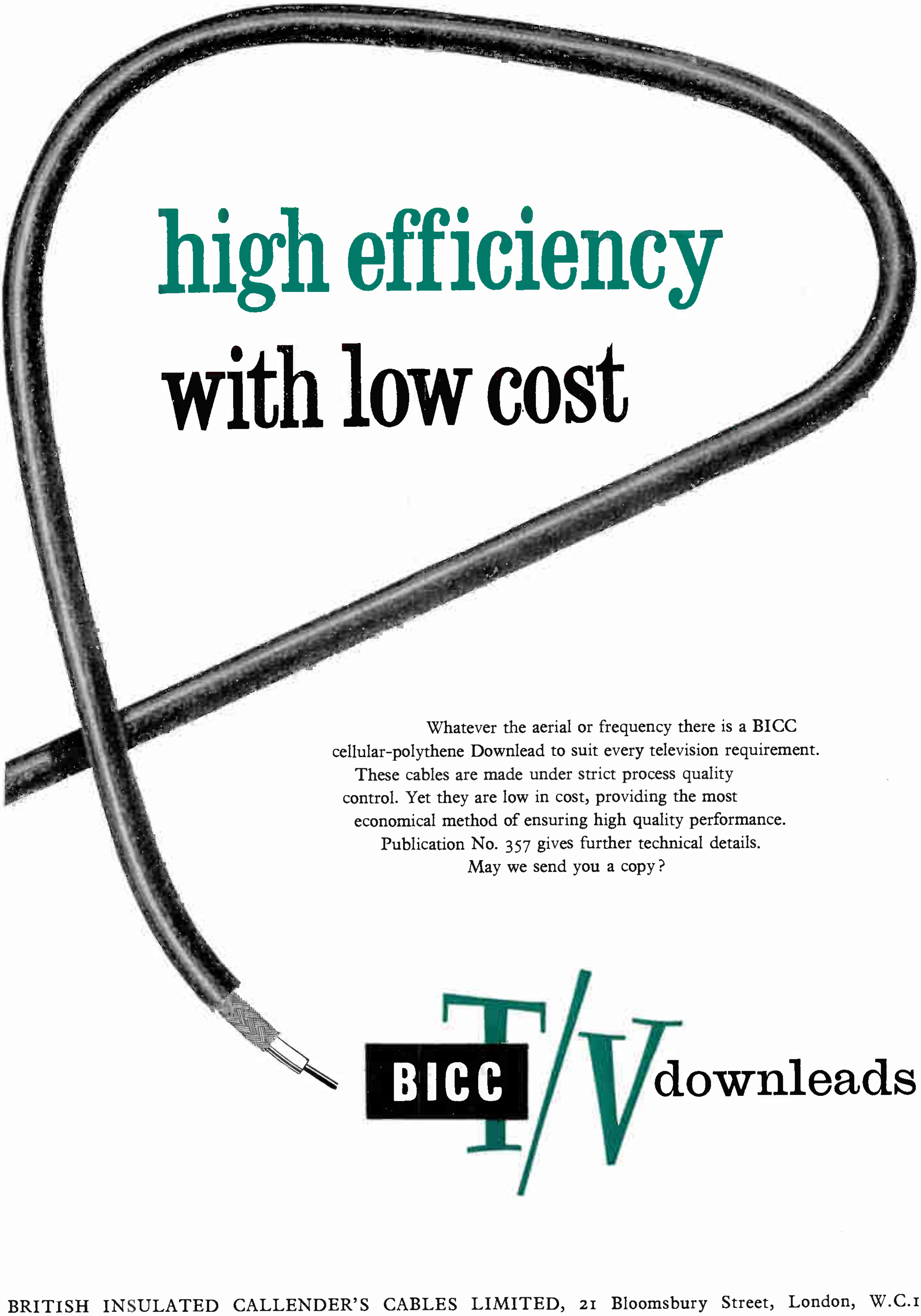
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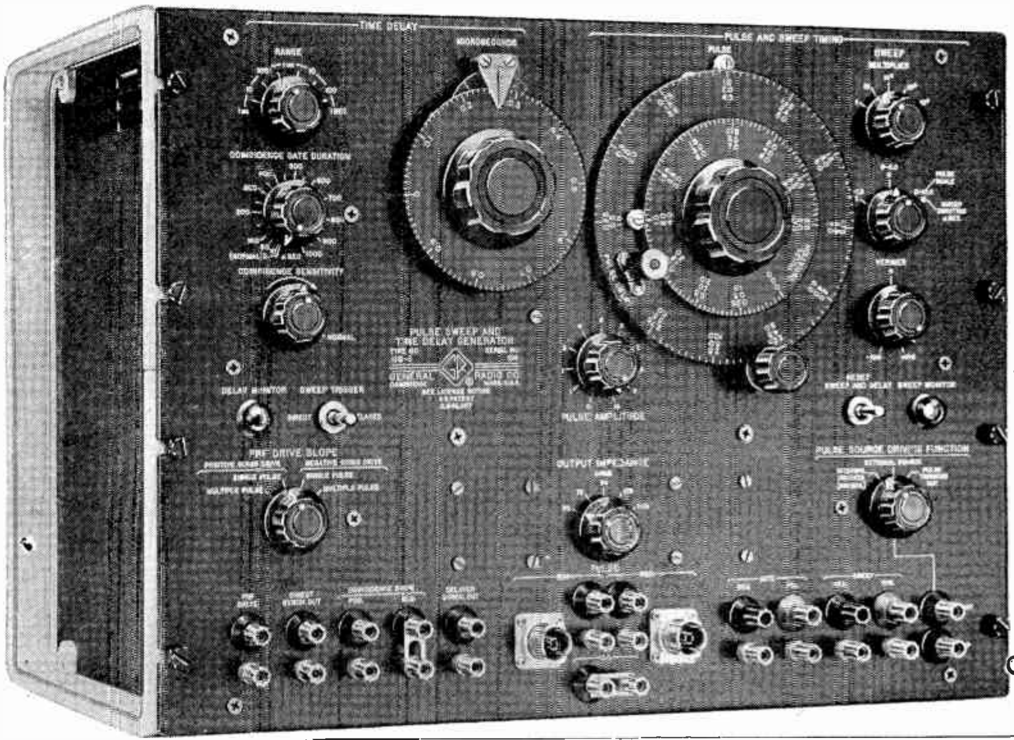

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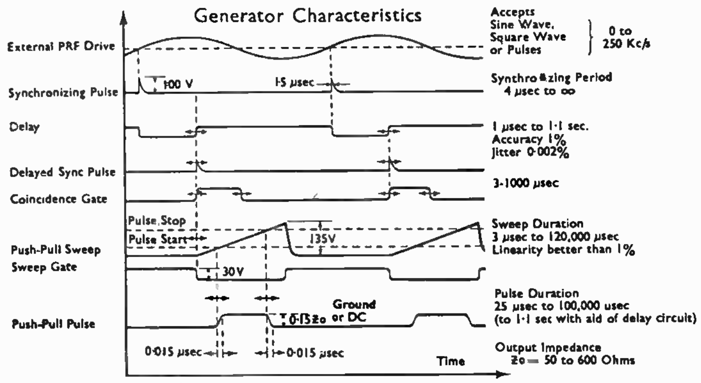

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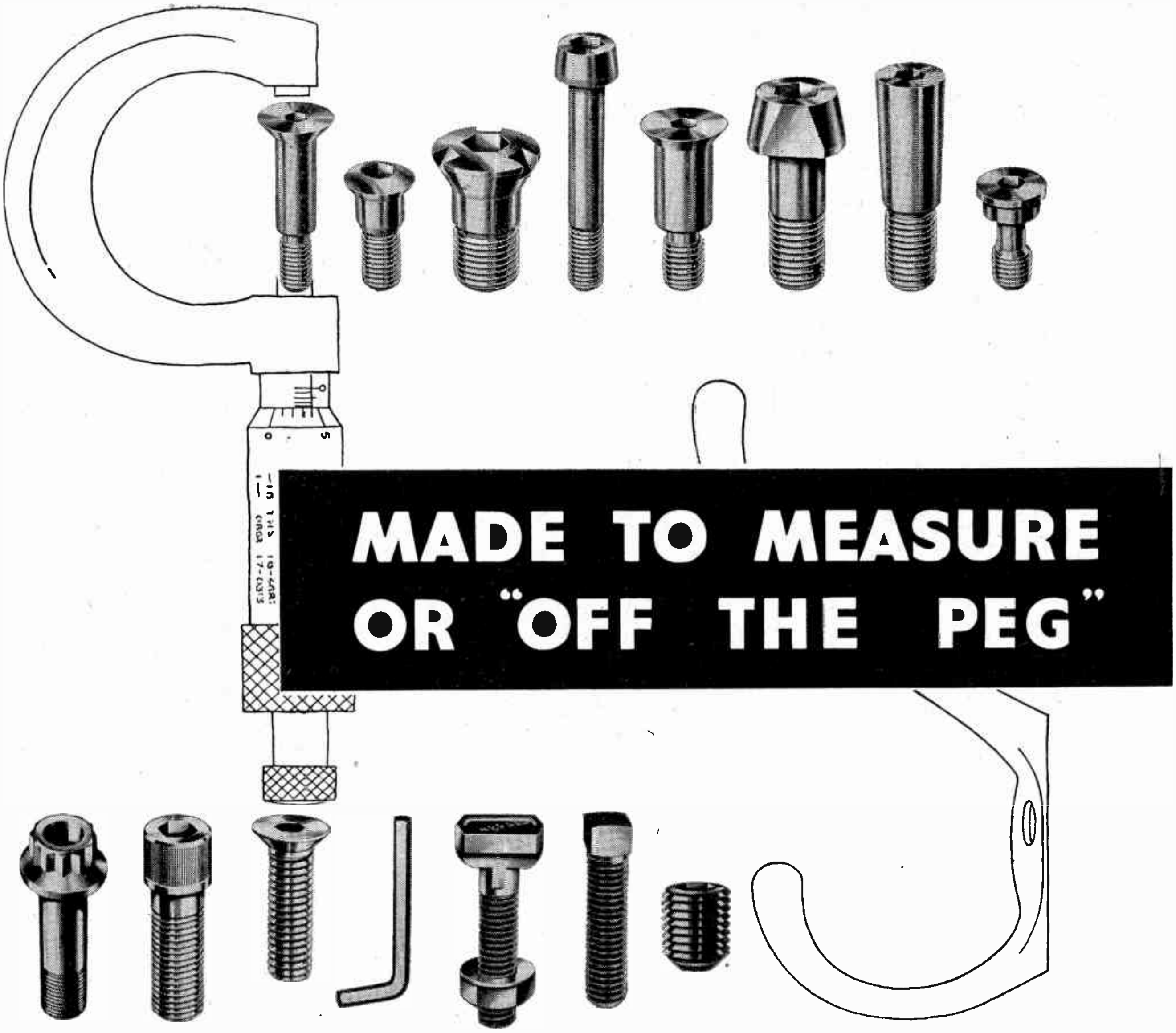
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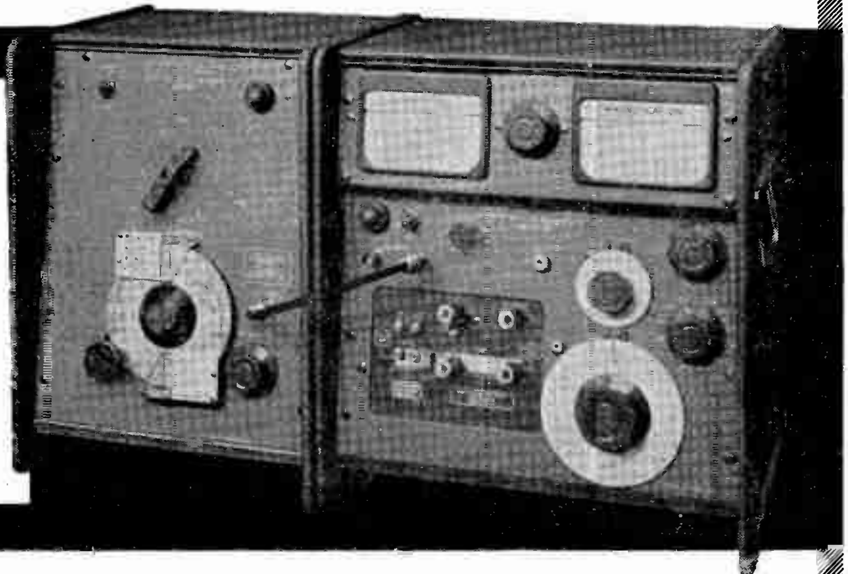

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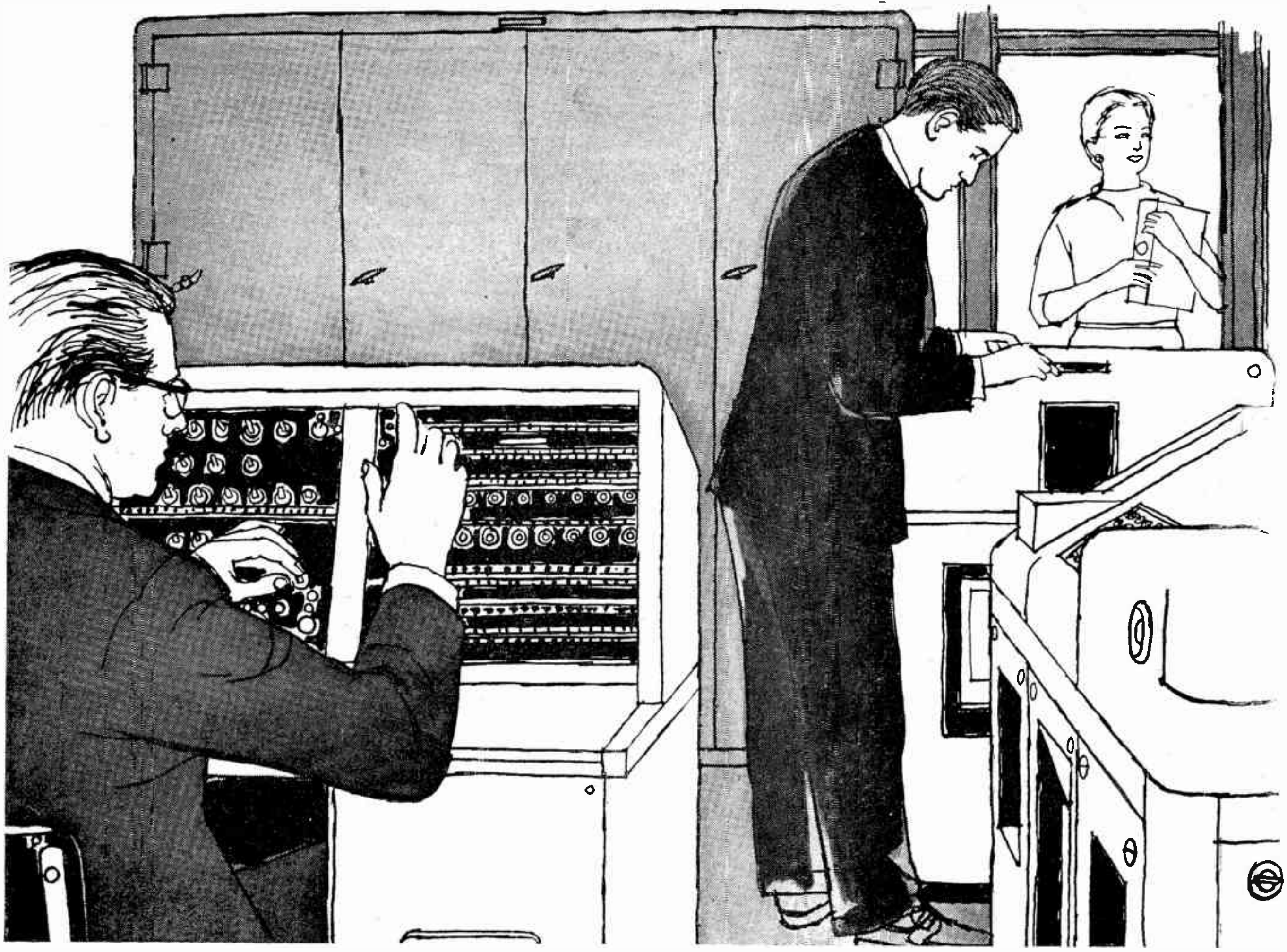

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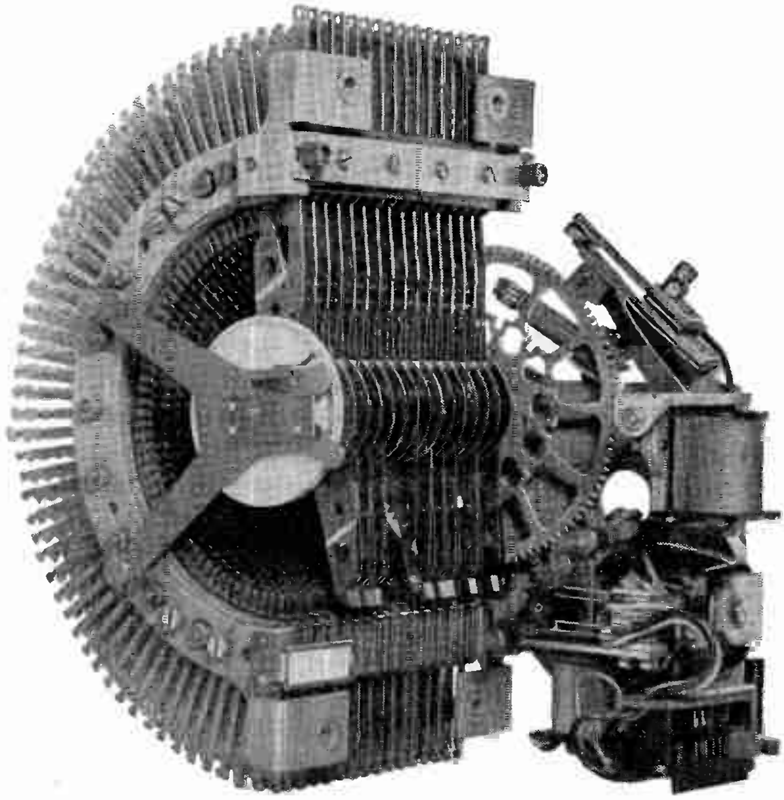

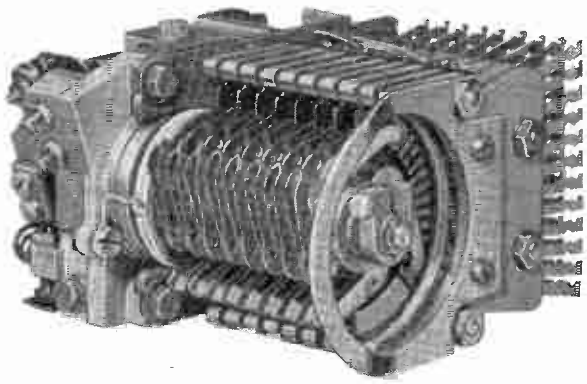

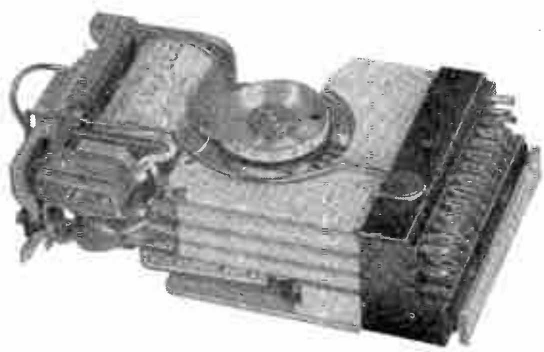

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Fields

IT is well known that field theory is considered difficult and there is no doubt that many students find it quite hard to grasp. We fancy that, in part at least, this may be due to the fact that the word 'field' is commonly used in different senses and that its meanings are not always adequately defined.

We have been reading an article in *Electrical Review* (10th April 1959, p. 672) by A. T. Craven, advocating a new method of teaching electrical engineering. He defines a field by saying that "The field is simply the region at any point of which the physical entity under consideration may be detected". He goes on to emphasize the everyday meaning of the word by an example: "a field of grass is simply a region at any point of which we may detect grass".

There is little doubt that this was the original meaning of field. Now, however, we attribute to it magnitude and, often, direction as well. We speak of the strength and direction of an electric field, for example. According to the definition given above, an electric field is a region in which a charged particle experiences a force. How can this region have magnitude and direction? It is, of course, the force on the particle which has magnitude and direction.

We do conventionally call this force itself a field and we have only to write that the magnitude and direction of an electric field varies from point to point in the field to realize that this word 'field' is being used in two distinct senses. The 'regional' definition is quite often given but comparatively rarely is the second more sophisticated meaning explained clearly. In spite of his stress on this primary meaning, Mr. Craven is not himself wholly consistent, for he refers to a magnetic field changing in magnitude, when he is obviously not meaning that a region of magnetic effects is changing its size.

We do feel that users of the word field would be well advised to think about what they mean by it, and to endeavour to employ it in a more consistent manner. Because it has two meanings, its careless use can lead to much misunderstanding.

Transistorized Pulse Amplifier

DESIGN OF A HIGH-SPEED TYPE

By J. N. Barry, M.Sc. and D. M. Leakey, Ph.D., B.Sc. (Eng.)., D.I.C., A.C.G.I., Grad. I.E.E.

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

SUMMARY. *The article describes the design of a transistorized pulse amplifier for use with diode logic circuits. With an effective overall gain exceeding 10, the total rise- and fall-times obtainable are each better than 0·05 microsecond, although there is also a short delay between the input and output pulses. The circuit employs two high-frequency drift transistors of a type specially developed for switching applications. Some of the problems associated with the use of such transistors are discussed briefly.*

During some recent work concerned with the design of diode logic circuitry the need arose for a transistorized interconnecting pulse amplifier with the following specification:

(i) To be driven from a 5-mA gate.
(ii) To be capable of driving ten 5-mA gates.
(iii) To be capable of maximum rise- and fall-times of better than 0·05 μsec.
(iv) To operate with 0·8-μsec pulses with a minimum spacing of 0·2 μsec.
(v) To operate with positive-going pulses of about 10 to 15 V amplitude.

It was decided to avoid if possible the use of pulse transformers owing to size and difficulty of manufacture. Since current and some voltage gain were both required and also no reversal of pulse polarity, a two-stage circuit had to be employed. Either two cascaded earthed-emitter stages or a combination of an earthed-collector and earthed-base stage could have been employed. The former arrangement was finally chosen as being more practicable with the transistors available.

The question of the necessary performance of the transistors will first be dealt with, followed by a description of the implied design requirements. The circuit details will then be discussed, concluding with a circuit of the complete amplifier.

Throughout the article reference is made only to transistors of the p-n-p type, or to semiconductor diodes having n-type base material.

List of Symbols

α, α_n	= Common-base current-gain factor, (complex).
α_0	= Low-frequency value of α.
α_I	= Value of α for inverse operation; i.e., with the functions of emitter and collector interchanged.
β	= Small-signal value of common-emitter current-gain factor.
β_l	= Large-signal value of common-emitter current-gain factor.
ΔE	= Electric field component.
f_α	= Small-signal common-base cut-off frequency.
f_1	= Frequency at which $\lvert\beta\rvert$ = unity.
I_B	= D.C. base current.
I_B'	= Transient base 'off'-to-'on' driving current.
I_B''	= Transient base 'on'-to-'off' reverse driving current.
I_c	= D.C. collector current in 'on' condition.
I_{c0}	= Collector–base leakage current.
k	= Boltzmann's constant.
ϕ	= Angular phase shift.
ρ_n	= n-type resistivity.
T	= Absolute temperature.
T_F	= Fall-time (not including storage time).
T_R	= Rise-time.
T_S	= Hole storage time.
V_{be}	= D.C. base-emitter forward voltage.
V_c	= D.C. collector voltage.
W_d	= Width of collector depletion layer.
ω_α, ω_n	= Common-base angular cut-off frequency.
ω_1	= Angular frequency at which $\lvert\beta\rvert$ = unity.
ω_I	= Common-base angular cut-off frequency for inverse operation.

Required Transistor Performance

Four distinct conditions of a non-linear, earthed-emitter, pulse amplifying stage can be defined. For each case there is the corresponding base driving condition. These are:

(a) The necessary reverse bias to hold the transistor effectively cut-off, ($I_c \approx 0$).
(b) The necessary base current to hold the transistor in the required 'on' state.
(c) The necessary transient 'on' drive current to switch the transistor from the 'off' to the 'on' condition in the required time.
(d) The necessary transient 'off' current to switch the transistor from the 'on' to the 'off' condition in the required time.

Condition (a) is normally satisfied by a small positive bias on the base relative to the emitter. Condition (b) can be found from a measurement of the large signal earthed-emitter current-gain β_l at the required collector current. A value of 20 or more is usually obtainable except possibly at very high collector currents. Conditions (c) and (d) can be found with reference to the transient response equations due to Moll[1]. These apply strictly only to conditions of constant current drive, but such conditions are approached in the circuits discussed later. As normally quoted, the equations are rather cumbersome but, fortunately, can be simplified if only approximate results are required. The condition for this simplification is that the transient drive current must be appreciably greater than the drive to hold the transistor in the 'on' condition. This will most certainly be true if transistors with cut-off frequencies of the order of 10 to 30 Mc/s are used, and a rise-time of about 0·04 μsec is required.

With this condition satisfied, the expression for the rise-time of the collector current for a given base driving current becomes, (see Appendix),

$$T_R \approx \frac{1}{\omega_\alpha} \left| \frac{I_c}{I_B'} \right| \qquad (1)$$

This result is optimistic in so far as the effects of the collector and external capacitances have been neglected. Giving a specific example for a transistor with an f_α of 16 Mc/s, then for a rise-time of 0·04 μsec the ratio of I_c to I_B' has to be about 4 to 1.

The fall-time for the earthed-emitter stage simplifies to a similar expression, (see Appendix),

$$T_F \approx \frac{1}{\omega_\alpha} \left| \frac{I_c}{I_B''} \right| \qquad (2)$$

Therefore, for the same rise- and fall-times $|I_B'| \approx |I_B''|$.

The direction of current flow in the two cases will, however, be reversed. It should also be noted that although a positive current is supplied to the base to drive the transistor 'off', the base is not necessarily driven positive until the transistor is actually cut-off.

From the above equations some idea of the necessary transistor cut-off frequency can be determined. If it is assumed that pulse transformers are not to be used, then for a 5-mA input gate there is available 2·5 mA to drive the first transistor 'on', and 2·5 mA to drive the transistor 'off'. The output stage must switch an amount in excess of 50 mA, say 70 mA. If only one stage were used (neglecting the phase reversal) this would require an f_α of at least 120 Mc/s for a 0·04-μsec rise-time. Using two stages in cascade (remembering that the second stage also requires drive in both directions), reduces this requirement to a minimum of about 30 Mc/s. This appears to be a more reasonable figure considering the currents to be switched.

As ordinary alloy-junction transistors with a cut-off frequency of 30 Mc/s and a reasonable power and voltage rating are very difficult to manufacture, it was necessary to resort to the newly developed graded base or drift-type switching transistors, which will now be discussed.

Drift Transistors for Switching

Transistor Characteristics

The idea of incorporating an accelerating or drift field in the base region of a transistor (in order to improve its frequency performance) was discussed originally by Kroemer[2]. Most of the drift transistors so far produced have been intended for use as small-signal amplifiers, in which application they have proved satisfactory in operation; however, when the use of drift transistors for switching applications is considered, two additional requirements arise, as follows:

(i) In devices of this type the cut-off frequency usually diminishes rapidly with a decrease in the collector voltage when the latter falls below a certain critical value. This is because the width of the collector depletion layer, W_d, which for a homogeneous material is given by

$$W_d = A. \, (\rho_n V_c)^{\frac{1}{2}}$$

where A = a constant and

ρ_n = base resistivity

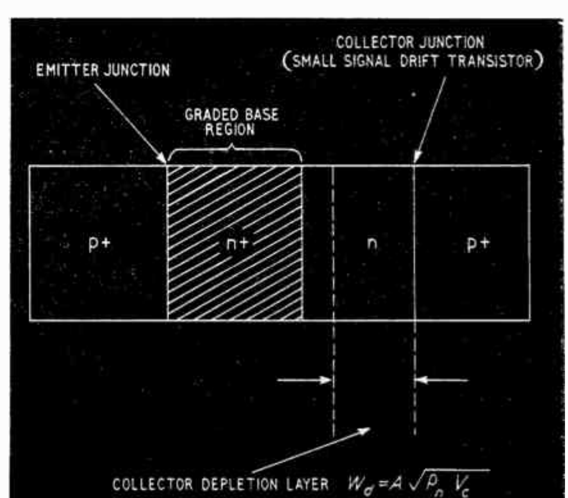
will no longer extend to the drift field region for collector voltages below a lower limiting value. This is illustrated in Fig. 1. Below this voltage the device acts as a combined drift and diffusion transistor and, as a result, the cut-off frequency drops. For many drift transistors this limiting voltage (sometimes called the 'starting' voltage) is in the region of 4–6 V.

When using a transistor as a switch it is necessary to operate it in the 'on' state with as low a collector voltage as possible, otherwise an excessive power dissipation level may be reached. This means that in order to achieve fast rise and fall times the frequency characteristics of the device must be maintained down to low collector voltages. In order to meet these requirements a modified drift transistor has been developed experimentally in which the cut-off frequency is maintained down to a collector voltage of 0·5 V.

(ii) The second requirement is that of maintaining the frequency characteristics at high levels of emitter current. This produces device design problems because the effect of the drift field tends to be swamped as the number of injected minority carriers increases, owing to the setting up of an opposing 'polarizing' field produced by the separation of the holes and electrons in the base region[2]. The result is to increase the effective hole transit-time and hence reduce the cut-off frequency. This difficulty is fundamental in this type of drift transistor. However, the base width may also be made very small as the graded base resistivity is effective in preventing punch-through until a reasonable collector voltage has been attained (say 15 V or more). In order to maintain the cut-off frequency reasonably constant over a wide range of emitter currents it appears that the effective drift field should not be too large.

Assuming the transistor can be designed to meet the above two conditions, the further requirements are that

Fig. 1. *Drift-transistor base-region structure*



the large-signal current-gain should be maintained up to high values of emitter current, and that the device should have a sufficient dissipation rating. The former can usually be met by ensuring a good emitter efficiency (e.g., by increasing the conductivity of the emitter) and, provided that the maximum current to be switched does not exceed about 50 mA, a dissipation rating of about 50 mW at the highest ambient temperature should be adequate.

An experimental switching transistor of the drift field type has been designed in an attempt to meet the above requirements and its application in practical circuits is discussed further in the section on 'Circuit Design'.

Effect of Drift Transistors on Switching Speeds

In deriving expressions, see Equs (1) and (2), for rise- and fall-times, Moll's equations[1] were used as a starting point. These equations are based on a frequency variation of the common-base current-gain factor α given by

$$\alpha = \frac{\alpha_0}{1 + j\omega/\omega_\alpha} \qquad \ldots \quad (4)$$

In the case of drift transistors, Equ. (4) is not very exact, but can still be used as a good approximation at frequencies appreciably below f_α if ω_α is replaced by a new parameter ω_1, the latter being defined as the angular frequency where the modulus of the common-emitter current-gain factor β is equal to unity. (Note that for a network obeying Equ. (4) exactly, $\omega_\alpha \approx \omega_1$.) Hence Equ. (1) should be re-written as

$$T_R = \frac{1}{\omega_1} \cdot \left| \frac{I_c}{I_B'} \right| \qquad \ldots \quad (5)$$

Similarly, Equ. (2) becomes

$$T_F = \frac{1}{\omega_1} \cdot \left| \frac{I_c}{I_B''} \right| \qquad \ldots \quad (6)$$

The relationship between ω_α and ω_1 for a drift transistor may be derived empirically as follows. The phase shift ϕ of the current-gain factor α may be shown to be[3]

$$\tan \phi = \frac{1{\cdot}22\ \omega}{\omega_\alpha} \left(\frac{\Delta E}{2kT} \right)^{\frac{1}{2}} \qquad \ldots \quad (7)$$

where $\Delta E/2kT$ is a measure of the drift field, and it is assumed that $\Delta E \geqslant 2kT$.

Now when $\omega = \omega_1$, to a good approximation† $\phi = 45°$. Hence:

$$\frac{\omega_\alpha}{\omega_1} \approx 1{\cdot}22 \left(\frac{\Delta E}{2kT} \right)^{\frac{1}{2}} \qquad \ldots \quad (8)$$

If we substitute for ω_1 in terms of ω_α in, for example, Equ. (5), we get the more accurate expression for the rise-time in terms of ω_α as

$$T_R = \frac{1{\cdot}22}{\omega_\alpha} \left| \frac{I_c}{I_B'} \right| \left(\frac{\Delta E}{2kT} \right)^{\frac{1}{2}} \qquad \ldots \quad (9)$$

For example, if we assume $\Delta E = 4kT$ then Equ. (9) indicates that for a rise-time of 0·04 μsec and a required current gain of 7·5 (these figures corresponding to the conditions outlined earlier in the article), the minimum value needed for f_α will be approximately 50 Mc/s.

Since it is easier in practice to measure the unity gain cut-off frequency parameter f_1, rather than f_α, it is likely to be more convenient to consider rise- and fall-times in terms of f_1 directly, and this may be done using Equs. (5) and (6).

A more accurate relationship between f_1 and f_α is given by

$$\frac{f_\alpha}{f_1} = \frac{\omega_\alpha}{\omega_1} = \frac{1{\cdot}22}{\tan \phi_{(f=f_1)}} \left(\frac{\Delta E}{2kT} \right)^{\frac{1}{2}} \qquad \ldots \quad (10)$$

This relationship is plotted in Fig. 2 and may be used if required to convert from one cut-off frequency parameter to the other. Results derived in this way are likely to be of sufficient accuracy for most practical purposes.

Circuit Design

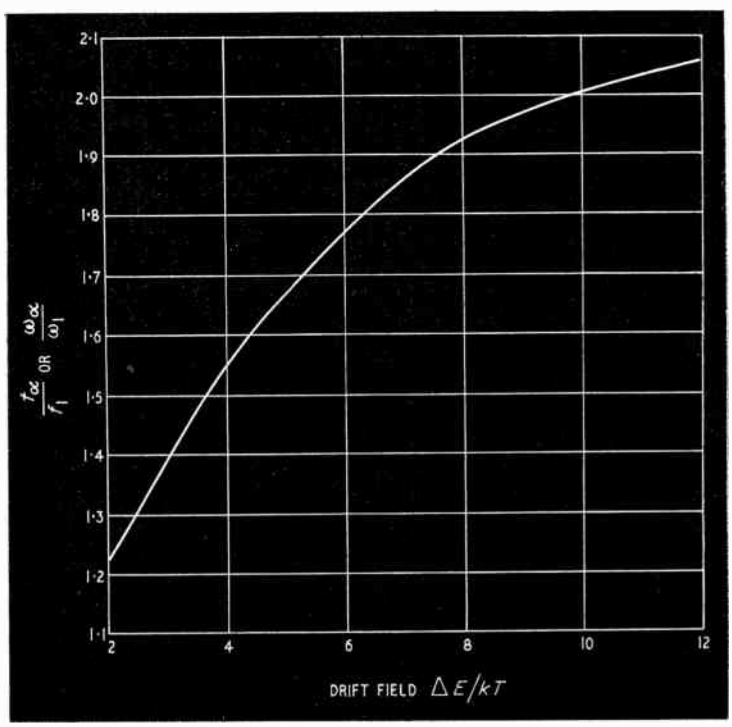
General Considerations

It has already been stated that an excess base-current drive (over that required to hold the transistor in the required 'on' state) must be provided to obtain the necessary transient time. This excess drive can be provided by the circuit shown in Fig. 3. The necessary reverse bias to hold the transistor in the 'off' state is provided by returning the generator to a small ($\frac{1}{2}$ to 1 V) positive bias. The current drive for the steady-state 'on' condition is determined by the input voltage V_g and the series combination of R_g and R_i.

$$\text{Thus } I_B = \frac{-V_g + E_B - V_{be}}{R_g + R_i} \approx \frac{-V_g}{R_g + R_i} \qquad \ldots \quad (11)$$

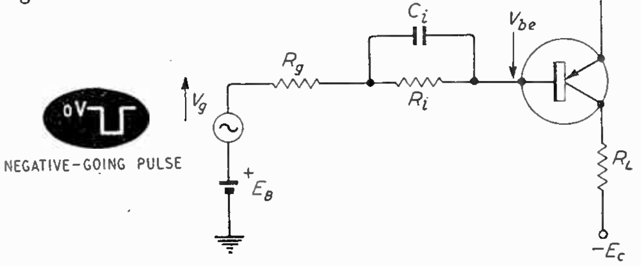
Owing to the presence of C_i however, the transient current drive is determined primarily by the value of R_g; i.e., for the 'off'-to-'on' transient condition,

Fig. 2. *Relationship of f_1 to f_α*



† This approximation becomes less valid for large values of drift field, a fact which is taken into account in Equ. (10).

Fig. 3. *Basic transistor drive circuit*



$$I_B' \approx \frac{-V_g + E_B - V_{be}}{R_g} \approx \frac{-V_g}{R_g} \quad .. \quad .. \quad (12)$$

assuming that $V_g \gg (E_B - V_{be})$.

Also, for the 'on'-to-'off' transient condition,

$$I_B'' \approx (+V_g + E_B - V_{be}) \cdot \frac{R_i}{R_g(R_i + R_g)} \approx + \frac{V_g}{R_g} \quad .. \quad (13)$$

providing $R_i \gg R_g$.

The optimum value of C_i is most conveniently found by experiment but its exact value is not usually critical. A compromise value is normally chosen which provides a reasonably fast transient time, coupled with a recovery time to the steady drive condition of I_B only, which is less than the pulse duration. Practical values usually lie between about 50 and 500 pF.

For a given drive and a given transistor the optimum values of R_g, R_i and C_i can be obtained as outlined. However, if a transistor with a higher value of β_l is now substituted, the transistor will be overdriven in the steady 'on' state with the result that the storage time, when the transistor is switched off, is liable to become appreciable and serious. The storage time is given approximately by the expression, (see Appendix)

$$T_S \approx \frac{1}{\omega_I(1-\alpha_I)} \cdot \frac{|I_B| - |I_c/\beta_l|}{|I_B''|} \ldots \text{(assuming } \omega_n \gg \omega_I) \quad .. \quad .. \quad (14)$$

The hole storage time is thus proportional to the difference between the steady-state base drive current I_B and that base current just sufficient to bottom the transistor (I_c/β_l). To limit this difference it is necessary to employ some form of clamping. The standard form of clamping (diode from collector to some suitable clamping voltage) is unsatisfactory since, with a given clamping level and a transistor with a very high β_l, the permissible transistor dissipation might be exceeded. Further, a diode passing a high current will itself suffer from hole storage effects. To avoid these troubles either

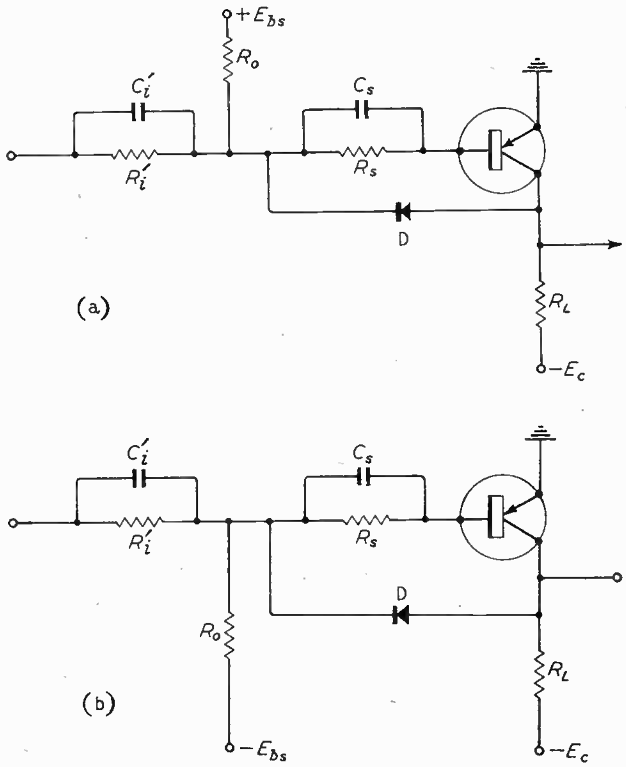
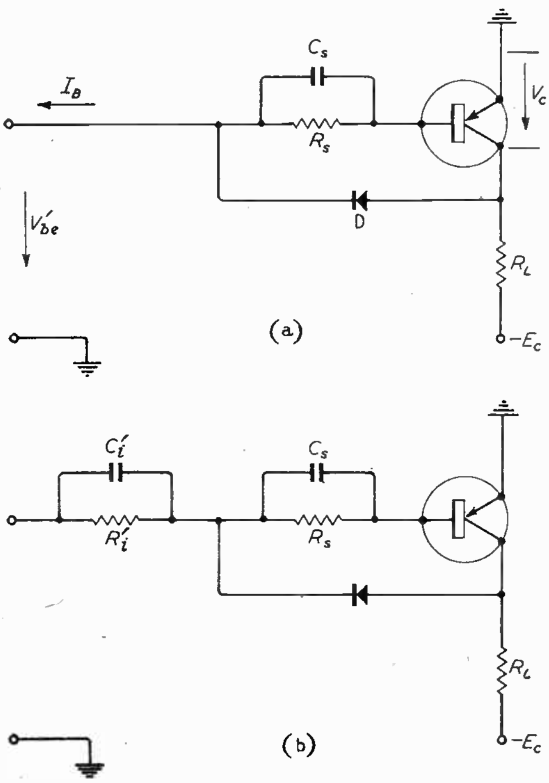


Fig. 5. *D.C.-coupled transistor drive circuits; (a) stage normally held 'off'; (b) stage normally held 'on'*

Fig. 4. *Shunt clamping circuit*



series or shunt feedback clamping, which limits the base drive, are preferable. The shunt method is most satisfactory for the present requirement. The circuit is illustrated in Fig. 4(a). When a base drive is applied V'_{be} will fall and V_c will rise towards earth. This will continue until V_c rises to above V'_{be} when the diode D will conduct and by-pass the excess drive. The point at which this clamping action occurs can be controlled by varying R_s. The higher the value chosen the farther will the collector voltage be clamped from earth. A mean value of about -1 V is normally satisfactory for graded base switching transistors such as those described previously (see section on Transistor Characteristics).

The capacitor C_s is present to avoid loss of the higher frequencies due to R_s and the input capacitance of the transistor.

With this method of clamping, excessive dissipation in the transistor is avoided. Further, since the diode only passes excess base drive current, hole storage in the diode is not so serious as in the conventional clamping circuit. However, hole storage in the diode can still be troublesome, and it is often necessary to include additional series resistance with a by-pass capacitor as shown in Fig. 4(b).

Given a diode having low forward resistance and small hole storage, it is usually possible to design a circuit to accept a 3-to-1 spread in β_l. A better, but more expensive circuit[4], using a diode in place of R_s, can be designed to accept a larger variation in β_l, but for the present purpose does not seem necessary.

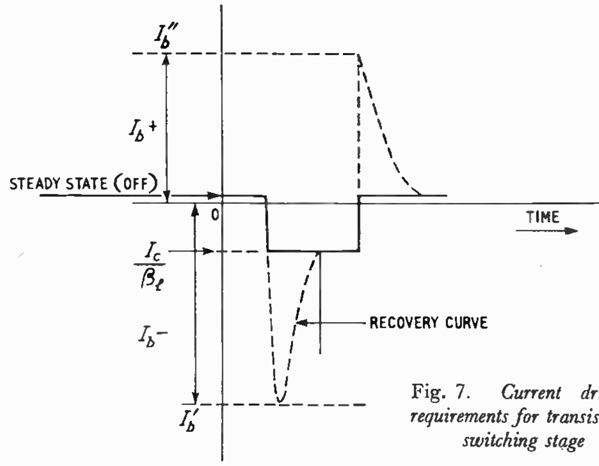


Fig. 7. Current drive requirements for transistor switching stage

Complete Single Stage Circuits

A d.c. coupled circuit is shown in Fig. 5(a). R_0 is fed from a high voltage (15 V) positive source and is adjusted just to cut-off the transistor. Normally, R_0 will be large and will not materially affect the rest of the circuit operation.

It should be mentioned that the circuit of Fig. 5(a) may require modification when used with graded base type transistors owing to the limited reverse emitter–base voltage rating of this type. This can be achieved as indicated below with reference to Fig. 6(a).

Fig. 6. *A.C.-coupled drive circuits; (a) stage normally held 'off'; (b) stage normally held 'on'*

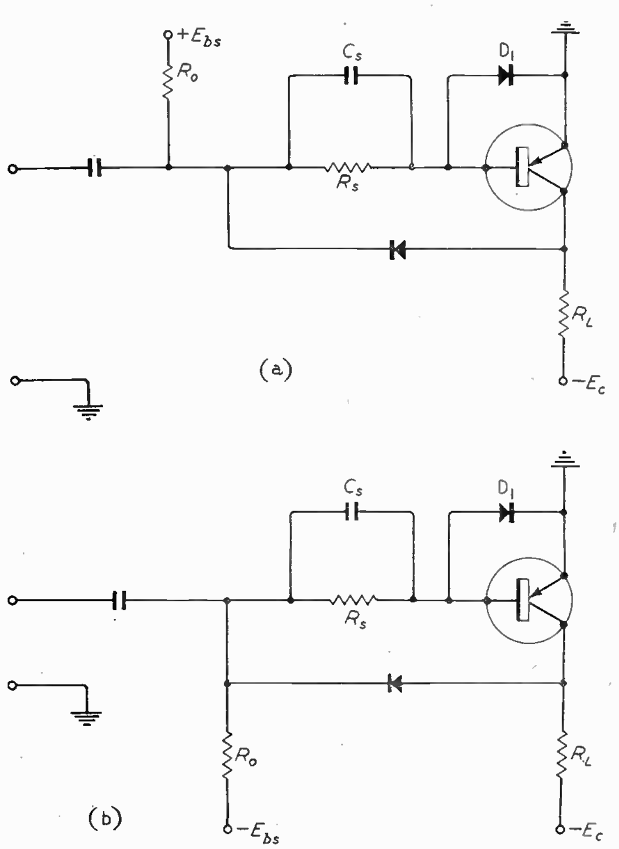


Fig. 5(b) shows a d.c. coupled circuit in which R_0 is returned to a negative voltage so that the transistor is normally held 'on'. The value is given approximately by

$$\frac{E_{bs}}{R_0}\beta_l = \frac{E_c}{R_L}$$

If the transistor is a.c. coupled to the previous stage a difficulty arises owing to the d.c. restoring action of the base-to-emitter junction. This results in a shifting of the operating point as the on-off ratio of the input signal is varied. In circuits where the on-off ratio is liable to alter, it is therefore necessary to include an additional diode to prevent the d.c. restoring action. This can be done as shown in Figs. 6(a) and (b). With such a circuit the input impedance of the transistor (plus diode) is low, both with the base positive and negative, thus preventing appreciable restoring action. The diode D_1 also prevents the base of the transistor from going too positive with respect to the emitter.

This may be an essential feature when using graded base transistors which normally have a maximum reverse emitter-base voltage rating of about 1 to 2 V.

Transistor Stages in Cascade

The total current swing which must be provided by one stage to drive a second stage having a resistive load, is approximately twice that needed to provide the required rise-time if a comparable fall-time is required. This is illustrated in Fig. 7.

It may be seen that if $I_B' \gg I_c/\beta_l$, then

$$I_B'' - I_B' \approx 2\, I_B' \quad \ldots \quad (15)$$

To take the specific example of the two-stage amplifier as envisaged, the 5-mA gate drive can be arranged to provide a $\pm$ 2·5-mA drive to the first transistor. Assuming a resistive collector load is used, then the first transistor must switch twice the transient current necessary to drive the second transistor from the 'off' to the 'on' state. Assuming also that the optimum condition is when both transistors provide the same transient current gain, (the ratio of I_B' or I_B'' to I_c), then the final values become

For V_1: $|I_B'| = |I_{B1}''| = 2{\cdot}5$ mA

$I_{c1} = 19$ mA. Gain 1st Stage $\approx 7{\cdot}5$

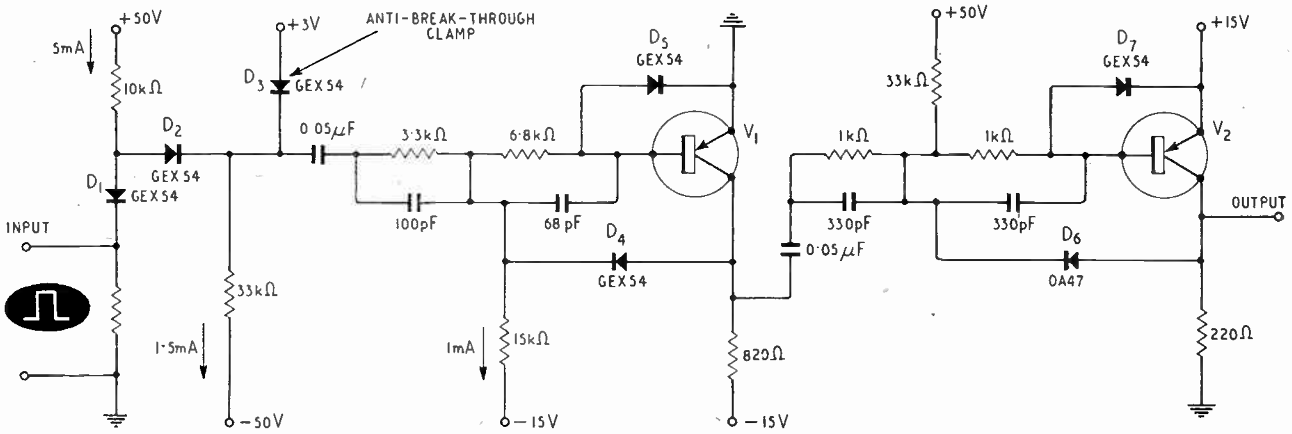


Fig. 8. *Complete two-stage circuit*

For V2: $|I_{B2}'| = |I_{B2}''| = 19/2 = 9{\cdot}5$ mA

$I_{c2} = 70$ mA. Gain 2nd Stage $\approx 7{\cdot}5$

The above analysis will enable approximate values to be assigned to the circuit components. In practice, however, the final load will probably also include a significant amount of shunt capacitance. This modifies the above theory owing to the fact that in switching the second transistor from the 'off' to the 'on' state additional drive current has to be provided to discharge this capacitance. Under these conditions the rise-time tends to be increased relative to the fall-time, and this is reflected in the practical values which are shown in Table 1.

It should be further noted that some improvement in the transient response times (in respect of both rise and fall-times) may be obtained by the use of compensating inductances in series with the collector load resistances.

TABLE 1

Performance of Transistor Pulse Amplifier

Performance of Transistors used

Position	f_1		β_l (70 mA)
	$V_c = -6$ V, $I_c = 1$ mA	$V_c = -3$ V, $I_c = 30$ mA	
V_1	30 Mc/s	24 Mc/s	>20
V_2	22 Mc/s	24 Mc/s	>20

Input Waveform to Gate 10 V, 0·6 μsec pulse.
10 % to 90 % rise-time ≈ 0·02 μsec.
90 % to 10 % fall-time ≈ 0·02 μsec.

Output Waveform Specification

Output Waveform (μsec)	220-Ω resistive load	220-Ω + 47-pF load	Load as in Fig. 9
Leading edge delay	0·03	0·035	0·075
10 % to 90 % rise ..	0·04	0·05	0·02
10 % to 90 % fall ..	0·045	0·05	0·02

Complete Two-Stage Circuit

The complete two-stage amplifier circuit designed from the above considerations is shown in Fig. 8. The first stage is normally 'on' and passes a standing current of about 15 mA. Clamping the collector at about −2 V relative to the emitter gives a steady dissipation of about 30 mW. The second stage is normally held 'off' by the positive potential developed across the diode D_7 by virtue of the current taken by the 33-kΩ resistor to +50 V. The transistor supplies are taken between earth and +15 V to provide the correct drive for other gates.

In the circuit, as shown in Fig. 8, a resistive load is shown for the second transistor. In practice, the output stage load would have to be arranged differently to provide an anti-break-through circuit as illustrated in Fig. 9. Since this is common practice this additional complication will not be discussed in detail, but note should be taken of the effect of hole storage in the collector clamping diodes, (see below).

The practical performance of the complete circuit for different load conditions is summarized in Table 1. The theoretical rise and fall-times using a resistive load

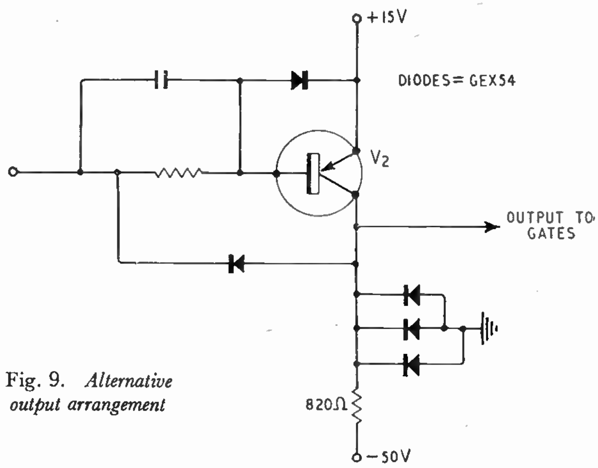


Fig. 9. *Alternative output arrangement*

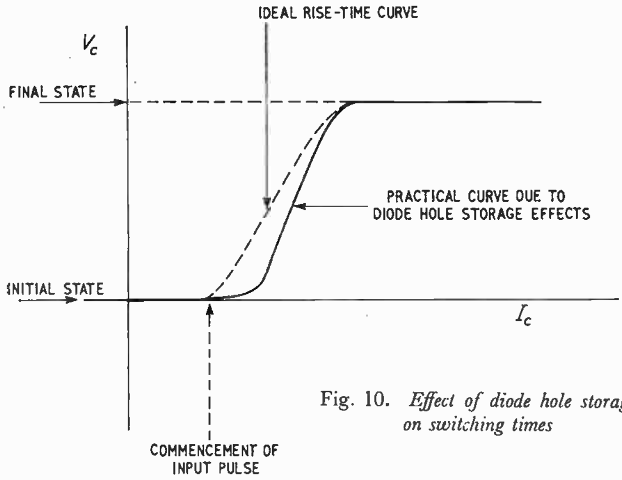


Fig. 10. *Effect of diode hole storage on switching times*

should be the same, and can be calculated for the practical circuit as follows

If T_{R0} = rise-time of input pulse,
T_{R1} = rise-time for 1st stage,
T_{R2} = rise-time for 2nd stage,

then

$$T_{R1} \approx T_{R2} = \frac{1}{\omega_1} \cdot \left| \frac{I_c}{I_B'} \right| \qquad \text{.. .. .. (16)}$$

The total rise-time may then be written

$$(T_R)_{total} \approx \sqrt{T_{R0}^2 + 2T_{R1}^2} \qquad \text{.. .. .. (17)}$$

If f_1 is taken to have a mean value of 25 Mc/s, then $T_{R1} = T_{R2} = 0{\cdot}045$ μsec, and if $T_{R0} = 0{\cdot}02$ μsec, then $(T_R)_{total} \approx 0{\cdot}065$ μsec.

It will be noted that the rise and fall-times obtained are rather better than the calculated values, and that an appreciable delay time is also present. It is considered that this behaviour is partly due to hole-storage effects in the clamping diodes D_4 and D_7 (Fig. 8), which will delay the start of the switching condition in transistors V_1 and V_2 respectively. When switching commences, the hole-storage current adds to that supplied from the generator and causes a transient current drive greater than that due solely to the other circuit components. The state of affairs may be represented by the conditions shown in Fig. 10. It will be seen that if such effects were not present the rise-time would lengthen.

If the diode clamping circuit of Fig. 9 is used, hole-storage effects are emphasized still more, and this can be seen in the increased delay time shown in the last column of Table 1.

A similar set of conditions also applies for the fall-time due to diodes D_5 and D_6 (Fig. 8).

Part of the delay is also caused by transit-time effects in the transistors.

The values of speed-up capacitors used were found to be optimum providing the pulse length was greater than about 0·4 μsec. For shorter pulses these values would have to be reduced to decrease the recovery time, as illustrated previously in Fig. 7, otherwise hole-storage effects become serious. Decreasing the capacitor values does however somewhat degrade the rise-time.

Photographs of the input and output pulse, after passing through the amplifier of Fig. 8, are shown in Figs. 11(a) to (c). These were taken using a Tektronix Type 545 oscilloscope and refer to conditions using a 220-ohm resistive load.

Conclusions

Using the recently introduced graded-base switching transistors, pulse amplifiers with a useful gain and fast rise-times can be constructed. Hole-storage in the transistors must be avoided unless considerable pulse widening can be tolerated. When using transistors with $f_1 \approx 25$ Mc/s, overdriving is necessary to obtain the required rise- and fall-times. This limits the useful stage gain to a figure well below that indicated by the low-frequency large-signal base-to-collector current gain. In order to obtain higher gains per stage and to use simpler circuits the cut-off frequency of the transistors used must be still further increased.

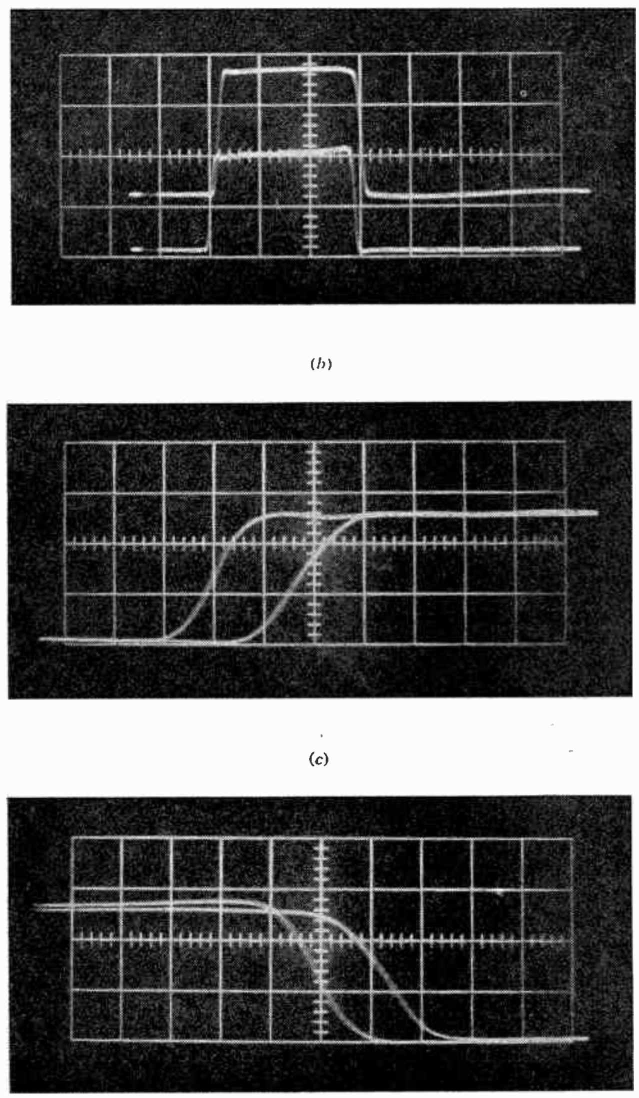
Another important feature which has been emphasized

Fig. 11. *Photographs of input waveform (a) and leading (b) and trailing (c) edges of output waveforms. Each square of the graticule denotes 10 V vertically and 0·2 µsec for (a) or 0·02 µsec for (b) and (c) horizontally*

(a)

(b)

(c)



by the work described in this article is the need to use diodes having low forward-voltage drop, combined with good recovery characteristics.

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APPENDIX

Derivation of Simplified Formulae for Rise-, Fall- and Storage-Times

Calculation of Rise-Time T_R

From Moll[1]

$$T_R = \frac{1}{(1-\alpha)\,\omega_\alpha} \cdot \log_e \frac{I_B'}{I_B' - \dfrac{0{\cdot}9\,(1-\alpha)}{\alpha}\, I_c}$$

writing for the large-signal condition:

$$\frac{1-\alpha}{\alpha} = \frac{1}{\beta_l} \text{ and } \frac{1}{1-\alpha} \approx \beta_l$$

$$T_R \approx \frac{\beta_l}{\omega_\alpha} \cdot \log_e \left[\frac{I_B'}{I_B' - 0{\cdot}9\, I_c/\beta_l}\right]$$

$$\approx \frac{\beta_l}{\omega_\alpha} \cdot \log_e \left[1 + \frac{0{\cdot}9\, I_c/\beta_l}{I_B' - 0{\cdot}9\, I_c/\beta_l}\right]$$

using expansion $\log_e (1+x) \approx x$. x small ($|x| \ll 1$)

$$\approx \frac{\beta_l}{\omega_\alpha} \cdot \frac{0.9\, I_c/\beta_l}{I_B' - 0{\cdot}9\, I_c/\beta_l}$$

Hence $$T_R \approx \frac{1}{\omega_\alpha} \cdot \frac{I_c}{I_B'} \text{ (when } 0{\cdot}9\, I_c/\beta_l = 0{\cdot}9\, I_B \ll I_B'\text{)}$$

Calculation of Fall-Time T_F

$$T_F = \frac{1}{(1-\alpha)}\, \omega_\alpha \log_e \left[\frac{I_c - \dfrac{\alpha}{1-\alpha} I_B''}{\dfrac{1}{10} I_c - \dfrac{\alpha}{1-\alpha} I_B''}\right]$$

Put $\dfrac{\alpha}{1-\alpha} = \beta_l$; $\dfrac{1}{1-\alpha} \approx \beta_l$

$$T_F = \frac{\beta_l}{\omega_\alpha} \log_e \left[\frac{I_c - \beta_l I_B''}{\dfrac{1}{10} I_c - \beta_l I_B''}\right]$$

$$\approx \frac{\beta_l}{\omega_\alpha} \log_e \left[1 + \frac{0{\cdot}9\, I_c}{0{\cdot}1\, I_c - \beta_l I_B''}\right]$$

using expansion $\log_e (1+x) \approx x$; x small ($|x| \ll 1$)

$$\approx \frac{\beta_l}{\omega_\alpha} \cdot \frac{0{\cdot}9\, I_c}{0{\cdot}1\, I_c - \beta_l I_B''}$$

$$\approx \frac{1}{\omega_\alpha} \cdot \frac{I_c}{-I_B''}, \text{ since } 0{\cdot}1\, I_c \ll \beta_l I_B''$$

Hence $$T_F \approx \frac{1}{\omega_\alpha} \left|\frac{I_c}{I_B''}\right|$$

Calculation of T_S, Storage Time

$$T_S = \frac{\omega_n + \omega_I}{\omega_n\, \omega_I\, (1 - \alpha_n \alpha_I)} \log_e \frac{I_B - I_B''}{[I_c (1-\alpha_n)/\alpha_n] - I_B''}$$

Putting $(1-\alpha_n)/\alpha_n = \beta_l$

$$T_S = \frac{\omega_n + \omega_I}{\omega_n\, \omega_I\, (1 - \alpha_n \alpha_I)} \log_e \frac{I_B - I_B''}{(I_c/\beta_l) - I_B''}$$

$$= \frac{\omega_n + \omega_I}{\omega_n\, \omega_I\, (1 - \alpha_n \alpha_I)} \log_e \left[1 + \frac{I_B - (I_c/\beta_l)}{(I_c/\beta_l) - I_B''}\right]$$

using expansion $\log_e (1+x) \approx x$; x small

$$\approx \frac{\omega_n + \omega_I}{\omega_n\, \omega_I\, (1 - \alpha_n \alpha_I)} \cdot \frac{I_B - (I_c/\beta_l)}{(I_c/\beta_l) - I_B''}$$

in general $\alpha_n \approx 1$ and $\omega_n \gg \omega_I$ ‡

Hence,

$$T_S \approx \frac{1}{\omega_I\, (1-\alpha_I)} \cdot \frac{I_B - (I_c/\beta_l)}{(I_c/\beta_l) - I_B''} \approx \frac{1}{\omega_I\, (1-\alpha_I)} \cdot \frac{|I_B| - |I_c/\beta_l|}{|I_B''|}$$

assuming $I_B'' \gg I_c/\beta_l$.

‡ This assumption is not valid in all cases (e.g., for symmetrical transistors), but is quite valid for the case of graded base transistors which are primarily being considered here.

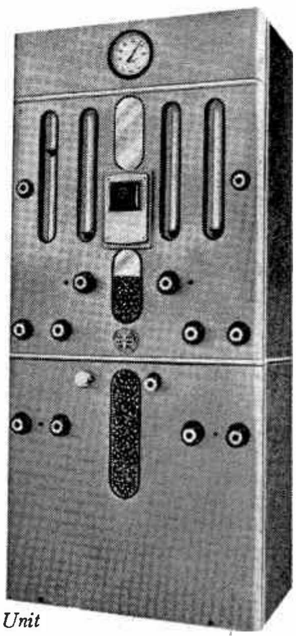
WATER DE-IONIZATION EQUIPMENT

It was discovered some years ago that certain synthetic resins have the ability to remove bases and radicles from water by an exchange of ions. Previously the removal of the bases calcium and magnesium by replacing them with the element sodium had been accepted as standard practice in the base exchange softening of water. It was ascertained however, that if water were passed through a column of one of the synthetic resins in the hydrogen form and then through a column of another resin in the hydroxyl form, the resultant exchange removed not only the bases calcium, magnesium and sodium but the sulphuric, hydrochloric and nitric acid acid radicles as well, thus producing a pure salt-free water without the necessity of evaporation and condensation.

In the Solobed system of de-ionization, which has been developed by the Paterson Engineering Company Ltd., of 129 Kingsway, London, a water of greater purity than distilled water is obtained, and, since the use of fuel in any form is not involved, the production costs are generally much less than those of distillation.

The two types of resins, in bead-shaped particles, are mixed together in one vessel. The water is metered on entering the unit and an instrument on the outlet measures its resistance to electrical current. This enables the purity of the water to be determined instantly and the instrument can be preset to give audible or visual warning when the resins are becoming exhausted. The resins can be regenerated with dilute acid and alkali solution.

For treating waters containing organic matter in solution, porous varieties of the resins are employed; these are unaffected by the organic impurities, which are washed out during regeneration. The smallest Solobed unit is capable of treating fifteen to twenty gallons per hour; there is no upper limit to capacity.



Solobed De-Ionization Unit

Overloading Effects with Cathode Compensation

The use of a cathode RC circuit to correct for the effects of an anode RC load circuit was discussed in some detail in a recent article[1]. This article contains an analysis of the circuit, shown here in Fig. 1, and both the steady-state and transient responses are given and expressed by families of curves.

The effect of the cathode circuit upon the output obtainable without overloading was not discussed, however, and this will be treated here. It will appear that it is inadvisable to use an RC cathode impedance to correct for an RC load in the anode circuit of the same or a subsequent valve. Instead, such a cathode impedance should normally only be used to correct for the effect of a preceding RC time constant. Such a use of the circuit does not invalidate the data of the previous article; it can all be readily applied, whatever the actual position of the time constant for which correction is needed.

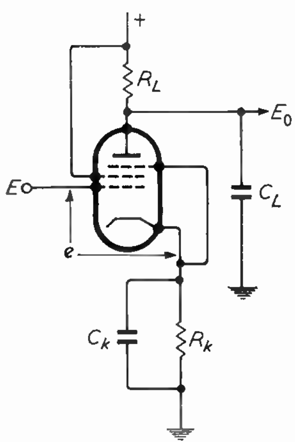


Fig. 1. *Basic circuit of cathode-compensated stage*

We shall adopt here the symbolism of the previous article and, in particular, we have, referring to Fig. 1, $\alpha = C_L R_L$, $a = C_k R_k / C_L R_L$ and $B = 1 + g' R_k$ where g' is the mutual conductance of the valve strapped as a triode. The anode current is

$$i_a = \frac{g_m E}{1 + g' Z_k} \quad \ldots \quad (1)$$

where g_m is the mutual conductance of the valve as a pentode.

The output voltage is

$$E_0 = -i_a Z_L \quad \ldots \quad (2)$$

Now $Z_k = R_k/(1 + pC_k R_k)$ and so

$$i_a = g_m E \frac{p + 1/\alpha a}{p + B/\alpha a} \quad \ldots \quad (3)$$

whence the response to a step of input voltage of amplitude E is

$$i_a = \frac{g_m E}{B}\left[1 + (B-1)e^{-tB/\alpha a}\right] \quad \ldots \quad (4)$$

When $t = 0$, we have $i_a = g_m E$ and when t is very large i_a tends to $g_m E/B$, the change of current between these limits being exponential with time constant $\alpha a/B = C_k R_k / B$. This variation of i_a with time is shown in Fig. 2 for the particular case of $B = 3$, which is a practical order of magnitude. When a step input is applied, the anode current jumps instantly to a maximum $g_m E$ and then drops back exponentially to $g_m E/3$.

Physically, the cathode voltage cannot change instantly because of the time constant $C_k R_k$. When the input step E is applied, therefore, it appears fully between grid and cathode of the valve and the anode current jumps to $g_m E$. The cathode current $g'E$ flows into C_k and charges it. The cathode voltage rises and so the grid–cathode voltage falls; as a result, the anode and cathode currents fall also.

The compensating action comes about as a result of this. If C_k were absent the anode current would be, in this example, $g_m E/3$ and of step form. As a result, the output voltage would change exponentially with time constant $C_L R_L$. With C_k present, the initial grid–cathode voltage step is three times as great, so that C_L starts to charge at three times the rate. The output voltage changes as if the anode time constant were $C_L R_L/3$ (assuming $C_k R_k = C_L R_L$).

The earlier article[1] considered the general case for any relation between these two time constants. For simplicity, the argument here will be restricted to the particular case when they are equal; that is, $a = 1$.

From equations (2) and (3), and taking $Z_L = R_L/(1 + p\alpha)$, we get

$$E_0 = -\frac{g_m R_L E}{\alpha} \frac{1}{p + B/\alpha}$$

$$= -\frac{g_m R_L E}{B}(1 - e^{-tB/\alpha}) \quad \ldots \quad (5)$$

The condition $B = 1$ corresponds to $R_k = 0$; that is, the equation then describes the performance of an uncompensated stage with an anode load $C_L R_L$. If, now, C_L is regarded as constant, the performance depends only on R_L/B. We can have $B = 1$ and $R_L = 3\ \text{k}\Omega$, say; or $B = 3$ and $R_L = 9\ \text{k}\Omega$. The step responses will be identical in the two cases and so will be the gains, $g_m R_L/B$.

It thus makes no difference at all whether we use cathode compensation or not; exactly the same step response and gain are obtained by adjusting R_L. This is, of course, for the particular case of $a = 1$; that is, equal anode and cathode time constants. For other values of a, the two cases are not so directly comparable.

There is a big difference in the anode-current waveforms, however. Without cathode compensation, the anode current is a step wave of amplitude g_mE. With compensation, it is a jump of g_mE followed by an exponential decay to $g_mE/3$ when $B = 3$, as shown in Fig. 2. As R_L is three times as great in the second case, the output voltage amplitudes are the same.

As the maximum changes of current are the same in the two cases, the maximum voltage outputs are also the same. Cathode compensation does not permit the attainment of any greater output voltage than that with an uncompensated stage of the same step response. We shall now show that, in certain cases, it cannot give as great an output.

Considering still the case of $B = 3$, let a positive-going step be applied to give a peak anode current change of I_a. As in Fig. 2 and in the first part of Fig. 3, the current jumps to $+ I_a$ and falls back to $+ I_a/3$.

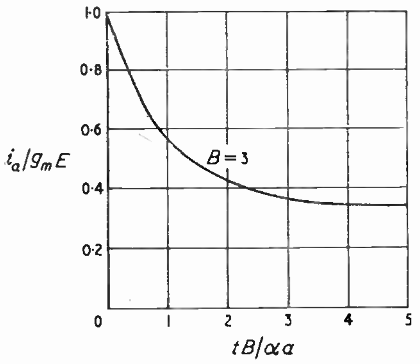


Fig. 2. *Anode current due to an input step voltage when $B = 3$*

When it has reached its steady-state, at $tB/\alpha a = 5$ say, let a negative input step of the same value be applied, so that the initial current jump is $- I_a$. This corresponds to a sudden termination of the input step, returning the input to zero, and so makes the whole waveform a rectangular pulse of duration $tB/\alpha a = 5$.

The initial change of current is from $+ I_a/3$ to $I_a/3 - I_a = - 2I_a/3$ and, from this, there is an exponential change of current to a final value of zero. The complete current waveform is shown in Fig. 3 by the solid line.

If compensation were not used, the current would be as shown dotted. The initial jump would still be I_a, but this would be maintained until $tB/\alpha a = 5$ and then the negative jump of I_a would occur, bringing the value to zero, at which it would remain.

In the uncompensated case, the total peak-to-peak current is I_a. In the compensated case with $B = 3$, it is 1·66 I_a. More generally, there is an initial current rise of g_mE followed by a fall to g_mE/B. There is then a change of $-g_mE$ at the end of the input pulse, taking the current to $g_mE(1/B - 1)$. The total change of current is thus from a positive peak of g_mE to a negative of $g_mE(1 - 1/B)$, making the change $g_mE(2 - 1/B)$ in amplitude. The total current change needed for the reproduction of a rectangular wave is $2 - 1/B$ times as great as when no compensation is used. This applies whether or not $a = 1$, for the current waveform does not depend on the anode time constant.

The current available is limited by the valve and it follows that with a given valve the voltage output obtainable is less with cathode compensation than without, and by the factor $1/(2 - 1/B)$. If B is large, this approaches 0·5. The effect is, of course, the same as the well-known one of a cathode follower with a capacitive load.

There is one mitigating thing in practice, which is that true step voltages do not occur. Any real voltage change must take some time to accomplish. In reality, therefore, the output limitation is not as severe as indicated here. It will, however, approach it if the rise time of the input voltage is small compared with the cathode time constant.

The overloading effect can be completely avoided, however, if the cathode circuit is used to compensate for a preceding time constant instead of a following one. The input to the compensating stage is then one with an exponential rise of time constant C_LR_L. The anode current, when $C_kR_k = C_LR_L$, is also exponential with time constant C_LR_L/B.

If the relationship between the time constants is such that overshoot occurs, the current response will then be oscillatory and the total current change will be increased. This is not likely to be very important because only small amounts of overshoot are normally permitted.

The data of the previous article[1] is readily applicable to cathode compensation of a preceding time constant. Referring to Fig. 4, V_1 has a mutual conductance g_{m1} and so, if a step E_1 is applied to its grid, the anode current is $g_{m1}E_1 = i_{a1}$ and the anode voltage is $-g_{m1}Z_LE_1$. The anode current of V_2 is then

$$i_a = \frac{- g_{m1}Z_Lg_mE_1}{1 + g'Z_k} = - i_{a1}\frac{g_mZ_L}{1 + g'Z_k}$$

or

$$\frac{- i_a}{i_{a1}} = \frac{Z_L}{1 + g'Z_k}$$

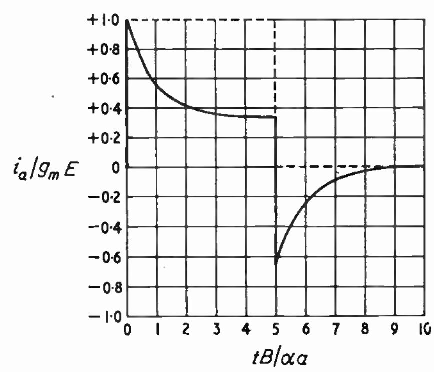
whereas for Fig. 1

$$\frac{- E_0}{g_mE} = \frac{Z_L}{1 + g'Z_k}$$

All the curves of the earlier article, which relate output to input voltages for Fig. 1, thus also relate the anode current of V_2 of Fig. 4 to the anode current of V_1.

The final valve of an amplifier invariably has shunt capacitance across its anode load. It has been pointed

Fig. 3. *Anode current of cathode-compensated stage for $B = 3$ (solid line) and for an uncompensated stage (dotted line). For both, the input is a rectangular pulse*



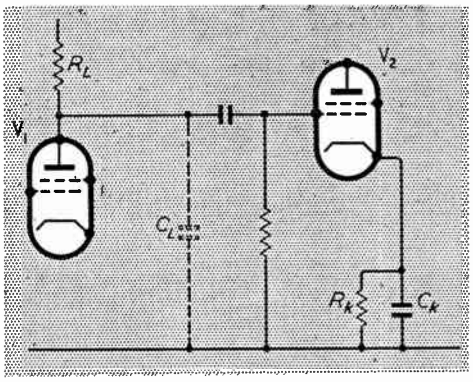


Fig. 4. Use of cathode components $C_k R_k$ to compensate a preceding time constant $C_L R_L$

out that it is better to choose the value of the load resistor to obtain the required step response than to use a higher-value resistor with cathode compensation. It is still better to use inductive compensation. The insertion of inductance in series with the load resistance enables a higher value of resistance to be used for a given rise-time and does not affect the current available from the valve; consequently, a larger output voltage can be obtained.

In practice, cathode-bias resistances are usually necessary and cannot be readily bypassed when the amplifier response must extend to very low frequencies. Unbypassed resistors cause appreciable loss of gain. In the writer's view, the proper procedure is for the anode circuit of the last stage to have inductive compensation and to use shunt capacitance on each cathode-bias resistor, choosing the values so that the cathode circuit of each valve compensates the immediately preceding anode circuit.

In a television receiver, there is normally only one video stage and the r.f., i.f. and detector commonly limit the rise-time to a greater degree than the anode circuit of the video stage. In that case, cathode compensation is unlikely to introduce any limitation of output. It may be thought of as compensating for the anode circuit, but it may equally well be considered as partially compensating for previous circuits.

Before concluding, it may be well to point out one particular case where cathode compensation is superior to almost any other method. This is when there is a preceding RC time constant of grossly excessive value which there is no way of reducing.

Suppose, for example, that one has 1 MΩ with 100 pF, a time constant of 100 μsec, when all that one can tolerate is 0·5 μsec, and that it is impossible to reduce either the resistance or the capacitance. Proper compensation will be obtained with $B = 200$ and $C_k R_k = 100$ μsec. Now $B = 200 = 1 + g' R_k \approx g' R_k$, so if $g' = 10$ mA/V, $R_k = 20$ kΩ and $C_k = 5{,}000$ pF.

Unless the anode load of the compensating stage can exceed some 20 kΩ it will attenuate, of course, but this may be unimportant compared with the advantage of effectively reducing an otherwise intractable time constant.

REFERENCE

[1] H. D. Kitchin, "Cathode-Compensated Video Stage," *Electronic & Radio Engineer*, April 1959, p. 122.

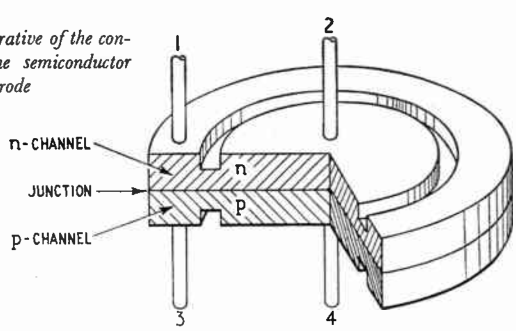
FIELD-EFFECT TETRODE

A new four-terminal semiconductor device will perform electronic functions which previously either could not be obtained at all, or only with extensive circuitry. It has been developed by Bell Telephone Laboratories.

The device, which is at present in the laboratory stage, is a field-effect tetrode. It can function as a transformer, gyrator, isolator, non-distorting modulator, or a short-circuit-stable negative resistance.

It comprises a disc of semiconductor with a diffused junction. A circular trench is cut and etched into each face of the disc to within about 1 mil of the junction on either side. Two leads are then attached to each face; one inside the trench, the other outside. When a voltage is applied across the junction, the thickness of the depletion layer adjacent to it is increased or decreased, depending on the direction of the biasing voltage. This, in turn, increases or decreases the resistance of each 'channel' between the bottom of the trench and the junction.

Diagram illustrative of the construction of the semiconductor tetrode



Functionally, the device has no analogue. It will function either as a transformer or a gyrator, depending on the polarity of the biasing voltage. As a transformer, it has a very decided size advantage for low-frequency use, although it does not afford d.c. isolation. As a gyrator, it should be of considerable circuit interest. One gyrator function is impedance inversion. For example, it should be possible to use the new device to convert the reactance of a miniature capacitor into that of a high-Q inductor.

If the device is biased properly, it will function as an isolator, allowing passage of alternating current in one direction only.

If a direct connection is made between the inner lead on one face and the outer lead on the other, the device will function as a two-terminal a.c. short-circuit-stable negative resistance. In the experimental models, this performance has been achieved over a range of about 30 to 250 V at 0·6 to 0·1 mA in a boron-doped silicon crystal with a phosphorous diffused junction.

Pulse and Square-Wave Generators

REVIEW OF MODERN INSTRUMENTS

SUMMARY. *Some basic circuits for generating rectangular pulses and square-waves are described and the performance obtainable is considered, with reference to commercially-available instruments.*

The last decade has been notable for a tremendous increase in the use of pulse circuitry. Apart from radar, pulse techniques are employed in digital computing, data transmission, nuclear particle detection and for a variety of biological, medical, and test purposes.

It follows that no one type of pulse or square-wave generator can reasonably be expected to meet all conceivable requirements and, in fact, there has been a tendency of late to produce specialized instruments for particular tasks. In some cases, the major requirement is for short pulses with rapid rises and falls; in others, precise control of pulse amplitude is essential and in some applications, notably data processing, there is a need for relatively complex trains of pulses rather than single pulses or simple repetitive ones.

Pulse or Square-Wave?

It becomes clear when one examines the specifications of pulse and square-wave generators that different people use these terms in rather different senses. To some, a square-wave means a periodic rectangular wave with unity mark–space ratio while, to others, any periodic wave of rectangular form is a square-wave irrespective of its mark–space ratio. A pulse may be a short unidirectional voltage or current impulse of any form, but frequently an approximation to a rectangular form is understood.

In this article we shall assume that a square-wave is one with unity mark–space ratio or, at any rate, one which is designed to have unity mark–space ratio and that a pulse has, ideally, a rectangular form. This latter is admittedly a restricted use of the term, since some pulses, notably sine-squared pulses, are of quite a different shape. However, in commercial parlance at least, an instrument described as a pulse-generator generally produces rectangular pulses and even when, as in the case of one instrument mentioned below, the pulses are deliberately made non-rectangular, the ideal rectangular shape remains useful as a basis of comparison.

Fig. 1. *Ideal and practical rectangular pulses, showing the usual meaning of terminology*

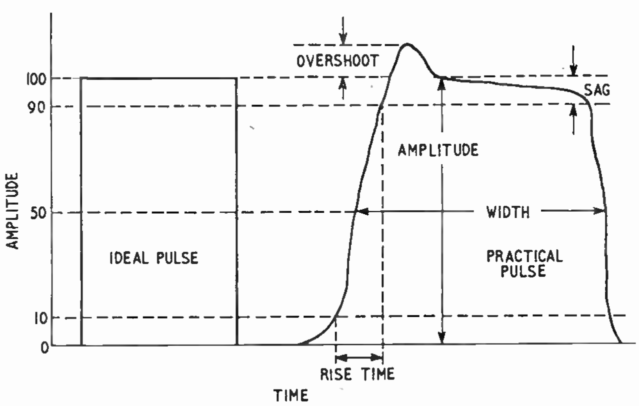
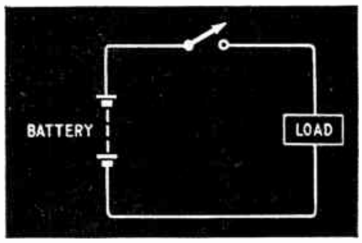


Fig. 2. *Elementary pulse generator*



Ideal and practical rectangular pulses are shown in Fig. 1, which illustrates what is usually meant by rise-time, overshoot, sag, amplitude, and duration (width). There are no hard and fast rules about terminology: for example, sag may be quoted as $\pm x\%$ about the mean pulse amplitude, and rise-time may be quoted as 5%–95%, etc.

Square waves are merely trains of pulses with unity mark–space ratio; and it follows that any pulse generator which has a 'duty ratio' of 0·5 can be used to generate square waves. An alternative method of obtaining square waves is to apply large-amplitude sine waves to a double limiter: a number of commercial sine-wave sources, such as those described in a previous article[1], incorporate the necessary extra circuitry.

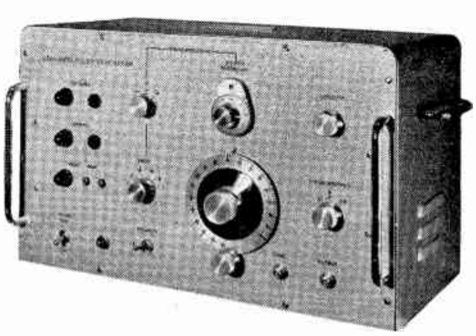
The Basic Pulse Generator

The ideal form of voltage-pulse generator is simply a perfect switch which connects a battery to a load, as in Fig. 2, for a period which determines the duration of the pulse. It will be noted that the nature of the load is not specified and, in practice, the usefulness of a pulse generator may depend to a large extent on its ability to deliver pulses of substantially the same shape and size to loads which vary appreciably in resistance and reactance.

If the switch is inertialess and has no contact resistance the rise-time is zero, and if the battery has no internal resistance the pulse will have a flat top, even if the load is capacitive. What happens to the voltage across the load when the switch contact opens at the end of the pulse then depends entirely on the nature of the load. Most loads are capacitive, so the trailing edge decays at a rate determined by an *RC* time constant. The rise and fall times quoted in commercial specifications frequently assume some standard load such as a resistance or a resistance shunted by a specified capacitance. It will be clear from Fig. 2 that the shape of the trailing edge could be controlled by arranging a second pair of contacts to short-circuit the load simultaneously with the opening of the other contacts, but commercial generators do not employ this artifice. The nearest approximation one gets is an output stage with a low impedance, such as a cathode-follower but, even here, the output impedance is often higher than one is accustomed to, because the cathode-follower valve may be switched off after the end of the pulse.

It is perhaps a chastening thought to the circuit designer that those commercial pulse generators which produce the most nearly perfect waveforms do, in fact, employ a variation on the ordinary mechanical switch. Such a generator, capable of producing pulses of very accurately known amplitude and with a rise-time of less than 1 mμsec, was described by O. H. Davie[2] in the September 1958 issue of this journal. The switch used is a magnetically-operated one and its contacts are wetted with mercury. The contact areas are therefore kept smooth and clean and the closure time is correspondingly short. A possible source of slow rise and fall times with mercury switches is the formation of arcs. The tendency to form an arc as the contacts approach one another is inhibited by filling the switch capsule with gas at a high pressure so that the voltage gradient necessary for breakdown is not reached until the contact surfaces are nearly physically touching.

Apart from the pulse generator already mentioned, at least two other British instruments make use of



Accuracy of pulse amplitude is a feature of the Sunvic Standard Pulse Generator. The main control is a ten-turn helical potentiometer giving 0–40 *V with an accuracy of* 80 *mV, while an incremental control provides* 0–1·2 *V with an accuracy of* 2 *mV*

mercury-wetted contacts. These are the E.M.I. Pulse Generator Type 1, which provides pulses of 4, 50, and 100 mμsec of either polarity, and also separate pulses for triggering an oscilloscope time-base, and the Sunvic Controls Standard Pulse Generator. In the latter, the

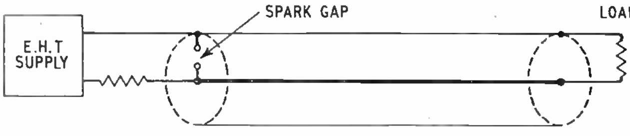
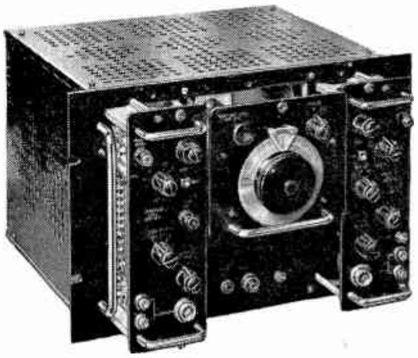


Fig. 3. *Simple high-voltage pulse generator using coaxial cable*

mercury relay is used to discharge a capacitor through an *RLC* network which produces an output pulse having a shape similar to that produced by typical pulse amplifiers. The amplitude of the pulses is known very accurately, so that the instrument is suitable for testing pulse-height analyzers.

The mechanical inertia of the mercury-wetted switch or relay imposes a severe restriction on the maximum



The Fleming Radio 1147*B Fast Multiple Pulse Generator provides pairs of output pulses with rise-times of about* 8 *mμsec at crystal-controlled recurrence frequencies. Two crystal oscillators are provided, with a frequency difference of* 1 *in* 10^4. *This enables pulse pairs to be produced which drift slowly in and out of coincidence, a useful arrangement for testing coincidence circuits*

pulse-repetition frequency obtainable (about 150 c/s). There is clearly a requirement for an inertialess device capable of achieving equally short rise-times. In general, this is not possible with thermionic valves or normal transistors because of the effects of stray capacitance. It might be mentioned that very fast very-high-voltage pulses can be obtained by charging an air-spaced concentric cable through a high resistance, and allowing breakdown to take place at a constriction in the cable (Fig. 3). This permits the cable capacitance to discharge through the load, which is normally made equal to the characteristic impedance of the cable. A flash of light is produced at practically the same time as the gas breakdown; this may be useful in certain applications.

At present, however, the most promising field for the development of fast switches is that of semi-conductors. If the voltage across a back-biased p-n junction is allowed to rise, a point is reached at which the junction breaks down, switching abruptly from a high resistance to a low one. The effect is caused by avalanche multiplication of current carriers and is analogous to fast gas discharges involving ionization by collision. It can occur at the collector junction of ordinary junction transistors and readers are referred for further information to a paper by Beale, Stephenson, and Wolfendale[3]. These authors used transistors to generate pulses with a rise-time of about 0·1 μsec at repetition frequencies of several megacycles. Visitors to this year's Physical Society's exhibition may have seen a strobing type of oscilloscope using an avalanche transistor producing pulses with a rise-time of about 1 mμsec. More recently, a special avalanche diode has been mentioned in an American journal[4]. This is an aluminium-silicon device with a computed rise-time of 50 μμsec. It is, of course, impossible to measure such a short time on an ordinary oscilloscope and, indeed, it should always be borne in mind when millimicrosecond pulses are applied to an oscilloscope that the effect of electron transit time through the Y plates may produce a distorted trace.

Complete Pulse Generators

So far, we have only considered switching devices, but a complete pulse generator must incorporate elements which determine the pulse-repetition frequency and the pulse width. Although it may be possible to control

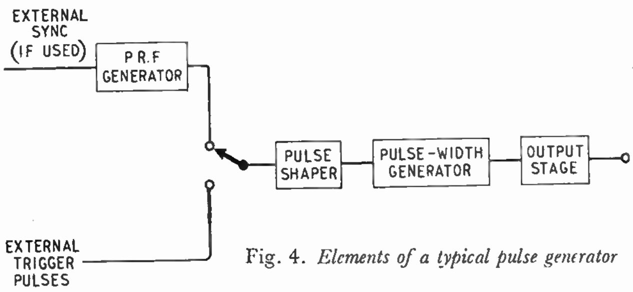


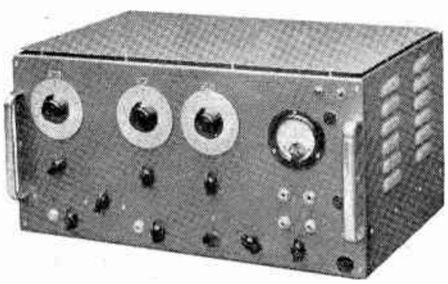
Fig. 4. *Elements of a typical pulse generator*

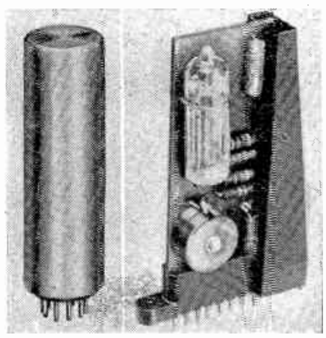
each of these by varying the parameters of one circuit, in practice the different functions are always performed by different circuits. The elements of a typical instrument are shown in Fig. 4. In some cases the pulses can only be initiated by an external triggering voltage, and no internal control of p.r.f. is provided but, in many others, facilities both for external triggering and for free running are provided.

P.R.F. Generators

The circuit most frequently encountered as a p.r.f.

The Kasama Type 302A generates pulses with repetition frequencies of 10 c/s to 1 Mc/s. A delay circuit is incorporated, with a range of 0·3 μsec to 10 msec. Pulse width is variable from 0·1 μsec to 10 msec, and the instrument can deliver 30-V pulses to a 75-Ω load. A square-wave output is also available





Showing two forms of Salford Electrical Instruments' packaged quartz-crystal controlled oscillator which provides a 100-kc/s output at 3 V (on open circuit) or 1·5 V (across a 20-kΩ load) when operated from a 6-V d.c. supply. It has a frequency tolerance of ± 0·015 % between −10 °C and 60 °C, a frequency stability of better than 5 parts in 10⁶ (with a 10 % change in supply voltage), and weighs just over an ounce. During operation it consumes about 5 mW of power only

generator is the ordinary multivibrator, consisting of a pair of valves with *RC* cross-couplings between grids and anodes. The time constants of the two *RC* networks are generally made equal and under these conditions the output waveform at the anodes approximates to a square wave. Triodes are generally used but, where an output with a short rise-time is required, pentodes may be used instead.

Cathode-followers may be used in the cross-coupling networks to provide a low-impedance drive circuit for the grids of the multivibrator valves proper. This is done in the Ultra Electric type 5004.

Of the two obvious alternatives to the multivibrator, namely the thyratron relaxation oscillator and the blocking oscillator, the former has gone out of use. Blocking oscillators are, however, occasionally employed as p.r.f. generators, for example in the Marconi Instruments TF675F, and they have the advantage that a high-current output pulse can be obtained from a low-power valve. If very high p.r.fs or short pulses are required the blocking-oscillator transformer can be wound on a ferrite core.

Pulse Shapers

Some kind of pulse-shaping circuit is often placed between the p r.f. generator and the circuit that deter-

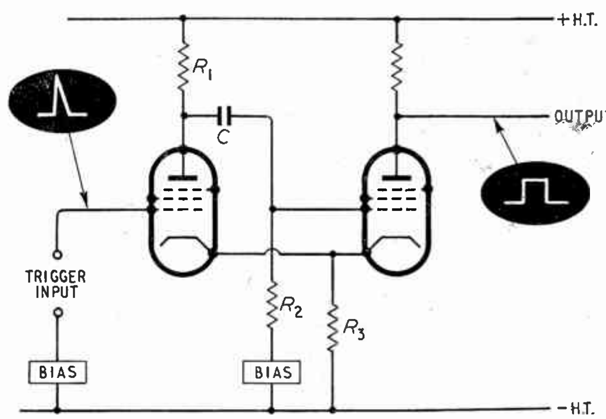


Fig. 5. *Cathode-coupled 'multivibrator'*

The Dawe Instruments type 412B provides twelve fixed p.r.fs, but by connecting an external resistor or capacitor p.r.fs. from 0 c/s to 100 kc/s can be obtained. Pulse durations are 0·3 μsec to 60 msec

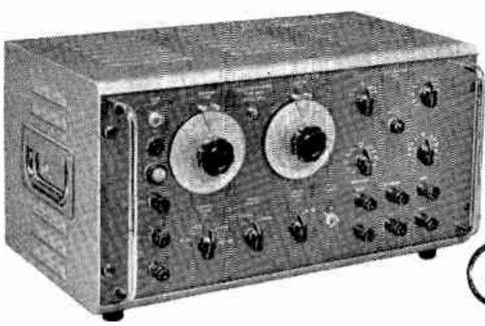
mines the pulse width. Its purpose is to ensure that triggering pulses applied to the 'pulse-width generator' are always of the same shape irrespective of the p.r.f. or the shape of the external triggering waveform. Also, the intermediate pulse shaper can generally be arranged to provide pulses with a faster rise than the p.r.f. generator. Since the pulse shaper is a driven device, some form of monostable circuit is used, generally a 'one-shot' blocking oscillator or a monostable 'multivibrator' such as the Schmitt trigger circuit.

Pulse-Width Generators

The latter kind of circuit (Fig. 5) is very commonly used for the 'pulse-width generator'. It has the advantages that one grid is free for injecting a triggering pulse and one anode is free for taking an output. In addition, there is only one *RC* timing network. Consideration of Fig. 5 will show that the operation of the circuit depends to a marked extent on the bias voltages applied to the two grids. In fact, the circuit can be made monostable or astable by suitable choice of bias voltages. When used as a pulse-width generator the circuit is arranged to be monostable and component values are chosen so that when V_1 switches off the resulting positive voltage step at its anode drives V_2 into grid current. This enables the charge on C to change rapidly, since the circuit time constant approximates to $C(R_1 + R_3)$, and R_1 and R_3 can be made small. When the circuit is in its other state, with V_1 switched on and V_2 off, there is no grid current and the time constant becomes $C(R_1 + R_2)$ and it is this latter time constant which is used to determine the pulse width. For this reason, in the quiescent state, V_1 is off and V_2 on. A positive triggering pulse is applied to the grid of V_1 and the resulting negative pulse at its anode cuts off V_2, a positive output pulse being obtained at the anode of V_2 having a duration determined by the timing circuit. In order to enable the circuit to revert quickly to its quiescent state at the end of the pulse it is common practice to connect a diode with its anode tied to the grid of V_2 and its cathode to a source of fixed potential. When this diode conducts, it effectively short-circuits the common cathode resistor R_3 and so reduces the time constant of the timing network. It also makes the circuit operation independent of grid current in V_2.

In order to obtain a fast rise on the leading edge of the output pulse a cathode-follower is sometimes used between the anode of V_1 and the timing capacitor. This is done, for example, in the Ericsson type 104A and in this case (Fig. 6) the cathode-follower is a 'stacked valve' circuit (White cathode-follower). This circuit is discussed below in connection with output stages.

In the circuit of Fig. 6 negative triggering pulses are applied to the anode of the first triode via a diode. This is one of several variations on the basic circuit, two others being shown in Fig. 7. In arrangement (a), the trigger pulse is fed to a grid through a diode, which is



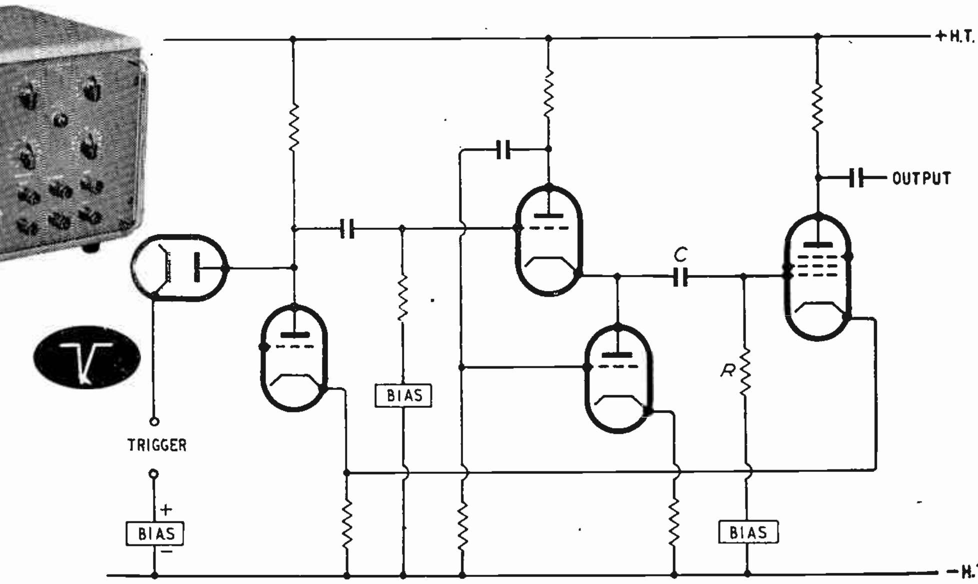
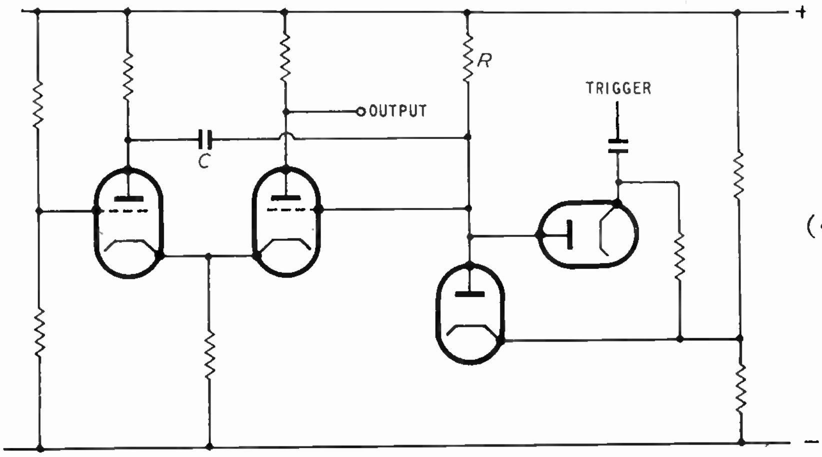
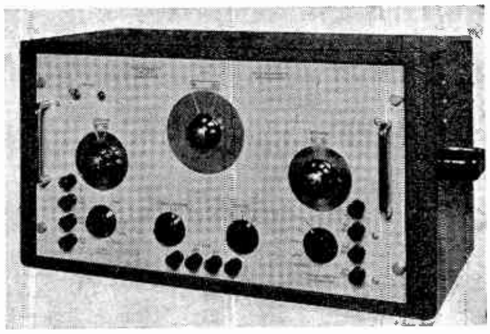


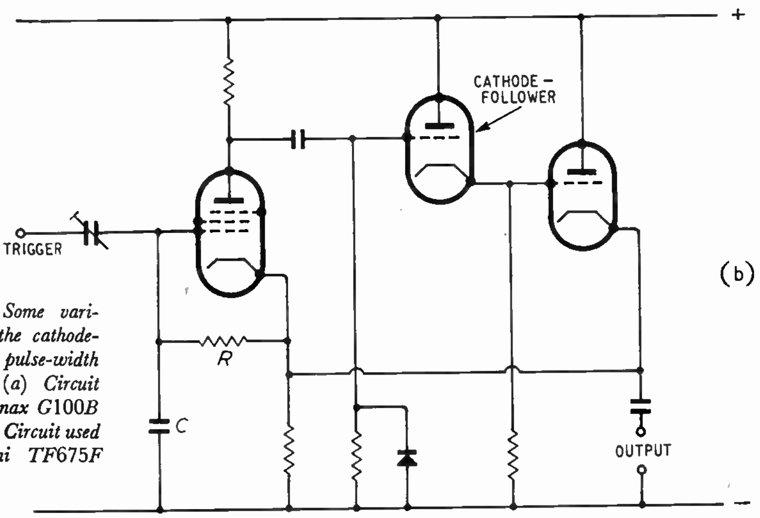
Fig. 6. *The circuit shows the essentials of the pulse-width generator in the Ericsson Type 104A. The instrument itself is shown in the photograph and it provides pulses at 10 c/s–100 kc/s with durations of 4–500 μsec. A delay generator designed to work with this instrument is also available, and provides a double-pulse facility*



(a)



The Panax G100B can provide single pulses, double pulses, or twin outputs of double pulses



(b)

Fig. 7. Some variations on the cathode-coupled pulse-width generator. (a) Circuit used in Panax G100B & H; (b) Circuit used in Marconi TF675F

The Marconi Instruments Pulse Generator Type TF675F

back-biased as soon as the timing cycle starts. At (b), a bootstrap connection for the timing circuit is used, the trigger pulse being introduced via a small capacitor.

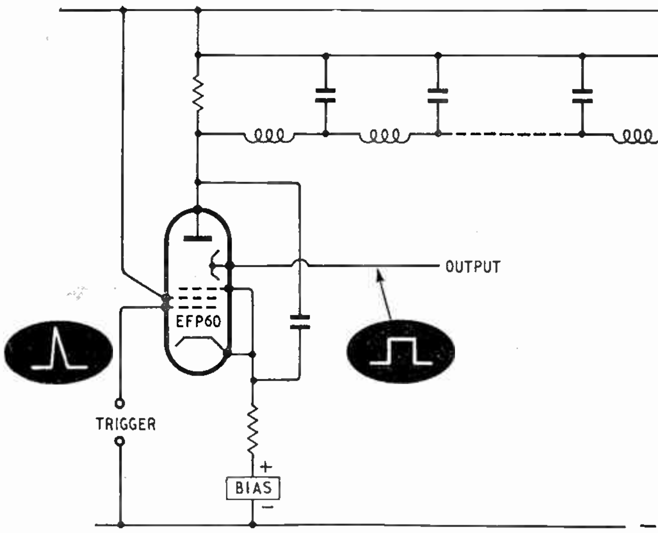
A radically different method of controlling pulse width is used in the British Physical Laboratories Model PG712 Mk. 2 pulse and square-wave generator (see Fig. 8). The circuit will be familiar to radar engineers. The basic timing device is a delay line, driven by a valve at one end and short-circuited at the other end. In fact, an electron multiplier is used in order to obtain a very high mutual conductance. The electron multiplier is normally almost cut-off by a positive bias applied to its cathode. It is switched on by a positive trigger impulse at the grid and this causes a negative-going voltage step at the anode. Positive feedback is applied from anode to cathode; since the valve is an electron multiplier it exhibits a current gain from cathode to anode greater than unity so that this type of feedback is capable of sustaining oscillation. The negative anode pulse propagates along the delay line in the anode circuit until it reaches the short-circuited end. The effect of the short-circuit is to cause a reflected pulse of opposite phase; that is, positive-going. When this arrives at the anode it is transferred to the cathode via the feedback capacitor and cuts off the valve again, causing it to revert to the quiescent state until the next trigger-pulse arrives.

A delay line in a similar type of circuit is used to form pulses in the Hewlett Packard type 212A. In this case the pulse is initiated by causing a thyratron valve to strike. The total delay is, however, fixed at 10 μsec, and the pulse duration is adjusted by making a second thyratron fire during the 10-μsec pulse and short-circuit the pulse as it passes to an output stage. By adjusting the relative times of firing of the two thyratrons, pulse durations of 0·07 to 10 μsec are available.

A delay cable is used by Ultra to generate pairs of short pulses from longer single pulses. The cable is connected as the anode load of a pentode, and shorted at the other end. The square pulse is applied to the grid, and the anode-voltage step which coincides with the leading edge goes down the cable and back, producing a short pulse. Another pulse of opposite polarity is produced by the trailing edge.

Output Stages

The output stages of pulse generators are somewhat unconventional. One reason is that they have to handle unidirectional signals and their operating point is, therefore, chosen to give the maximum available current swing in the selected direction, instead of being chosen so that equal swings can be obtained in either direction as in an ordinary amplifier. Another factor which influences design is the rapid rise-time or fall-time of the



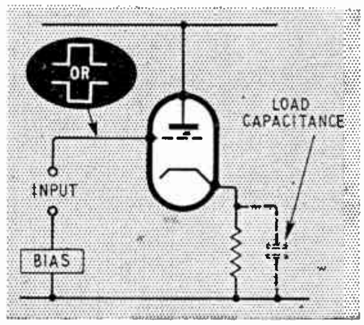
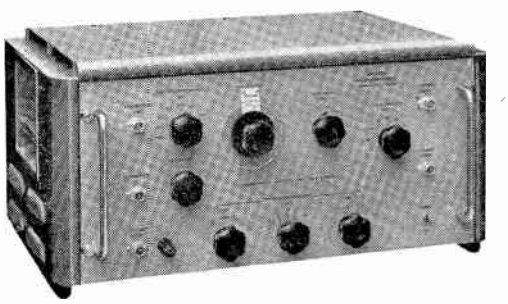


Fig. 9. *Cathode-follower with a capacitive load*

Fig. 8. *Pulse generation by short-circuited delay line (British Physical Laboratories PG712 Mk. 2)*

signals. The cathode-follower would seem at first sight to be the obvious choice since the required large driving voltage is easily obtained and since it has a large bandwidth. Unfortunately, the cathode-follower, when used as a pulse amplifier, does not necessarily follow. Suppose that the load is capacitive as in Fig. 9. When there is no signal, the load capacitance is charged to whatever voltage appears across the cathode resistance. If a



100-volt pulses are provided by the Solartron OPS100C at eight spot frequencies. The main pulse can be delayed 2 μsec to 4 msec with respect to a trigger output

positive-going voltage step is applied between grid and earth the presence of the load capacitance means that the cathode potential cannot change instantaneously. The instantaneous grid-to-cathode voltage is therefore of the same magnitude as the input-voltage step and, without grid current, the valve cannot handle any larger signal than it could handle as a straightforward amplifier. In practice, of course, the grid-to-earth voltage does not change instantaneously, so some degree of following action takes place, with a corresponding increase in signal-handling capacity. However, the effect may be very serious if rise-times are short and load capacitances large. The only thing the designer can do is to drive the cathode-follower from a low impedance source capable of supplying grid current without deterioration of waveform.

The situation is no better if the input pulse is negative-going. In this case, the valve will have been biased so as to take current in the no-signal condition. The effect of the load capacitance is then to maintain the bias after the arrival of a negative input-voltage step, so that the valve is promptly cut-off. A trick frequently resorted to when a negative output is required is the two-valve circuit shown in Fig. 6. The upper valve is the cathode-follower proper but, in addition to its cathode load, it is given an anode load. The arrival of a negative input step causes a positive-going output to appear at the anode of the upper valve. This is passed to the grid of the lower valve, which therefore draws more current and its anode potential falls. Since its anode is connected to the cathode of the cathode-follower, the result is that the output voltage is taken in the required direction. This circuit is used in the Panax types G100B & H, the Cawkell PE31PG Mk. 2, and also the Mullard L141. As an example of a more straightforward approach to the problem of capacitive loads, one might cite the Solartron type OPS100C. Here a large power pentode drives the cathode-follower, which is another pentode of the same type. The drive to the cathode-follower is positive-going and the resistance in the cathode circuit is kept low. The output valve has an inherently large signal-handling capacity and can supply a large current so that the load capacitance is charged quickly.

A rather unusual type of output stage, which also acts

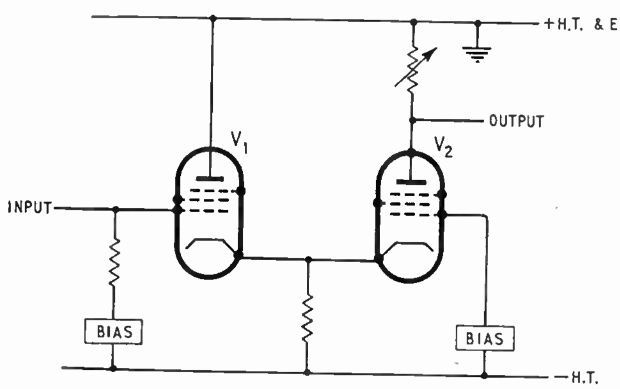
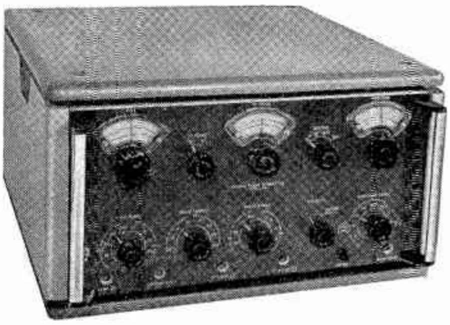


Fig. 10. *Cathode-coupled output stage*



Mullard Precision Pulse Generator L141. Double pulses, and a trigger pre-pulse, are provided. Amplitude and pulse separation are accurately determined

as a limiter, is found in the Ericsson 104A. Two high-slope pentodes are connected in a cathode-coupled circuit (Fig. 10). The output is taken across a low resistance in the anode circuit of the second valve. This amplifier can deliver pulses of either polarity. When positive-going pulses are required, the standing biases on the valves are arranged so that V_1 is cut-off and V_2 conducting. The application of a positive pulse to the grid of V_1 causes the latter to take current and the resulting voltage drop across the common cathode resistor switches off V_2. At the end of the pulse, the circuit reverts to its quiescent state. It will be noted that the circuit can operate as a slicer, clipping the

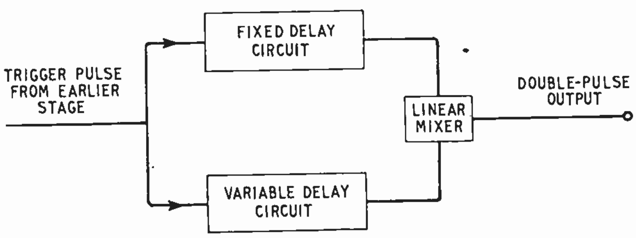


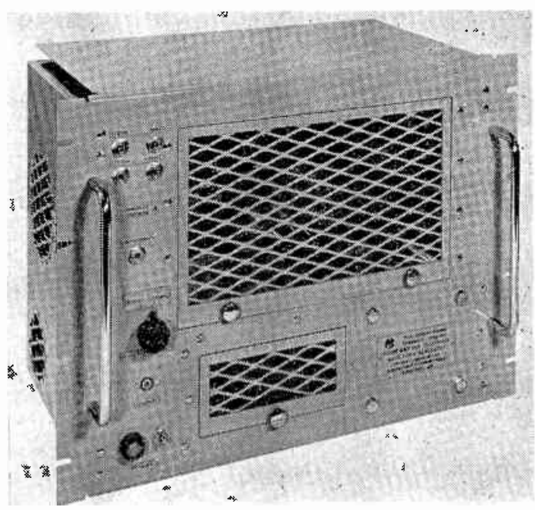
Fig. 11. *Essentials of double-pulse generator using fixed and variable delay circuits*

positive and negative peaks of the input pulse. This is an advantage if the applied pulse has its fastest rates of rise and fall in the middle of each edge, as is commonly the case. If negative output pulses are required, it is arranged that in the quiescent state V_1 conducts and V_2 is cut-off. A negative pulse applied to V_1 switches it off and allows V_2 to conduct, producing a negative-going output.

A somewhat similar output stage is used in the Nagard type 5002 double-pulse generator. In this, however, the common cathode resistor is replaced by a pentode valve. The current through this pentode, and therefore the output current, is adjusted by varying the grid bias. Thus the pulse amplitude is determined in the output stage. This particular instrument is capable of generating long pulses with a duration of up to two seconds. (In fact the range of pulse widths is from 0·2 μsec to 2 sec.) If a coupling capacitor of conventional size were used at the output an appreciable amount of sag would occur with normal values of load resistance. This is avoided by earthing the h.t. supply to the output stage and taking the output direct from the anode without going through a capacitor. With most other types of output stage, such a direct connection is impracticable, either because the output would be superimposed on a highly steady potential or because accidental short-circuiting of the output terminals could cause damage to an output valve. The Mullard L141 uses a White cathode-follower as a direct-coupled output stage.

Multiple-Pulse Generators

Generators of pairs of pulses with a known time separation are useful as standards for measuring short time intervals. A typical application is measuring the paralysis time of a pulse-counting circuit. In this case, one might start with a long pulse interval, observing the output of the counting circuit, and progressively reduce the interval until the second pulse is not counted. Another requirement which is almost universal in work involving pulses is for a triggering pulse, produced before the main pulse, for triggering an oscilloscope on which the main pulse, or some other waveform initiated by it, is to be displayed. More complicated trains of pulses may be required for simulating systems employing pulse-code modulation, such as radar beacons, or for synthesizing computer 'words'. The requirement for a prepulse for oscilloscope triggering can be met by including a delay circuit in the chain of circuits which make up the complete generator. If only a short delay is required between the prepulse and the main pulse it can readily be obtained from a delay line and this has the advantage that being a passive circuit stability of delay is high. If the delay required is more than a few microseconds an active device is used instead. This frequently takes the form of an auxiliary 'pulse-width



The Pye PTC1201/3 Pulse and Bar Waveform Generator produces a waveform for qualitative and quantitative assessment of channels for the transmission of television video signals. The waveform consists of a very short positive-going sine-squared pulse, a long positive rectangular pulse, and a negative-going sync pulse

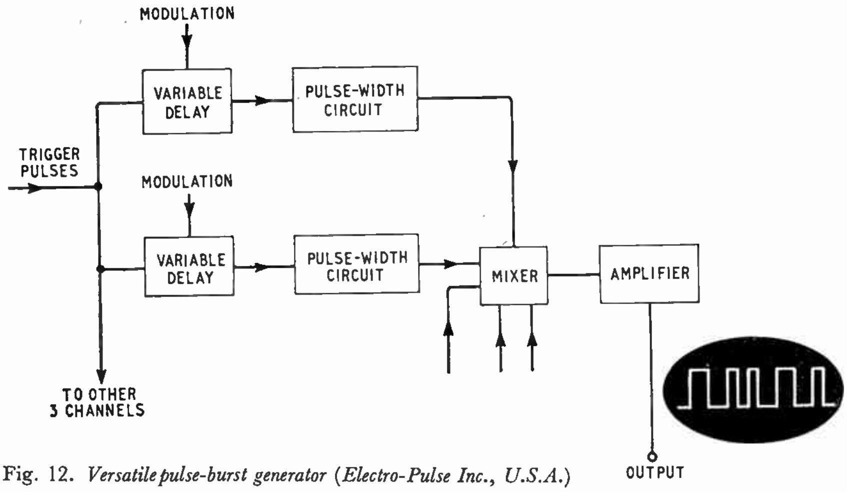


Fig. 12. *Versatile pulse-burst generator* (*Electro-Pulse Inc., U.S.A.*)

generator'. The leading edge of the output waveform provides the required prepulse while the trailing edge triggers the main pulse generator. Thus the delay obtained is equal to the duration of the auxiliary pulse. Such an arrangement is used, for example, in the Nagard 5002, where the range of delays is 0·2 μsec to 2 seconds. With this sort of arrangement, the width of the main pulse can be varied without affecting the timing of the prepulse.

A more general arrangement is to split the generator, providing two channels, one with a fixed delay and the other with a variable delay as in Fig. 11. This is done in the Mullard L141, where the fixed delay is provided by delay lines, and in the Panax G100B and G100H where all the delays are provided by pulse-width generators. In the latter instrument there are two output stages, each of which is capable of delivering pairs of pulses.

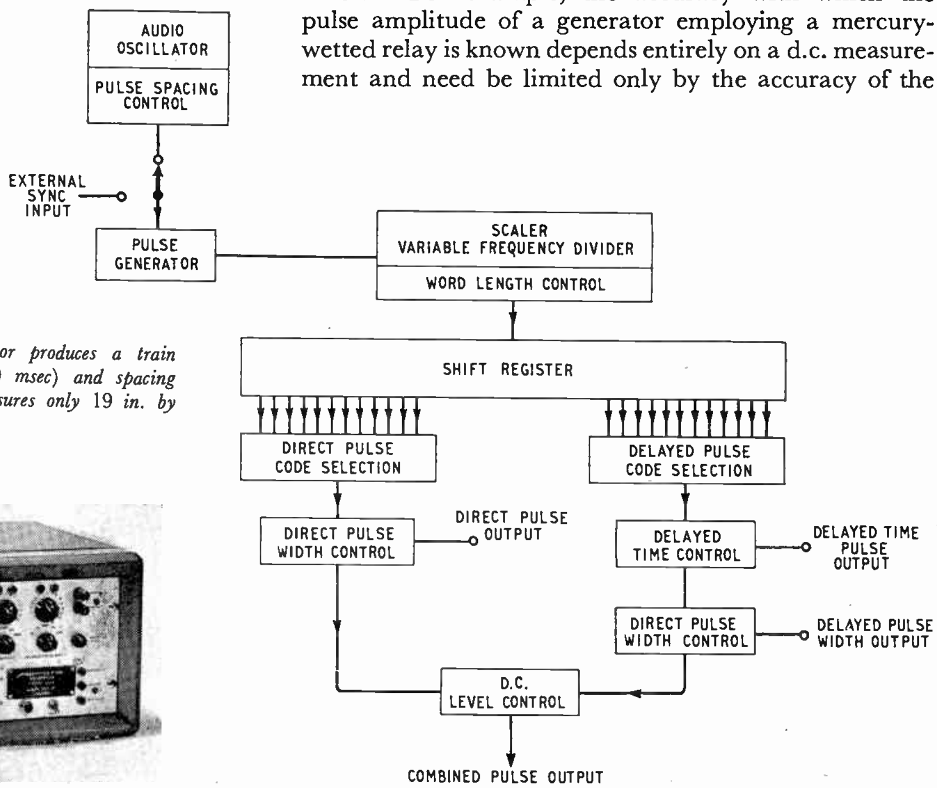
The principle can be extended to produce longer pulse trains; an American equipment, the Electro-Pulse 2150A, provides bursts of up to 5 pulses by means of the arrangement shown in Fig. 12. The width of each pulse is variable and the delay to each pulse can be varied over a small range by means of an external modulating signal.

A very versatile 'programmed pulse generator' made by Lintronic uses computer techniques to provide trains of up to 24 pulses. The arrangement used is shown in Fig. 13; the pulse generator, scaler, and shift register employ magnetic circuits based on square-loop materials and semiconductor diodes. The pulse spacing is variable between 8 μsec and 100 msec, the p.r.f. is 10 c/s to 125 kc/s and the pulse width 1 μsec to 150 msec. The purpose of the scaler/frequency divider is to fix the number of pulses fed into the shift register. (Any number between 1 and 64 is possible.) Up to 12 pulses can be selected directly or after a time delay caused by their passing down the shift register, and the two trains of 12 can then be combined, after a width adjustment, to form a 24-pulse train.

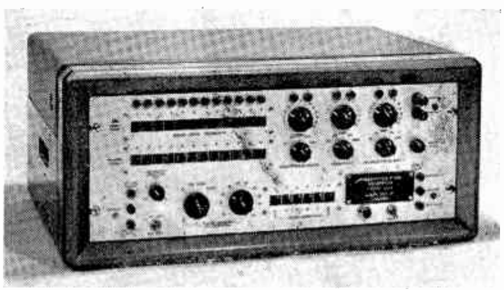
Performance

The modern pulse generator is a precision instrument, the pulse amplitude, duration and frequency being controllable within a few per cent. The limits of performance depend on the type of circuit, those circuits which excel in one direction being less outstanding in others. For example, the accuracy with which the pulse amplitude of a generator employing a mercury-wetted relay is known depends entirely on a d.c. measurement and need be limited only by the accuracy of the

Fig. 13. *Block diagram of Lintronic programmed pulse generator*



The Lintronic programmed pulse generator produces a train of pulses of variable width (1 μ*sec*–150 *msec*) *and spacing* (8 μ*sec*–100 *msec*). *The instrument measures only* 19 *in. by* 7 *in. by* 10 *in. deep*



measuring instrument. But the maximum p.r.f. is too low for many purposes. The following notes give a general indication of the best performances available in commercial instruments, but it must be understood that no one instrument combines all possible virtues.

The maximum p.r.f. is typically around 100 kc/s, but generators of short pulses may have p.r.fs of several megacycles. The Nagard 5002 can be operated at 2·5 Mc/s and a Rutherford (American) instrument goes to 10 Mc/s. Stability of p.r.f. is typically 5–10% but it can of course be improved by synchronizing the generator with an external signal of higher stability. The Fleming type 1147B incorporates crystal control of p.r.f. with an accuracy of 1 part in 10^4. Jitter may be as low as 1 part in 10^4 with the best instruments.

A figure for sag is not often quoted but is 5% of peak amplitude in the Hewlett-Packard 212A and Dawe 412 and 2% in the Mullard L141. Overshoot should not exceed 5–10%.

In delayed-pulse generators the accuracy of delay is often better than 5%; in the General Radio 1391-A the absolute accuracy is 2%, the incremental accuracy 1% and the resolution of the delay dial 1 part in 8,800.

Pulse durations as low as 5 mμsec are available from the Ultra 5004/B, which uses thermionic valves, while the E.M.I. type 1 goes down to 4 mμsec with a mercury relay and delay cable.

The amplitude of pulses tends to be rather low with the fast-pulse generators using valves, but the mercury relay type scores heavily here; for example, the E.M.I. instrument delivers 150-V pulses into 75 Ω. The Hewlett-Packard 212A employs a large double triode to deliver 50 V in 50 Ω, and the Kasama 302A gives a 30-V output in 75 Ω. The accuracy of amplitude calibration for generators using thermionic valves may be as high as 2%. Rise-time varies between about 1 μsec and 1 mμsec. The valve-type generators go down to about 5 mμsec but a fair average of rise and fall times is about 10–15 mμsec for the fastest pulses. Some instruments give rise and fall times which are independent of pulse duration.

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MATHEMATICAL TOOLS

By Computer

Integration for Engineers – 1

Last month we considered differentiation, and various useful tricks, such as Leibritz' Theorem, which simplify the differentiation of the types of function likely to occur in engineering practice. We now have to consider the reverse process of integration, that is, finding a function $f(x)$ whose gradient or derivative $f'(x)$ is given.

Now the first point we have to notice is that whereas when we are given $f(x)$, $f'(x)$ is uniquely determined, when $f'(x)$ is given, $f(x)$ is not uniquely determined: the addition of any constant C to $f(x)$ will not affect $f'(x)$. The geometrical significance of this is made clear in Fig. 1. Here we suppose that the curve drawn is $y = f(x)$; we are given the slope $f'(x)$ of the tangent at every point (of which P is typical, at a distance $x = OM$ to the right of Oy). But this slope will be unaltered if we replace the original x-axis Ox by $O'x'$ parallel to Ox and a distance C below it, the y-axis being unaltered so that the new origin is O′. The complete specification of $f(x)$ therefore requires an additional item of information besides the value of the slope $f'(x)$ of the tangent to $y = f(x)$ at each point P. Now this additional information may be lacking, in which case we have to include the term $+ C$ in $f(x)$ and accept the fact that $f(x)$ is not unique*. Thus, if we know only that $f'(x) = x^2$, all we can deduce (with the help of Table 1) is that

$$f(x) = \tfrac{1}{3} x^3 + C \quad \ldots \quad (1)$$

If, however, we also know that $f(0) = 3$, we deduce from Equ. (1) that $C = 3$ and thus $f(x)$ is now completely specified for all other relevant values of x. Alternatively, the additional information might be given in the form that $f(x)$ was zero for some value α of x. In this case, putting α for x in Equ. (1) tells

TABLE 1

Simple Cases of Integration

$f'(x)$	$x^n \; (n \neq -1)$	$\cos(kx + \alpha)$	$\exp(ax)$
$f(x)$	$x^{n+1}/(n+1) + C$	$\frac{1}{k}\cos\left(kx + \alpha - \frac{\pi}{2}\right) + C$	$\frac{1}{a}\exp(ax) + C$

* This has the disadvantage that in an operation involving a series of integrations and differentiations, the order in which the integrations and differentiations are performed will matter. For some purposes (such as the solution of differential equations) it is very desirable to ensure that integration is so performed that these arbitrary 'constants of integration' are taken as zero, so that the processes of integration and differentiation are completely reversible.

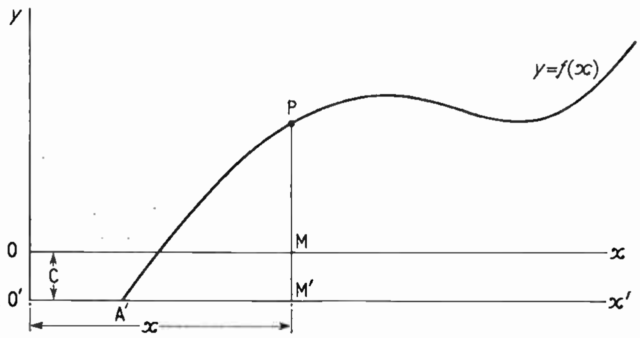

Fig. 1. *Derivation of $f(x)$ given $f'(x)$*

us that $C = -\frac{1}{3}\alpha^3$ so that

$$f(x) = \tfrac{1}{3}(x^3 - \alpha^3) \quad \ldots \quad (2)$$

From last month's article, we can draw up Table 1 showing the corresponding values of $f(x)$ when $f'(x)$ is given.

To these must be added another (which is in fact deducible from the last entry in Table 1), namely:

$$\text{if } f'(x) = 1/x, \text{ then } f(x) = \log_e x + C \quad \ldots \quad (3)$$

Now the notation "if $f'(x)$ is given, then $f(x)$ has a certain value" which we have so far used is cumbersome even if comprehensible, and we need an explicit notation for integration. But this notation will be clearer if we first consider Fig. 2. Suppose that $A(x)$ is the shaded area bounded on the left by the ordinate AM at $x = \alpha$, on the right by the ordinate PK where $x = \text{OK}$, below by the axis of x and above by the curve $y = \phi(x)$. Then it can be shown that

$$A'(x) = \phi(x) \quad \ldots \quad (4)$$

so that the process of integration, which we first regarded as the opposite or inverse of differentiation is, in fact, also related to finding the area under a given curve.†

From Table 1 it therefore follows that if $\phi(x) = x^2$, then

$$A(x) = \tfrac{1}{3}x^3 + C \quad \ldots \quad (5)$$

and since $A(x)$ is zero when $x = \alpha$, we have $C = -\frac{1}{3}\alpha^3$ and therefore

$$A(x) = \tfrac{1}{3}(x^3 - \alpha^3) \quad \ldots \quad (6)$$

Thus the case considered in Equs (2) and (6) is one which arises naturally when we wish to deal with areas bounded on the left by a particular ordinate like AM in Fig. 2. The notation for integration takes this into account. We write

$$f(x) = \int f'(x)dx + C\,;\quad A(\beta) = \int_\alpha^\beta \phi(x)\,dx \quad \ldots \quad (7)$$

The sign $\int$ is called the 'sign of integration', and the expression dx can be regarded simply as a short-hand for 'with respect to x'.‡ Thus the first of Equs. (7) reads "$f(x)$ is the integral of $f'(x)$ with respect to x"; the constant C of integration is often omitted though its presence is always implied. The second of Equs. (7) reads "The area $A(\beta)$ [bounded by the ordinates AM, BN, the axis of x and the curve $y = \phi(x)$ in Fig. 2] is obtained by integrating $\phi(x)$ with respect to x from α to β". This means that we have to determine $I(x)$ where

$$I(x) = \int \phi(x)\,dx + C \quad \ldots \quad (8)$$

and we may allow C to have any value we please; $A(\beta)$ is defined to mean $I(\beta) - I(\alpha)$. The quantities α and β are called the lower and upper limits of the 'definite' integral $A(\beta)$, and it should be noted that $A(\beta)$ depends only upon these limits, not upon the variable (x) with respect to which the integration was performed. When there are no limits and an arbitrary constant of integration, as in Equ. (8), the integral $I(x)$ is said to be 'indefinite'. The process of evaluating the definite integral (second of Equs. 7) can be summed up

$$A(\beta) = \int_\alpha^\beta \phi(x)\,dx = \Big[I(x)\Big]_\alpha^\beta = I(\beta) - I(\alpha) \quad \ldots \quad (9)$$

the square brackets in the third member of Equ. (9) indicating that the result of substituting α for x in $I(x)$ is to be subtracted from the result of substituting β for x. Thus if $\phi(x) = x^2$, Equ. (9) becomes

$$A(\beta) = \int_\alpha^\beta x^2\,dx = \left[\frac{x^3}{3}\right]_\alpha^\beta = \tfrac{1}{3}(\beta^3 - \alpha^3) \quad \ldots \quad (10)$$

Now it is an unfortunate fact that while we can nearly always formulate explicitly the gradient $f'(x)$ of a given function $f(x)$ [by means of Equ. (1) of last month's article] it is not by any means always possible to integrate a function explicitly. We are therefore fortunate if we can express the function we wish to integrate as a combination of functions whose integrals are known from, say, Table 1. We shall next consider a few simple cases in which this can be done, and thereafter the process of changing the variable, which in effect enables us to add greatly to the list of functions which could be included in Table 1. The procedure in more difficult cases will have to be left until next month.

Now from Table 1 (omitting the constant of integration)

$$\int \cos\theta\,d\theta = \sin\theta;\ \int \cos 3\theta\,d\theta = \tfrac{1}{3}\sin 3\theta \quad \ldots \quad (11)$$

But it is well known that

$$\cos 3\theta = 4\cos^3\theta - 3\cos\theta \quad \ldots \quad (12)$$

so that combining Equs. (11) and (12) we can evaluate

$$\int \cos^3\theta\,d\theta = \int \tfrac{1}{4}(3\cos\theta + \cos 3\theta)\,d\theta = \tfrac{1}{4}(3\sin\theta + \tfrac{1}{3}\sin 3\theta) \quad \ldots \quad (13)$$

which is not in the original list. There are many trigonometrical identities which can be used in the same sort of way as Equ. (12) to transform a trigonometrical integrand into a form more amenable to integration.

Now consider three integrals

$$\int_a^b x(c^2 - x^2)\,dx\,;\ \int_a^b \frac{x\,dx}{(c^2 - x^2)^2}\,;\ \int_a^b \frac{dx}{(c^2 - x^2)^{\frac{1}{2}}} \quad \ldots \quad (14)$$

where $0 < a < b < c$.

The first of these is straightforward from Table 1; we have

† Fig. 2 has been drawn for the simplest case when K is to the right of O, and x and $\phi(x)$ are positive. As is usually the case when functions formulated mathematically are illustrated geometrically, the "signs take care of themselves" in other cases provided that any shaded area below Ox is counted negative.

‡ We shall see later that there is a simple rule for dealing with dx if we wish to simplify the integration by changing the variable x into say y.

$$\int (c^2x - x^3)\,dx = \tfrac{1}{2}c^2x^2 - \tfrac{1}{4}x^4 = \tfrac{1}{4}[c^4 - (c^2 - x^2)^2]$$

so that the integral reduces to

$$\left[\tfrac{1}{2}c^2x^2 - \tfrac{1}{4}x^4\right]_a^b = \tfrac{1}{2}c^2(b^2 - a^2) - \tfrac{1}{4}(b^4 - a^4)$$
$$= \tfrac{1}{4}(b^2 - a^2)(2c^2 - a^2 - b^2) \quad (15)$$

A clue to the second integral is obtained by differentiating $(c^2 - x^2)^{-1} = y$ for if y has this value

$$(c^2 - x^2)\,y = 1 \quad \ldots \quad (16)$$

and differentiating both sides of Equ. (16) with respect to x as in last month's article

$$(c^2 - x^2)\frac{dy}{dx} - 2\,xy = 0 \quad \ldots \quad (17)$$

so that

$$\frac{dy}{dx} = \frac{2xy}{(c^2 - x^2)} = \frac{2x}{(c^2 - x^2)^2} = \text{twice integrand} \quad (18)$$

Hence

$$\int \frac{x\,dx}{(c^2 - x^2)^2} = \frac{1}{2\,(c^2 - x^2)} + C;$$

$$\text{integral} = \left[\frac{1}{2\,(c^2 - x^2)}\right]_a^b = \tfrac{1}{2}\left[\frac{1}{c^2 - b^2} - \frac{1}{c^2 - a^2}\right] \quad \ldots \quad (19)$$

But there does not seem to be any satisfactory way of finding any expression which when differentiated gives the integrand of the third of the integrals (14), and we were clearly somewhat fortunate that the differentiation of Equ. (16) gave us a method of doing the second. A more general method of attack is needed, and this is provided by the method of changing the variable. Not even this (or any method to be described in this or next month's article) is infallible; it is, however, worth trying and sometimes spectacularly successful.

In the integrals (14) let $x = c \sin\theta \; (-\pi/2 < \theta < \pi/2)$ and consider the last integral first as we have failed hitherto to do it at all. We have

$$\frac{dx}{d\theta} = c\cos\theta \quad \ldots \quad (20)$$

and the rule for dealing with dx in this situation is

Replace dx by $\dfrac{dx}{d\theta}\,d\theta$

so that
$$\int_{x=a}^{x=b} \frac{dx}{(c^2 - x^2)^{\frac{1}{2}}} = \int_{\theta=\sin^{-1}(a/c)}^{\theta=\sin^{-1}(b/c)} \frac{c\cos\theta\,d\theta}{c\cos\theta} = \int_{\theta=\sin^{-1}(a/c)}^{\theta=\sin^{-1}(b/c)} 1\,d\theta$$
$$= \sin^{-1}(b/c) - \sin^{-1}(a/c) \quad \ldots \quad (21)$$

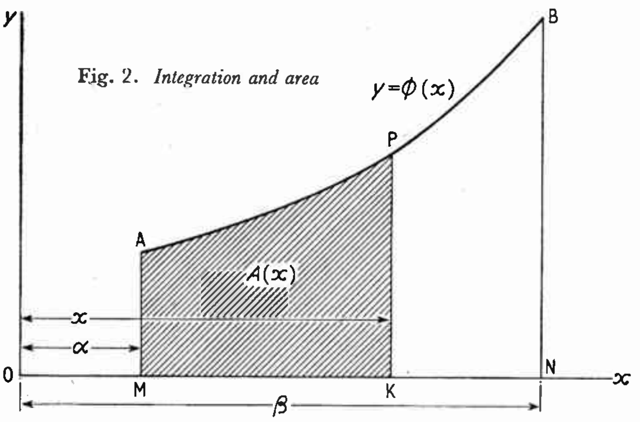

Fig. 2. *Integration and area*

Hence the substitution technique has been successful; the basis of its success was that when we made a change of variable designed to simplify $(c^2 - x^2)^{\frac{1}{2}}$ (which was the most awkward expression involved in the integrand), the factor $(dx/d\theta)$ associated with the rule for replacing dx introduced something into the θ-integrand, namely $c\cos\theta$, which was also simple and fitted in conveniently. There is no guarantee that the factor $(dx/d\theta)$ may not introduce something more complicated into the integrand. Applying the same technique to the second of the integrals (14), it reduces to

$$\frac{1}{c^2}\int_{\sin^{-1}(a/c)}^{\sin^{-1}(b/c)} \frac{\sin\theta\,d\theta}{\cos^3\theta} \quad \ldots \quad (22)$$

and in this case a second substitution $\cos\theta = y$ is the easiest way to evaluate the integral in terms of y, because differentiating this with respect to y gives

$$-\sin\theta\,\frac{d\theta}{dy} = 1 \quad \ldots \quad (23)$$

so that, replacing $d\theta$ by $(d\theta/dy)dy$ we have that the integral reduces to

$$\frac{1}{c^2}\int_\alpha^\beta -\frac{1}{y^3}\,dy = \frac{1}{2c^2}\left(\frac{1}{\beta^2} - \frac{1}{\alpha^2}\right) \quad \ldots \quad (24)$$

where $\alpha = \cos\theta$ when $\sin\theta$ is a/c. It follows that $\alpha = \{1 - (a^2/c^2)\}^{\frac{1}{2}}$ and similarly $\beta = \{1 - (b^2/c^2)\}^{\frac{1}{2}}$.

It is easily seen that Equs. (19) and (24) give the same result, and wisdom after the event suggests that the integral could have been done in one step by substituting $y = (1 - x^2/c^2)^{\frac{1}{2}}$. In this kind of situation, however, it is often very much easier to see and understand by taking two reasonably obvious steps rather than to make a great effort to replace these by one more difficult step.

The first of the integrals (14) could also have been done by means of the substitution $x = c\sin\theta$ but, in this case, direct integration with respect to x is obviously the correct method.

Finally, we have to consider the situation if either the integrand or one of the limits becomes infinite. In the last of the integrals (14), for example, if $b = c$ the integrand becomes infinite when $x = b = c$. But Equ. (21) is still correct; even if $b = c$, it merely means that $\sin^{-1}(b/c)$ can be replaced by $\pi/2$. Now the general rule for dealing with this sort of situation is to examine the situation carefully if the critical value likely to cause difficulty is closely approached. The substitution of $c\sin\theta$ for x shows that in this case no difficulty occurs. As a rough general guide it may be mentioned that if

$$I = \int_\alpha^\beta \frac{dx}{x^n} \quad \ldots \quad (25)$$

then if $n < 1$, there is no difficulty when α, β are on opposite sides of zero and if $n > 1$ there is no difficulty when $\alpha \to -\infty$, $\beta < 0$ or when $\alpha > 0$, $\beta \to \infty$. If the integrand is $\phi(x)$, and we suspect a difficulty because $\phi(x)$ tends to infinity when $x = 0$, then if $x^n\,\phi(x)$ is small for x sufficiently small and n greater than n_0 (n_0 being less than but *not* equal to 1) we can ignore the fact that $\phi(x)$ tends to infinity. Again, if $x^n\phi(x)$ is small for x

sufficiently large, n being greater than n_1 where n_1 is greater than but not equal to 1, we need not treat an infinite upper limit in any special way different from that appropriate to a finite limit. In this connection it should be noted that $\exp(-\alpha x)$ can be regarded as decaying very rapidly to zero as x tends to infinity when α is positive, because $x^n \exp(-\alpha x)$ tends to zero as x tends to infinity however large n may be and however small α may be, provided that n is finite and α is not actually equal to zero.

We have thus found that there are cases when integration can be easily performed because we can recognize the integrand as the derivative of a known function or a combination of known functions, and that a change of variable may reduce the integral to this condition when originally it was not obvious. If the integrand or the limits become infinite, there are a large number of cases in which no special difficulties are introduced; sometimes integration is possible even when the restrictions just mentioned are not satisfied, but such cases are a mathematician's job rather than an engineer's. In next month's article we shall consider a general method known as 'integration by parts', which sometimes replaces an initially intractable integral by one which can be evaluated explicitly; we shall also consider the possibility of replacing the integrand by a more manageable approximation and, in particular, the approximation known as 'Simpson's Rule'.

Design of 'Optimum' Arrays for Direction-Finding

By N. F. Barber, M.Sc.*

The study of radio astronomy has called for detection systems with special directional properties. They must be able to determine the direction and strength of each of a large number of small sources. Alternatively, they must be able to plot the distribution of power over some distributed source such as the face of the sun. Directional systems of a variety of designs have been made and used. The large steerable paraboloids[1,2] are often called 'radio telescopes' because their action is similar to that of a reflecting telescope. The present discussion is aimed more at those devices that are aggregates of many separate detectors (aerials) whose individual outputs are added or subtracted to create directional properties for the array as a whole.

One well-known method is to set a large number of similar detectors at equal intervals along a straight line and to observe the sum of their signals. The total signal is negligible, except for those radio waves whose direction of travel is normal or nearly normal to the line. If suitable phase shifts are introduced into the signals, the array can be made to accept waves coming from some specified direction other than the normal.

The directional distribution of power in the incident radiation is only approximately discovered by any actual experiment and Bracewell[3] pointed out that the errors present in the measurements made with a uniform array could be partly removed by a graphical treatment of the experimental curve. Arsac[4] emphasized the nature of the 'optimum' sensitivity pattern and showed that a close approach to it could be obtained by certain asymmetric arrays in which the detectors were set at unequal space intervals.

Arrays usually employ a large number of detectors in order to provide high directivity, but McReady, Pawsey and Payne-Scott[5] used a single detector of small directivity mounted on a high cliff above the sea to observe radiation from the sun, the detector and its reflected image acting in a manner similar to Michelson's stellar interferometer. Bolton and Stanley[6] used the same method to estimate the width of the radio source in Cygnus, and Ryle and Smith[7] used two widely-spaced detectors on land to estimate in a similar way the diameter of a radio source near the zenith. It is possible, however, to use two detectors to resolve a large number of sources. Thus McReady et al.[5] pointed out that a pair of detectors at a known separation measures one Fourier harmonic of the distribution of power. Stanier[8] made measurements at a variety of separations in order to measure many harmonics and so to build up by computation the distribution of power in one dimension. O'Brien[9] extended the idea to two dimensions, pointing out that it was essential to make phase comparisons of the two signals in order to gain correct information. The author[10] used the principle of movable detectors to make an experimental measure of the directional distribution in waves

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generated by wind on the sea. More recently Blythe[11] has combined a uniform array with a single movable detector to synthesize the distribution of radio noise over the celestial sphere.

Ryle[12] devised an improved form of interferometer. It incorporates a switch that alternately takes the sum and the difference of the signals from the two detectors. In effect, the system records the mean product of the two signals. Mills and Little[13] applied the technique to two long arrays set at right angles and so achieved a system that discriminated in two dimensions. The technique has been used more recently by Christiansen and Mathewson[14] and by Mills, Little, Sheridan and Slee[15]. It has also been employed by Covington and Broten[16] in a compound interferometer whose separate detectors comprise a long slotted waveguide and a pair of widely-spaced dishes. It can be shown that these later systems using Ryle's technique possess the 'optimum' qualities required by Arsac.

Nodtvedt[17] pointed out that if detectors are made to accept a wide band of radio frequencies, then a single pair of detectors has a sensitivity pattern similar to that of a complete array that uses a single narrow band of frequencies. The possibilities of this technique have been re-emphasized by Koch and Stone[18].

Super-directive arrays have received considerable attention as transmitting systems and the principles apply equally well to receiving systems. A discussion with full references to the literature has been given by Bloch, Medhurst and Pool[19]. Detailed treatment of these arrays will not be given here.

The purpose of the present article is to show that although the various devices mentioned above appear very different from one another, the same principle is common to all. In most cases a simple rule serves to show whether the various detectors are effectively placed and whether any are redundant. It provides a simple formula from which to calculate the sensitivity pattern of the particular system.

The Directional Spectrum

Radio noise may be regarded as the sum of an indefinitely large number of coherent plane wave trains having differing wavelengths and coming from differing directions. The direction and wavelength of any single wave train can be deduced if its rates of change of phase with distance are measured in three rectangular directions. If attention is restricted to waves of a certain wavelength it is necessary to measure only two of these rates of change. Thus, the signal from a coherent wave train varies sinusoidally with position.

$$\text{Signal} = a \cos 2\pi\,(lx + my + nz + ft + \alpha) \quad .. \quad (1)$$

where l, m and n are the rates of change of phase along the x, y and z axes. The wavelength L of this wave is given by

$$\frac{1}{L} = +\,(l^2 + m^2 + n^2)^{\frac{1}{2}} = k \quad .. \quad .. \quad .. \quad (2)$$

and the direction cosines of the wave normal are

$l/k,\ m/k,\ n/k$

If L is known from the frequency and wave velocity, and if l and m are measured, the direction cosines can be deduced. Doubt about the sign of n in Equ. (2) does not lead to ambiguity so long as the z axis is vertical, for the radio sources are known to be above the horizon, not below it.

The problem is to decide how the power of the incident radiation varies with direction; i.e., with the values of l and m when waves of only a single frequency are being examined. On a rectangular plot of l and m, illustrated in Fig. 1, all the coherent wave trains whose phase rates lie in the narrow ranges $l \pm \frac{1}{2}\delta l$ and $m \pm \frac{1}{2}\delta m$ are represented by the points lying in a small region of area $\delta l.\delta m$. The total power (mean-square signal) from all these wave trains may be written formally as

$E\,(l, m).\delta l.\delta m$

so that $E\,(l, m)$ is a coefficient of power density on this plot. The aim of the experiments is to find the values of power density over the whole of the plot. Its distribution may be indicated by detailed contours. This is called the directional spectrum.

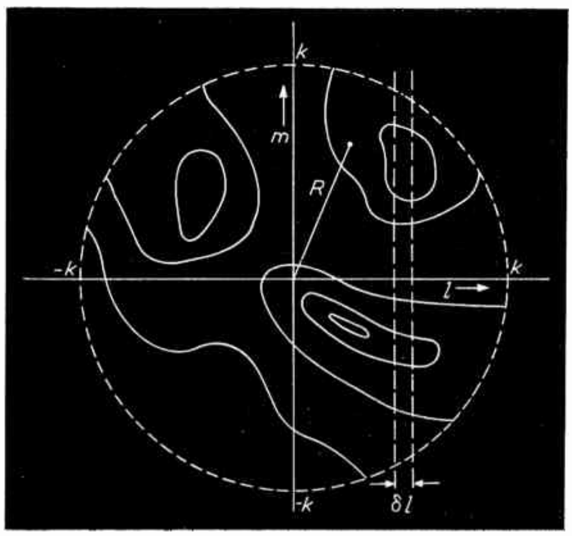
It should be observed that no incident waves can have values of l or m greater than k, Equ. (2). All the power, therefore, lies within a circle of radius k. Points on this circle correspond to radio sources on the horizon, while the origin corresponds to a source at the zenith. A point at a radial distance R from the origin corresponds to a direction whose zenith angle is $\sin^{-1} R/k$. The bearing of radio sources can be read directly from the plot if the l and m axes are drawn in the x and y directions.

The spectrum presented in Fig. 1 is of direct use in the discussions that follow. It will be appreciated, however, that when the whole of the celestial sphere is to be mapped, the results are better presented in some standard mapping form such as 'plan carrée'.

The Basic System

The element from which all the direction-finding systems are built up is the following. A pair of detectors is set at a known distance apart. The significant

Fig. 1. *The directional power spectrum. This indicates the distribution of power over the celestial hemisphere for a selected wavelength* $1/k$ *using as rectangular co-ordinates the phase rates* l *and* m



measurement is the product of the signals from these detectors, averaged over a long time.

It may seem unlikely that such average products can arise in using an ordinary array in which the signals from the various detectors are merely added so that the signal itself is merely the sum

$$V = u_1 + u_2 + u_3 + \ldots \qquad (3)$$

The signal comes, however, not from some man-made transmitter but from a source of radio 'noise'. Even though the receivers select quite a narrow frequency range, the selected signal still has the character of noise in that its amplitude is not constant. Some estimate is needed of the mean signal strength. The most usual practice is to measure the mean 'power' of the signal; i.e., to rectify the signal using a circuit with a square-law response and to record the mean d.c. level of its output. Using this method, the quantity that is recorded can be written as

$$\begin{aligned} \overline{V^2} &= \overline{(u_1 + u_2 + u_3 + \ldots)^2} \\ &= \overline{u_1^2} + \overline{u_1u_2} + \overline{u_1u_3} + \ldots \\ &\quad + \overline{u_2u_1} + \overline{u_2^2} + \overline{u_2u_3} + \ldots \\ &\quad + \overline{u_3u_1} + \overline{u_3u_2} + \overline{u_3^2} + \ldots \text{etc.} \ldots \end{aligned} \qquad (4)$$

Evidently, it is the sum of many mean squares and mean products of the signals from individual detectors.

It must, therefore, be asked how the mean product of the signals from two detectors depends upon the direction of the source of radiation. It is evident from first principles that if a coherent wave train travels in a direction at right angles to the line joining the detectors, it produces in them signals that are exactly in phase. Then the mean product of these signals is necessarily positive. On the other hand, if radiation arrives at some finite angle of incidence it may produce signals differing in phase by just half a cycle, so that the two are always of opposite sign; their mean product is then negative. At still larger angles of incidence the phase difference may be a whole cycle, so that the mean product of signals is again positive, or may be $1\frac{1}{2}$ cycles, when the product is again negative, and so on.

The following argument concerns the case where radiation is arriving simultaneously from many directions.

Suppose that two detectors, each with unit sensitivity, are set at positions separated by an interval X parallel to the x axis and Y parallel to the y axis. It will be supposed that each detector is equally sensitive to all radiation whatever its direction of travel. If each coherent wave train in the radiation gives a signal like that quoted in Equ. (1), the total signal from a detector sited at a position (x, y) may be expressed as,

$$u_1 = \sum_{1}^{r=\infty} a_r \cos 2\pi \,(l_r x + m_r y + f_r t + \alpha_r) \qquad (5)$$

The same radiation acting on the second detector whose position is $(x + X, y + Y)$ will produce a signal

$$u_2 = \sum_{1}^{r=\infty} a_r \cos 2\pi \,(l_r x + m_r y + f_r t + \alpha_r + l_r X + m_r Y) \qquad (6)$$

Before estimating the mean product of these two signals, it should be remarked that although radio noise may be regarded as an interference pattern between the inter-crossing wave trains, the noise amplitude being momentarily larger at some places than at others, observations show that the interference pattern is not stationary in space. This behaviour can be taken account of by supposing that the frequencies of the individual wave trains are all slightly different. Consider then the mean product of the above expressions for u_1 and u_2. It involves many mean products between sinusoidal signals coming from different wave trains, but if their frequencies differ such cross products disappear when a long time-average is taken. It can readily be verified that the mean product of u_1 and u_2 reduces to

$$\overline{u_1u_2} = \sum_{1}^{r=\infty} \tfrac{1}{2} a_r^2 \cos 2\pi \,(l_r X + m_r Y) \qquad (7)$$

This expression is merely the sum of the powers $\frac{1}{2}a_r^2$, of all the wave trains, each weighted by a factor that varies sinusoidally with the characters l_r and m_r that indicate the wave direction, the rates of variation being directly proportional to X and Y, the two components of spatial separation of the detectors.

It is convenient to express the result in terms of the power density in the spectrum. Because all the wave trains whose rates lie in the range $l \pm \frac{1}{2}\,\delta l$, $m \pm \frac{1}{2}\,\delta m$ have a total power $E\,(l, m)\,.\,\delta l\,.\,\delta m$, Equ. (7) may be written in the integral form

$$\overline{u_1u_2} = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} E(l, m) \cos 2\pi \,(lX + mY)\; dl\,.\,dm \qquad (8)$$

In short, when the power spectrum of the incoming radiation is pictured in the form shown in Fig. 1, a detector-pair records a weighted total of this power, the weighting factor being the sinusoid $\cos 2\pi\,(lX + mY)$. On the plot of Fig. 1 the alternate positive and negative zones of this weighting factor lie at right angles to the direction in which the detectors are separated. The spacing of the zones is inversely proportional to the separation of the detectors. In the special case where the two detectors coincide in position, the zones are so wide that the weighting factor is unity over the whole plot; in other words the mean square of the signal from a single detector is merely the total of the incident power.*

This is the main conclusion of the present section. Two minor points deserve attention however. First it will be appreciated that if the circuit that rectifies the signal V in Equ. (3) does not obey a square law, the expression for the measured output will be a more complicated expression than appears in Equ. (4). It can be shown that the sensitivity curve of the array of detectors (as defined in the sections that follow) is modified if the rectifier is made to obey a modified law.

Secondly, in order to make a proper measurement of the mean output from a detector-pair or from a whole array, the average must extend over a sufficiently long time interval, whose length in turn depends upon the width of the frequency band to which the detectors are sensitive. If the signal V is observed during a time T it may be regarded as the sum of harmonics whose frequencies are multiples of $1/T$. If the accepted frequencies lie in a bandwidth δ it means that the number of harmonics being received is $T\delta$. Because the

* From another point of view it may be remarked that $\overline{u_1u_2}$ regarded as a function of X and Y is the space correlogram of the incoming radiation while Equ. (8) indicates that the space correlogram and the power spectrum are Fourier transforms.

signal has the character of random noise these harmonics have unrelated amplitudes though the variance of each (throughout a large number of observations) is the same. The mean-square signal observed during the interval T is equal to the sum of the squares of these amplitudes, times $1/T$, and its probable error depends on the number of harmonics that have been accepted, being inversely proportional to the square root of this number. To achieve 5% accuracy, about 400 harmonics are needed. This means that the averaging time T must be about $400/\delta$. In radio studies at several megacycles/second a bandwidth of 10 kc/s is acceptable and the averaging time is then less than one-twentieth of a second. But in other studies it may be much longer. In acoustics one may wish to select a bandwidth of 100 c/s and the averaging time then rises to 4 seconds. In the study of waves on the sea, since their frequencies are very low, say between 0·05 c/s and 0·2 c/s, one would wish to use a bandwidth of perhaps 0·02 c/s in order to distinguish the different kinds of waves. The averaging time then extends to several hours.

The Action of a Line Array

This section will discuss some simple arrays in which the detectors are set out along a straight line. It will be convenient to suppose that all the detectors have the same sensitivity, taken to be unity. In setting up arrays in practice it is well known that considerable difficulty can arise in making all the detectors equally sensitive because the tuned aerial of each detector modifies the radio wave in its neighbourhood and so affects the response of the aerials nearby, altering both the amplitude and phase of the signals they receive. Such difficulties can be overcome partly by design and partly by experiment (see for instance Sections 5.20 and 5.22 in reference 20). They will not be further discussed here.

When the detectors are set out on a straight line, say the x axis, every different pair of detectors shows a separation that is merely in the x direction. Then according to the findings of the previous section, every detector-pair weights the power spectrum pictured in Fig. 1 by a sinusoidal factor whose zones lie parallel to the m axis. In effect, no distinction is made between incident radiations that have different m values providing their l values are the same. To discuss the action of a line array it is simplest to replace the two-dimensional spectrum of Fig. 1 by a single curve showing how the power distribution varies with l alone. Using a new coefficient $E(l)$ the total power of all radiation from a strip between the limits $l \pm \frac{1}{2}\delta l$ may be written as

$$\delta l \int_{-\infty}^{\infty} E(l, m)\, dm \equiv E(l) \, . \, \delta l$$

The curve $E(l)$ against l is the power distribution which a line array can aim to measure. Also from Equ. (8) it is now seen that the output of a detector-pair separated by an interval X only (Y being zero) can be written as

$$\overline{u_1 u_2} = \int_{-\infty}^{\infty} E(l) . \cos 2\pi\, l\, X . dl \qquad \ldots \quad \ldots \quad (9)$$

In this new point of view, a detector-pair takes a weighted total of the power indicated by the curve $E(l)$, the weighting factor being the sinusoid $\cos 2\pi l X$. Various line arrays will now be discussed with respect to Equ. (9).

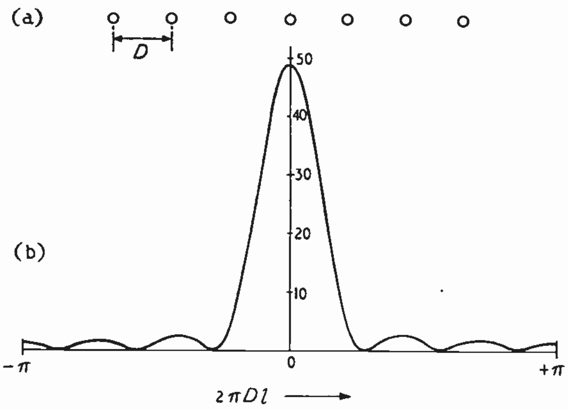


Fig. 2. *A uniform array of seven detectors (a) and its sensitivity curve (b)*

Fig. 2(a) illustrates an array of seven detectors, set at uniform intervals, D. The output of this array, the mean square of the sum of the signals is

$$\overline{V^2} = \overline{(u_1 + u_2 + \ldots u_7)^2}$$

If this is expanded, each term can be expressed in the form of Equ. (9). Seven of the terms are mean squares, for which the separation X must be written as zero. Twelve of the terms are products between signals from detectors at a separation D. Ten terms are products corresponding to a separation $2D$, and so on. The output of the array becomes

$$\overline{V^2} = \int_{-\infty}^{\infty} E(l)\, [7 + 12 \cos 2\pi Dl + 10 \cos 4\pi\, Dl + \ldots + 2 \cos 12\pi\, Dl]\, dl \quad \ldots \quad \ldots \quad (10)$$

The output of the array is the integral of the power coefficient $E(l)$ weighted by the factor in square brackets. This factor varies with the wave direction l. Its greatest value occurs at $l = 0$. It is in fact, the sensitivity function of the array in Fig. 2(a), and its form is sketched in Fig. 2(b). Evidently the array is most sensitive to radiation whose l values are close to zero; i.e., radiation incident within a fairly small range of angles about the normal to the line of the array. For example, if the basic interval D is made equal to half a wavelength, or $1/2k$, then it may be seen from Fig. 2 that the higher sensitivities lie approximately in a range where l/k is between $\pm 1/6$. Since l/k is the direction cosine, such radiation is incident at angles between $\pm 10°$ about the normal to the array. By analogy with transmitting systems this zone of high sensitivity is often called the 'beam' of the array.

It should also be noted that Fig. 2 only illustrates the sensitivity curve in the range of l between $\pm 1/2D$. It could, of course, be plotted beyond the range and it would be found to have a repetitive form. Peaks of high sensitivity occur near to l values of $\pm 1/D$ and $\pm 2/D$ and so on. If radiation with such l values were incident on the array, the array would accept all of them; in fact, a badly-designed array may accept radiation from many different directions; it may have multiple beams. It is usual, however, to tune the detectors so that they only accept radiation close to a certain frequency, whose wavelength is near to a certain value $1/k_0$. Whatever its direction of travel, such radiation cannot show a phase rate along the line of detectors that exceeds k_0. To avoid multiple beams corresponding to l values of $\pm 1/D$, $\pm 2/D$ and so on, the array must be designed so that the spacing D is less than $1/k_0$, a wavelength. In some cases it is advisable to make D equal to half a wavelength.

This example shows that the sensitivity curve of an array can readily be written down from a knowledge of the positions of the detectors. Every different arrangement of detectors produces an array with a different sensitivity curve because each new arrangement involves detector-pairs having new separations and giving rise to new sinusoids in the curve.

It is of interest to compare this uniform array with an unsymmetrical array used by Arsac, where the second, fourth and fifth detectors are absent. This is shown in Fig. 3(a). The output of this array is

$$\overline{V^2} = \overline{(u_1 + u_3 + u_6 + u_7)^2}$$

If this is expanded and the various terms are expressed by Equ. (9) the expression for the output is readily found to be

$$\overline{V^2} = \int_{-\infty}^{\infty} E(l)\,[4 + 2\cos 2\pi Dl + 2\cos 4\pi Dl + \ldots + 2\cos 12\pi Dl]\,dl \quad \ldots \quad \ldots (11)$$

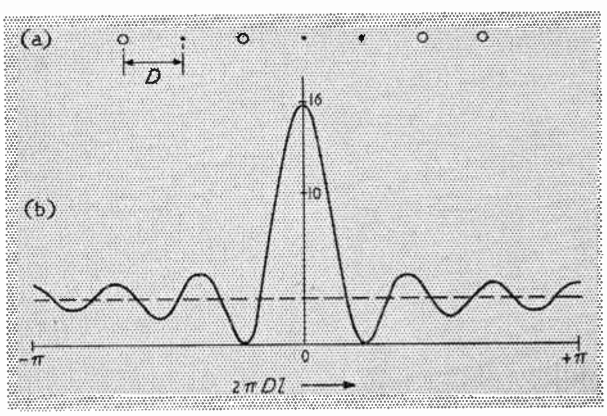
The sensitivity function of this array is the expression in square brackets and it is sketched as a curve in Fig. 3(b). Its central peak is narrower than that of the uniform array but the side lobes are larger. The advantages which can result from this different form of curve will be discussed in the next section.

The improved interferometer of Ryle[12] can be examined in a similar way. His arrangements are indicated in Fig. 4(a). The two elements of the interferometer are aerials with a moderate directivity. They are represented in Fig. 4 as uniform arrays of seven detectors so that each has the sensitivity curve shown in Fig. 2(b). Ignoring for the moment the switch and the phase-sensitive rectifier, the signals V_1 and V_2 from the two arrays are added and are then passed to a circuit which provides the mean square of their sum, namely

$$\overline{(V_1 + V_2)^2} = \overline{V_1^2} + 2\,\overline{V_1 V_2} + \overline{V_2^2} \quad \ldots \quad \ldots (12)$$

When the switch reverses the sign of V_2 it also reverses

Fig. 3. *An unsymmetrical array of four detectors (a) and its sensitivity curve (b)*



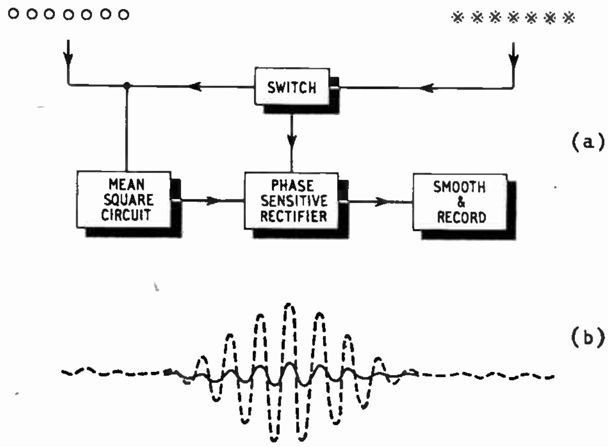


Fig. 4. *Ryle's interferometer (a) and the signal from a point source (b)*

the sign of the output from the phase-sensitive rectifier, so the output becomes

$$-\overline{(V_1 - V_2)^2} = -\overline{V_1^2} + 2\,\overline{V_1 V_2} - \overline{V_2^2} \quad \ldots (13)$$

When the switch is in continuous operation, the smoothed output appearing on the recorder is the mean sum of these two quantities; i.e.,

$$-\overline{(V_1 - V_2)^2} + \overline{(V_1 + V_2)^2} = 4\,\overline{V_1 V_2} \quad \ldots (14)$$

The record is therefore proportional to the mean product of the two signals. The system acts like a single detector-pair whose elements are the directive arrays. If the two elements of the pair were non-directive, Equ. (9) shows that the system would have a sensitivity curve of $4 \cos 2\pi lX$ which oscillates rapidly between positive and negative values because X is large. When this curve is multiplied by the sensitivity curve of the individual arrays, it gives the curve shown as a broken line on Fig. 4(b) which has a number of alternate positive and negative lobes.

The advantages of Ryle's device appear when it is used to view a weak radio 'star' among a broad source of radio noise. Because the lobes of the sensitivity curve have alternate signs the broad distribution of noise produces a negligible signal, but as the weak radio 'star' moves through the beam of the interferometer it produces a weak signal, sketched as a full line in Fig. 4(b), and this signal can, if necessary, be amplified to give a measurable record. If on the other hand the switch were not used, the recorded output would be given by Equ. (12). This involves three terms and, using Equ. (9) for each one, the sensitivity curve is readily found to be

$$2\,(1 + \cos 2\pi lX) \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots (15)$$

This is essentially positive. A broad source of radio noise would produce a large deflection in the recorder, and the signal from a weak star would be present only as an additional small oscillation.

To conclude this section it may be remarked that a similar treatment can be used to find the sensitivity curves of arrays whose detectors do not have equal sensitivity or may introduce phase changes into the signals they produce. Arguments very similar to those given in the section on 'The Basic System' can be used to show that if two detectors sited at positions x_1 and x_2 have sensitivities C_1 and C_2 and introduce phase reductions equal to p_1 and p_2 fractions of a cycle then the output of this detector-pair is the following weighted total of the incident power,

$$\overline{u_1 u_2} = \int_{-\infty}^{\infty} E(l) \,.\, C_1 C_2 \cos 2\pi \,[l\,(x_1-x_2)-p_1+p_2]\, dl \quad .. \; (16)$$

The output of any array can be found by adding appropriate expressions of this type. One particular use of phase shifts is to swing the beam of an array. The three arrays that have already been discussed are most sensitive to radiation that travels approximately at right angles to the line of detectors; i.e., in directions for which l is small or zero. It may be desired however to select a wave which is not incident normally but exhibits a phase rate l_0 along the line of detectors. The signal produced by such a wave in a detector at position x would have a phase exceeding by $l_0 x$ cycles the phase of the signal produced in a detector at the origin. Obviously, one might take account of this by introducing phase reductions p_1, p_2, etc., into all the signals, which are numerically equal to $l_0 x_1$, $l_0 x_2$, etc. The signals developed by all the detectors would then be in phase. Substituting these values in Equ. (16) it appears that a typical detector-pair would then give an output

$$\overline{u_1 u_2} = \int_{-\infty}^{\infty} E(l) \,.\, C_1 C_2 \cos 2\pi \,(l-l_0)\,(x_1-x_2) \quad .. \; (17)$$

This output is greatest when $l = l_0$; the array accepts the wanted wave.

This particular system of phase reductions does not alter the form of the sensitivity function as illustrated in Figs. 2 and 3, but merely translates it along the axis of l so that its major peak falls at a value l_0 instead of at zero. The beam of the array then points in this new direction. To explore radiation coming from many directions the phase reductions may be slowly changed, varying l_0, so that the beam moves over the whole range of directions. A record of the output plotted against l_0 then nominally represents the distribution of incident power though, in fact, it will only be an approximation to the true distribution curve. It must be recalled that the sensitivity curve is repetitive and has a number of major peaks separated by intervals $1/D$ on the scale of l. On the other hand the incident radiation, whose wavelength is say $1/k$, may show phase rates l which lie anywhere in the range $\pm k$. If no two of the major peaks of the sensitivity curve are to fall at the same time within this significant range (and if they do so the array will have two beams and accept radiation from two directions at once) the interval $1/D$ must be greater than $2\,k$. In other words, an array that is to be used in this way must be designed so that the basic interval D is less than half a wavelength.

In general, changes in the sensitivity or changes in the phase of individual signals cause changes in the form of the sensitivity curve. Thus Schelkunoff[26] showed that side lobes could be avoided by using sensitivities proportional to the coefficients of a binomial series. For instance, if the seven detectors pictured in Fig. 2 were given sensitivities in the ratio 1 : 6 : 15 : 20 : 15 : 6 : 1, the new sensitivity curve of the array would show no side lobes at all, though its central lobe would be appreciably wider than that pictured in Fig. 2. Dolph[21] showed how to create arrays which minimized the width of the central lobe of the sensitivity curve when the maximum height of the side lobes was specified. The theory of 'super-directive' arrays has shown that the directivity of an array of a given overall length can be increased without limit at the expense of considerable practical difficulties.

The results that can be achieved by changing the phases and the amplitudes of the signals from the various detectors have been explored very fully in the literature (see for instance Sections 5 and 6 in reference 20). These techniques do not, however, lead to arrays with the 'optimum' qualities discussed in the next section, and for these reasons no further space will be given to them in the present article.

The 'Optimum' System

It is a matter of some importance to decide what form of sensitivity curve is best when so many different forms are possible in directional systems. In general one would wish the curve to have as narrow a central peak as possible, and at first sight it would seem that side lobes are undesirable. Arsac has given a specific answer to this question and it may seem to be an unexpected one. The sensitivity curve which he finds most desirable is sketched in Fig. 5. It possesses many side lobes, half of them being negative.

The argument is as follows. The purpose of a directional array is to measure the power distribution $E(l)$ as accurately as possible. It is known that the power must be zero for values of l outside the range $\pm k$ if $1/k$ is the accepted wavelength. Within this range the

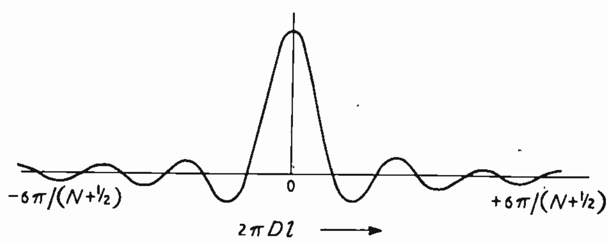


Fig. 5. *The central portion of an 'optimum' sensitivity curve*

curve $E(l)$ may be written as a Fourier series provided that its repetition interval in l is greater than $2k$. It is convenient to adopt $1/D$ where D is the basic spacing of the detectors in the array. Such a series has the form

$$E\,(l) = A_0 + A_1 \cos 2\pi\, Dl + A_2 \cos 4\pi\, Dl + \ldots + B_1 \sin 2\pi\, Dl + B_2 \sin 4\pi\, Dl \ldots \quad (18)$$

Now the output of a detector-pair that is phased to accept waves with a phase rate l_0 has been given in Equ. (17). Suppose that the space interval between the detectors is some integral multiple of D, say nD, and that they have unit sensitivity. Using Equs. (17) and (18) the output is found to be

$$\overline{u_1 u_2} = \int_{-1/2D}^{1/2D} E(l) \cos 2\pi\,(l-l_0)\, nD \,.\, dl$$
$$\begin{cases} = \dfrac{1}{2D}\,(A_n \cos 2\pi\, l_0\, nD + B_n \sin 2\pi\, l_0\, nD);\ \text{if } n \neq 0 \\ = A_0/D;\ \text{if } n = 0 \quad .. \quad .. \quad .. \; (19) \end{cases}$$

Thus if the phasing constant l_0 is slowly varied over the whole range from $-k$ to $+k$ it appears that a single detector-pair will generate a sinusoidal curve which is identical in amplitude and phase (apart from the factor $1/2D$) with the nth harmonic of the power distribution.

An ideal array would be one which combined the outputs of a large number of detector-pairs at separations 0, D, $2D$, etc., so that its output was the sum of all the harmonics of $E(l)$ and was, in consequence, an exact copy of the $E(l)$ distribution. In general, this is impossible because an array of overall length ND cannot represent harmonics of a higher order than the Nth. Only when these are absent (so that the distribution of power is fairly smooth) can the ideal be attained.

The aim would be to make the array give the minimum error, using harmonics of order up to N, and a minimum mean-square error is a reasonable objective. It is a well-known property of the Fourier series that the amplitudes and phases it prescribes for the first N harmonics are the best way in which this limited number of sinusoids can be combined to approximate to the true curve. Arsac's conclusion is that the array should merely sum the first N harmonics of the series. According to Equ. (19) this calls for one detector-pair at zero separation and two detector-pairs at all separations D, $2D$ . . . to ND.

Such an array would have the 'optimum' sensitivity function.

$$1 + 2\sum_{1}^{n=N} \cos 2\pi nDl = \frac{\sin 2\pi\,(N+\frac{1}{2})\,Dl}{\sin \pi Dl} \qquad .. \quad (20)$$

The shape of this sensitivity curve tends to the form in Fig. 5 when N is large so that the array is long. It is worth emphasizing that this is only called an 'optimum' for the purpose for which it is designed; i.e., to explore a continuous distribution of incident radiation with the minimum mean-square error. It is not the best curve for a transmitting system, indeed it is an impossible one, for it would require negative power to be transmitted in a number of directions. As transmitters, the systems devised by Dolph[21] are called optimum systems with good reason. Even for receiving arrays the sensitivity curve in Fig. 5 is not always the best. When an experimenter has good reason to believe that he is studying a collection of point-sources of radiation his object is to guess from the height and position of the peaks in his experimental curve, the strengths and directions of the sources. The numerous side lobes of an 'optimum' system make it unsuitable for such work. The arguments in favour of the 'optimum' curve no longer hold good because the experimenter is engaged in reading into the experimental curve information which it does not contain. But it should be noted that the record of power of distribution given by an 'optimum' array contains more, not less, information than would be given by any other system with the same overall length. By a suitable smoothing of the final record an 'optimum' array can be made to act as if it were a non-optimum system with side lobes small or absent.

The next sections describe some realizations of the 'optimum' system.

Some Optimum Arrays in One Dimension

The asymmetric array in Fig. 3 is easily converted to an 'optimum' system. Equ. (11) shows that the constant term in the sensitivity function is too large, and that the output of the array contains the extra power $3\int E(l).dl$. This is merely three times the power given by a single detector and it does not vary as the signals are phased to explore the various directions (or as the celestial radio sources move past the beam of the array). To convert this array to an 'optimum' system it is only necessary to subtract this constant amount from the recorded curve of power.

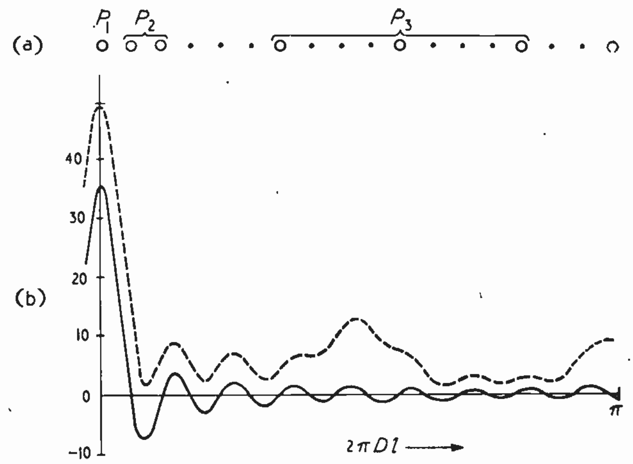


Fig. 6. *An unsymmetrical array of seven detectors (a) and the 'optimum' sensitivity curve (b) obtained by using it as a compound array with output* $P - P_1 - P_2 - P_3$

With so few detectors as 4 one must either be content with a very-low order resolution, or else restrict attention to a very small arc of directions by using detectors that individually are very directive. Arsac has used this method to measure the distribution of radio noise across the face of the sun.

Asymmetric arrays can be designed with a much larger number of detectors but they are not so readily brought to the 'optimum' form. Fig. 6(a) shows an example using seven detectors in an array 17-units long. If it were an optimum system its sensitivity curve would be given by Equ. (20), writing N as 17. This is sketched as the full line in Fig. 6(b). Inspection shows however that there are a number of supernumerary pairs at intervals 0, 1, 4 and 8 units. The sensitivity function of the array exceeds the optimum by terms

$$6 + 2\cos 2\pi Dl + 4\cos 8\pi Dl + 2\cos 16\pi Dl$$

and, by addition, it is obtained as the broken line in Fig. 6(b). The excess terms can be grouped, however, as

$$1 + [2 + 2\cos 2\pi Dl] + [3 + 4\cos 8\pi Dl + 2\cos 16\pi Dl]$$

which are the sensitivity functions of the three sub-arrays indicated in the diagram. If outputs can be developed simultaneously and in separate channels from the whole array P and from the sub-arrays P_1, P_2 and P_3, the combination $P - P_1 - P_2 - P_3$ is an output having the optimum curve shown in Fig. 6(b) as a full line. Some further arrangements of this kind have been suggested by the author[22].

The 'sum and difference' method used by Ryle is a fruitful source of optimum systems. Before discussing these systems it should be noted that an alternative expansion for the power distribution in the range $\pm 1/2D$ in l is

$$E(l) = C_1 \cos \pi Dl + C_2 \cos 3\pi Dl/2 + C_3 \cos 5\pi Dl/2 + \ldots$$
$$+ D_1 \sin \pi Dl + D_2 \sin 3\pi Dl/2 + D_3 \sin 5\pi Dl/2 + \ldots \quad (21)$$

and it leads to the expression

$$\sum_{1}^{n=N} \cos 2\pi \,(n - \tfrac{1}{2})\, Dl = \frac{\sin 2\pi\, NDl}{\sin \pi\, Dl} \quad (22)$$

as an alternative form for the 'optimum' sensitivity function.

Consider then the various dual systems indicated in Fig. 7. In each case the asterisks represent the detectors of one array and the full circles represent the detectors of the companion array. In each case a switching system is to be used after the manner of Ryle. As explained in the section on 'The Action of a Line Array', this is equivalent to recording the mean product of the voltage signals from the two arrays.

In example 7(a) there is only one detector in the first array. The mean product of signals can be written

$$\overline{u_0\,(u_1 + u_3 + \ldots + u_{23})}$$

The various pairs in this expansion have separations ranging from $\frac{1}{2}D$, $1\frac{1}{2}D$, etc., to $11\frac{1}{2}D$. Each contributes a sinusoid to the sensitivity function and it turns out that this is the 'optimum' function quoted in Equ. (22), writing N as 12.

In example 7(b) there are two detectors in the first array. The mean product of signals is

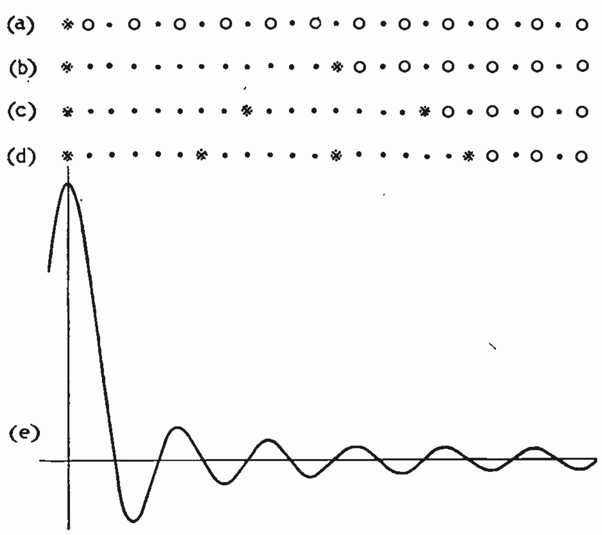
$$\overline{(u_0 + u_{12})\,(u_{13} + u_{15} + \ldots u_{23})}$$

When the various terms are examined it is found once again that the system has the optimum sensitivity function. In fact, all the four systems have the same optimum sensitivity curve, which is sketched in Fig. 7(e).

Arrangement 7(b) in fact corresponds to the arrangement described by Covington and Broten[16], the isolated detectors being dish receivers and the uniform array being a long slotted waveguide.

It may be noted that arrangement 7(a) calls for 13 detectors, arrangement 7(b) for 8, and arrangements 7(c) and 7(d) only require 7. Where individual detectors are expensive and an arrangement of this type is being planned it would be economical to use about the same number of detectors in each of the two arrays. However, if an arrangement similar to that in Fig. 6 can be adopted, the same number of detectors can provide an 'optimum' array of greater overall length.

Fig. 7. *Various dual arrays, a, b, c and d which all have the same 'optimum' sensitivity curve (e)*



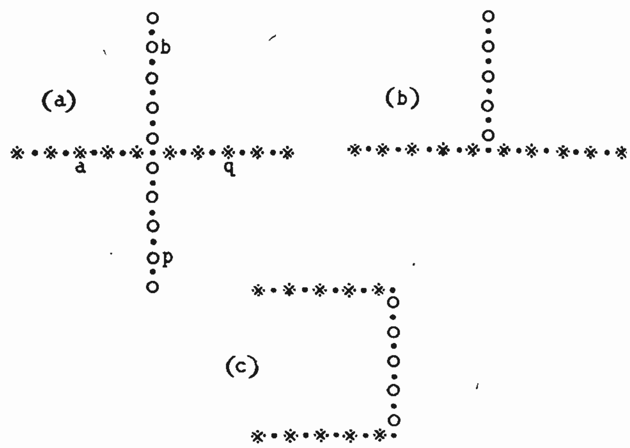


Fig. 8. *Dual systems which all have the same 'optimum' sensitivity curve in two dimensions; the Mills cross (a) and some alternatives (b) and (c)*

Some Optimum Arrays in Two Dimensions

The Mills cross[13] is a well-known example of an array which gives resolution in two dimensions and is said to have a 'pencil' beam. On examination it turns out to be an 'optimum' system in two dimensions. The arrangement is illustrated diagrammatically in Fig. 8. Two uniform arrays are placed at right-angles. In this illustration it is supposed that each array carries an even number of detectors so that their distances from the central point of the arrangement are all expressible as $(n - \frac{1}{2})\, D$, where n is an integer and D is the spacing. The two arrays are used in a 'sum and difference' way, after the manner of Ryle, and the measured quantity is the mean product of the signals from the two arrays. As before, this can be expanded into the sum of many mean products between pairs of individual signals. The spatial separation of these various pairs must now be expressed in two co-ordinates corresponding to X and Y in Equ. (8). If there are $2N$ detectors in each array (i.e., N in each arm) the X separations and the Y separations range independently between $\pm\,(N - \frac{1}{2})\, D$ in unit steps of D. Using Equ. (8) it is readily seen that the gain curve of the system is

$$\sum_{-N+\frac{1}{2}}^{r=N-\frac{1}{2}} \; \sum_{-N+\frac{1}{2}}^{s=N-\frac{1}{2}} \cos 2\pi\,(rl + sm)\, D\,; \; r \text{ and } s \text{ integers}$$

By analogy with Equ. (22) in the previous section this is seen to have the 'optimum' form.

On further examination it is seen that the terms in this gain curve all occur in pairs, since cos α and cos $(-\alpha)$ are identical. Indeed it is evident that a pair of detectors such as 'a' and 'b' in Fig. 8(a) has just the same separation as the pair 'p' and 'q'. It appears that some of the detectors are redundant. Fig. 8(b) shows a modified form of the Mills cross; one of the four arms (negative y) has been omitted and the signal strength is reduced to one half, but the form of the gain curve is the same as that of the complete cross in 8(a).

The arrangement 8(c) uses the same number of detectors as 8(b) and has the same optimum sensitivity curve as the other two arrangements, the intervals presented by the various pairs of detectors being the same as before. It occupies a slightly smaller space than does arrangement 8(b). A disadvantage of arrangement 8(c) is that the side lobes of the sensitivity curve cannot be reduced by attenuating the signals of the detectors that are remote from the centre of the system. Arrangement 8(c) calls for some automatic smoothing of the recorded curve of power if it is to be adopted for resolving point-sources (see for instance reference 3).

Synthesizing the Power Distribution

It has been recognized for some time that not more than two detectors are essential in plotting the distribution of power in as great detail as can be produced with an extensive array. This great economy in apparatus is obtained at the expense of considerable computational work on the part of the experimenter.

The principle is as follows. It has been seen that the power distribution can be written as a Fourier series. Providing that this distribution does not change during the course of the measurements, a detector-pair can be used to determine in succession the various coefficients of the series. These being known, the distribution can be computed.

Thus Equ. (19) shows that if a pair of detectors is set at a separation nD and no phase changes are introduced into the signals, $l_0 = 0$, the output of the detector pair is $A_n/2D$.

If, on the other hand, the phase of the second detector, the one at the greater x co-ordinate position, is reduced by 1/4 cycle, the output of the detector pair becomes $B_n/2D$.

These two observations therefore provide measures of the two Fourier coefficients A_n and B_n. By repeating the observations at all intervals from 0, D, etc., to ND, harmonics up to the Nth order can be determined and an 'optimum' representation of the power distribution can be got by calculation.

The early use of this principle by Stanier[8] was defective because he did not introduce phase shifts to discover the coefficients B_n. His results, therefore, show only the symmetrical part of the actual distribution. The need for such phased measurements was pointed out by O'Brien[9] who extended the idea to two dimensions in work on radio astronomy. A more convenient application of the principle to radio astronomy is described by Blythe.[11] He uses a 'sum and difference' switch to combine the signal of a long uniform array with the signal from a single detector which on successive nights occupies positions corresponding to those of the second array in a Mills cross. In effect, the long array focuses attention on a strip of sky and the outputs of the dual system record the varying values of A_n and B_n as the radio sources cross the strip of sky. The records are then converted by computation to a plot of power density on the celestial sphere.

It may be noted that these two measurements of A_n and B_n can be regarded as the measurement of a single complex quantity $(A_n + jB_n)$ which is one value of the complex space correlogram of the incoming radiation. The method is then an example of the Weiner-Khintchine theorem (see for instance Section 6.2 of reference 23) which may be interpreted as showing that the space correlogram and the directional power spectrum are Fourier transforms. It may be convenient experimentally to measure the complex correlation coefficient from the form of a scatter diagram presented on the screen of a cathode-ray tube[24].

Wide-Band Systems

Sources of noise usually produce radiation covering a wide range of frequencies. All the systems that have been discussed so far are intended to select for study quite a narrow-frequency band. Nodtvedt[17] appears to have been the first to point out the possible advantages in radio or acoustic work of accepting a wide range of frequencies. The principle has been used experimentally by Vitkevich[25].

If a wide band is accepted more energy is available for study but, apart from this, it appears that in favourable circumstances a single pair of spaced detectors may give as good a resolution as a complete array of the same overall length working at the highest frequency in the range.

The previous discussion has classified incident radiation according to its rates of change of phase (l and m) along the rectangular x and y axes, but interest really lies not in the phase rates of the radiation but in its direction of travel in space. Wide-band systems become useful in those studies where all the various frequencies show the same distribution in space.

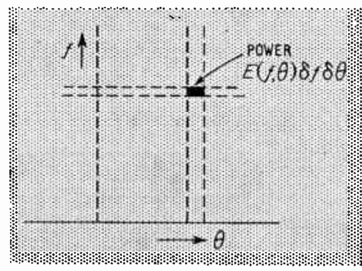


Fig. 9. *The plot of power distribution in both frequency and direction*

It has been seen, Equ. (9), that a detector-pair with separation X weights the power of the incoming radiation by a sensitivity factor cos $2\pi xl$. The phase rate l can be expressed in terms of the wave velocity, v, the wave frequency f and the angle of incidence θ. Providing θ is small l may be written as

$$l = f\theta/v \qquad \text{.. .. ..} \quad (23)$$

The sensitivity factor becomes

$$\cos 2\pi x f \theta/v \qquad \text{.. .. ..} \quad (24)$$

On a scale of directions θ this sinusoid varies rapidly for radiation whose frequency is high but varies only slowly for radiation whose frequency is low. It is not surprising, therefore, that when a detector-pair is made to accept all radiation in a wide band of frequencies, its effective sensitivity curve in favourable circumstances may approximate to the sum of very many sinusoids of different rates and may even resemble the 'optimum' sensitivity curve in Fig. 5.

To discuss the use of wide-band systems more exactly, one must consider the distribution of power with both direction and frequency as it might be pictured in some rectangular (f, θ) plot such as Fig. 9.

Write the power density in this (f, θ) plot as $E(f, \theta)$. The power in any small region $\delta f \delta \theta$ can then be written alternatively as

$$E(l)\,\delta l \equiv E(f, \theta)\,\delta f \delta \theta \quad .. \quad .. \quad .. \quad (25)$$

where l, f and θ are related by Equ. (23). Making these substitutions in Equ. (9), the output of a detector-pair that accepts all frequencies from 0 to f_0 becomes

$$\overline{u_1 u_2} = \int_0^{f_0} \int_{-\alpha}^{\alpha} E(f, \theta) \cos 2\pi x f \theta / v \,.\, df\, d\theta \quad .. \quad (26)$$

The angular limits $\pm\,\alpha$ are arbitrary so long as they are greater than the width of the sensitivity function which appears later in Equ. (28). If it can be assumed that the power density varies only with direction and is the same for all frequencies in the accepted range, the integral becomes

$$\overline{u_1 u_2} = \int_{-\alpha}^{\alpha} f_0 E(f, \theta) \frac{\sin 2\pi X f_0 \theta / v}{2\pi X f_0 \theta / v} \,.\, d\theta \quad .. \quad (27)$$

Now $f_0 E(f, \theta)$ is the total energy per unit angle in the accepted range of frequency and Equ. (27) shows that the sensitivity function has the 'optimum' form

$$\frac{\sin 2\pi X f_0 \theta / v}{2\pi X f_0 \theta / v} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (28)$$

Comparing this with Equ. (21) it appears that the detector-pair working on a wide band has a sensitivity function closely resembling that of a complete 'optimum' array of the same overall length, working at the maximum accepted frequency. The comments made by Koch and Stone[18] on this matter appear to be in error.

Nodtvèdt also calculates sensitivity patterns for intermediate cases in which the detectors accept less than the whole range of frequencies. Though the method must be restricted to studies in which the power distribution is very similar over a wide range of frequency, its possibilities need to be borne in mind.

The Linear Scale Size of an Array: Super-Directive Arrays

Except for wide-band systems, the arrays discussed so far have sensitivity curves that are repetitive in l. For example, the uniform array of seven detectors shown in Fig. 2 has a sensitivity curve which shows peaks at values of l equal to every integral multiple of $1/D$. This is illustrated in Fig. 10(a). These extra peaks do not matter so long as they do not correspond to any possible values of l in the incident radiation. The significant values of l are limited to $\pm\,k$ if $1/k$ is the wavelength corresponding to wave trains whose angles of incidence are $\pm\,90°$. One way of using an array is to introduce phase shifts into the signals, as described in the section on line array action. The sensitivity curve is then shifted bodily along the scale of l, so that the central peak falls at l_0 instead of at zero, and the array then accepts the waves incident at an angle which gives them a phase rate l_0. If the full range of possible wave directions is to be explored in this way, it is clear that the main peaks of

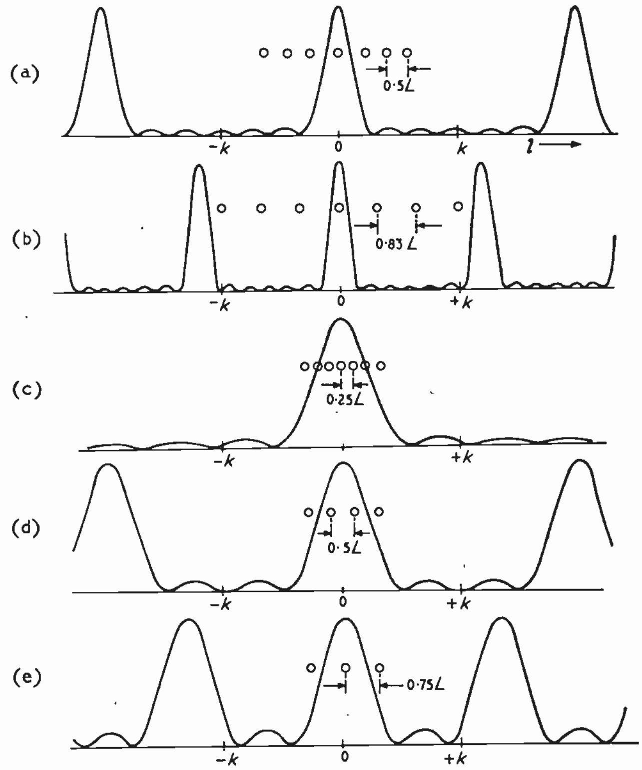


Fig. 10. *A uniform array of seven detectors (a), at wide spacing (b), and close spacing (c). The latter is ineffective in the significant range of l and other systems using four detectors (d), or even three detectors (e), are as good*

the sensitivity curve must be separated by an interval of l greater than $2k$ or else two of the the peaks may fall in the range $\pm\,k$ so that the array accepts waves from two main directions at the same time. This condition means that $1/D \geqslant 2\,k$ showing that D, the unit spacing in the array, must not be greater than $1/2\,k$ or half a radio wavelength. This is the arrangement illustrated in Fig. 10(a).

In radio astronomy, however, the celestial radio sources continually pass overhead and their distribution on the celestial sphere can be examined without shifting the beam of the array. An improvement in directivity can then be obtained by increasing the dimensions of the array without altering its form. An increase in D brings the peaks closer together because the repetition interval is $1/D$. The central peak remains at the origin and to avoid ambiguity it is only necessary that the repetitive peaks should fall just outside the significant limits $\pm\,k$, as illustrated in Fig. 10(b). This condition means that $1/D > k$ or that the spacing D shall be somewhat less than a radio wavelength. This change nearly doubles the length of the array and its directivity or resolving power is increased by the same factor, the central peak being narrower than it was before.

It may be, however, that the array has to be set up in a limited space. What happens then if the scale of the array is reduced so that D is much smaller than half a wavelength? Evidently the repetition interval $1/D$ is increased and the angular resolving power of the

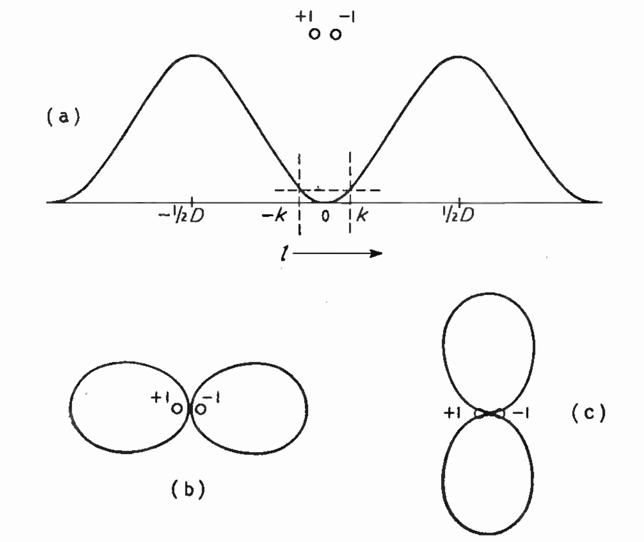


Fig. 11. *Super-directive systems have sensitivity curves which take large values outside the significant range of l. Curve (a) is the simplest example, the curve of a pair of detectors closely spaced. Used as an end-fire system the polar diagram is (b). As a broadside system the polar diagram is (c)*

array is less than before. The case where D is one quarter of a wavelength is illustrated in Fig. 10(c). The array now has an overall length of only three half-wavelengths. Compressing the array in this way makes it ineffective. The central peak is wide, and the seven detectors are being used to make the sensitivity curve low over large intervals of l outside the significant range $\pm k$. The values of the sensitivity curve outside the range $\pm k$ do not matter. Equally good directivity in the range $\pm k$ could be obtained by four detectors spaced at intervals of half a wavelength. The sensitivity curve is shown in Fig. 10(d). Even three detectors spaced at intervals of three-quarters of a wavelength are as good; their sensitivity curve is shown in Fig. 10(e).

But if the only significant part of the sensitivity curve is that within the range $\pm k$, further possibilities appear. An array might show extremely large values of sensitivity outside the range $\pm k$ and yet be acceptable. This is the basis of the 'super-directive' systems. It has been seen that the sensitivity curve of an array is the sum of sinusoids, and super-directive systems are no exception to this rule. The simplest super-directive system consists of two detectors whose sensitivities are $+1$ and -1. The sensitivity curve of this arrangement may be written down by inspection in the manner used for the previous arrays. It is $2\,(1 - \cos 2\pi Dl)$, and is shown in Fig. 11(a). When the detectors are very close together the significant range of l only includes a small region around the minimum of the curve. In this region the sensitivity is approximately $(2\pi Dl)^2$. It is an 'end-fire' system because the sensitivity is greatest when l equals $\pm k$ corresponding to waves travelling along the array in either direction. In terms of the actual direction, θ

$$l = k \sin \theta$$

so the sensitivity can be written as $(2\pi Dk)^2 \sin^2 \theta$. It is shown as a polar diagram in Fig. 11(b).

If arrangements are also made to measure the power output from a single detector, and a small fraction of this output is subtracted from the output of the array, this is equivalent to subtracting a constant from the sensitivity curve. Let this constant be $(2\pi Dk)^2$. The overall output is then $-(2\pi Dk)^2 \cos^2 \theta$ and the arrangement acts as a broadside system. Its polar diagram is shown in Fig. 11(c).

In both arrangements the maximum sensitivity is $(2\pi Dk)^2$ which is much smaller than that of a single detector if the separation D is only a small fraction of the wavelength $1/k$. Super-directive systems all tend to have a sensitivity that is much lower within the significant range of l than outside it.

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Network Characteristics

SOURCE RESISTANCE, NETWORK STRUCTURE AND TRANSFER FUNCTIONS

By J. T. Allanson.*

Griffiths and Mole[1] have suggested that there is "a common network transformation" by which "the same response is obtained from networks of the same structure working between (a) a source of zero resistance and a finite load resistance, and (b) source and load both of finite resistance." In this article it will be shown that networks exist for which such a transformation is not possible and that, therefore, there can be no general theorem of this nature. Furthermore, it can be shown that it is not possible to make simple distinctions between

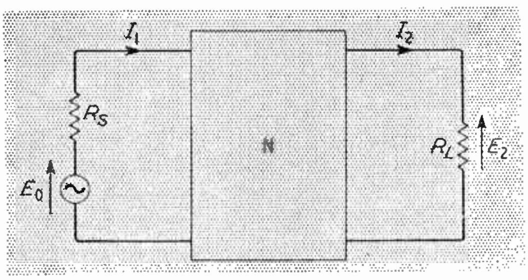


Fig. 1. *Two terminal-pair networks N with associated source and load resistances*

all networks for which the theorem is true and all those for which it fails. Clearly, the theorem is true for networks having a purely-resistive input impedance of constant value, or such that a resistance may be extracted and placed in series with the source. However, this is not a comprehensive description of the networks for which the transformation is valid and the networks discussed by Griffiths and Mole do not fall into this category.

The notation used is shown in Fig. 1. The term 'transfer function' will be used to denote the functions of the complex frequency p given by either E_2/E_0 or I_2/E_0. If two networks have transfer functions differing only by a scalar multiplying constant, the networks and transfer functions will be referred to as being 'similar'. Networks are regarded as having the same structure if one may be transformed into the other by changing the value, but not the nature or connections of any of the elements.

The suggested theorem might be held to be valid in only one way; i.e., it is always possible to transform from zero to finite source resistance but not to transform from finite to zero resistance, or vice versa. These transformations will be considered in turn, in order to dispose of this possibility. Finally, attention is given to the relationships between transfer function, source impedance and network structure for some functions realizable by RC networks, as a simple example illustrative of the more general problem.

Transformation from a Finite to a Zero Source Resistance

Suppose a network to have a branch of impedance $Z_s(p)$ connected in shunt with the source. If the source has finite internal resistance, the zeros of the transfer function will include the zeros of $Z_s(p)$. However, if the source is of zero internal impedence, the branch Z_s has no effect on the transfer function and may be removed without effect. Thus it is possible to construct a class of networks including shunt branches of such a nature

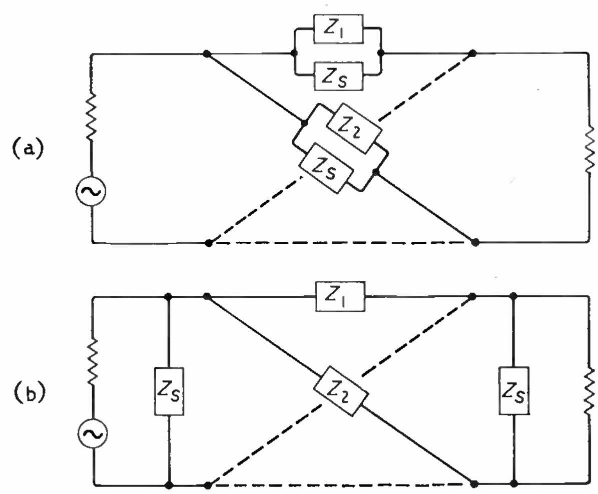
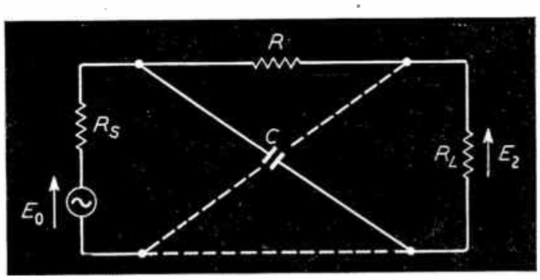


Fig. 2. *Lattice network (a) equivalent to circuit (b)*

Fig. 3. *RC lattice network with source resistance R_s and load resistance R_L*



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that the transition from a finite to a zero source resistance changes the nature of the transfer function.

In certain cases a network has an equivalent which will contain a shunt arm of the kind described above. The lattice of Fig. 2(a) is of this type, since it is equivalent[2] to the circuit of Fig. 2(b), and values may be given to Z_1, Z_2 and Z_3 to make the suggested source

If R_s is zero, this expression reduces to

$$\frac{E_1}{E_0} = \frac{kZ_1 Z_b (Z_c + Z_a)}{kZ_1 Z_b (Z_c + Z_a) + Z_a [kZ_c (Z_1 + Z_b) + Z_1 Z_b]} \quad \ldots \quad (3)$$

Equ. (2) shows that the function E_1/E_0 will necessarily possess the zeros of Z_c, but if $R_s = 0$ Equ. (3) shows that this property will not necessarily be maintained. A simple example of this general circuit is given in Fig. 5. The transfer function of this network is given by

$$\frac{E_2}{E_0}(p) = \frac{R_L (1 + p\, C_1 R_1)}{p^3 R_s L_1 C_2 R_1 C_1 + p^2 [LC_2(R_1 + R_s) + LC_1 R_1 + C_1 C_2 R_1 R_L R_s] + p[L + C_1 R_1 (R_L + R_s) + C_2 R_L (R_1 + R_s)] + (R_1 + R_s + R_L)} \quad \ldots \quad (4)$$

transformation impossible.

In addition, there are other networks for which this transformation is invalid. The voltage ratio for the circuit of Fig. 3 is given by

$$\frac{E_2}{E_0}(p) = \frac{R_L (1 - pCR)}{(2R + R_s + R_L)\left[1 + pCR \dfrac{R(R_s + R_L) + 2R_s R_L}{2R^2 + R(R_s + R_L)}\right]} \quad \ldots \quad (1)$$

Suppose that with R_s finite, $R_s R_L \geqslant R^2$. The transfer function will then be of the form $\dfrac{h(1 - pT)}{(1 + apT)}$, $a \geqslant 1$. If now R_s is made zero, the transfer function will be given by the same expression, but with $a < 1$ for any set of values of R, C and R_L. Thus no scalar multiplying factor can make the transfer functions identical. A more general treatment of this type of circuit is given later in this article, and it can be shown that there is a large class of networks to which the transformation under discussion does not apply.

In general, it is clear that it would only be possible to formulate a theorem along the lines suggested[1] by excluding many types of networks, and that the distinctions between these networks and those to which the theorem might apply are complex and obscure.

Transformation from a Zero to a Finite Source Resistance

The circuits discussed in the previous section are not relevant to the transformation from zero to finite source resistance. First it would only be of academic interest to construct a circuit to work from a zero-impedance source but to contain a branch in shunt with that source. Furthermore if $R_s = 0$ in the circuit of Fig. 3, thus restricting a to be less than unity, an equivalent circuit with finite source resistance can be made to give an identical value of a by ensuring that $R_s\, R_L < R^2$.

The network of Fig. 4 is, however, typical of a class for which the transformation under discussion, as well as the one previously discussed, may fail. Suppose the input impedance of the network to be Z_1. Then the voltage developed across the input terminals E_1 will be given by the expression

$$\frac{E_1}{E_0} = \frac{k\, Z_c\, Z_1\, Z_b\, (Z_c + Z_a)}{Z_c\{k\, Z_1\, Z_b\, (Z_c + Z_a) + Z_a\, [kZ_c(Z_1 + Z_b) + Z_1\, Z_b]\} + R_s\, (Z_c + Z_a)\, [k\, Z_c\, (Z_1 + Z_b) + Z_1\, Z_b]} \quad \ldots \quad (2)$$

If a change is made from $R_s = 0$ to $R_s \neq 0$, the denominator of the transfer function changes from a quadratic in p to a cubic in p. It is not profitable to continue the search for exceptions, since enough has already been said to show that there cannot be any general theorem to include all possible structures. It is of some interest, however, to consider whether there may not exist some

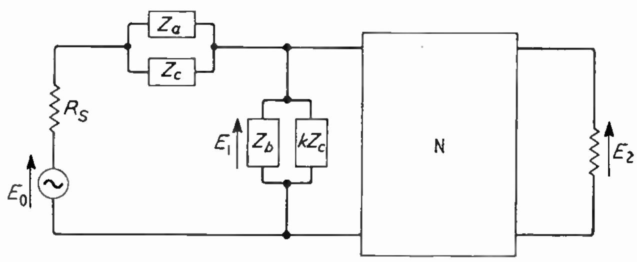


Fig. 4. *General circuit used in the transformation analysis of zero to finite source resistance*

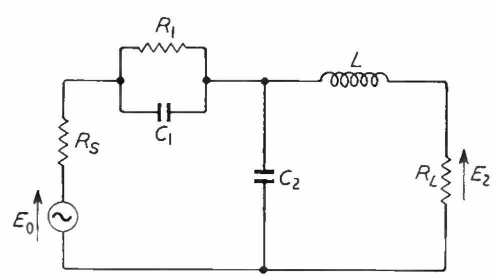


Fig. 5. *A simple RLC network of the form shown in Fig. 4*

particular structure or structures which will enable any given transfer function to be realized with the source resistance either zero or finite.

Network with Terminal Resistances of Any Ratio

This problem is best considered by treating a simple case, since the relationship between a transfer function and a structure capable of realizing it is one of great complexity. In most methods of network synthesis the type of structure to be used is specified at the outset.

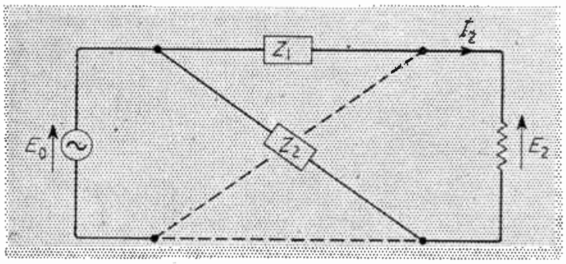


Fig. 6. *The voltage transfer function of this lattice network is derived in the appendix*

Thus, it may be decided to synthesize a function by means of a lattice, or by a set of ladder networks in parallel, or by some other form of circuit. Although the details of the structure will be determined by the transfer function to be synthesized, this function does not determine which type of structure shall be used. There is one method of synthesis in which the structure is apparently determined by the transfer function itself[3]. In this case, however, the general character of the structure is specified implicitly in the method of degenerating the transfer function. Therefore, it is reasonable to restrict attention to particular structures and, for the purposes of illustration, we shall consider synthesizing a simple transfer function by a resistively-terminated *RC* lattice. The function considered is similar in form to that discussed by Griffiths and Mole.

Consider a transfer admittance function given by

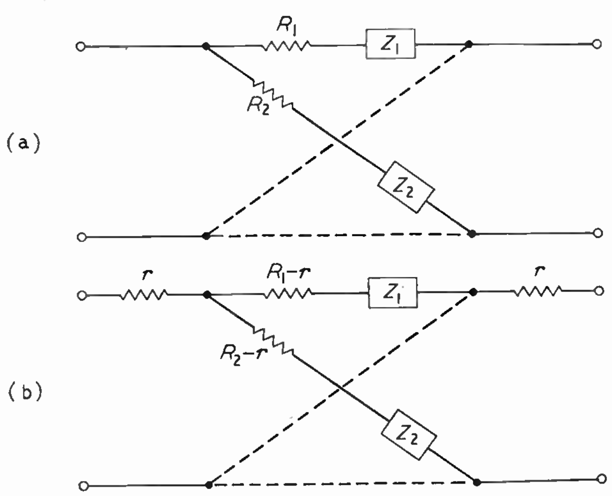
$$\frac{I_2}{E_0} = h\,\frac{(1-ap)}{(1+bp)} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

where h, a, b are positive real constants.

It is shown in the appendix, by a simple variant on the method of Bower and Ordung[2], that this transfer function may be obtained by the network of Fig. 6, in which

$$Z_1(p) = \frac{R\,[(a+c)-(b-c)]}{2\,(b-c)} \quad \ldots \quad \ldots \quad (6)$$

Fig. 7. *Lattice network (a) modified to include equal source and load resistances r*



$$Z_2(p) = \frac{cR[(a+c)-(b-c)]}{2\,(a+c)\,(b-c)} + \frac{R[(a+c)-(b-c)]}{2p\,(a+c)\,(b-c)} \quad (7)$$

The constant c is a positive real number not exceeding b.

At this stage, the transfer function is being realized with zero source resistance. It is possible to make the source resistance finite since the impedances Z_1 and Z_2 both contain a resistive component and the two networks of Fig. 7 are equivalent. Therefore, by extracting from Z_1 and Z_2 a resistance not greater than $\dfrac{cR\,[(a+c)-(b-c)]}{2\,(a+c)\,(b-c)}$, we obtain a network operating between finite resistances. If we wish the source and load resistances to be equal, the original load resistance R must be made zero. If $R = (b-c) = 0$, then

$$Z_1\,(p) = (a+c)/2 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

$$Z_2\,(p) = \frac{c}{2} + \frac{1}{2p} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

In the more general case let

$$R_s = \frac{kc\,R\,[(a+c)-(b-c)]}{2\,(a+c)\,(b-c)} \quad \ldots \quad \ldots \quad (10)$$

where $0 < k < 1$.

Then $R_L = R + R_s$, and it is obvious that by suitable adjustment of c, k and R we may obtain any finite

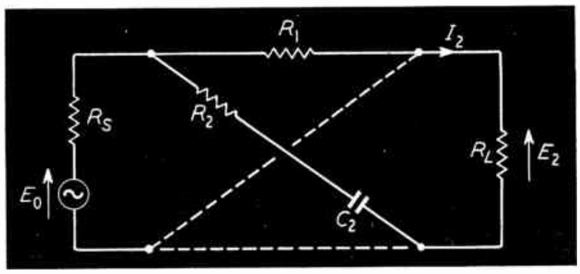


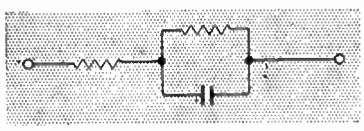
Fig. 8. *RC lattice network. When $R_1 > R_2$, R_2 may be excluded from the arms of the lattice*

ratio of $R_s : R_L$ of value less than unity. Ratios of greater than unity may now be obtained by interchanging source and load resistances, since such an interchange does not affect the transfer admittance. The circuit used in all cases, is that of Fig. 8. It can easily be shown that the voltage transfer function of this circuit, for $R_s = 0$, $R_L = \infty$ is given by

$$\frac{E_2}{E_0} = \frac{1 - pC_2\,(R_1 - R_2)}{1 + pC_2\,(R_1 + R_2)} \quad \ldots \quad \ldots \quad \ldots \quad (11)$$

Thus for $b \geqslant a$, a similar transfer function may be obtained by the same structure for $R_L = \infty$. It is shown in the appendix that it would have been possible to include this case without restriction on the ratio of

Fig. 9. *Showing the form of each arm of network in Fig. 8 if arms were developed from Equ.* (14) *in the appendix*



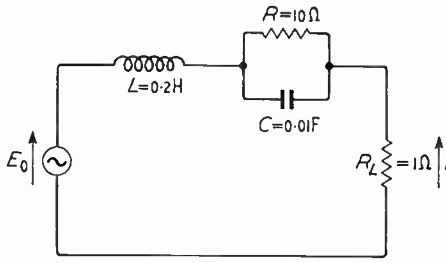


Fig. 10. *Simple RLC circuit with zero source resistance as derived from transfer functions given by Equ.* (12)

$b : a$ if the arms of the lattice were developed from Equ. (14) of the appendix, rather than Equ. (15), and left in general form. Each arm would then have had the form shown in Fig. 9.

It will be noted that an attempt to maintain the same structure throughout leads to the use of a network which is more complicated than it need be for some cases. Even if we exclude from consideration the case $R_L = \infty$, it is clear that since $R_1 > R_2$ in the circuit of Fig. 8, the resistance R_2 may be extracted from both pairs of arms in the lattice. Hence, for finite source and load resistances, the given transfer function may be obtained by a lattice, two arms of which contain resistors only, and the other two capacitors only. There is no virtue in rejecting this circuit in favour of one containing two additional resistors merely to conserve a particular structure.

A practical problem often arises when it is known that a particular circuit will yield a given transfer function when operated from a source of zero resistance, and it is required to obtain the function using a real source of finite internal resistance. If the original network is an *RLC* lattice, it is possible to develop general rules for the conversion but the calculation involved is at least as complicated as that required to synthesize a suitable structure directly from the transfer function. It is probable that the same would be true for any other type of structure and, therefore, there is little practical value to be obtained from such conversion procedures. As an example, consider the transfer function produced by the network of Fig. 5 for the case $R_s = 0$. This function is given by

$$\frac{E_2}{E_0}(p) = \frac{R_L(1+pC_1R_1)}{p^2LR_1(C_1+C_2) + p[L+R_1R_L(C_1+C_2)] + (R_1+R_L)} \quad \ldots \quad (12)$$

Assume that in a particular case the values of the components after frequency normalization are $R_L = 1\,\Omega$, $R_1 = 10\Omega$, $C_1 = 10^{-2}$F, $C_2 = 10^{-2}$F and $L = 10^{-1}$H.

Then

$$\frac{E_2}{E_0}(p) = \frac{0{\cdot}1p + 1}{0{\cdot}02p^2 + 0{\cdot}3p + 11}$$

$$= \frac{1}{0{\cdot}2p + \dfrac{0{\cdot}1p + 11}{0{\cdot}1p + 1}}$$

$$= \frac{1}{0{\cdot}2p + 1 + \dfrac{10}{0{\cdot}1p + 1}} \quad \ldots \quad (13)$$

Clearly the required response may be obtained by the simple circuit of Fig. 10, and it is obvious that some of the load resistance may be associated with the source if required.

Conclusion

It has been shown that the network transformation suggested by Griffiths and Mole does not apply to all networks. It is suggested that for simple transfer functions a new network with appropriate source and load resistances may often be obtained more easily by direct synthesis than by means of network transformations.

APPENDIX

The voltage transfer function of the circuit of Fig. 6 is given by

$$\frac{E_2}{E_0} = \frac{Y_1 - Y_2}{2G + Y_1 + Y_2} \quad \ldots \quad (14)$$

where $Y_1 = 1/Z_1$, $Y_2 = 1/Z_2$, $G = 1/R$.

The ratio of load current to source voltage is thus

$$\frac{I_2}{E_0} = \frac{G(Y_1 - Y_2)}{2G + Y_1 + Y_2} = \frac{m(1 - ap)}{n(1 + bp)} \quad \ldots \quad (15)$$

By the procedure of Bower and Ordung

$$G(Y_1 - Y_2) = m\left(\frac{1 - ap}{1 + cp}\right) \quad \ldots \quad (16)$$

where $0 < c \leqslant b$

$$2G + Y_1 + Y_2 = n\left(\frac{1 + bp}{1 + cp}\right) \quad \ldots \quad (17)$$

N.B.—If c is made equal to b, then R must be made zero.

Thus

$$Y_1 - Y_2 = mR\left[1 + \frac{-(a + c)p}{1 + cp}\right] \quad \ldots \quad (18)$$

$$Y_1 + Y_2 = \left(n - \frac{2}{R}\right) + \frac{n(b - c)p}{1 + cp} \quad \ldots \quad (19)$$

$$\therefore Y_1 = \tfrac{1}{2}\left\{mR + n - \frac{2}{R} + \frac{p[-mR(a+c) + n(b-c)]}{1 + cp}\right\} \quad (20)$$

$$Y_2 = \tfrac{1}{2}\left\{n - \frac{2}{R} - mR + \frac{p\,[n(b-c) + mR(a+c)]}{1 + cp}\right\} \quad (21)$$

In order to reduce the number of elements required, let

$$n(b - c) = mR(a + c) \quad \ldots \quad (22)$$

$$n = \frac{2}{R} + mR \quad \ldots \quad (23)$$

Therefore

$$m = \frac{2(b - c)}{R^2[(a + c) - (b - c)]} \quad \ldots \quad (24)$$

$$n = \frac{2(a + c)}{R\,[(a+c) - (b-c)]} \quad \ldots \quad (25)$$

Hence

$$Y_1 = \frac{2(b - c)}{R\,[(a+c) - (b-c)]} \quad \ldots \quad (26)$$

$$Y_2 = \frac{2(a+c)\,(b-c)}{R\,[(a+c) - (b-c)]} \cdot \frac{p}{1+cp} \quad \ldots \quad (27)$$

Therefore

$$Z_1 = \frac{R\,[(a+c) - (b-c)]}{2\,(b - c)} \quad \ldots \quad (28)$$

$$Z_2 = \frac{R\,[(a+c) - (b-c)]}{2\,(a+c)\,(b-c)} \cdot \frac{1 + cp}{p}$$

$$= \frac{Rc\,[(a+c) - (b-c)]}{2\,(a+c)\,(b-c)} + \frac{R[(a+c) - (b-c)]}{2p\,(a+c)\,(b-c)} \quad \ldots \quad (29)$$

An alternative procedure is to be found by starting with Equ. (14).

Let

$$\frac{E_2}{E_0} = \frac{Y_1 - Y_2}{2G + Y_1 + Y_2} = \frac{r(1 - ap)}{s(1 + bp)} \quad \ldots \quad (30)$$

$$Y_1 - Y_2 = \frac{r(1-ap)}{(1+cp)} = r - \frac{r(a+c)p}{1+cp} \quad \ldots \quad (31)$$

$$Y_1 + Y_2 = -2G + \frac{s(1+bp)}{(1+cp)} = -2G + s + \frac{s(b-c)p}{1+cp} \quad \ldots \quad (32)$$

Then

$$Y_1 = s - 2G + r + \frac{p[s(b-c) - r(a+c)]}{1 + cp} \quad \ldots \quad (33)$$

$$Y_2 = s - 2G - r + \frac{p[s(b-c) + r(a+c)]}{1 + cp} \quad \ldots \quad (34)$$

These functions are similar to those of Equs. (17) and (18) except that whereas s, r and n have the dimensions of admittance, m has the dimensions of (admittance)^2. Furthermore, whereas in the earlier equations Y_1 and Y_2 would both become infinite for $R = \infty$, this set of admittances remains finite in such circumstances.

The constants s, r and c must be such that

$$s \geqslant 2G + r \quad \ldots \quad (35)$$

$$s \geqslant \frac{r(a + c)}{(b - c)} \quad \ldots \quad (36)$$

If the admittance functions are to correspond to RC networks, and if these inequalities are satisfied, the networks will be as shown in Fig. 9. This is also true for the admittance functions of Equs. (20) and (21).

It is not possible to use this method of synthesis for the case $R_s = R_L$ since, under such circumstances, $G = \infty$ and, hence, $Y_1 = Y_2 = \infty$. It is for this reason, because it will permit any finite values to R_s and R_L, that the argument in the article is based on the first method of synthesis.

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Principles of Transistor Circuits
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Proprietes et Applications des Transistors
By J. P. VASSEUR, Ingénieur E.C.P., Docteur ès-Sciences. Pp. 479. Société Française de Documentation Electronique, 12 rue Carducci Paris 19e. Price Fr. 5,540.

This book, published in French, deals almost entirely with the fundamental characteristics, properties and circuit applications of the junction type of transistor. It is intended for the student or technician who is well versed in the general principles of radio engineering but who has only an elementary knowledge of mathematics.

NATIONAL BUREAU OF STANDARDS

Techniques for Accurate Measurement of Antenna Gain

By H. V. COTTONY. Pp. 10. N.B.S. Circular No. 598. Price 19 cents.

Methods are described for measuring (with a minimum of experimental error) aerial gain at frequencies ranging from 400 Mc/s down to 50 Mc/s.

Research Highlights of the National Bureau of Standards: Annual Report 1958

Pp. 138. N.B.S. miscellaneous publication No. 226. Price 57 cents.

Describes a wide range of scientific studies, laboratory experiments, instrument developments and technical publications.

The above publications are obtainable from the Superintendent of Documents, U.S. Government Printing Office, Washington 25, D.C., U.S.A.

BRITISH STANDARDS

Enamelled and Cotton-Covered Copper Conductors (Oleo Resinous Enamel)

B.S. 1815: Part 3: 1959. Round wire. Metric units. Price 4s. 6d.

Varnish-Bonded Glass-Covered Copper Conductors

B.S. 1933: Part 3: 1959. Round wire. Metric units. Price 4s. 6d.

Enamelled and Silk-Covered or Rayon-Covered Copper Conductors

B.S. 2479: Part 3: 1959. Round wire. Metric units. Price 5s.

Silk-Covered and Rayon-Covered Copper Conductors

B.S. 2480: Part 3: 1959. Round wire. Metric units. Price 5s.

Radio-Frequency Cables for use with Domestic Television and V.H.F. Receiving Aerials

B.S. 3040: 1958. Pp. 14. Deals essentially with the constructional and electrical requirements of coaxial and balanced twin cables having a nominal impedance of 75 ohms. Price 4s. 6d.

Television and V.H.F. Broadcast Receiving Aerial Feeder Connectors

B.S. 3041: 1958. Pp. 8. Gives details of two types of feeder connector (coaxial and polarized two-pin types), suitable for use with the aerial feeder cables specified in B.S. 3040 above. Price 4s.

Basic Dimensions for Printed Wiring

B.S. 3081: 1959. Specifies the basic dimensions for printed wiring (forming part of a printed circuit) as used in telecommunication and allied electronic equipment. Price 3s.

Fixed Electrolytic Capacitors (Aluminium Electrodes) for use in Telecommunication and Allied Electronic Equipment

B.S.2134: Part 1 (General Requirements and Use): 1959. Price 7s. 6d.

In using this standard, reference must be made to B.S.2011 which deals with climatic and durability tests for components.

Part 2 of B.S.2134 (to be published later) will specify sizes, ratings, etc., of a standard range of electrolytic capacitors.

Slotted Grub Screws

B.S. 768: 1958. A revised edition of the B.S. 768 Standard, "Grub Screws". Price 4s. 6d.

Standard Test Finger: For Checking Protection against Electric Shock

B.S. 3042: 1958. Pp. 6. Illustrated publication giving the form and method of use of a test finger to determine whether metal parts (which may become electrically 'live') are adequately protected from human contact. Price 3s.

The Relation between the Sone Scale of Loudness and the Phon Scale of Loudness Level

B.S.3045: 1958. Price 3s.

The above Standards are obtainable from the British Standards Institution, Sales Branch, 2 Park Street, London, W.1.

"INSTRUMENT CONSTRUCTION"

This new journal is an English translation of the Russian monthly *Priborostroenie*. It is produced for the Department of Scientific & Industrial Research by the British Scientific Instrument Research Association and published by Taylor & Francis Ltd., Red Lion Court, Fleet Street, London, E.C.4.

The annual subscription is £6 ($17.10, U.S.A. and Canada). The issues are being dated to correspond with the Russian publication, but will appear two or three months later.

INSTITUTION OF ELECTRONICS

The Annual Exhibition and Convention of the Northern Division of the Institution of Electronics will be held from 9th-15th July at the Manchester College of Science and Technology.

Tickets can be obtained from the exhibitors or from the Honorary General Secretary, 78 Shaw Road, Rochdale, Lancs.

OBITUARY

Sir Stanley Angwin, K.C.M.G., K.B.E., D.S.O., M.C., died on 31st March at the age of 75. He entered the engineering department of the Post Office in 1906, and, when the Territorials were inaugurated, he raised the Lowland Division Signal Company. He served in World War I and afterwards returned to the Post Office to join the wireless section of the engineering department, becoming Assistant Engineer-in-Chief in 1932, Deputy Engineer-in-Chief in 1935 and Engineer-in-Chief in 1939.

Sir Stanley became chairman of Cable & Wireless Ltd. in 1947, but resigned in 1951 to become chairman of the Commonwealth Telecommunications Board until his retirement in 1956.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations from nominal frequency for April* 1959

Date 1959 April	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	−186	−9
2	−183	−11
3	NM	−13
4	−186	NM
5	−187	NM
6	−186	−13
7	−184	−10
8	−183	−12
9	−181	−8
10	−182	−8
11	−184	NM
12	−182	NM
13	−182	−9
14	−180	−7
15	−184	−5
16	−181	+1
17	−181	−1
18	−184	−2
19	−183	−1
20	−182	0
21	−181	+2
22	−181	+1
23	−180	+1
24	−180	+1
25	−180	NM
26	−180	NM
27	−180	+1
28	−178	+4
29	−180	+4
30	−178	+5

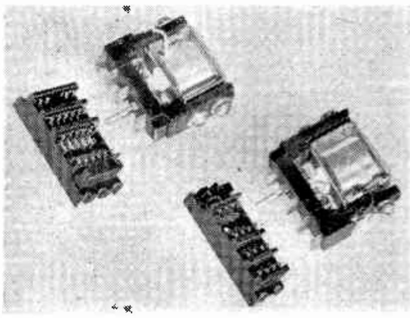
* *Nominal frequency is defined to be that frequency corresponding to a value of* 9 192 631 770 *c/s for the N.P.L. caesium resonator. NM = Not Measured.*

New Products

Wire Contact Relay

This high-speed multi-pole relay, introduced by the Woden Transformer Co. Ltd., and T.S. Precision Engineering Ltd., is suitable for all light-current switching applications, especially in computer and video switching circuits.

The basic unit comprises a 4-pole change-over relay with twin-wire moving contacts which are readily interchangeable. The unit and all fixed-contacts (silver alloy) are



contained in a single piece of moulding of high dielectric strength material. The design of the relay is such that high-speed operation is obtainable with operating and release times down to 3 msec.

The unit has an expected contact life of three million operations which can be considerably extended by using quenching circuits or by operating at lower ratings.

Four, six and twelve-pole models are available with coil resistances from 100 Ω to 10 kΩ and in operating voltages from 6 to 150 V d.c. Each model is suitable for use in maximum operating temperatures of 50° C, although this may be higher for special applications.

Details from the makers' specification for a 4-pole model (shown on the right of the photograph) include:

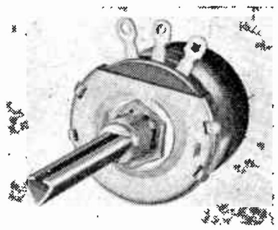
Coil resistance	170 Ω (at 20°C)
Operating voltage	10 V
Pull-in current	36 mA
Drop-out current	18 mA
Operating time	5 msec
Release time	3 msec
Temperature rise after 1 hour (max.)	10° C
Dimensions (overall)	1¾ in. × ⅝ in. × 1½ in. (approx.)

Woden Transformer Co. Ltd.,
Moxley Road,
Bilston, Staffs.

Hollow-Spindle Potentiometer

A triangular-spindle potentiometer in which the spindle is formed from flat brass strip has been produced by The Plessey Co. Ltd.

The design of the spindle is such that the need for machining special flats and knob fixings, etc., has been obviated, all locations and fixings being achieved by bending or forming the hollow brass spindle to accom-



modate a polystyrene ring which, in turn, holds the brush gear. The limit stop and switch operating lug are also incorporated in the polystyrene ring.

Another advantage of the new method of construction relates to the control knob. This merely needs a triangular plain hole slightly smaller than the spindle size to give positive location in three alternative positions. Furthermore, inherent flexibility in the spindle prevents straining of the control.

The bush and back plate are a one-piece zinc-based die casting, which gives great rigidity and accuracy of location. Insulation resistance in the potentiometer is extremely good by virtue of the use of polystyrene.

The Plessey Co. Ltd.,
Ilford, Essex.

Valve Voltmeter

This compact portable instrument, type 614 C, is designed to measure voltages from 80 μV up to 300 V at frequencies between 5 c/s and 100 kc/s. Alternating voltages with superimposed direct current may also be measured provided that a combined peak level of 1 kV is not exceeded.

The basic circuit comprises a high-gain amplifier which feeds a full-wave rectifying indicating meter through a two-stage attenuator. Each attenuator stage (one at the input and the other separating the twin amplifier stages) is ganged by a 13-position range switch to obtain full-scale deflections from 300 μV up to 300 V. The attenuator stage following the first amplifier section is brought in first to improve the signal-to-noise ratio at higher input levels. High calibration accuracy and stability are ensured by applying negative feedback over the two amplifier stages and by incorporating a stabilized h.t. power supply.

The meter is calibrated in r.m.s. values for a sinusoidal input, the indications representing the full-wave average value. The logarithmic voltage scale gives equal reading accuracy (better than ± 2% of indication with a sinusoidal input) for all voltages. The frequency error is less than ± 3% over the stipulated frequency range. In addition to the voltage scales, an auxiliary linear decibel scale is calibrated from − 80 to + 50 dB with reference to 1 V.

An output jack is provided for monitoring the signal or to enable the instrument to be used as an amplifier with a gain of up to 88 dB.

Measuring 6½ in. × 10 in. × 9½ in. and weighing only 10 lb., the instrument may be operated from a 110-V or 200–250-V 50–60-c/s mains supply.

Dawe Instruments Ltd.,
99 *Uxbridge Road, London, W.*5.

Peak Accelerometer

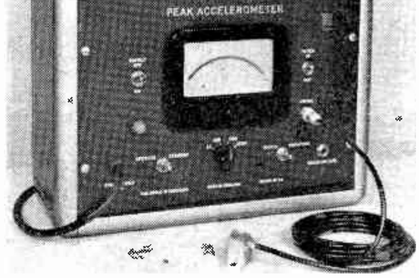
This portable mains-operated instrument has been designed for the measurement of shock inside packages and the cushioning properties of packaging materials. It can also be used for measuring peak '*g*'* values in vibration testing.

The instrument comprises two main units, a probe for fitting on the sample under test and an indicator unit. The latter incorporates a cathode follower, an a.c. amplifier, a low-pass filter with a cut-off frequency of 675 c/s, and an output socket for connecting a cathode-ray oscilloscope. It is also fitted with a double-scale 3½-in. flush-mounted meter giving four ranges of *g* units from 0–25, 0–100, 0–250 and 0–1,000 *g*. The minimum readable *g* is 3. Provision is made for vibration readings from 8 c/s to 750 c/s.

The acceleration-sensitive element or probe consists of a barium titanate crystal accelerometer fitted in a metal housing with a screened output lead, nominally 12 ft. long; a 60-ft. length of special low-capacitance coaxial cable can be supplied if required. The probe can either be attached directly to a rigid member of the sample under test or to the inner cargo container (for package testing).

In operation, the probe is attached to any

* Acceleration quoted in *g* units implies the number of times the weight of a body apparently increases, when compared to its weight at rest.



rigid member of the equipment in such a manner that the probe drops on one plane on to the packaging material which is sufficiently thick to prevent bottoming; i.e., full compression of the material. After a reading has been obtained on the instrument (which will be retained within 2% for 30 seconds even although the actual shock pulse may be as short in duration as a thousandth of a second), the equipment is examined for visible and functional damage, and the probe is then disconnected and connected to another plane of the sample under test. The drop test can then be repeated at different heights.

Drop tests like this simulate a similar order of time duration of shock to those which equipment would be subjected to when transported in a crate.

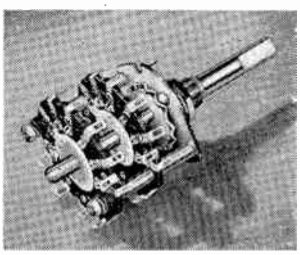
Tests are normally conducted with the filter out of circuit but, should readings be higher than those expected be obtained, the test could be repeated with the high-frequency filter in circuit. If a lower reading was obtained, this would indicate bottoming of the packaging material or, alternatively, that equipment in the crate was insufficiently clamped within the inner cargo housing.

Ferranti Ltd.,
Hollinwood, Manchester, Lancs.

Rotary Wafer Switch

This NSF-OAK rotary wafer switch, model JK, provides up to 22 contact positions on a single section without the use of insulated back contacts.

The clips of the contacts are made of silver-plated spring brass and the rotor blades of silver-plated hard brass.



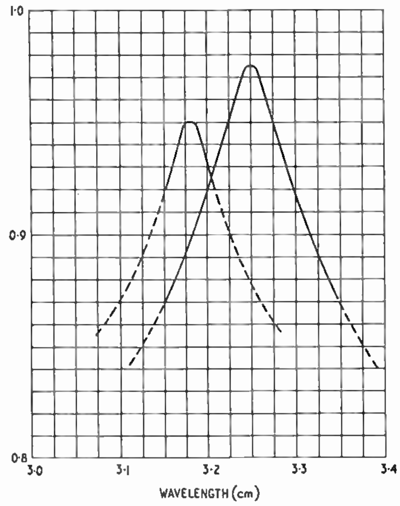
Details from the makers' specification include:

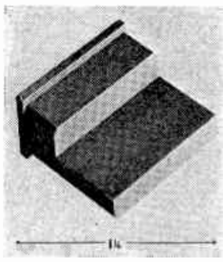
Maximum working voltage:	300 V a.c./d.c.
Proof voltage:	1 kV (r.m.s.)
Rating:	1 A; 28 V d.c. resistive load 50 mA; 300 V d.c. 100 mA; 250 V a.c. (r.m.s.) non-reactive load
Current-carrying capacity: (not breaking)	5 A
Contact resistance:	<0·003 Ω

N.S.F. Ltd.,
31–32 *Alfred Place, London, W.C.*1.

Caslite Cores

Caslite (powdered iron) cores, suitable for the dielectric loading of concentric conductors and waveguides operating at centi-





metre wavelengths, have been produced by The Plessey Co. Ltd.

A typical core and performance curves for 'X' band dummy loads are shown in the illustration.

The Plessey Co. Ltd.,
Ilford, Essex.

Transistorized Conductivity Bridge

This compact portable instrument, type BC1, has been designed to measure solution conductivities over the range 0·5 μmho to 1 mho.

A 1·5-kc/s transistor oscillator is used to energize the a.c. bridge circuit, balance being indicated on a built-in meter. The conductivity of the solution under test is then read off the calibrated dial (located on the left of the instrument).

Normal temperature compensation is provided for measurements between 5 and 1,000 μmhos (at the rate of 2 % per °C) and a built-in variable capacitor is incorporated for balancing out cell and cable capacitances.

A standard conductivity cell, type 7407, having a constant of approximately 0·45 is supplied with the instrument. Two other types with unity and variable cell constants can be supplied, if required.

A. M. Lock & Co. Ltd.,
Prudential Buildings, Oldham, Lancs.

Zener Regulators and Diodes

Texas Instruments Ltd. have introduced a new range of silicon zener voltage regulators, types 1S501–1S516 as well as a new series of silicon zener-voltage reference diodes, types 1S207–1S218.

The voltage regulators, rated at 8 W, are stud-mounted and cover the voltage range of 22 to 91 V in 16 steps. They are available with zener-voltage tolerances of 5% or 10%, and can be supplied with anode to stud, cathode to stud, or as double-anode clippers.

The zener diodes cover the range 3·6 to 10 V in 12 steps. They are available with zener-voltage tolerances of 5% or 10% and are hermetically sealed in small molybdenum-glass sleeves, similar to the makers' range of 400-mA silicon rectifiers.

Texas Instruments Ltd.,
Dallas Road, Bedford.

Spot-Welding Machine

Triangle Products Ltd. has produced a 7·5 kVA general-purpose spot welder which can be easily fitted to any bench or fabricated stand. It will automatically produce precision spot welds on foils as thin as 0·003 in., on steel sheet up to 16 s.w.g., and on steel wire up to 6 s.w.g.

It is a completely self-contained unit and incorporates an electronic timer, six selective heat stages and a built-in water circulatory pump to save the necessity of a water-mains connection.

Two models are available, one for 200/250 V, 50-c/s single-phase mains, and the other for 380/440 V, 50-c/s three-phase mains.

Triangle Products Ltd.,
Hyde, Cheshire.

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

534.522.1 1751

Diffraction of Light by Ultrasonic Waves—Oblique Incidence and Sound Intensity.—S. Parthasarathy & C. B. Tipnis. (*Nature, Lond.*, 27th Dec. 1958, Vol. 182, No. 4652, pp. 1795–1796.) The diffraction is considerably modified if the intensity of the sound is increased above that observed in the previous experiment (1049 of April).

534.612-8 : 537.228.1 1752

Piezoelectric Crystal Probe for the Measurement of Ultrasonic Power and the Investigation of an Ultrasonic Field.—N. Ségard, J. Cassette & F. Cocquerez. (*C. R. Acad. Sci., Paris*, 22nd Sept. 1958, Vol. 247, No. 12, pp. 873–876.) An X-cut quartz crystal is used in a triode circuit.

534.78 1753

Experiences Gained in the Development of a Vocoder and the Measurements of Intelligibility Achieved with it.—G. Krohm. (*Z. angew. Phys.*, Feb. 1958, Vol. 10, No. 2, pp. 56–65.) An intelligibility of about 80 % was achieved with the equipment described using a nominal bandwidth of 300 c/s.

534.79 1754

Subjective Measurements of the Influence of Peak Content in Band-Pass Noise on the Sensation of Loudness.—H. Niese & J. Köhler. (*Hochfrequenztech. u. Elektroakust.*, March 1958, Vol. 66, No. 5, pp. 150–160.) Tests were made with two types of noise of variable bandwidth, (*a*) filtered white noise, and (*b*) noise generated using a f.m. method to eliminate amplitude fluctuations which might give rise to conflicting assessments of loudness. See also 3381 of 1957 (Zwicker et al.).

534.86 : 621.396.712.3 1755

The Acoustic and Technical Characteristics of Reverberation Plates.—W. Kuhl. (*Rundfunktech. Mitt.*, June 1958, Vol. 2, No. 3, pp. 111–116.) A device for producing artificial reverberation is described which consists of a large thin sheet of tin-plate electrodynamically excited in the flexural mode, the oscillations being picked up by a piezoelectric microphone. The reverberation time may be varied over the range 0·8 – 5 sec by altering the distance between the oscillating plate and a movable porous damping plate.

534.861 : 621.396.712.3 1756

The Acoustic Design of the New Studio in Karlsruhe.—L. Keidel. (*Rundfunktech. Mitt.*, June 1958, Vol. 2, No. 3, pp. 106–110.) Details are given of the construction adopted to eliminate traffic noise.

621.395.61 1757

Theory of First- and Second-Order Gradient Receivers.—C. Smetana. (*Hochfrequenztech. u. Elektroakust.*, March 1958, Vol. 66, No. 5, pp. 143–150.) The characteristics of differential microphones are derived with particular reference to their noise-suppression capabilities.

621.395.613.32 1758

Life Tests of the Microphone Carbon in Practical Uses.—H. Hirabayashi, H. Toyoda, H. Shibata & T. Ayakawa. (*Rep. elect. Commun. Lab., Japan*, June 1958, Vol. 6, No. 6, pp. 211–220.) A study of the effects of different methods of preparing the carbon powder.

621.395.623.7 : 537.523.3 1759

The Operation of Ionic Loudspeakers.—G. Bolle. (*Nachrichtentech. Z.*, April 1958, Vol. 11, No. 4, pp. 172–178.) The theory and design of corona-wind loudspeakers are given, and their performance is compared with that of other types.

621.395.623.7 : 621.318.2 1760

Permanent Magnets in Audio Devices.—R. J. Parker. (*Trans. Inst. Radio Engrs*, Jan./Feb. 1958, Vol. AU-6, No. 1, pp. 15–21. Abstract, *Proc. Inst. Radio Engrs*, July 1958, Vol. 46, No. 7, p. 1440.)

621.395.623.7.001.4 1761

Procedures for Loudspeaker Measurements.—P. J. A. H. Chavasse & R. Lehmann. (*Trans. Inst. Radio Engrs*, May/June 1958, Vol. AU-6, No. 3, pp. 56–67.) Translation of paper abstracted in 8 of 1954.

621.395.623.7.002 1762

Progress in the Construction of Loudspeakers.—F. K. Schröder. (*Nachrichtentech. Z.*, April 1958, Vol. 11, No. 4, pp. 169–172.) The improvement of performance characteristics by impregnation of the diaphragm and the fitting of copper rings in the air gap is discussed.

621.395.623.7.012 : 681.142 **1763**

Loudspeaker Enclosure Calculations.—M. V. Callendar. (*Wireless World*, April 1959, Vol. 65, No. 4, pp. 162–163.) An analogue network is described for studying the performance of a loudspeaker in a bass-reflex cabinet.

681.84 : 534.851 **1764**

Pickup for Low Record Wear.—J. Walton. (*Wireless World*, April 1959, Vol. 65, No. 4, pp. 182–185.) A design for a crystal pickup to track within the elastic limit of the record is described. The effective mass is 0·6 mg. See also 2330 of 1957 (Barlow).

AERIALS AND TRANSMISSION LINES

621.372.2 : 538.569.2/.3 **1765**

Transmission Lines with Inhomogeneous Attenuation for the Low-Reflection Absorption of Electromagnetic Waves.—K. L. Lenz. (*Z. angew. Phys.*, Jan. 1958, Vol. 10. No. 1, pp. 17–25.) Measurements were made on a recurrent network of 32 Π-sections to determine by an analogue method the optimum distribution of losses in an absorber for a given minimum residual reflection. For calculations on homogeneous lines see 1068 of April (Lenz & Zinke).

621.372.8 **1766**

Nonlocal Reflection in Waveguides of Variable Cross-Section.—V. Pokrovskiĭ, F. Ulinich & S. Savvinykh. (*Dokl. Ak. Nauk S.S.S.R.*, 11th Jan. 1959, Vol. 24, No. 2, pp. 304–306.) A brief mathematical analysis of the reflection and scattering due to irregularities in the shape of a waveguide considered as a whole. See also 3326 of 1958.

621.372.8 : 621.3-71 **1767**

Waveguides for Use in Low-Temperature Cryostats.—A. Caistor, S. J. Fray & W. C. Hopper. (*J. sci. Instrum.*, March 1959, Vol. 36, No. 3, p. 144.) A design which minimizes the conduction of heat into the cryostat.

621.372.831 **1768**

Reflection of Tapered Waveguides.—K. Matsumaru. (*Rep. elect. Commun. Lab., Japan*, June 1958, Vol. 6, No. 6, pp. 235–239.) An approximate study of the characteristics of conically tapered guides and rectangular-to-circular tapers.

621.372.832.8 **1769**

Low-Loss L-Band Circulator.—F. R. Arams & G. Krayer. (*Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, p. 442.) Insertion loss averages 0·3 dB between 1 200 and 1 450 Mc/s when the magnetic field around the ferrite is optimized for each frequency. Reverse isolation is $\geqslant$ 30 dB, and input voltage s.w.r. $\leqslant$ 1·11.

621.372.852.22 **1770**

Modes in Rectangular Guides Filled with Magnetized Ferrite.—G. Barzilai & G. Gerosa. (*Nuovo Cim.*, 1st March 1958, Vol. 7, No. 5, pp. 685–697. In English.) The analysis is based on the characteristic equation for a rectangular waveguide with a ferrite slab placed against one side wall, the slab being magnetized in a direction parallel to the wall and perpendicular to the waveguide axis.

621.372.852.22 **1771**

Rectangular Guide Ferrite Phase Shifters employing Longitudinal Magnetic Fields.—P. A. Rizzi & B. Gatlin. (*Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, pp. 446–447.)

621.372.852.3 1772

The Homogeneous Rectangular Waveguide with Attenuating Foil.—H. Buseck & G. Klages. (*Arch. elekt. Übertragung*, April 1958, Vol. 12, No. 4, pp. 163–168.) The influence of an axial foil on the waveguide transmission characteristics is analysed assuming metallic contact between foil and waveguide walls. Discrepancies between measured and calculated values are attributed to the absence of electrical contact between the surfaces.

621.396.67 : 621.315.668.2 **1773**

Aalen Television Tower of the Süddeutscher Rundfunk.—(*Rundfunktech. Mitt.*, June 1958, Vol. 2, No. 3, pp. 143–144.) A reinforced-concrete tower of 80-m height supports a 61-m mast for the v.h.f. and television aerials.

621.396.676 **1774**

Slot-Antenna Array for Missiles and Aircraft.—E. J. Wilkinson. (*Electronics*, 27th Feb. 1959, Vol. 32, No. 9, pp. 56–57.) Circular polarization is achieved by combining a folded dipole and a slot radiator. Impedance properties and radiation patterns are given.

621.396.677 : 621.397.62 **1775**

A Second Band-III Programme?—The Aerial Problem.—F. R. W. Strafford. (*Wireless World*, April 1959, Vol. 65, No. 4, pp. 171–174.) Problems associated with the use of existing aerials for the reception of programs separated in frequency by no more than three channels are examined.

621.396.677.5 **1776**

Calculated Radiation Resistance of an Elliptical Loop Antenna with Constant Current.—J. Y. Wong & S. C. Loh. (*J. Brit. Instn Radio Engrs*, Feb. 1959, Vol. 19, No. 2, pp. 89–91.) Extension of the analysis given in 20 of January (Loh & Wong) to derive formulae for loops comparable in size to the wavelength.

621.396.677.73 **1777**

The Construction of Horn-Type Aerials with Parabolic Reflectors.—L. Calligaris. (*Alta Frequenza*, June/Aug. 1958, Vol. 27, Nos. 3/4, pp. 401–432.) Design and constructional details and methods of erection are given for the aerials used in the radio links Milan–Palermo and Rome–Pescara.

621.396.677.83 : 621.396.65 **1778**

Some Remarks on Passive Repeaters.—F. Cappuccini. (*Alta Frequenza*, June/Aug. 1958, Vol. 27, Nos. 3/4, pp. 263–268.) The use of the curves obtained by Jakes (1243 of 1953) for calculating the attenuation o reflectors in radio links is described.

AUTOMATIC COMPUTERS

681.142 **1779**

Digital Computers Available in Britain.—C. H. Lees. (*Brit. Commun. Electronics*, Dec. 1958, Vol. 5, No. 12, pp. 942–949.) Information on 27 computers is given in tabulated form.

681.142 **1780**

Digital Differential Analysers.—G. C. Rowley. (*Brit. Commun. Electronics*, Dec. 1958, Vol. 5, No. 12, pp. 934–938.) Principles of operation are described and details are given of the design of the Avro D.D.A. machine ADDAM II.

681.142 : 537.227 **1781**

Ferroelectrics and Computer Storage.—M. Prutton. (*J. Brit. Instn Radio Engrs*, Feb. 1959, Vol. 19, No. 2, pp. 93–99. Discussion, pp. 100–102.) The polarization reversal process in ferroelectric single crystals and its application to data storage is reviewed. An optical system for reading information from a ferroelectric store is described.

681.142 : 538.221 **1782**

Magnetics for Computers—a Survey of the State of the Art.—J. A. Rajchman. (*RCA Rev.*, March 1959, Vol. 20, No. 1, pp. 92–135.) A review of the application of magnetic materials to storage and switching devices.

681.142 : 621.314.7 **1783**

Transistorized-Core Memory.—R. E. McMahon. (*Trans. Inst. Radio Engrs*, June 1957, Vol. I-6, No. 2, pp. 157–160.)

681.142 : 621.374.33 : 621.314.7 **1784**

Transistors Provide Computer Clock Signals.—S. Schoen. (*Electronics*, 27th Feb. 1959, Vol. 32, No. 9, pp. 70–72.) Switching circuits capable of high speed and controlling peak currents up to 5 A are described.

681.142 : 621.395.625.2 **1785**

Digital Storage on Punched Tape.—M. E. Theis. (*Trans. Soc. Instrum. Technol.*, Dec. 1958, Vol. 10, No. 4, pp. 178–182.)

681.142 : 621.395.625.3 **1786**

The Storage and Processing of Digital Data on Magnetic Tape.—D. W. Willis. (*Trans. Soc. Instrum. Technol.*, Dec. 1958, Vol. 10, No. 4, pp. 182–189.)

681.142 : 621.396.822 **1787**

The Noise Problem in a Coincident-Current Core Memory.—F. McNamara. (*Trans. Inst. Radio Engrs*, June 1957, Vol. I-6, No. 2, pp. 153–156.)

CIRCUITS AND CIRCUIT ELEMENTS

621.3.049.75 **1788**

The Printed Circuit.—C. Brinkmann. (*Elektrotech. Z., Edn B*, 21st Dec. 1958, Vol. 10, No. 12, pp. 461–467.) Summary of manufacturing and assembly techniques.

621.316.5 **1789**

Circuits for Ternary Switching Variables.—E. Mühldorf. (*Arch. elekt. Übertragung*, April 1958, Vol. 12, No. 4, pp. 176–182.) Applications of a ternary switching algebra (see 1636 of May) are considered and the synthesis of a ternary adder is described.

621.318.4 : 621.318.134 **1790**

Design of Toroidal Coils with Ferrite Cores Operating in the Audio-Frequency Range.—L. I. Rabkin & Z. I. Novikova. (*Radiotekhnika i Elektronika*, June 1957, Vol. 2, No. 6, pp. 762–768.) A method of determining the optimum relation between core dimensions is outlined. This gives the minimum coil size for a given Q-factor, or the maximum Q-factor for a given size.

621.318.435.34 **1791**

Auto Self-Excited Transductors.—E. L. Clarke. (*Instrum. Practice*, Oct. 1958, Vol. 12, No. 10, pp. 1093–1100.) Basic circuits are examined nonmathematically, considering only resistive loads.

621.318.57 : 621.396.963.3 **1792**

Coincidence Diodes gate Electronic Switch.—J. B. Beach. (*Electronics*, 20th Feb. 1959, Vol. 32, No. 8, pp. 66–68.) A transistor switching circuit for radar indicators is described. Six channels are used in each coordinate axis of a c.r.o. presentation.

621.319.4 : 537.56 : 538.63 **1793**

Hydromagnetic Capacitor.—Anderson, Baker, Bratenahl, Furth & Kunkel. (See 1853.)

621.319.4.011.4 **1794**

Accurate Determination of the Capacitance of Rectangular Parallel-Plate Capacitors.—D. K. Reitan. (*J. appl. Phys.*, Feb. 1959, Vol. 30, No. 2, pp. 172–176.) The subarea method is recast and applied to derive a universal curve for a square-plate capacitor. Values calculated by other methods are compared.

621.372.01 **1795**

Elements of Electronic Circuits: Part 1—Time Constant and Differentiation.—J. M. Peters. (*Wireless World*, April 1959, Vol. 65, No. 4, pp. 156–158.) First of a series of articles reviewing basic electronic circuits, with emphasis on physical explanations of their operation.

621.372.049 : 621.314.7 **1796**

Analogous Transistor System Design and Nodal Methods of Construction with Applications to Research Equipment and Prototype Evaluation.—R. F. Treharne. (*Proc. Instn Radio Engrs, Aust.*, July 1958, Vol. 19, No. 7, pp. 319–347.) Transistor action and circuits are described in terms of the thermionic valve analogy. A modular technique of circuit construction based on this principle is described, in which stages are assembled individually using a method which simplifies their interconnection.

621.372.4 : 621.372.5 **1797**

On the Reactance Theorem.—H. Wolter. (*Arch. elekt. Übertragung*, April 1958, Vol. 12, No. 4, pp. 158–162.) Any passive quadripole with Foster-type short-circuit and open-circuit impedances at either end transforms any Foster-type two-terminal network into a Foster-type two-terminal network again. Polygons consisting of Foster-type two-terminal networks are also considered. See also 3006 of 1958.

621.372.5 **1798**

Group Delay and Group Velocity.—W. P. Wilson. (*Electronic Radio Engr*, April 1959, Vol. 36, No. 4, pp. 145–146.) The concepts are defined and their relation to the transfer function of a network is given. See also 2004 of 1958 (Gouriet).

621.372.5 **1799**

Radio Engineering Use of the Minkowski Model of the Lorentz Space.—E. F. Bolinder. (*Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, p. 450.)

621.372.5 **1800**

The Condition of Passivity for the Linear Quadripole with Complex Characteristic Impedances.—E. R. Berger. (*Arch. elekt. Übertragung*, April 1958, Vol. 12, No. 4, pp. 149–157.) Reciprocal and non-reciprocal two-port networks are considered.

621.372.5.029.6 : 621.317.341 **1801**

An Analysis of Lossy Symmetrical Quadripoles in the Decimetre and Centimetre Wavelength Ranges using Voltage Node Displacements.—F. Gemmel. (*Arch. elekt. Übertragung*, April 1958, Vol. 12, No. 4, pp. 169–172.) See also 1114 of April.

621.372.54 : 621.397.62 **1802**

A Combined Pulse-Width Filter for Television Receivers.—W. Schröder. (*Elektronische Rundschau*, April 1958, Vol. 12, No. 4, pp. 115–118.) Synchronization by means of a combined integrating and differentiating filter network is described. Performance data are tabulated.

621.372.543.2 **1803**

Intermediate-Frequency Circuits with Three Coupled Resonators.—G. B. Stracca. (*Alta Frequenza*, June/Aug. 1958, Vol. 27, Nos. 3/4, pp. 304–346.) The operation of triple-tuned wide-band band-pass filters is analysed and design formulae are tabulated. For data on double-tuned filters see 3063 of 1957.

621.373.421.13 **1804**

Theory of the Crystal-Controlled Inductive Three-Terminal Circuit.—G. Becker. (*Arch. elekt. Übertragung*, April 1958, Vol. 12, No. 4, pp. 183–191.) Conditions of oscillation, equivalent circuits, and methods of compensation for the crystal-controlled Hartley oscillator are discussed.

621.373.421.13 : 621.314.7 : 538.569.4 **1805**

Transistorized, Crystal-Controlled Marginal Oscillator.—R. L. Garwin, A. M. Patlach & H. A. Reich. (*Rev. sci. Instrum.*, Feb. 1959, Vol. 30, No. 2, pp. 79–80.) Circuit details of a small, non-microphonic oscillator unit for nuclear-magnetic-resonance observations.

621.373.431.1 **1806**

Operating Conditions of the Symmetrical Multivibrator.—N. A. Zheleztsov & M. I. Feigin. (*Radiotekhnika i Elektronika*, June 1957, Vol. 2, No. 6, pp. 751–761.) An approximate method for the division of a multidimensional phase space into sub-spaces is described and applied to analyse the operation of a symmetrical multivibrator taking account of parasitic capacitance and grid current. Three modes of operation are considered.

621.373.431.1 **1807**

Multivibrator with Negative Feedback.—I. Sh. Libin. (*Radiotekhnika i Elektronika*, June 1957, Vol. 2, No. 6, pp. 809–810.) A description is given of a multivibrator circuit with very good frequency stability.

621.373.52 **1808**

Improved *RC* Oscillator.—L. H. Dulberger. (*Electronics*, 6th March 1959, Vol. 32, No. 10, p. 62.) A modified version of the bridged-T circuit described by Sulzer (2943 of 1951), this oscillator operates at a single frequency in the range 4 c/s–350 kc/s.

621.373.52 **1809**

The Conditions for the Onset of Oscillation in Transistor Oscillators.—R. J. Paul. (*NachrTech.*, March 1958, Vol. 8, No. 3, pp. 109–116.) The condition for self-oscillation is established and oscillators are considered in two groups, with frequency either dependent on or independent of transistor parameters.

621.373.52 : 538.569.4 **1810**

Transistorized Nuclear-Resonance Magnetic-Field Probe.—J. R. Singer & S. D. Johnson. (*Rev. sci. Instrum.*, Feb. 1959, Vol. 30, No. 2, pp. 92–93.) Description of a marginal-oscillator type of nuclear-resonance detector.

621.373.52 : 621.395.44 **1811**

Frequency-Stable Transistor Oscillators in Carrier-Frequency Techniques.—W. Hüfner. (*NachrTech.*, March 1958, Vol. 8, No. 3, pp. 117–122.) The design of a Meacham-bridge oscillator is given.

621.374.32 : 621.387 **1812**

Reversible Dekatron Counter.—W. K. Hsu. (*Wireless World*, April 1959, Vol. 65, No. 4, pp. 190–192.) A counter circuit is described which has two inputs, one of which allows addition to and the other subtraction from an existing count.

621.374.34 **1813**

The Static Characteristics of the Cathode-Coupled Limiter (Clipper).—J. Schulz. (*Frequenz*, April 1958, Vol. 12, No. 4, pp. 114–117.)

621.375.223.029.33 : 621.397.62 **1814**

Cathode Compensation.—H. D. Kitchin. (*Electronic Radio Engr*, April 1959, Vol. 36, No. 4, pp. 122–128.) The design of a cathode-compensated pentode video stage is discussed, particularly the selection of values for the cathode resistor and capacitor and the use of a 'bleed' resistor.

621.375.226.012.6 **1815**

Response of Cascaded Double-Tuned Circuits.—Y. Peless. (*Electronic Radio Engr*, April 1959, Vol. 36, No. 4, pp. 134–140.) The transient and steady-state responses are developed in terms of the location of the poles of the transfer function. The results can also be applied to networks with a response asymmetrical about the band centre, but with a narrow relative bandwidth.

621.375.4 **1816**

Design of Transistor *RC* Amplifiers.—R. P. Murray. (*Trans. Inst. Radio Engrs*, May/June 1958, Vol. AU-6, No. 3, pp. 67-76. Abstract, *Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, pp. 1888–1889.)

621.375.4.018.7 **1817**

Transistor Nonlinearity : Dependence on Emitter Bias Current in *P-N-P* Alloy-Junction Transistors.—D. R. Fewer. (*Trans. Inst. Radio Engrs*, March/April 1958, Vol. AU-6, No. 2, pp. 41–44. Abstract, *Proc. Inst. Radio Engrs*, Oct. 1958, Vol. 46, No. 10, p. 1774.)

621.375.9 : 538.569.4 **1818**

Radiation Damping Effects in Two-Level Maser Oscillators.—A. Yariv, J. R. Singer & J. Kemp. (*J. appl. Phys.*, Feb. 1959, Vol. 30, No. 2, p. 265.) Analytical note on the modulation effects occurring in a spontaneously radiating inverted two-level spin system.

621.375.9 : 621.3.011.23 **1819**

Phase-Distortionless Limiting by a Parametric Method.—A. E. Siegman. (*Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, pp. 447–448.) Nearly ideal limiting can be obtained by using the signal to be limited as the 'pump'-frequency signal in a parametric device.

621.375.9 : 621.3.011.4 **1820**

Nonlinear-Capacitance Amplifiers.—L. S. Nergaard. (*RCA Rev.*, March 1959, Vol. 20, No. 1, pp. 3–17.) An account of the physical principles and the design of variable-capacitance amplifiers. The effective noise temperatures achieved are compared with those of other low-noise amplifiers and of terrestrial and extraterrestrial noise sources.

621.375.9 + 621.372.632] : 621.385.029.6 **1821**

A Three-Frequency Electron-Beam Parametric Amplifier and Frequency Converter.—Louisell. (See 2065.)

621.375.9 : 621.385.029.6 **1822**

Gain, Bandwidth and Noise in a Cavity-Type Parametric Amplifier using an Electron Beam.—Wade & Heffner. (See 2066.)

621.375.9 : 621.385.029.6 : 621.372.2 **1823**

Travelling-Wave Couplers for Longitudinal Beam-Type Amplifiers.—Gould. (See 2067.)

621.375.9.029.6 : 537.311.33 **1824**

The Physical Principles of a Negative-Mass Amplifier.—H. Krömer. (*Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, pp. 397–406.) Negative effective masses for relatively low energies may be obtained if the energy contours are re-entrant near the edge of the frequency band, as is the case for heavy holes in germanium and certain other semiconductors with degenerate band edges. Operation at frequencies up to 1 000 kMc/s (0·3 mm λ) is envisaged. See also 2354 of 1958.

621.376.22.029.64 : 621.318.134 **1825**

Microwave Ferrite Modulators for High Signal Frequencies.—A. L. Morris. (*J. Brit. Instn Radio Engrs*, Feb. 1959, Vol. 19, No. 2, pp. 117–129.) Methods are suggested for avoiding ferromagnetic resonance. Skin effects are overcome by using the waveguide as a modulating helix. Performance details of two experimental modulators for X-band frequencies are given.

621.376.32 : 538.569.4 **1826**

Frequency Modulator for a Marginal Oscillator.—D. A. Jennings & W. H. Tanttila. (*Rev. sci. Instrum.*, Feb. 1959, Vol. 30, No. 2, pp. 137–138.) A voltage-sensitive capacitor with d.c. bias is used in the tank circuit of the oscillator.

621.376.4 : 621.314.7 : 621.398 **1827**

The Accuracy Obtainable with Transistors in Pulse Amplitude and Pulse Width Modulation.—E. Schenck. (*Nachrichtentech. Z.*, April 1958, Vol. 11, No. 4, pp. 191–196.) The effects of ambient temperature and power supply fluctuations on transistor p.m. circuits for telemetry applications are analysed. The design of a p.a.m./p.w.m. converter is given.

GENERAL PHYSICS

535.215 : 535.34 **1828**

Photoconductivity as a Function of Optical Absorption.—A. M. Goodman. (*J. appl. Phys.*, Feb. 1959, Vol. 30, No. 2, pp. 144–147.) A theoretical analysis of the dependence of photoconductivity on optical absorption, based on the concept of a 'schubweg' or mean range for the optically liberated charge carriers of each sign.

537.122 **1829**

Correlation Function for a System of Particles Carrying Like Charges.—V. P. Galaĭko & L. E. Pargamanik. (*Dokl. Ak. Nauk S.S.S.R.*, 21st Dec. 1958, Vol. 123, No. 6, pp. 999–1002.) A description of a mathematical method for the determination of correlation functions for systems of particles carrying like charges. This method could be extended to charges of different sign and also to the kinetic theory of charged particles. See also 3825 of 1957 (Tyablikov & Tolmachev).

537.122 : 537.2 **1830**

A Dielectric Formulation of the Many-Body Problem : Application to the Free Electron Gas.—P. Nozières & D. Pines. (*Nuovo Cim.*, 1st Aug. 1958, Vol. 9, No. 3, pp. 470–490. In English.)

537.226 : 539.2 **1831**

Theory of the Contribution of Exciton to the Complex Dielectric Constant of Crystals.—J. J. Hopfield. (*Phys. Rev.*, 1st Dec. 1958, Vol. 112, No. 5, pp. 1555–1567.)

537.291 **1832**

Graphical-Analytical Construction of the Space Trajectory of Charged Particles in Magnetic Fields.—N. I. Shtepa. (*Radiotekhnika i Elektronika*, June 1957, Vol. 2, No. 6, pp. 790–795.) Two methods are described for plotting the trajectory of relativistic charged particles: an acceleration method and the 'radii of curvature' method (see 2720 of 1957).

537.525 **1833**

Effect of Space Charge in Cold-Cathode Gas Discharges.—A. L. Ward. (*Phys. Rev.*, 15th Dec. 1958, Vol. 112, No. 6, pp. 1852–1857.) Townsend's basic ionization equations for cold-cathode discharges between parallel plates are modified by Poisson's equation to account for space-charge effects. Numerical calculations have been made for argon.

537.525.029.6 : 551.510.52 **1834**

High-Frequency Breakdown in Air at High Altitudes.—A. D. MacDonald. (*Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, pp. 436–441.) Breakdown field is computed for 100 Mc/s, 3, 10, 20 and 35 kMc/s, from atmospheric data, the validity of which is discussed. Considerably more power per unit area can be transmitted at the higher frequencies.

537.533 **1835**

Longitudinal Oscillations of Electron-Ion Beams.—P. V. Polovin & N. L. Tsintsadze. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2615–2623.) Investigation of the stability of electron-ion beams leads to a differential equation that is difficult to solve. A qualitative method based on self-conjugate differential operators is used, which avoids solution of this equation.

537.533 **1836**

Compensation of Space Charge in an Electron Beam.—V. I. Volosok & B. V. Chirikov. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2624–2630.) Measurements of the electric field of the space charge in the beam were carried out to evaluate the lifetime of the virtual cathode.

537.533.74 **1837**
Calculation of the Spatial and Angular Distribution of a Stream of Particles with Multiple Scattering.—F. Lenz. (*Z. angew. Phys.*, Jan. 1958, Vol. 10, No. 1, pp. 31–34.) Scattering of electron beams is considered taking account of absorption and retardation effects.

537.534.8 **1838**
Energy Spectrum of Secondary Electrons Emitted by a Metal under the Action of a Fast Ion Beam.—F. Pradal & R. Simon. (*C. R. Acad. Sci., Paris*, 28th July 1958, Vol. 247, No. 4, pp. 438–441.) Spectra have been analysed by means of a magnetic spectrograph.

537.54 : 538.6 : 523.75 **1839**
Solar Proton Stream Forms with a Laboratory Model.—Bennett. (See 1864.)

537.56 : 538.56 **1840**
A Transit-Time Relation for Plasma Electron Oscillations.—K. G. Emeleus & D. W. Mahaffey. (*J. Electronics Control*, Dec. 1958, Vol. 5, No. 6, pp. 559–560.)

537.56 : 538.566 **1841**
Microwave Propagation in Hot Magneto-plasmas.—J. E. Drummond. (*Phys. Rev.*, 1st Dec. 1958, Vol. 112, No. 5, pp. 1460–1464.) The refractive indices for circularly polarized waves propagating along the magnetic field in an ionized gas at high temperature are calculated. They are found to depend sensitively on electron density and temperature.

537.56 : 538.566 **1842**
Conductivity of Plasmas to Microwaves.—E. A. Desloge, S. W. Matthysse & H. Margenau. (*Phys. Rev.*, 1st Dec. 1958, Vol. 112, No. 5, pp. 1437–1440.) A new derivation of an expression previously obtained for plasma conductivity [see 1702 of 1958 (Margenau)], and an alternative expression which avoids previous difficulties with negative conductivities.

537.56 : 538.63 : 621.319.4 **1843**
Hydromagnetic Capacitor.—O. Anderson, W. R. Baker, A. Bratenahl, H. P. Furth & W. B. Kunkel. (*J. appl. Phys.*, Feb. 1959, Vol. 30, No. 2, pp. 188–196.) Very high dielectric constants can be achieved with an ionized gas in a strong magnetic field. When an orthogonal electric field is applied, resultant particle drift stores electrical energy. Dielectric constants from 10^6 to 10^8 have been measured in a coaxial capacitor using a rotating plasma disk. Potential use in fast-discharge work is considered.

538.56 : 621.372.413 : 537.122 **1844**
On the Question of Quantum Effects in the Interaction of Electrons with High-Frequency Fields in Resonators.—V. L. Ginzburg & V. M. Faĭn. (*Radiotekhnika i Elektronika*, June 1957, Vol. 2, No. 6, pp. 780–789.) It is shown that the quantum effect corresponds to the interaction of electrons with a neutral oscillatory field in the resonator. Methods of quantum mechanics are avoided by using the Nyquist formula [see e.g. *Phys. Rev.*, 15th March 1956, Vol. 101, No. 6, pp. 1620–1626 (Weber)].

538.56.029.66 **1845**
The Band between Microwave and Infrared Regions.—I. Kaufman. (*Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, pp. 381–396.) Difficulties that have hitherto prevented microwave generation in the 300–3 000 kMc/s region are discussed. Schemes which might overcome these difficulties are considered.

538.566 **1846**
Multistage Resonance Absorbers for Centimetre Electromagnetic Waves.—H. J. Schmitt & W. Futtermenger. (*Z. angew. Phys.*, Jan. 1958, Vol. 10, No. 1, pp. 1–7.) A three-stage dipole resonance absorber is derived from the two-stage absorber described in 1064 of 1957 (Schmitt) by the addition of a second dipole grid at a distance of $\lambda/8$ in front of the metal surface. The frequency characteristics of the reflection coefficient are calculated and compared with the results of measurements.

538.566 **1847**
The Absorption of Centimetre Electromagnetic Waves in Artificially Anisotropic Media.—R. Pottel. (*Z. angew. Phys.*, Jan. 1958, Vol. 10, No. 1, pp. 8–16.) Two methods of equalizing the complex permeability and dielectric coefficient of a medium to obtain absorption at cm λ are discussed. In one case the medium consists of thin parallel layers, in the other use is made of gyromagnetic resonance in ferrite material subjected to a static magnetic field. Experimental results are given.

538.566 **1848**
The Group Velocity of Damped Waves.—L. A. Vaĭnshteĭn. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2606–2614.) The group velocity is shown to be equal to S_z/w where S_z is the component of the Poynting vector along the direction of propagation and w the energy density. The electromagnetic energy and group velocities are calculated for a simple medium, a plasma and a dielectric with bound electrons.

538.566 **1849**
Kinematics of Growing Waves.—P. A. Sturrock. (*Phys. Rev.*, 1st Dec. 1958, Vol. 112, No. 5, pp. 1488–1503.) Analytical treatment of the problem of distinguishing between amplifying and evanescent waves.

538.566 : 535.42] + 534.26 **1850**
Asymptotic Development of Double Integrals encountered in Diffraction Theory.—N. Chako. (*C. R. Acad. Sci., Paris*, 28th July 1958, Vol. 247, No. 4, pp. 436–438.)

538.566 : 535.42 **1851**
A New Method for the Solution of a Problem of Diffraction of Electromagnetic Plane Waves at an Unlimited Rectilinear Slit, and Related Problems.—G. A. Grinberg. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2595–2605.) Using the method described a good approximation to the exact solution of the diffraction problem can be obtained for normally incident plane waves of wavelength much smaller than or equal to the width of the slit.

538.569.4 **1852**
Radio-Frequency Spectra of Hydrogen Deuteride in Strong Magnetic Fields.—W. E. Quinn, J. M. Baker, J. T. LaTourrette & N. F. Ramsey. (*Phys. Rev.*, 15th Dec. 1958, Vol. 112, No. 6, pp. 1929–1940.)

538.569.4 **1853**
Steady-State Free Precession in Nuclear Magnetic Resonance.—H. Y. Carr. (*Phys. Rev.*, 1st Dec. 1958, Vol. 112, No. 5, pp. 1693–1701.) Description of a new technique for observing nuclear magnetic resonance.

538.569.4 : 535.33 **1854**
An X-Band Spectrometer for the Demonstration of Paramagnetic Resonances.—W. Stieler. (*Z. angew. Phys.*, Feb. 1958, Vol. 10, No. 2, pp. 89–95.) A bridge-type spectrometer with frequency stabilization is described.

538.569.4 : 537.311.62 **1855**
Theory of Cyclotron Resonance in Metals.—S. Rodriguez. (*Phys. Rev.*, 1st Dec. 1958, Vol. 112, No. 5, pp. 1616–1620.) Analysis of the low-temperature case in which the mean free path of the electrons is greater than the skin depth.

538.569.4 : 538.222 **1856**
Multiple Quantum Transitions in Paramagnetic Resonance.—P. P. Sorokin, I. L. Gelles & W. V. Smith. (*Phys. Rev.*, 1st Dec. 1958, Vol. 112, No. 5, pp. 1513–1515.) At high r.f. field strengths the normal paramagnetic resonance spectrum of Mn^{++} in cubic MgO is found to be modified by additional absorption lines, which are interpreted as double quantum transitions ($\Delta M = 2$) occurring between nearly equally spaced energy levels at high r.f. power. Calculation of the relative absorption intensity for such transitions agrees with experimental results.

538.569.4 : 621.375.9 : 535.61-1/2 **1857**
Infrared and Optical Masers.—A. L. Schawlow & C. H. Townes. (*Phys. Rev.*, 15th Dec. 1958, Vol. 112, No. 6, pp. 1940–1949.) Theoretical aspects of maser-like devices for wavelengths much shorter than 1 cm are discussed. The short-wavelength limit for practical devices is examined. Design principles are illustrated by a system for the infrared region using potassium vapour.

539.2 : 537.311.31 **1858**
The Effective Radius of the Electron in Crystal Lattices.—A. F. Kapustinskiĭ. (*Dokl. Ak. Nauk S.S.S.R.*, 21st Feb. 1959, Vol. 124, No. 6, pp. 1265–1266.) The effective radius of electrons in metals determined by different methods is found to be $0 \cdot 78 \pm 0 \cdot 02$ Å.

539.2 : 537.311.31 **1859**
Relation between the Constants Characterizing the Interaction of Electrons with Phonons and Impurities in

Metals.—V. L. Bonch-Bruevich. (*Dokl. Ak. Nauk S.S.S.R.*, 21st Feb. 1959, Vol. 124 No. 6, pp. 1233–1235.) A mathematical analysis.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.165 **1860**
Unusual Cosmic-Ray Intensity Fluctuations Observed at Southern Stations during October 21–24, 1957.—K. G. McCracken & N. R. Parsons. (*Phys. Rev.*, 1st Dec. 1958, Vol. 112, No. 5, pp. 1798–1801.) The fluctuations observed have unusual features which suggest the existence of a short-lived and highly directional anisotropy of the primary radiation during the period immediately preceding a Forbush-type decrease.

523.165 : 061.3 **1861**
International Convention on Cosmic Rays.—(*Nuovo Cim.*, 1958, Vol. 8, Supplement No. 2, pp. 125–804. In English.) The text is given of over 90 papers presented at a convention held at Varenna, 21st–26th June 1957. The subject matter is divided into the following groups: (*a*) solar and geomagnetic effects on cosmic rays, (*b*) problems of origin, (*c*) composition of primary radiation, (*d*) air showers, and (*e*) interactions of cosmic radiation.

523.165 : 523.75 **1862**
Latitude Variation of 27-Day Cosmic-Ray Intensity Decreases.—R. R. Brown. (*Nuovo Cim.*, 16th July 1958, Vol. 9, No. 2, pp. 197–207. In English.) The latitude variation of decreases in cosmic-ray intensity due to the modulation effects of a geocentric nebula of disordered magnetic fields has been measured. Experimental data provided by neutron monitor observations during the period of intense solar activity show that equatorial variations exceed calculated values by a factor of two or more.

523.3 : 621.396.96 **1863**
Radio Observations of the Lunar Surface.—J. K. Hargreaves. (*Proc. phys. Soc.*, 1st March 1959, Vol. 73, No. 471, pp. 536–537.) Information about the moon's surface can be obtained from a statistical consideration of radar scattering from the surface.

523.75 : 537.54 : 538.6 **1864**
Solar Proton Stream Forms with a Laboratory Model.—W. H. Bennett. (*Rev. sci. Instrum.*, Feb. 1959, Vol. 30, No. 2, pp. 63–69.) A 'Störmertron' tube is described for simulating the streams of charged particles in the earth's dipole magnetic field. Photographs of stream forms and contacts are presented. The same techniques can be used to study stream forms in other complicated magnetic fields.

550.385 : 523.75 **1865**
Recurrent Geomagnetic Storms and Solar Prominences.—R. T. Hansen. (*J. geophys. Res.*, Jan. 1959, Vol. 64, No. 1, pp. 23–25.) An examination of the association of prominence areas with days of recurrent storms during the period 1917–1944. The identification of *M* regions with solar prominences is not confirmed.

550.389.2 : 629.19 **1866**
Motion of a Satellite around an Unsymmetrical Central Body.—T. E. Sterne. (*J. appl. Phys.*, Feb. 1959, Vol. 30, No. 2, p. 270.) Comment on 1541 of May (Newton).

550.389.2 : 629.19 **1867**
A Determination of the Coefficient *J* of the Second Harmonic in the Earth's Gravitational Potential from the Orbit of Satellite 1958 β2.—M. Lecar, J. Sorenson & A. Eckels. (*J. geophys. Res.*, Feb. 1959, Vol. 64, No. 2, pp. 209–216.)

550.389.2 : 629.19 **1868**
Vanguard I I.G.Y. Satellite (1958 Beta).—R. L. Easton & M. J. Votaw. (*Rev. sci. Instrum.*, Feb. 1959, Vol. 30, No. 2, pp. 70–75.) The instrumentation used, the measurements made and the uses of the satellite are described. Information on satellite temperatures and rotation rate is given. Orbital data are being used for measuring the oblateness of the earth and for correcting mapping errors.

550.389.2 : 629.19 **1869**
Information by Radio from the Satellites.—J. A. Ratcliffe. (*J. Instn elect. Engrs*, Nov. 1958, Vol. 4, No. 47, pp. 603–608.) An account of some early results obtained mainly from observations of satellite 1957α.

550.389.2 : 629.19 : 061.3 **1870**
I.G.Y. Conference in Moscow: Soviet Papers Presented at the Rocket and Satellite Symposium.—J. W. Townsend, Jr. (*Science*, 9th Jan. 1959, Vol. 129, No. 3341, pp. 80–84.) Summary of preliminary data presented at the 5th General Assembly of C.S.A.G.I., Moscow, 30th July–9th August 1958.

550.389.2 : 629.19 : 551.510.53 **1871**
Densities and Temperatures of the Upper Atmosphere Inferred from Satellite Observations.—G. F. Schilling & T. E. Sterne. (*J. geophys. Res.*, Jan. 1959, Vol. 64, No. 1, pp. 1–4.) The atmospheric density between 180 and 400 km altitude appears to be appreciably higher than that derived from rocket data. The temperature in this region must therefore be higher than that given by present model atmospheres. The density is still below that derived from meteor data.

550.389.2 : 629.19 : 551.510.53 **1872**
Investigation of the Upper Atmosphere by means of the Third Artificial Earth Satellite.—V. I. Krasovskiĭ. (*Priroda, Mosk.*, Dec. 1958, No. 12, pp. 71–78.) A brief description of geophysical investigations carried out in U.S.S.R. The temperature variation and the distribution of the electron density with height up to about 500 km are discussed, and the collision of an artificial satellite with micrometeorites is considered.

550.389.2 : 629.19 : 551.510.535 **1873**
The Ionosphere and Artificial Earth Satellites.—Ya. L. Al'pert. (*Priroda, Mosk.*, Oct. 1958, No. 10, pp. 71–77.) Two methods for investigating the ionosphere are briefly described; the Doppler shift of emitted signals and the observation of the times of 'radio rise' and 'radio setting' of the satellites. From the difference between the optical and the radio observations of the rising and setting, the electron concentration can be calculated.

550.389.2 : 629.19 : 551.510.535 **1874**
Satellite Doppler Measurements and the Ionosphere.—J. A. Thomas & F. H. Hibberd. (*J. atmos. terr. Phys.*, Feb. 1959, Vol. 13, Nos. 3/4, pp. 376–379.) Simultaneous Doppler measurements of satellite 1957α at frequencies of 20 and 40 Mc/s were shown to be in agreement with an approximate model of the ionosphere for the same period.

550.389.2 : 629.19 : 621.396.43 **1875**
Transoceanic Communication by Means of Satellites.—Pierce & Kompfner. (See 2012.)

551.510.53 : 551.524.7 **1876**
Temperatures in the High Atmosphere.—F. S. Johnson. (*Ann. Géophys.*, Jan./March 1958, Vol. 14, No. 1, pp. 94–108. In English.) Temperature distribution is discussed in relation to rocket and radio observations on the assumption that heating of the ionosphere is due primarily to solar radiation and that temperature distribution is controlled by thermal conduction. A model is presented which indicates a very rapid temperature rise between 100 and 200 km. This would produce a density distribution in agreement with obversations.

551.510.53 : 551.593 **1877**
The Temperature in the Atmospheric Region Emitting the Nightglow OI 5577 Line and in Regions above Faint Auroral Arcs.—E. B. Armstrong. (*J. atmos. terr. Phys.*, Feb. 1959, Vol. 13, Nos. 3/4, pp. 205–216.)

551.510.535 **1878**
Effect of Vertical Drifts on the Nocturnal Ionization of the Lower Ionosphere.—M. N. Rao & A. P. Mitra. (*J. atmos. terr. Phys.*, Feb. 1959, Vol. 13, Nos. 3/4, pp. 271–290.) Vertical ionic drifts present in quiet conditions and enhanced during a magnetic storm result in a vertical redistribution of ionization and thus change the apparent recombination coefficient at a given height. The resulting *N* (*h*) profiles are calculated and shown to explain the sudden cessation of night-time echoes of l.f. radio waves at times of magnetic disturbance reported by Lindquist (724 of 1954).

551.510.535 **1879**
A Survey of the Present Knowledge of Sporadic-E Ionization.—J. A. Thomas & E. K. Smith. (*J. atmos. terr. Phys.*, Feb. 1959, Vol. 13, Nos. 3/4, pp. 295–314.) Discussion of the various techniques of observing E_s, and the classification of the

phenomenon by geographical zones (auroral, temperate and equatorial), and by type, as evidenced by the trace recorded on an ionogram. The diurnal and seasonal variations in the three zones are shown to be quite distinct. Various theories of the nature of the ionization formations and the agencies producing them are considered, and the subjects requiring special experimental and theoretical attention are indicated. Over 100 references.

551.510.535 **1880**

Lunar Tides in the Sporadic-E Layer at Ibadan.—R. W. Wright & N. J. Skinner. (*J. atmos. terr. Phys.*, Feb. 1959, Vol. 13, Nos. 3/4, pp. 217–221.) "An analysis of the lunar semi-diurnal tides in fE_s and $h'E_s$ is made for Ibadan. The results are presented in the form of harmonic dials. Comparisons are made between these results and those of other stations."

551.510.535 **1881**

Ionospheric Measurements made at Halley Bay.—W. H. Bellchambers & W. R. Piggott. (*Nature, Lond.*, 6th Dec. 1958, Vol. 182, No. 4649, pp. 1596–1597.) Graphs of the monthly median values of f_0F_2 at Halley Bay (75°31′S, 26°36′W) for July (winter), September (equinox) and December (summer) are given. A large diurnal variation of electron density occurs in winter. The seasonal maximum of electron density during the day is found at the equinoxes.

551.510.535 **1882**

The Diurnal and Annual Variations of f_0F_2 over the Polar Regions.—S. C. Coroniti & R. Penndorf. (*J. geophys. Res.*, Jan. 1959, Vol. 64, No. 1, pp. 5–18.) The variations have been studied for about 15 stations over a three year period and the following results have been found: (*a*) the diurnal variation is largest in winter and smallest in summer; the maximum values of f_0F_2 occur at noon in the northern and at midnight in the southern hemisphere; (*b*) there is a regular annual variation for a given time of day; f_0F_2 in northern winter has deep minima in the morning and evening; (*c*) in summer the latitude differences in f_0F_2 are small but become large in winter, up to 0·85 Mc/s per degree; (*d*) the lines of equal critical frequency lie between circles of geographic and geomagnetic latitude.

551.510.535 : 523.78 **1883**

Ionospheric Changes at Singapore during the Solar Eclipse of 14th December 1955.—C. M. Minnis. (*J. atmos. terr. Phys.*, Feb. 1959, Vol. 13, Nos. 3/4, pp. 346–350.) The eclipse occurred in the late afternoon and no simple interpretation of the changes in the E and F_1 layers can be given. The critical frequency of the F_2 layer did not change but there were considerable eclipse effects in the lower part of the F layer.

551.510.535 : 621.396.11 **1884**

An Analysis of Drifts of the Signal Pattern associated with Ionospheric Reflections.—D. G. Yerg. (*J. geophys. Res.*, Jan. 1959, Vol. 64, No. 1, pp. 27–31). A statistical treatment to determine the drift and random velocity components from the signals observed on three spaced receivers. Particular attention is given to the random motion in order to examine the detailed movement of the pattern. The results suggest a ruled pattern undergoing fluctuations in contour spacing as the pattern drifts across the receiver site.

551.510.535 : 621.396.11 **1885**

Triple Splitting of the F Echoes.—R. Satyanarayana, K. Bakhru & S. R. Khastgir. (*J. atmos. terr. Phys.*, Feb. 1959, Vol. 13, Nos. 3/4, pp. 201–204.) Polarization measurements of triple-split echoes at Banaras (25°N, 83°E) agree with measurements at high latitudes. The process concerned in producing triple-splitting at low latitudes is discussed.

551.510.535(98) **1886**

Arctic Measurements of Electron Collision Frequencies in the D Region of the Ionosphere.—J. A. Kane. (*J. geophys. Res.*, Feb. 1959, Vol. 64, No. 2, pp. 133–139.) Simultaneous measurement of refractive indices and the difference in the absorption of the two magneto-ionic components of a 7·75-Mc/s c.w. signal transmitted from a rocket, allowed the electron collision frequency profile to be determined. The results of two midday flights indicate that the frequencies are lower by a factor of three than the unpublished theoretical values calculated by Nicolet.

551.594.5 **1887**

Electric Field Theory of Aurorae.—G. C. Reid. (*Nature, Lond.*, 27th Dec. 1958, Vol. 182, No. 4652, pp. 1791–1792.) A qualitative description of the development of a typical auroral display following the growth of an electric field in the ionosphere.

551.594.5 **1888**

An Artificial Aurora.—P. H. Fowler & C. J. Waddington. (*Nature, Lond.*, 20th Dec. 1958, Vol. 182, No. 4651, p. 1728.) The auroral display at Apia reported by Cullington (1569 of May) could not have been caused by direct radiation from the explosion of a nuclear bomb at Johnston Island (17°N, 169°W), which is approximately 2 200 miles away, but was probably due to charged particles from the explosion.

551.594.5 : 621.396.11.029.62 **1889**
: 551.510.535

A Bistatic Radio Investigation of Auroral Ionization.—Collins & Forsyth. (See 2001.)

LOCATION AND AIDS TO NAVIGATION

621.396.933 **1890**

B.O.A.C.'s Comet: Communications and Navaids.—(*Brit. Commun. Electronics*, Jan. 1959, Vol. 6, No. 1, pp. 34–35.) A general report of the radio equipment used.

621.396.933.1 **1891**

A Low-Drain Distress Beacon for a Crash Position Indicator.—D. M. Makow, H. R. Smyth, S. K. Keays & R. R. Peal. (*J. Brit. Instn Radio Engrs*, March 1959, Vol. 19, No. 3, pp. 135–147.) A pulsed transmitter with trickle-charged batteries and an internal capacitor antenna operates at 243 Mc/s for approximately 100 h. Cased in shock-absorbing foam, the device forms part of an aerofoil designed to fall at a safe speed and land clear of wreckage when automatically released from a crashing aircraft.

621.396.933.23 **1892**

B.L.E.U.'s Automatic Landing System.—(*Brit. Commun. Electronics*, Dec. 1958, Vol. 5, No. 12, p. 927.) A system developed by the Blind-Landing Experimental Unit of the Royal Aircraft Establishment. See 464 of February.

621.396.962.25/.3 : 621.396.969.11 **1893**

A Comparison between Pulse and Frequency Modulation Echo-Ranging Systems.—L. Kay. (*J. Brit. Instn Radio Engrs*, Feb. 1959, Vol. 19, No. 2, pp. 105–113.) The f.m. system is more flexible in its design parameters than the pulse system. Where the echo/background ratio is not important the f.m. system can provide a higher information rate than the pulse system.

621.396.962.3 : 621.396.665 **1894**

The Effects of Automatic Gain Control Performance on the Tracking Accuracy of Monopulse Radar Systems.—J. H. Dunn & D. D. Howard. (*Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, pp. 430–435.) An analysis including practical data shows that a short-time-constant fast-acting a.g.c. will minimize tracking noise. Servo bandwidth should be as small as possible.

621.396.962.38 **1895**

A Practical Application of Phase-Measuring Techniques to Precision Angle and Distance Measurements.—W. J. Thompson. (*Trans. Inst. Radio Engrs*, March 1957, Vol. I-6, No. 1, pp. 12–17. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 898.)

621.396.965 : 621.318.134 **1896**

Volumetric Scanning of a Radar with Ferrite Phase Shifters.—F. E. Goodwin & H. R. Senf. (*Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, pp. 453–454.)

621.396.969.11 **1897**

VOR — Compatible Doppler Omnirange, Design Considerations.—P. G. Hansel. (*Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, pp. 443–444.)

621.396.969.14 : 656.1 **1898**

The Telefunken Traffic Radar.—H. Lueg, W. Schallehn & H. Toedter. (*Elektrotech. Z., Edn B*, 21st Oct. 1958, Vol. 10, No. 10, pp. 385–390.) Description of the design and operation of Doppler radar equipment for the measurement of the speed of road vehicles.

MATERIALS AND SUBSIDIARY TECHNIQUES

533.5 **1899**

Theory and Design of Getter-Ion Pumps.—L. Holland. (*J. sci. Instrum.*, March 1959, Vol. 36, No. 3, pp. 105–116.) Includes a survey of the sorption and deposition properties of a range of getter materials. 63 references.

535.215 **1900**

Gain-Bandwidth Product for Photoconductors.—A. Rose & M. A. Lampert. (*RCA Rev.*, March 1959, Vol. 20, No. 1, pp. 57–68.) This product is shown to be proportional to an enhancement factor *M* dependent on the trap distribution. For many materials the maximum value of *M* is unity; much greater values may however be obtained under certain circumstances. See also 1901 below.

535.215 **1901**

Properties of Deep Traps derived from Space-Charge-Current Flow and Photoconductive Decay.—R. W. Smith. (*RCA Rev.*, March 1959, Vol. 20, No. 1, pp. 69–78.) The gain-bandwidth product is evaluated from measurements on a CdS single crystal. By using different light levels and a range of voltages, the Fermi level is moved to scan the deep-lying states.

535.215 **1902**

Gains, Response Times, and Trap Distributions in Powder Photoconductors.—H. B. DeVore. (*RCA Rev.*, March 1959, Vol. 20, No. 1, pp. 79–91.) Measurements have been made on CdS and CdSe powders and the data compared with the analytic expression given by Rose & Lampert (1900 above) relating the gain-bandwidth product to the dielectric relaxation time.

535.215 : 546.47-31 **1903**

The Field Effect in Insulating ZnO Powder.—W. Ruppel. (*Z. Phys.*, 28th July 1958, Vol. 152, No. 2, pp. 235–241.) Lifetime and trap concentration are calculated from the results of field-effect measurements on photoconductive ZnO powder. See also 877 of March (Ruppel et al.).

535.215 : 546.482.21 **1904**

The Spectral Distribution of Photoconductivity in CdS Single Crystals.—K. W. Böer & H. Gutjahr. (*Z. Phys.*, 28th July 1958, Vol. 152, No. 2, pp. 203–213.) The distribution of photoconductivity is investigated in relation to crystal structure and as a function of applied voltage.

535.215 : 546.482.21 : 548.5 **1905**

Vaporization-Crystallization Method for Growing CdS Single Crystals.—D. R. Boyd & Y. T. Sihvonen. (*J. appl. Phys.*, Feb. 1959, Vol. 30. No. 2, pp. 176–179.)

535.215 : 546.482.31 **1906**

Characteristics of the Increased Conductivity of Cadmium Selenide Single Crystals under X-Ray Excitation: Parts 1 & 2.—S. V. Svechnikov. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2492–2506.)

535.215 : 546.482.41 : 539.23 **1907**

High-Voltage Photovoltaic Effect.—B. Goldstein & L. Pensak. (*J. appl. Phys.*, Feb. 1959, Vol. 30, No. 2, pp. 155–161.) See 1756 and 1757 of 1958.

535.37 : 546.472.21 **1908**

On Certain Chromatic Aspects of the Photoluminescence of a ZnS-Cu Phosphor.—J. P. Leroux & P. Thureau. (*C. R. Acad. Sci., Paris*, 29th Sept. 1958, Vol. 247, No. 13, pp. 924–926.) The colour of the light emitted by a ZnS-Cu phosphor depends on the intensity of excitation and on the surface density of the irradiated powder.

535.376 **1909**

Electroluminescence.—V. E. Oranovskiĭ. (*Priroda, Mosk.*, Nov. 1958, No. 11 pp. 17–22.) Description of the process of luminescence of substances such as Zn_2SiO_4, BN, $BaTiO_3$, $SrTiO_3$ and TiO_2 and particularly ZnS and ZnSe and their application to television screens.

535.376 : 546.281.26 **1910**

Impurity Bands and Electroluminescence in SiC *p-n* Junctions.—L. Patrick & W. J. Choyke. (*J. appl. Phys.*, Feb. 1959, Vol. 30, No. 2, pp. 236–248.) A study of the electroluminescence of certain SiC *p-n* junctions, between 77°K and 830°K, and over a range of 10^4 in current density, has been used to verify and extend a three-part model of the junctions derived from electrical measurements.

537.227 **1911**

Anomalous Polarization in Ferroelectrics and Other Oxides.—J. D. Hurd, A. W. Simpson & R. H. Tredgold. (*Proc. phys. Soc.*, 1st March 1959, Vol. 73, No. 471, pp. 448–454.) It is shown that a number of metal oxides have a large relaxation polarization which cannot be explained in terms of ferroelectricity or of the Maxwell-Wagner effect. Experimental results indicate that the materials behave as solid-state secondary cells.

537.227 **1912**

Nonferroelectric Phase Transitions in Solid Solutions Formed in (Ca, Sr) (Ti, Zr)O_3 and Na (Nb, Ta)O_3 Systems.—G. A. Smolenskiĭ, V. A. Isupov, A. I. Agranovskaya & E. D. Sholokhova. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2528–2534.) An investigation of the temperature dependence of permittivity.

537.227 : 546.431.824-31 **1913**

Effect of γ-Ray and Pile Irradiation on the Coercive Field of $BaTiO_3$.—I. Lefkowitz & T. Mitsui. (*J. appl. Phys.*, Feb. 1959, Vol. 30, No. 2, p. 269.)

537.311.31 **1914**

Solution of Bloch's Integral Equation for Metal Electrons in an Electric Field for the Whole Temperature Range.—D. Langbein. (*Z. Phys.*, 11th July 1958, Vol. 152, No. 1, pp. 123–142.)

537.311.33 **1915**

Decay of Excess Carriers in Semiconductors.—K. C. Nomura & J. S. Blakemore. (*Phys. Rev.*, 1st Dec. 1958, Vol. 112, No. 5, pp. 1607–1615.) "A discussion is given of the nonlinear differential equations which govern the decay of excess carriers with arbitrary densities. The form of decay is explored for situations where the Fermi level is in the same half of the energy gap as the recombination level; criteria are established for both strong and weak trapping in addition to recombinative action. Analytic results are augmented and illustrated by numerically computed decay curves for a variety of circumstances. The separate solutions for holes and electrons are combined to show various kinds of behaviour for photoconductive lifetime."

537.311.33 **1916**

Effects of Carrier Injection on the Recombination Velocity in Semiconductor Surfaces.—G. C. Dousmanis. (*J. appl. Phys.*, Feb. 1959, Vol. 30, No. 2, pp. 180–184.) Predictions of theory based on the Shockley-Read model are illustrated with curves of surface recombination velocity as a function of the fractional excess carrier density where, for Ge, experimentally determined surface parameters are used.

537.311.33 **1917**

Narrowing the Energy Gap in Semiconductors by Compensation.—F. Stern & J. R. Dixon. (*J. appl. Phys.*, Feb. 1959, Vol. 30, No. 2, pp. 268–269.) The addition of large and equal numbers of both donor and acceptor impurity atoms has the effect of lowering the bottom of the conduction band and raising the top of the valence band, thus narrowing the energy gap while maintaining a low carrier concentration. This has been verified by doping InAs with S and Zn.

537.311.33 **1918**

Change of Semiconductor Properties with Fusion.—A. I. Gubanov. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2510–2516.) The variations in width of the forbidden zone and effective masses of the charge carriers on fusion are considered in relation to the variations produced by deformation of the crystal.

537.311.33 **1919**

Distribution of Non-equilibrium Charge Carriers in the Base Region of a *p-n* Junction with a High Injection Coefficient.—M. I. Iglitsyn, Yu. A. Kontsevoĭ & A. I. Sidorov. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2458–2460.) Stationary conditions of *p-n* junctions in semiconductors with an arbitrary injection level and lifetime depending upon this level are expressed by a system of equations, and the distribution of minority-carrier concentration in the base region is shown graphically.

537.311.33 **1920**

Vitreous Semiconductors.—T. N. Vengel' & B. T. Kolomiets. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2484-2491.) Eleven compounds in the system As_2Se_3–As_2Te_3 have been investigated, and their photoelectric and thermoelectric properties are examined in relation to their chemical composition. See also 2796 of 1957 and 1766 *n* of 1958 (Goryunova & Kolomiets).

537.311.33 : 535.215 **1921**

Measurement of Lifetime by the Photoconductive Decay Method.—B. K. Ridley. (*J. Electronics Control*, Dec. 1958, Vol. 5, No. 6, pp. 549–558.) General equations have been derived considering both steady-state and pulsed initial distributions. The decay is exponential only for zero surface recombination. With high surface recombination velocities the loss of exponential nature of the decay leads to an error in the measured value of filament lifetime. This error has been found numerically and graphically for various lifetimes and penetration depths.

537.311.33 : 537.533 **1922**

On the Electrostatic Electron Emission of Semiconductors.—I. I. Gofman, B. G. Smirnov, G. S. Spirin & G. N. Shuppe. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2662-2663.) Note of an experimental investigation of W_2C showing that the field-emission current/voltage characteristic is in qualitative agreement with theory [see 717 of 1956 (Stratton)].

537.311.33 : 538.63 **1923**

The Theory of Electrical and Photoelectric Effects for Three Carriers in a Magnetic Field.—A. K. Walton & T. S. Moss. (*Proc. phys. Soc.*, 1st March 1959, Vol. 73, No. 471, pp. 399–412.) "Formulae for the Hall, magnetoresistance, photoelectromagnetic and Dember effects due to electrons and slow and fast holes are derived for the cases of energy-independent relaxation time and lattice scattering. The results are discussed for the particular case of germanium."

537.311.33 : 546.26-1 : 537.32 **1924**

The Thermoelectric Power of a Semiconducting Diamond.—H. J. Goldsmid; C. C. Jenns & D. A. Wright. (*Proc. phys. Soc.*, 1st March 1959, Vol. 73, No. 471, pp. 393–398.) The thermoelectric power of *p*-type semiconducting diamond was measured between 220°K and 700°K. The phonon-drag component had a value of about 2·5 mV/°K at room temperature and varied with temperature approximately as $T^{-3\cdot6}$.

537.311.33 : 546.28 **1925**

Diffusion of Impurities into Evaporating Silicon.—R. L. Batdorf & F. M. Smits. (*J. appl. Phys.*, Feb. 1959, Vol. 30, No. 2, pp. 259–264.) A diffusion technique in a vacuum system where the only ambient is the vapour of a diffusing impurity is described and results of measurements are quoted.

537.311.33 : 546.28 **1926**

Mechanism of the Formation of Donor States in Heat-Treated Silicon.—W. Kaiser, H. L. Frisch & H. Reiss. (*Phys. Rev.*, 1st Dec. 1958, Vol. 112, No. 5, pp. 1546–1554.) A mechanism is proposed which accounts quantitatively for existing kinetic and extra-kinetic data on the system. It depends on reactions with atomically dissolved oxygen introduced during the process of crystal growing.

537.311.33 : 546.28 **1927**

A Comparison of the Theory of Impact Ionization with Measurements on Silicon *p-n* Junctions.—F. W. G. Rose. (*J. Electronics Control*, Jan. 1959, Vol. 6, No. 1, pp. 70–73.) There is a voltage range over which the inverse current rises linearly with voltage on a log-log graph. It is shown that over this range the current is mainly due to impact ionization.

537.311.33 : 546.289 **1928**

Investigation of the Field Effect and Surface Recombination in Germanium Samples.—A. V. Rzhanov, Yu. F. Novototskiĭ-Vlasov & I. G. Neizvestnyĭ. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2440–2450.) The investigation shows that the action of ozone on Ge gives rise to 'fast' surface states some of which are recombination states. The density and energy positions of all states introduced by ozone are estimated. The cross-section of capture of an electron by recombination states is also evaluated. Preliminary data are derived concerning the dependence of surface recombination velocity on the electrostatic surface potential, from which the ratio of the effective cross-sections of capture of an electron and a hole are obtained.

537.311.33 : 546.289 **1929**

The Lifetime of Non-equilibrium Charge Carriers in Germanium with Arbitrary Injection Levels.—M. I. Iglitsyn, Yu. A. Kontsevoĭ & A. I. Sidorov. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2461–2468.) An investigation of the dependence of lifetime on the concentration of non-equilibrium charge carriers at different temperatures for Ge specimens alloyed with Sb. At room temperature, the lifetime decreases with increasing injection level for samples of high resistance, and increases for samples of low resistance. The type of level (donor or acceptor), their position in the forbidden band, the energy of ionization and the ratio of probabilities of recapture of electrons and holes can be determined.

537.311.33 : 546.289 **1930**

Measurement of Short Lifetimes of Charge Carriers in Germanium.—L. S. Smirnov. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2469–2471.) Measurements were carried out on Ge plates with a large-area *p-n* junction. The non-equilibrium carriers were excited by a monochromatic light source near the surface of the semiconductor, and the lifetime of the charge carriers, in the range $2 \times 10^{-6} - 10^{-8}$ s, was determined from the short-circuit current and the number of light quanta absorbed. The temperature dependence of lifetime and the position of recombination levels were also evaluated.

537.311.33 : 546.289 **1931**

The Influence of an Intense Electric Field on Germanium-Diode Transparency.—Yu. I. Ukhanov & S. G. Shul'man. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2507–2509.) Measurements were made on a Ge *p-n* junction with an infrared beam directed perpendicularly to the applied electric field using 1–3-μs pulses to give current densities up to 20 mA/mm². The infrared transparency varied proportionally with the reverse current. Cooling the specimen to 78°K had no effect on the transparency. With a forward current a decrease in transparency was observed.

537.311.33 : 546.289 **1932**

Dependence of the Lifetime of Injected Current Carriers on the Concentration of Antimony Impurity in Germanium.—V. E. Lashkarev, V. G. Litovchenko, N. M. Omel'yanovskaya, P. N. Boudarenko & V. I. Strikha. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2437–2439.) Report of an investigation carried out close to the limit of solubility of Sb in Ge. Results indicate that the Sb impurity atoms are not directly effective as recombination centres but that the recombination is due to deeply embedded and uncontrolled impurities originally in the Ge or introduced with the Sb.

537.311.33 : 546.289 **1933**

Investigation of the Recombination of Current Carriers in Germanium with Iron Impurity.—K. D. Glinchuk, E. G. Miselyuk & N. N. Fortunatova. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2451–2457.) The acceptor level situated at 0·27 eV from the conduction band in *n*-type Ge can be removed by annealing at 450°–500°C. This results in a sharp increase in the lifetime of non-equilibrium charge carriers. It may be explained by the deactivation of iron atoms following their expulsion from Ge lattices.

537.311.33 : 546.289 **1934**

The Effect of Annealing on Local Levels and the Lifetime of Non-equilibrium Current Carriers in Germanium with Iron Impurity.—K. D. Glinchuk, E. G. Miselyuk & N. N. Fortunatova. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2666–2667.)

537.311.33 : 546.289 **1935**

Orientation Control for Germanium Wafers.—B. J. Coughlin, G. L. Davis & R. L. Kingsnorth. (*J. sci. Instrum.*, March 1959, Vol. 36, No. 3, pp. 144–145.) A method for the accurate mounting of ingots for cutting and for determining the orientation of small wafers.

537.311.33 : 546.289 **1936**

Precision Measurement of the Lattice Constant of Germanium Single Crystals by the Method of Kossel and van Bergen.—G. Mack. (*Z. Phys.*, 11th July 1958, Vol. 152, No. 1, pp. 19–25.) See also 1937 below.

537.311.33 : 546.289 **1937**

Precision X-Ray Investigations on Germanium-Indium *p-n* Alloy Junc-

tions.—G. Mack. (*Z. Phys.*, 11th July 1958, Vol. 152, No. 1, pp. 26–33.) Report of measurements made by the method of Kossel and van Bergen.

537.311.33 : 546.289 **1938**

Light-Induced Plasticity in Germanium.—G. C. Kuczynski & R. H. Hochman. (*J. appl. Phys.*, Feb. 1959, Vol. 30, No. 2, p 267.)

538.311.33 : 546.681.19 **1939**

Piezoresistance in *n*-Type GaAs.—A. Sagar. (*Phys. Rev.*, 1st Dec. 1958, Vol. 112, No. 5, p. 1533.) "Piezoresistance and elastoresistance coefficients of *n*-type GaAs were determined at room temperature. The results are consistent with a spherical energy-band model as predicted by Callaway 1181б of 1957] from theory."

537.311.33 : 546.682.86 **1940**

Band Structure of InSb.—T. Igo, E. Yamaka & M. Yatani. (*Rep. elect. Commun. Lab., Japan*, June 1958, Vol. 6, No. 6, pp. 205–210.) A short discussion on the calculation of the band structure.

537.311.33 : 546.817.241 **1941**

Investigation of the Thermoelectric Properties of Lead Telluride.—E. Z. Gershteĭn, T. S. Stavitskaya & L. S. Stil'bans. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2472-2483.) An investigation of the properties of degenerate and non-degenerate samples shows the dependence of the length of the free path of electrons on temperature and energy. The possibility of a correlation between mobility and the temperature dependence of the width of the forbidden zone is also examined.

537.311.33 : 546.873.241 **1942**

The Electrical Conductivity and Hall Coefficient of Bismuth Telluride.—B. Yates. (*J. Electronics Control*, Jan. 1959, Vol. 6, No. 1, pp. 26–38.) These quantities have been examined, for a wide range of doping concentrations, over the temperature range 1·3°– 660 °K. The results cannot be explained by simple temperature-dependent scattering in the conduction or the valence band, nor are they in complete agreement with a model based on two scattering mechanisms operating in one band. A qualitative explanation of the results for *n*-type specimens, over a restricted temperature range, is possible on the basis of an impurity-band model.

537.311.33 : 621.923.7 **1943**

Improved Machine for Lapping Very Thin Slices of Semiconductor Materials.—D. Baker. (*J. sci. Instrum.*, March 1959, Vol. 36, No. 3, pp. 145–147.)

537.312.8 : 539.234 **1944**

Magnetic Resistance Variation in Vapour-Deposited Nickel Films as a Function of Film Thickness and Structure.—W. Hellenthal. (*Z. Phys.*, 2nd June 1958, Vol. 151, No. 4, pp. 421–430.) A quantitative link with conditions observed in solid material is obtained if account is taken of the increase in resistivity as a function of film thickness and structure, and the decrease of spontaneous magnetization.

537.581 : 546.77 : 538.63 **1945**

Effect of a Magnetic Field on Thermionic Emission from Molybdenum.—J. Greenburg. (*Phys. Rev.*, 15th Dec. 1958, Vol. 112, No. 6, pp. 1898–1900.) Experiment shows that an applied field of up to 6 000 G has no effect on the saturation current density.

538.22 **1946**

Antiferromagnetism of $CuF_2.2H_2O$.—G. S. Verma & K. Tokunaga. (*Phys. Rev.*, 1st Dec. 1958, Vol. 112, No. 5, pp. 1521–1522.) "The perpendicular and parallel magnetic susceptibilities have been calculated for $CuF_2 . 2H_2O$ on the basis of Nakamura's theory. The computed values of molar susceptibility for the same compound have been compared with the recent measurements of Bozorth & Nielsen [3163 of 1958] and are found to be in good agreement."

538.22 : 538.569.4 **1947**

Evidence for Antiferromagnetism in $Cu_3(CO_3)_2(OH)_2$.—R. D. Spence & R. D. Ewing. (*Phys. Rev.*, 1st Dec. 1958, Vol. 112, No. 5, pp. 1544–1545.) Proton resonance measurements at low temperatures are reported. At 1·86 °K a transition takes place and the resonance pattern below this temperature indicates an antiferromagnetic state.

538.22 : 538.569.4 **1948**

Magnetic Resonance Line Shapes at the Onset of Saturation.—D. F. Holcomb. (*Phys. Rev.*, 1st Dec. 1958, Vol. 112, No. 5, pp. 1599–1603.) Results are given of measurements of magnetic resonance absorption line shapes and widths in Li metal and CaF_2 crystals, as a function of r.f. power level, up to the region of appreciable saturation.

538.22 : 538.569.4 **1949**

Paramagnetic Resonance of Fe^{3+} in Sapphire at Low Temperatures.—G. S. Bogle & H. F. Symmons. (*Proc. phys. Soc.*, 1st March 1959, Vol. 73, No. 471, pp. 531–532.)

538.221 **1950**

Search for New Heusler Alloys.—D. P. Morris, R. R. Preston & I. Williams. (*Proc. phys. Soc.*, 1st March 1959, Vol. 73, No. 471, pp. 520–523.) A brief report is given of investigations on silver and gold ternary alloys.

538.221 : 534.213–8 **1951**

Ultrasonic Wave Propagation in a Nickel Single Crystal.—J. de Klerk. (*Proc. phys. Soc.*, 1st March 1959, Vol. 73, No. 471, pp. 337–344.) An improved pulse technique has been used to investigate the dynamic elastic constants and energy losses with and without an applied magnetic field. Energy losses are substantially reduced when the material is magnetized to saturation.

538.221 : 538.632 **1952**

Spin-Orbit Coupling and the Extraordinary Hall Effect.—C. Strachan & A. M. Murray. (*Proc. phys. Soc.*, 1st March 1959, Vol. 73, No. 471, pp. 433–447.) Quantum-mechanics transport theory is used to evaluate the magnitude of the extraordinary Hall coefficient.

538.221 : 538.632 **1953**

Two Hall Effects of Iron-Cobalt Alloys.—F. P. Beitel, Jr, & E. M. Pugh. (*Phys. Rev.*, 1st Dec. 1958, Vol. 112, No. 5, pp. 1516–1520.) The ordinary and extraordinary Hall coefficients, and the resistivity, of Fe-Co alloys have been measured at 77°K, 169°K and room temperature. The results are analysed in terms of two models for the electronic structure.

538.221 : 538.652 : 621.372.41 **1954**

Ferrites for Magnetostriction Oscillators in Filter Circuits—S. Schweizerhof. (*Nachrichtentech. Z.*, April 1958, Vol. 11, No. 4, pp. 179–185.) The performance of ferrite rings with improved temperature characteristics and high *Q*-values is discussed.

538.221 : 539.23 **1955**

Magnetization Reversal by Rotation and Wall Motion in Thin Films of Nickel-Iron Alloys.—E. M. Bradley & M. Prutton. (*J. Electronics Control*, Jan. 1959, Vol. 6, No. 1. pp. 81–96.) Measurements of the 400-c/s hysteresis loops on uniaxial films indicate that both coherent rotation and domain wall motion can occur depending on film orientation and thickness.

538.221 : 621.318.134 **1956**

Domain Behaviour in some Transparent Magnetic Oxides.—R. C. Sherwood, J. P. Remeika & H. J. Williams. (*J. appl. Phys.*, Feb. 1959, Vol. 30, No. 2, pp. 217–225.) Magnetic domains were observed by means of the Faraday effect, and by the Bitter technique in a number of compounds with the spinel, magnetoplumbite and perovskite-like structures.

538.221 : 621.318.134 **1957**

Some Investigations on Li-Zn Ferrites.—Kh. S. Valeev, N. G. Drozdov & A. L. Frumkin. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2517–2527.) As a result of an investigation of the influence of composition and temperature on the magnetic properties of Li-Zn ferrites, materials were obtained having low losses in the range 20–75 Mc/s. The permeability at a wavelength of 3·2 cm was found to be less than 1.

538.221 : 621.318.134 **1958**

Domain Wall Motion and Ferrimagnetic Resonance in a Manganese Ferrite.—J. F. Ditton, Jr, & H. E. Earl, Jr. (*J. appl. Phys.*, Feb. 1959, Vol. 30, No. 2, pp. 202–213.) Very simple domain walls were driven through single crystals of a high-resistivity Mn ferrite. A considerable temperature range was covered, through which the material constants varied substantially, as did the losses encountered. Ferrimagnetic resonance experiments are reported for the same material over the same temperature range.

538.221 : 621.318.134 : 621.357.7 **1959**

Preparing Ferrites by Continuous Electrolytic Co-precipitation.—H. B. Beer & G. V. Planer. (*Brit. Commun. Electronics*, Dec. 1958, Vol. 5, No. 12, pp. 939–941.) A new method is outlined having the advantage of greater economy and simplicity, with increased homogeneity and chemical purity of the product.

621.318.2 **1960**

Permanent Magnet Stability.—J. E. Gould. (*Instrum. Practice*, Oct. 1958, Vol. 12, No. 10, pp. 1083–1091.) Consideration is given to the influence on magnetization of external magnetic fields, thermal effects, mechanical shock and nuclear radiation. Under normally steady conditions, magnetization changes occur which are proportional to the logarithm of time and can be minimized by choice of material, working permeance and prestabilization.

621.318.2-492 **1961**

Powdered Magnets.—G. Sideris. (*Electronics*, 27th Feb. 1959, Vol. 32, No. 9, p. 69.) The properties of sintered alloy, sintered oxide and pressed-powder permanent magnets are tabulated.

669.71 : 621.795 **1962**

Aluminium Finishes for Use in Electronics.—W. E. Pocock. (*Electronics*, 20th Feb. 1959, Vol. 32, No. 8, pp. 58–59.) A survey of the properties and applications of various surface finishes.

MATHEMATICS

517.93 **1963**

Parametric Excitation.—N. Minorsky. (*C. R. Acad. Sci., Paris*, 28th July 1958, Vol. 247, No. 4, pp. 406–408.) Simplified solutions of the equation

$$\ddot{x} + b\dot{x} + x + (a - cx^2)\, x \cos 2t + cx^3 = 0$$

are obtained by a development of the method described earlier (2951 of 1951).

518.1 **1964**

Some Properties of Strongly Connected Graphs.—B. Roy. (*C. R. Acad. Sci., Paris*, 28th July 1958, Vol. 247, No. 4, pp. 399–401.) Two general theorems are stated concerning flow-diagram analyses.

MEASUREMENTS AND TEST GEAR

531.76 : 621.374.32 **1965**

Vernier Chronotron times Nuclear Particle Flight.—H. W. Lefevre & J. T. Russell. (*Electronics*, 6th March 1959, Vol. 32, No. 10, pp. 44–47.) A time interval analyser with a resolution better than 10^{-9} sec is described.

621.317.2 : 621.373.4 **1966**

Timed-Signal Generator with Flexible Output.—D. E. Minow. (*Electronics*, 6th March 1959, Vol. 32, No. 10, pp. 52–53.) Details of a portable unit with two output channels delivering pulses of controllable duration, amplitude, carrier content and repetition rate.

621.317.3 : 537.311.33 : 621.3-71 **1967**

Cryostat for Measuring the Electrical Properties of High-Resistance Semiconductors at Low Temperatures.—W. H. Mitchell & E. H. Putley. (*J. sci. Instrum.*, March 1959, Vol. 36, No. 3, pp. 134–136.) Resistances of up to $10^{11}\ \Omega$ may be measured at temperatures down to 2°K, using a vibrating-reed electrometer. The lower limit for Hall mobility is 10 cm²/V. sec.

621.317.34.029.63/.64 : 621.372.5 **1968**

Three-Point Method of Measuring U.H.F. Quadripoles.—J. Smejkal : L. Mollwo. (*Hochfrequenztech. u. Elektroakust.*, March 1958, Vol. 66, No. 5, pp. 167–169.) Comment on 1213 of 1958 and author's reply.

621.317.35 : 621.372.54 **1969**

Calculation of the Resolving Power of Automatic Frequency Analysers.—N. V. Terpugov. (*Radiotekhnika i Elektronika*, June 1957, Vol. 2, No. 6, pp. 796–806.) A method is described for calculating the dynamic frequency characteristics of filter systems. Factors and coefficients for evaluating the characteristics are given and results of an experimental investigation of nine types of filter are tabulated.

621.317.373 **1970**

Phase-Angle Measurement.—P. Kundu. (*Electronic Radio Engr*, April 1959, Vol. 36, No. 4, pp. 150–154.) The signals are applied to a heptode mixer whose differential anode current with respect to the reference value for quadrature inputs is a measure of phase.

621.317.373 : 621.3.05 **1971**

Measurement of Phase Difference on Long Power Transmission Lines.—G. Zito. (*Alta Frequenza*, June/Aug. 1958, Vol. 27, Nos. 3/4, pp. 378–400.) The apparatus described can be used for determining, with an error not exceeding ± 1%, the phase relations between different power supply systems feeding a network of television stations.

621.317.42 : 537.311.33 **1972**

A New Method of Measuring Magnetic Field Intensity.—G. E. Pikus & O. V. Sorokin. (*Zh. tekh. Fiz.*, Nov. 1957, Vol. 27, No. 11, pp. 2647–2651.) This method is based on the change of concentration of charge carriers in a thin semiconductor specimen located in a magnetic field and passing an alternating current. By virtue of a linear relation between the applied voltage and the magnetic field intensity the latter can be estimated in the range $5 \times 10^3 - 10^{-5}$ oersteds.

621.317.44 : 538.632 **1973**

Magnetic Field Probe of High Sensitivity and Resolution.—B. Kostyshyn & D. D. Roshon, Jr. (*Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, p. 451.) Note on the performance characteristics of a miniature Hall effect probe of Bi.

621.317.71 : 621.314.7 **1974**

Transistor Junction Temperature.—H. Sutcliffe & D. J. Matthews. (*Electronic Radio Engr*, April 1959, Vol. 36, No. 4, pp. 143–144.) A circuit is described for measurement of the temperature-dependent base leakage current in a class-C transistor stage.

621.317.742 : 621.317.755 **1975**

4 000-Mc/s-Band Wide-Band V.S.W.R. Scanner.—Y. Ninomiya, N. Miyamoto & A. Yanagi. (*Rep. elect. Commun. Lab., Japan*, May 1958, Vol. 6, No. 5, pp. 154–157.) Voltage s.w.r. is displayed on an oscilloscope over the frequency range 3 600–4 200 Mc/s.

621.317.75 : 621.374 **1976**

Amplitude Slicer for Signal Analysis.—T. A. Bickart. (*Electronics*, 27th Feb. 1959, Vol. 32, No. 9, pp. 64–65.) Description of a circuit providing a rectangular output pulse whose width is proportional to the time during which the input signal lies between specified voltage levels.

621.317.755 **1977**

The Cathode-Ray Oscilloscope: a Survey.—J. F. Golding. (*Brit. Commun. Electronics*, Jan. 1959, Vol. 6, No. 1, pp. 27–33.) Abridged specifications of oscilloscopes available in U.K. are tabulated.

621.317.755.087.5 **1978**

Automatic Recorder for Cathode-Ray Oscillography.—H. Lindner & U. Gladhorn. (*Arch. tech. Messen*, April & Aug. 1958, Nos. 267 & 271, pp. 79–82 & 171–174.) An oscilloscope camera for film or drum recording is described. With an adaptor the camera can be used at right angles to the c.r. tube screen. Details are also given of an electronic control unit for the provision of triggering and brightening pulses.

621.317.763 : 621.314.7 **1979**

Transistorized Absorption Wavemeter.—G. W. Short. (*Wireless World*, April 1959, Vol. 65, No. 4, pp. 193–196.) A description of a wavemeter incorporating a modulating oscillator and covering the frequency range 1–100 Mc/s.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

526.2 : 621.396.9 **1980**

Electronic Principles of the Tellurometer.—T. L. Wadley. (*Trans. S. Afr. Inst. elect. Engrs*, May 1958, Vol. 49, Part 5, pp. 143–161. Discussion, pp. 161–172.) A detailed description of the instrument. See also 3250 of 1957 and 874 of 1958 (Hammond).

531.767 : 621.396.96 **1981**

Radar Meter helps enforce Traffic Laws.—J. Barker. (*Electronics*, 6th March 1959, Vol. 32, No. 10, pp. 48–49.) A battery-powered system is described, converting Doppler shift to give a direct reading of vehicle speed.

538.569.4 : 621.372.8 **1982**

Low-Power Microwave Reflection Bridge.—R. C. Rempel & H. E. Weaver. (*Rev. sci. Instrum.*, Feb. 1959, Vol. 30, No. 2, p. 137.) A standard microwave bridge for detection of electron paramagnetic resonance is modified by introducing an adjustable ferrite isolator in the sample arm.

621.313.334 : 621.318.57 : 621.314.7 **1983**

Four-Transistor Inverter drives Induction Motor.—W. H. Card. (*Electronics*, 20th Feb. 1959, Vol. 32, No. 8, pp. 60–61.) Direct-current motors used in low-pressure or explosive environments can be replaced with induction motors by employing transistors as controlled switches to provide two-phase square-wave output from a single d.c. source.

621.384.6 **1984**

Longitudinal Space-Charge Effects in Particle Accelerators.—C. E. Nielsen & A. M. Sessler. (*Rev. sci. Instrum.*, Feb. 1959, Vol. 30, No. 2, pp. 80–89.) "The modification of the single-particle theory of particles subject to r.f. acceleration caused by electrostatic repulsion between particles is calculated."

621.384.611 **1985**

A Fixed-Frequency Cyclotron with One Dee.—R. Bock, A. Doehring, J. Jänecke, O. Knecht, L. Koester, H. Maier-Leibnitz, C. Schmelzer & U. Schmidt-Rohr. (*Z. angew. Phys.*, Feb. 1958, Vol. 10, No. 2, pp. 49–55.) The design and construction of a 12-MeV deuteron accelerator installed at Heidelberg are described.

621.384.7 : 537.54 **1986**

High-Current Ion Source.—R. G. Meyerand, Jr, & S. C. Brown. (*Rev. sci. Instrum.*, Feb. 1959, Vol. 30, No. 2, pp. 110–111.) An ion source of simple design and construction is described capable of producing a pulsed ion beam of ½ A.

621.385.833 **1987**

A High-Resolution Emission Microscope for Viewing Surfaces with Electrons Released by Ultraviolet Radiation.—W. Koch. (*Z. Phys.*, 11th July 1958, Vol. 152, No. 1, pp. 1–18.) The 40-kV microscope described has an electron-optical magnification of 700 with a resolution of 1 000 Å.

621.385.833 **1988**

Electrostatic Charging of the Photosensitive Material in Electron Microscopes.—E. Kinder. (*Z. angew. Phys.*, Feb. 1958, Vol. 10, No. 2, pp. 95–98.) Methods of eliminating or reducing excessive charges are discussed.

621.387.4 : 621.395.625.3 **1989**

Magnetic Recording of Pulse-Amplitude Data.—J. Baumgardner. (*Rev. sci. Instrum.*, Feb. 1959, Vol. 30, No. 2, pp. 134–135.) The limitations and advantages of the direct recording of pulses on magnetic tape are considered.

621.387.462 **1990**

Diamond Conduction Counters with Small Electrode Separations.—F. C. Champion & S. B. Wright. (*Proc. phys. Soc.*, 1st March 1959, Vol. 73, No. 471, pp. 385–392.) Measurements of the charge pulse-height/applied-field curve at electrode spacings from 10 μ to 1 mm show variations in the characteristics of the curves which cannot be explained in terms of trapping field distribution or field distortion due to dark current.

621.397.3 : 621.39 **1991**

Automatic Character Recognition.—K. Steinbuch. (*Nachrichtentech. Z.*, April & May 1958, Vol. 11, Nos. 4 & 5, pp. 210–219 & 237–244.) Detailed investigation of the problems of recognizing written or printed characters, particularly numerals. A number of scanning methods are reviewed. 30 references including patents.

621.398 : 621.376.55 **1992**

Telemetry Demodulator using Modified AND Gate.—L. Weisman. (*Electronics*, 20th Feb. 1959, Vol. 32, No. 8, pp. 54–57.) A description of a pulse-position telemetry system, with details of the demodulator channels.

PROPAGATION OF WAVES

621.396.11 : 523.53 **1993**

Observations of Direction of Arrival of Long-Duration Meteor Echoes in Forward Scatter Propagation.—T. Hagfors & B. Landmark. (*J. geophys. Res.*, Jan. 1959, Vol. 64, No. 1, pp. 19–22.) The angular distribution of enduring meteor bursts is shown to be similar to that observed for short-duration specular reflections and in marked contrast to that of the turbulent background component. It is concluded that the long-duration echoes are from specularly reflecting meteor trails and not from trails broken up by atmospheric turbulence.

621.396.11 : 551.311.122 : 537.226 **1994**

The Electromagnetic Properties of Glacier Ice.—M. Lafargue & R. Millecamps. (*C. R. Acad. Sci., Paris*, 22nd Sept. 1958, Vol. 247, No. 12, pp. 884–886.) Experiments were made on e.m. wave propagation at frequencies in the range 50–400 kc/s in a glacier. The ice appears to behave as a low-loss dielectric. At 150 kc/s reception was possible at a distance of 5 km from a 1 W transmitter on the glacier.

621.396.11 : 551.510.52 **1995**

Calculation of Multiple Dispersion in U.S.W. Scatter Propagation in the Troposphere.—D. M. Vysokovskiĭ. (*Radiotekhnika i Elektronika*, June 1957, Vol. 2, No. 6, pp. 807–809.) An approximate evaluation of energy loss due to scatter.

621.396.11 : 551.510.535 **1996**

Propagation of Electrical Waves along a Plasma Layer Bounded by a Dielectric with a Longitudinal Magnetic Field.—W. O. Schumann. (*Z. angew. Phys.*, Jan. 1958, Vol. 10, No. 1, pp. 26–31.) See also 1881 of 1957.

621.396.11 : 551.510.535 **1997**

An Analysis of Drifts of the Signal Pattern associated with Ionospheric Reflections.—Yerg. (See 1884.)

621.396.11 : 551.510.535 **1998**

The Effects of Ionospheric Irregularities and the Auroral Zone on the Bearings of Short-Wave Radio Signals.—H. A. Whale. (*J. atmos. terr. Phys.*, Feb. 1959, Vol. 13, Nos. 3/4, pp. 258–270.) For sunlit paths less than 15 000 km most of the daily variation of bearing of F_2-propagated signals is said to arise from refraction in the F_1 region. During night-time, obstructing patches of E-region ionization are suggested as the cause. For paths longer than 15 000 km the direction of the transmitter beam may be relatively unimportant in determining the incoming bearing, and large observed bearing changes are supposed to arise from absorbing and scattering processes in the auroral regions. A method for plotting the shape of the absorbing parts of the auroral zones is described.

621.396.11 : 621.396.67 **1999**

Transmission of Power in Radio Propagation.—J. R. Wait. (*Electronic Radio Engr*, April 1959, Vol. 36, No. 4, pp. 146–150.) As the transmission of power between aerials can be influenced appreciably by the immediate neighbourhood of the aerials, it is suggested that the total transmission loss should be divided into two parts, one of which should account for this effect.

621.396.11.029.6 **2000**

Symposium on Long-Distance Propagation above 30 Mc/s.—(*Proc. Instn elect. Engrs*, Part B, 1958, Vol. 105, Supplement No. 8, pp. 1–191.) The following papers were read at the I.E.E. Symposium held in London, 28th January 1958.

Ionospheric Forward-Scatter Propagation:

(*a*) Survey of the Gibraltar-United Kingdom Ionospheric Scatter Measurements.—F. A. Kitchen & G. Millington (pp. 2–6).

(*b*) A Scatter-Signal Analyser.—P. H. Cutler & D. Williams (pp. 7–11).

(*c*) The Choice of Aerial Height for Ionospheric Scatter Links.—E. Fitch & R. Ruddlesden (pp. 12–18).

(*d*) The Structure of High-Frequency Ionospheric Scatter Signals.—D. Williams (pp. 19–26).

(*e*) Radio Interference as a Factor in Ionospheric Scatter Communication.—G. A. Isted (pp. 27–35).

(*f*) Analysis of Gibraltar–United Kingfom Ionospheric Scatter Signal Recordings.—G. A. Isted (pp. 36–44).

(*g*) Polar - Diagram Requirements for Aerials for Communication by Ionospheric Scatter.—D. H. Shinn (pp. 45–52).

(*h*) The Angular Distribution of Energy Received by Ionospheric Forward Scattering at Very High Frequencies.—W. C. Bain (pp. 53–55).

(*i*) The Direction and Amplitude of Reflections from Meteor Trails and Sporadic-E Ionization on a 1 740-km North–South Path at Very High Frequencies.—R. W. Meadows (pp. 56–64).

(*j*) Short Bursts of Amplitude of a 50-Mc/s Wave Received over a Distance of 480 km.—G. S. Kent (pp. 65–69).

(*k*) Amplitude of Very-High-Frequency Signals Reflected from the Sporadic-E Layer in North-West Europe.—P. J. Brice (pp. 70–72).

Discussion (pp. 73–78).

Tropospheric Propagation Beyond the Horizon:

(*l*) Guglielmo Marconi and Communication Beyond the Horizon: a Short Historical Note.—G. A. Isted (pp. 79–83).

(*m*) A Survey of Tropospheric Wave Propagation Measurements by the B.B.C., 1946–1957.—R. A. Rowden, L. F. Tagholm & J. W. Stark (pp. 84–90).

(*n*) The Measurement and Prediction of V.H.F. Tropospheric Field Strengths at Distances Beyond the Horizon.—J. K. S. Jowett (pp. 91–96).

(*o*) The Effects of Atmospheric Discontinuity Layers up to and including the Tropopause on Beyond-the-Horizon Propagation Phenomena.—B. J. Starkey, W. R. Turner, S. R. Badcoe & G. F. Kitchen (pp. 97–105).

(*p*) Some Investigations of Metre-Wave Radio Propagation in the Transhorizon Region.—F. A. Kitchen, E. G. Richards & I. J. Richmond (pp. 106–116).

(*q*) The Reduction of Threshold by the Use of Frequency Compression.—A. J. Buxton & M. O. Felix (pp. 117–121).

(*r*) Propagation Measurements at 3 480 Mc/s over a 173-Mile Path.—B. C. Angell, J. B. L. Foot, W. J. Lucas & G. T. Thompson (pp. 128–142).

(*s*) Some Tropospheric Scatter Propagation Measurements and Tests of Aerial Siting Conditions at 858 Mc/s.—G. C. Rider (pp. 143–152).

(*t*) The Long-Range Propagation of Radio Waves at 10-cm Wavelength.—W. R. R. Joy (pp. 153–157).

(*u*) Radio Propagation Far Beyond the Horizon at about 3·2-cm Wavelength.—W. R. R. Joy (pp. 158–164).

(*v*) A Review of Tropospheric Scatter Propagation Theory and its Application to Experiment.—M. A. Johnson (pp. 165–176).

(*w*) The Estimation of Transmission Loss in the Transhorizon Region.—E. G. Richards (pp. 177–183).

Discussion (pp. 122–126, 184–188).

621.396.11.029.62 : 551.510.535 : 551.594.5 **2001**

A Bistatic Radio Investigation of Auroral Ionization.—C. Collins & P. A. Forsyth. (*J. atmos. terr. Phys.*, Feb. 1959, Vol. 13, Nos. 3/4, pp. 315–345.) The scattering of radio waves in the upper atmosphere at times of auroral disturbance has been studied by means of some twenty 30–50-Mc/s radio systems in Canada, each having transmitter and receiver about 1 000 km apart. At least four different kinds of auroral events are distinguishable. Of these, two appear to be associated with different phases of visible aurora, the third with a later stage in the auroral process which is not observed visually, and a fourth with the recurrent daytime absorption which often precedes auroral disturbance. In these four events evidence is found for three separate scattering mechanisms, each of which has been proposed previously as the principal source of radar echoes from aurora.

621.396.11.029.63 : 621.396.812.3 **2002**

Propagation Tests at 1 000 Mc/s with Diversity Reception between Monte Penice and Monte Venda.—P. Quarta. (*Alta Frequenza*, June/Aug. 1958, Vol. 27, Nos. 3/4, pp. 219–225.) Analysis and discussion of test results obtained in 1954 over a 196-km path. See also 3522 of 1956 (Vecchiacchi).

621.396.8.029.62 **2003**

Long-Distance V.H.F. Reception.—H. V. Griffiths. (*Wireless World*, April 1959, Vol. 65, No. 4, pp. 179–181.) An analysis of observations made since 1946 shows that interference on Band I is due to three modes of propagation, F-layer, sporadic-E and tropospheric. The intensity of the interfering signal has not been measured, but the frequency of occurrence is correlated with the sunspot number for reception via the F layer and sporadic E.

RECEPTION

621.376.23 **2004**

The Optimum Detector with Log I_0 Characteristic for the Detection of Weak Signals in Noise.—B. S. Fleishman. (*Radiotekhnika i Elektronika*, June 1957, Vol. 2, No. 6, pp. 726–734.) The characteristics of the optimum log I_0 detector [see e.g. 2782 of 1953 (Middleton)] are calculated. Inaccuracies in earlier calculation of the I_0 expansion are pointed out and the application of the derived expression to radar is considered.

621.376.3 **2005**

Passage of Random Noise Signals through a Detector considering Biasing and Limiting Effects.—E. G. Logachev. (*Radiotekhnika i Elektronika*, June 1957, Vol. 2, No. 6, pp. 735–750.) A general expression is derived for the correlation function of the noise current at the output of a detector in the low-frequency region.

621.396.621.54 : 621.385.029.6 **2006**

Carcinotron Harmonics boost Receiver Range.—C. H. Currie. (*Electronics*, 27th Feb. 1959, Vol. 32, No. 9, pp. 58–61.) A continuous frequency coverage from 30 Mc/s to 75 kMc/s is obtained by using harmonic mixing. The carcinotron local oscillator operates in the band 2–4 kMc/s and supplies two separate r.f. sections.

STATIONS AND COMMUNICATION SYSTEMS

621.376.018.78 **2007**

The Amplitude and Frequency of a Modulated Carrier Wave.—A. Ditl. (*Hochfrequenztech. u. Elektroakust.*, March 1958, Vol. 66, No. 5, pp. 160–167.) Both a.m. and f.m. systems are considered. Signal distortion due to linear distortion of the carrier, the distortion in s.s.b. systems, the effect of short pulses on the output signal, the interference between f.m. carriers, and the effect of reflection in f.m. systems are investigated with examples.

621.391 **2008**

Discrimination between Several Signals in Telecommunication.—P. Béthoux. (*C. R. Acad. Sci., Paris*, 28th July 1958, Vol. 247, No. 4, pp. 412–415.) Mathematical treatment of signal discrimination in noise.

621.391 : 621.376.56 **2009**

Signal/Noise Ratio in Pulse-Code Modulation Systems: Use of the 'Ideal Observer' Criterion.—J. W. R. Griffiths. (*J. Brit. Instn Radio Engrs*, March 1959, Vol. 19, No. 3, pp. 183–186.) The 'ideal observer' criterion is applied to determining the probability of error in selecting a single pulse in a background of noise. The results are similar to those obtained by a method due to Flood (3970 of 1958).

621.391.1 : 621.396.14 **2010**

Theoretical Considerations about the Merits of the Normal Binary Telegraph Code and the So-Called Gaussian Code, and Special Methods of Detection for Both Codes.—K. Posthumus. (*Tijdschr. ned. Radiogenoot.*, March 1958, Vol. 23, No. 2, pp. 55–82.)

621.396.2 : 621.394.4 **2011**

Multichannel V.F. Telegraph Systems for H.F. Networks.—J. V. Beard. (*Point to Point Telecommun.*, Oct. 1958, Vol. 3, No. 1, pp. 29–48.) A general description of a two-tone frequency-diversity system which gives under conditions of selective fading, a tenfold reduction in error rate compared with a comparable f.m. system using the same bandwidth.

621.396.43 : 550.389.2 : 629.19 **2012**

Transoceanic Communication by means of Satellites.—J. R. Pierce & R. Kompfner. (*Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, pp. 372–380.) "A satellite in a polar orbit at a height of 3 000 miles would be mutually visible from Newfoundland and the Hebrides for 22·0 per cent of the time and would be over 7·25° above the horizon at each point for 17·7 per cent of the time. Out of 24 such satellites, some would be mutually visible over 7·25° above the horizon 99 per cent of the time. With 100-foot-diameter spheres, 150-foot-diameter antennas, and a noise temperature of 20°K, 85 kW at 2 000 Mc/s or 9·5 kW at 6 000 Mc/s could provide a 5-Mc/s base band with a 40-dB signal-to-noise ratio."

621.396.65 **2013**

Radio-Link Equipment for 60–120 Channels.—G. Strocchi. (*Alta Frequenza*, June/Aug. 1958, Vol. 27, Nos. 3/4, pp. 269–291.) Three types of Italian equipment are described, including that used in the radio link Milan–Palermo [2230 of 1958 (Peroni)].

621.396.65 **2014**

Convention on Radio Links.—(*Alta Frequenza*, June/Aug. 1958, Vol. 27, Nos.

3/4, pp. 177–432.) Second issue covering the proceedings of a convention held in Rome 5th–8th June 1957. First issue: 2229 of 1958. Abstracts of some of the papers are given individually; titles of others are as follows:

(*a*) The Transappennine Radio Link.—G. Monti-Guarnieri (pp. 179–218).

(*b*) Realization and Future Application of Pulse Techniques in Radio Communication Networks.—R. Cabessa (pp. 226–235, in French).

(*c*) Diversity Systems and their Influence on the Economic Operation of Radio Links.—P. Clavier (pp. 236–244, in French).

(*d*) Some Design Problems in F.M. Broad-Band Microwave Systems.—B. Håård (pp. 245–262, in English).

(*e*) The Evaluation of Transmission Quality in Multichannel F.M. Radiotelephony Links.—I. Medici (pp. 347–362).

(*f*) Field-Strength Recordings and Performance of Very-Short-Wave Radio Links.—J. A. Smale (pp. 363–377, in English).

621.396.71.029.55 **2015**

The Olifantsfontein and Derdepoort Radio Stations of the Department of Posts and Telegraphs.—A. Birrell. (*Trans. S. Afr. Inst. elect. Engrs*, June 1958, Vol. 49, Part 6, pp. 177–228. Discussion, pp. 228 231.) A detailed description of the Olifantsfontein transmitting and Derdepoort receiving stations and of their part in the radio telephone and telegraph services of the Union of South Africa is given. A v.h.f. radio link operating at 100 Mc/s for line-of-sight transmission of standard-frequency and time signals from the Union Observatory to Olifantsfontein is under construction; the signals are to be broadcast on 5, 10, 15 and 20 Mc/s.

621.396.712.3 **2016**

Broadcasting Equipment at the New Studio in Karlsruhe.—W. Hoffmann. (*Rundfunktech. Mitt.*, June 1958, Vol. 2, No. 3, pp. 100–105.) Details are given of the control room, recording and distribution installations.

621.396.712.3 : 534.861 **2017**

The Acoustic Design of the New Studio in Karlsruhe.—Keidel. (See 1756.)

621.396.73 **2018**

The Megacoder—a High-Speed, Large-Capacity Microminiature Decoder for Selective Communication.—H. Kihn & W. E. Barnette. (*RCA Rev.*, March 1959, Vol. 20, No. 1, pp. 153–179.) The device can be preset to respond to any one of a million possible code combinations for use in f.m. personal paging systems.

SUBSIDIARY APPARATUS

621.3.087.9 : 621.318.4 **2019**

I.R.E. Standards on Static Magnetic Storage: Definitions of Terms, 1959.—(*Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, pp. 427–430.) Standard 59 I.R.E. 8.S1.

621.3.087.9 : 621.395.625.3 **2020**

Magnetic Head reads Tape at Zero Speed.—M. E. Anderson. (*Electronics*, 6th March 1959, Vol. 32, No. 10, pp. 58–60.) Design details of a system enabling recorded h.f. signals to be played back at speeds low enough for the output to be fed to a pen-recorder without deterioration in signal/noise ratio.

621.314.58 : 621.314.7 **2021**

Development of the Transistor Inverter at 20 kc/s using Power Transistors.—W. A. Martin. (*Trans. Inst. Radio Engrs*, June 1957, Vol. I-6, No. 2, pp. 118–122).

621.314.63 : 621.39 **2022**

Physical and Electrical Properties of Silicon Rectifiers for Communications Applications.—H. L. Rath. (*Elektronische Rundschau*, April 1958, Vol. 12, No. 4, pp. 119–122.) Small junction-type diodes with a current-carrying capacity of up to 1 A are considered.

621.352 : 541.135.6 **2023**

Current Integration with Solion Liquid Diodes.—R. N. Lane & D. B. Cameron. (*Electronics*, 27th Feb. 1959, Vol. 32, No. 9, pp. 53–55.) The construction and characteristics of electrochemical diodes using iodine/potassium-iodide solution are described (see 1563 of 1958). These can be used as electrical current integrators and flow or pressure detectors.

TELEVISION AND PHOTOTELEGRAPHY

621.397.611 : 778.5 **2024**

The Film Recording of Television Transmissions in the German Federal Republic.—J. Goldmann & H. Funk. (*Rundfunktech. Mitt.*, June 1958, Vol. 2, No. 3, pp. 129–136.) Review of methods used and description of installations.

621.397.611.2 **2025**

A Vidicon Camera for Industrial Colour Television.—I. L. P. James. (*J. Brit. Instn Radio Engrs*, March 1959, Vol. 19, No. 3, pp. 165–180. Discussion, pp. 181–182.) The main features of a simultaneous colour system employing three vidicons are described and signal amplifiers and line and field scanning circuits are discussed. The picture quality obtained is adequate for 625-line broadcast standards and the equipment is universally applicable for general industrial use.

621.397.611.2 **2026**

Contribution on the Problem of Portable Television Cameras for Outside Broadcasts.—H. Fix. (*Rundfunktech. Mitt.*, June 1958, Vol. 2, No. 3, pp. 120–128.) The design and application of portable television cameras are discussed and three cameras, including the French Type CP103 [see 3650 of 1958 (Polonsky)], are compared.

621.397.62 **2027**

Inexpensive Sound for Television Receivers.—R. B. Dome. (*Electronics*, 27th Feb. 1959, Vol. 32, No. 9, pp. 66–68.) The system provides a.m. compression, a high-level a.f. output, f.m. detection and cancellation of the a.m. fundamental frequency.

621.397.62 : 621.376.33 : 621.314.7 **2028**

TV Sound Detector uses Drift Transistor.—M. Meth. (*Electronics*, 20th Feb. 1959, Vol. 32, No. 8, pp. 62–64.) Circuit details of a sensitive, oscillating linear-slope detector giving improved performance at low signal levels compared with a passive detector.

621.397.621 : 535.623 **2029**

Results with an Experimental Colour Television System using Controlled Colour Filters.—V. A. Babits. (*Brit. Commun. Electronics*, Jan. 1959, Vol. 6, No. 1, p. 15.) In a brief report, reference is made to problems to be overcome. For an account of similar work see 621 of February (Wells).

621.397.7 (71/73) **2030**

Television Station List.—M. I. Schiller. (*Radio-Electronics*, Jan. 1959, Vol. 30, No. 1, pp. 106–107.) A list of U.S., Canadian and Mexican stations correct to 1st December 1958 giving call sign, location and channel number.

621.397.8 **2031**

A Method for the Measurement of Random Fluctuations in Television.—D. Waechter. (*Rundfunktech. Mitt.*, June 1958, Vol. 2, No. 3, pp. 117–119.) In the comparison method described adjustable random noise is superimposed on a small area of the picture under test.

621.397.8.083 : 535.623 **2032**

A Simple Method of Mixing a Colour Subcarrier of Variable Frequency with a Black and White Picture.—G. Bolle. (*Frequenz*, April 1958, Vol. 12, No. 4, pp. 103–108.) Test equipment is described for investigating the interference caused by the colour subcarrier of the N.T.S.C. system in monochrome picture reproduction. A constant-amplitude carrier is continuously variable in the frequency ranges 1·5–3·0 Mc/s and 3·0–4·5 Mc/s.

621.397.9 : 629.136.3 **2033**

Single-Line-Scan Television.—F. H. Harris & J. Ainsworth. (*Rev. sci. Instrum.*, Feb. 1959, Vol. 30, No. 2, pp. 76–78.) Description of equipment constructed to show the practicability of using a vidicon camera tube for measuring the space orientation of spin-stabilized rockets.

TRANSMISSION

621.376.32 **2034**

A Reactance-Valve Frequency Modulator.—F. Carassa. (*Alta Frequenza*, June/Aug. 1958, Vol. 27, Nos. 3/4, pp. 292–

303.) A portable modulator unit used for feeding television outside-broadcast programs into a radio-link network is described.

VALVES AND THERMIONICS

621.314.63 2035

The Problem of Representation of a Semiconductor Diode in the Form of a Series Connection of Two Nonlinear Inertia Elements and the Applicability of the Pulse Method of Voltage Division.—Yu. K. Barsukov. (*Zh. tekh. Fiz.*, Oct. 1957, Vol. 27, No. 10, pp. 2262–2267.) The relation of the overall *I/V* characteristic of a Ge diode to the *I/V* characteristic of the *p-n* junction and the volume of Ge is considered, and the applicability of a pulse method for determining the division of voltage in the circuit representation of the diode is discussed.

621.314.63 2036

The Inductive Behaviour of *p-n* Rectifiers under High Forward-Current Loads.—E. Spenke. (*Z. angew. Phys.*, Feb. 1958, Vol. 10, No. 2, pp. 65–88.) A simplified model of a *p-n* junction and its equivalent circuit are used as a basis for detailed calculations of the a.c. characteristics, to account for the inductive component of the rectifier impedance.

621.314.63 2037

Semiconductor–Semiconductor 'Point-Contact' Diode.—A. Levitas & I. Ladany. (*J. appl. Phys.*, Feb. 1959, Vol. 30, No. 2, pp. 267–268.) Description of a technique developed for the fabrication of a junction device made from a single-crystal Ge bar containing a grown *p-n* junction, with an external appearance similar to a point-contact diode.

621.314.63.012.6 : 621.317.6 2038

Evaluating Logarithmic Diodes.—A. Gill. (*Electronics*, 6th March 1959, Vol. 32, No. 10, pp. 64 . . 67.) Note on a method of determining the low-level characteristic of a semiconductor diode, using a sawtooth input voltage and obtaining a c.r.o. trace showing *di/dt* as a function of current *i*.

621.314.63 : 621.316.722.1 2039

Characteristics of Silicon Junction Diodes as Precision Voltage-Reference Devices.—K. Enslein. (*Trans. Inst. Radio Engrs*, June 1957, Vol. I-6, No. 2, pp. 105–118.)

621.314.63 : 621.318.57 2040

Two-Terminal Solid-State Switches.—T. P. Sylvan. (*Electronics*, 27th Feb. 1959, Vol. 32, No. 9, pp. 62–63.) Characteristics of commercial *p-n-p-n* and *p-n-p-m* semiconductor diodes are tabulated. See e.g. *Trans. Inst. Radio Engrs*, Jan. 1958, Vol. ED-5, No. 1, pp. 13–18 (Philips & Chang).

621.314.7 + 621.385.3 2041

Simple General Analysis of Amplifier Devices with Emitter, Control, and Collector Functions.—E. O. Johnson & A. Rose. (*Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, pp. 407–418.) The photoconductor, unipolar and bipolar transistors, vacuum triode, analogue transistor and beam-deflection tube are considered. The characteristics are compared and discussed for particular applications.

621.314.7 2042

Contribution on the Representation of the A.C. Characteristics of the Earthed-Base Transistor.—H. Schneider. (*Nachr Tech.*, March 1958, Vol. 8, No. 3, pp. 126–129.)

621.314.7 2043

On the Lifetime and Diffusion Constant of the Injected Carriers and the Emitter Efficiency of a Junction Transistor.—S. Deb & A. N. Daw. (*J. Electronics Control*, Dec. 1958, Vol. 5, No. 6, pp. 514–530.) Experimental data are given for three types of *p-n-p* alloy-junction transistor. Variations of minority-carrier lifetime and diffusion coefficient with emitter current, temperature and carrier injection level are examined in relation to theory.

621.314.7 : 621.3.012.029.63 2044

U.H.F. Transistor Data.—H. Tulchin. (*Electronics*, 6th March 1959, Vol. 32, No. 10, p. 57.) Characteristics are presented for eight commercially available transistors with operating frequencies above 300 Mc/s.

621.314.7 : 621.317.71 2045

Transistor Junction Temperature.—Sutcliffe & Matthews. (See 1974.)

621.314.7 : 621.318.57 2046

Solid-State Thyratrons Available Today.—T. P. Sylvan. (*Electronics*, 6th March 1959, Vol. 32, No. 10, pp. 50–51.) Characteristics of three-terminal switching transistors are tabulated.

621.314.7 : 621.396.822 2047

The Influence of Inductive Source Reactance on the Noise Figure of a Junction Transistor.—E. R. Chenette. (*Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, pp. 448–449.) Theoretical predictions for the equivalent noise resistance agree well with measurements, but the correlation reactance does not, except at low frequencies. Reasons for this are given.

621.314.7.001.4 2048

Complete Linear Characterization of Transistors from Low through Very High Frequencies.—H. G. Follingstad. (*Trans. Inst. Radio Engrs*, March 1957, Vol. I-6, No. 1, pp. 49–63. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 898.)

621.314.7.002.2 2049

Techniques of Transistor Production.—R. N. Wheaton. (*Proc. Instn Radio Engrs, Aust.*, July 1958, Vol. 19, No. 7, pp. 358–369.) Requirements of a transistor suitable for h.f. operation are discussed and fabrication methods are reviewed, in particular alloying and diffusion techniques. Characteristics of different types of transistor are tabulated.

621.314.7.012.8 2050

Unified Representation of Junction-Transistor Transient Response.—A. Harel & J. F. Cashen. (*RCA Rev.*, March 1959, Vol. 20, No. 1, pp. 136–152.) A general mathematical formula is derived which is applicable to any circuit configuration of the transistor.

621.314.7.016.35 2051

Methods of Calculation for the Stabilization of Transistor Circuits at Variable Temperature.—K. Lunze. (*Nachr Tech.*, March 1958, Vol. 8, No. 3, pp. 98–108.)

621.314.7.078 : 621.316.825 2052

Application of Negative-Temperature-Coefficient Resistors to Temperature Stabilization of Transistor Circuits.—C. Wright. (*Proc. Instn Radio Engrs, Aust.*, July 1958, Vol. 19, No. 7, pp. 374–376.)

621.383.032.217.2 2053

Interference Photocathodes of Increased Yield with Freely Variable Maximum Spectral Response.—K. Deutscher. (*Z. Phys.*, 1st July 1958, Vol. 151, No. 5, pp. 536–555.) Full report of an investigation noted earlier (2271 of 1958).

621.383.27 2054

The Linearization of Multipliers at High Anode Currents.—H. J. Kopp & W. Petzold. (*Z. angew. Phys.*, Jan. 1958, Vol. 10. No. 1, pp. 34–36.) A method of obtaining a linear relation in the measurement of light intensity by means of photomultipliers is described.

621.383.5 2055

Bismuth-Tellurium Photovoltaic *p-n* 'Sandwich' Layer.—T. Piwkowski. (*Nature, Lond.*, 27th Dec. 1958, Vol. 182, No. 4652, pp. 1793–1794.) A layer of Te evaporated on to a layer of Bi backed by a glass plate was found to act as a barrier-type photocell.

621.383.5 2056

Influence of Selenium Microstructure on Photocell Characteristics.—T. K. Lakshmanan. (*J. appl. Phys.*, Feb. 1959, Vol. 30, No. 2, pp. 265–266.)

621.385.029.6 2057

The Cut-Off Characteristics of Magnetrons (Static Regime).—W. Fulop. (*J. Electronics Control*, Dec. 1958, Vol. 5, No. 6, pp. 531–548.) Experiments show strong emission dependence and indicate that the anomalous current flow arises from electron interaction. The electron ensemble is shown to be far removed from thermal equilibrium and to approach this condition, though never reaching it, only at extremely high emission currents. Indications are given of theoretical trends needed to account for the experimental results.

621.385.029.6 2058

On Space-Charge Waves.—D. H. Trevena. (*J. Electronics Control*, Jan. 1959, Vol. 6, No. 1, pp. 50–64.) The existence is shown of four sets of space-charge waves in a uniform electron beam focused by an axial magnetic field with arbitrary uniform cathode flux. The magnetic field affects the esults only through the cathode flux. Expressions for the plasma frequency reduction factors are given.

621.385.029.6 **2059**

The Theory of the Formation of Electron Beams.—V. T. Ovcharov. (*Radiotekhnika i Elektronika*, June 1957, Vol. 2, No. 6, pp. 696–704.) A method is described for calculating the electric field inside an electron beam with prescribed trajectories and magnetic field taking into account the charge of the beam. By an appropriate choice of orthogonal curvilinear coordinates for the trajectory an ordinary second-order differential equation is obtained.

621.385.029.6 **2060**

The Effect of the Inclination of the Focusing Electrodes on Electron-Beam Formation.—R. J. Lomax. (*J. Electronics Control*, Jan. 1959, Vol. 6, No. 1, pp. 39–49.) It is shown that variations of the angle at which Pierce electrodes meet the cathode give rise to variations in the angle at which the beam emerges, and that the current density at the periphery of the cathode depends markedly on this angle.

621.385.029.6 **2061**

Periodic Electrostatic Focusing of Laminar Parallel-Flow Electron Beams. —W. W. Siekanowicz & F. E. Vaccaro. (*Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, pp. 451–452.)

621.385.029.6 **2062**

Electron Waves in Retarding Systems. Nonlinear Equations for Travelling-Wave Valves.—L. A. Vaĭnshteĭn. (*Radiotekhnika i Elektronika*, June 1957, Vol. 2, No. 6, pp. 688–695.) Generalization of the results of an earlier analysis (338 of 1957) with application to the nonlinear theory of travelling-wave valves.

621.385.029.6 **2063**

The Engineering of Low-Noise Travelling-Wave Tubes.—F. J. Bryant, R. B. Coulson & J. K. Fowler. (*Brit. Commun. Electronics*, Jan. 1959, Vol. 6, No. 1, pp. 20–25.) Design features and operating characteristics of C-band and L-band amplifiers are given.

621.385.029.6 **2064**

The Exponential Gun—a Low-Noise Gun for Travelling-Wave-Tube Amplifiers.—A. L. Eichenbaum & R. W. Peter. (*RCA Rev.*, March 1959, Vol. 20, No. 1, pp. 18–56.) The gun section between the cathode region and the circuit input is treated as a transmission-line matching transformer. The exponential gun is shown to be the best solution for the gun requirements of a low-noise travelling-wave valve.

621.385.029.6 **2065**
: [621.375.9 + 621.372.632

A Three-Frequency Electron-Beam Parametric Amplifier and Frequency Converter.—W. H. Louisell. (*J. Electronics Control*, Jan. 1959, Vol. 6, No. 1, pp. 1–25.) Analysis of Louisell & Quate (2273 of 1958) is generalized for the case of an electron beam in which the pump frequency ω need not be twice the signal frequency ω_1. If $\omega_1 \neq \omega_2$, where ω_2 is the idler frequency generated in the beam, growing and decaying fast space-charge waves can be excited equally, independent of the phase of the pump relative to the signal. The threshold of modulation needed to produce gain increases with beam size. If $\omega = \omega_1 - \omega_2$, the device acts as a frequency converter.

621.385.029.6 : 621.375.9 **2066**

Gain, Bandwidth and Noise in a Cavity-Type Parametric Amplifier using an Electron Beam.—G. Wade & H. Heffner. (*J. Electronics Control*, Dec. 1958, Vol. 5, No. 6, pp. 497–509.) A modulated beam flowing across the gap of a resonant cavity changes the gap capacitance at the modulating frequency. It is shown that complete cancellation simultaneously of the two uncorrelated noise sources in the beam, while feasible in principle, is virtually impossible in practice. Conflicting requirements of large beam current for acceptable capacitance variation and large plasma wavelength for optimum noise cancellation lead to practical minimum noise figures of about 3 dB. Design data are given for an amplifier with pump frequency 2 kMc/s, a gain of about 15 dB at 500 Mc/s and bandwidth 43 kc/s. A noise figure of 3·4 dB could be achieved with some difficulty.

621.385.029.6 : 621.375.9 : 621.372.2 **2067**

Travelling-Wave Couplers for Longitudinal Beam-Type Amplifiers.—R. W. Gould. (*Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, pp. 419–426.) The theory is developed and applied to the design of couplers for parametric amplifiers. Matrices for velocity jumps and drift regions are given, and conditions for the removal of beam noise from the fast space-charge wave are derived.

621.385.032.213 **2068**

Development of Thermionic Cathodes.—B. M. Tsarev. (*Radiotekhnika i Elektronika*, June 1957, Vol. 2, No. 6, pp. 675–687.) The general requirements of thermionic cathodes are discussed in relation to current applications in vacuum tubes. A classification is given of different types of cathode in use and under investigation, and the characteristics of cathode materials are tabulated.

621.385.032.213.13 **2069**

A Study of the Moulded Nickel Cathode.—C. P. Hadley, W. G. Rudy & A. J. Stoeckert. (*J. electrochem. Soc.*, July 1958, Vol. 105, No. 7, pp. 395–398.) "Research work on the moulded nickel cathode is described. Results are given regarding the effects on emission and life of variations in nickel powder, alkaline-earth carbonates, reducing agents, sintering, and aging. Data on pulsed emission are presented."

621.385.032.213.63 **2070**

On the Energy Distribution of Electrons from Antimony-Caesium Cathodes.—A. I. Pyatnitskiĭ. (*Radiotekhnika i Elektronika*, June 1957, Vol. 2, No. 6, pp. 714–725.) Results of measurements of the photocurrent and secondary emission of a Cs-Sb cathode and a Cs-Ag layer show the basic difference between their I/V characteristics, confirming the presence in Cs-Sb of low-energy secondary electrons. The number of these electrons is dependent on the quantity of Cs in the cathode.

621.385.3 : 621.365.5 **2071**

Intermittent Use of Oscillator Valves in R.F. Heating Generators.—E. G. Dorgelo & J. C. van Warmerdam. (*Electronic Applic.*, April 1958, Vol. 18, No. 2, pp. 41–47; *Mullard tech. Commun.*, Dec. 1958, Vol. 4, No. 36, pp. 173–178.)

621.385.3 : 621.365.5 **2072**

Output and Load Resistance of Oscillating Triodes in R.F. Heating Generators.—E. G. Dorgelo. (*Electronic Applic.*, Jan. 1958, Vol. 18, No. 1, pp. 19–26; *Mullard tech. Commun.*, Dec. 1958, Vol. 4, No. 36, pp. 179–185.) A method is described for calculating the output power and other operating conditions of triode oscillators as a function of their load resistance.

621.385.832 : 621.396.662 **2073**

Electron-Beam Voltage-Indicator Tube EM84.—A. Lieb. (*Elect. Commun.*, 1958, Vol. 35, No. 2, pp. 76–82.) A specially designed tuning indicator with a ZnO phosphor forming two fluorescent bands along the tube axis is described, which can be used in conjunction with colour filters or a printed scale to measure voltages to an accuracy within 5–15%.

MISCELLANEOUS

538.569.2.047 **2074**

Health Hazards from Powerful Radio Transmissions.—D. H. Shinn. (*Nature, Lond.*, 27th Dec. 1958, Vol. 182, No. 4652, pp. 1792–1793.) Field strength contours in decibels for a paraboloidal aerial in free space are given, and the danger area is calculated on the basis of the theory of Schwan & Li (537 of 1957).

538.569.2.047 **2075**

Researching Microwave Health Hazards.—F. Leary. (*Electronics*, 20th Feb. 1959, Vol. 32, No. 8, pp. 49–53.) A summary of the effects of high-intensity microwave radiation on the human body.

621.38.004.15 **2076**

Designing for Reliability in Electronic Instrumentation.—R. E. Fischbacher. (*J. Electronics Control*, Nov. 1958, Vol. 5, No. 5, pp. 471–482.) Discussion of the main factors which contribute to the design of reliable electronic instruments.

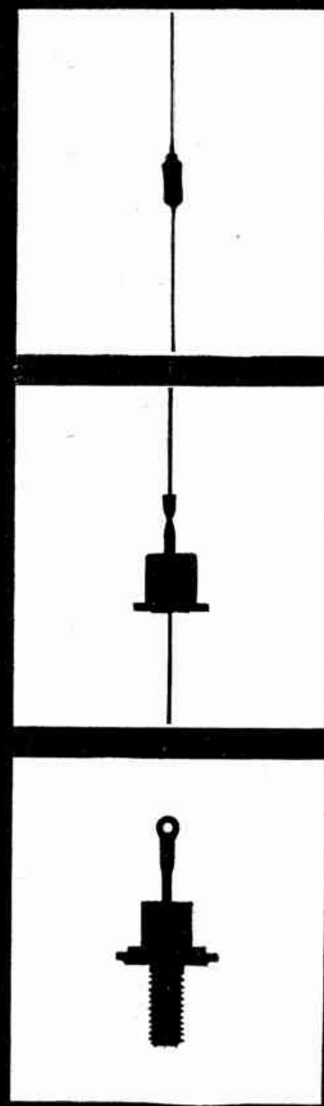
001.891 : 621.396 **2077**

Radio Research 1957: The Report o the Radio Research Board and the Report of the Director of Radio Research. [Book Review]—Publisers: H.M. Stationery Office, London, 1958, 43 pp., 3s. 6d. (*Nature, Lond.*, 6th Dec. 1958, Vol. 182, No. 4649, pp. 1558–1559.)

SILICON RECTIFIERS

up to 600 Volts P.I.V.

3 amps



TEXAS SILICON RECTIFIERS offer excellent high temperature characteristics. For example the stud-mounted type gives an output of 3 amperes at 50°C and will still provide 1 ampere at 150°C with no change in the rated peak inverse voltage. The rugged, welded housing with glass-to-metal seal provides high resistance to shock and vibration. Listed in the table below are the characteristics of representative devices of each type.

Have you received your copy of the latest Texas Application Report "D.C. Power Supply Circuits using Silicon Rectifiers"? If not, or if you require fuller details of Texas Rectifiers, please write your name and address in the margin and return this advertisement to us.

	Symbol	1S001	1S005	1S111	1S115	1S401 1S401R	1S405 1S405R
MAXIMUM RATINGS							
Peak Inverse Voltage at −65°C to +150°C	PIV	200V	600V	200V	600V	200V	600V
Average Rectified Forward Current at +50°C	I_o	750mA	750mA	† 400mA	† 400mA	*3A	*3A
Average Rectified Forward Current at +150°C	I_o	250mA	250mA	150mA	150mA	*1A	*1A
Recurrent Peak Forward Current at +50°C	i_i	† 2·5A	† 2·5A	† 1·25A	† 1·25A	*10A	*10A
Surge Current for 10 Milliseconds	I_{PK}	16A	16A	6A	6A	33A	33A
Operating Temperature, Ambient	T_A	−65°C to +150°C					
SPECIFICATIONS							
Minimum Breakdown Voltage at +150°C	V_z	240V	720V	240V	720V	240V	720V
Maximum Reverse Current at P.I.V. at +25°C	LI_b	10μA	10μA	0·2μA	0·2μA	10μA	10μA
Maximum Forward Voltage Drop at +25°C	E_b	1·0V	1·0V	1·0V	1·0V	1·1V	1·1V
		(I_o=500mA)		(I_o=400mA)		(I_b=1Amp)	

* Rectifier mounted on 2" x 2" x $\frac{1}{16}$" aluminium Heat Sink † @ 25°C

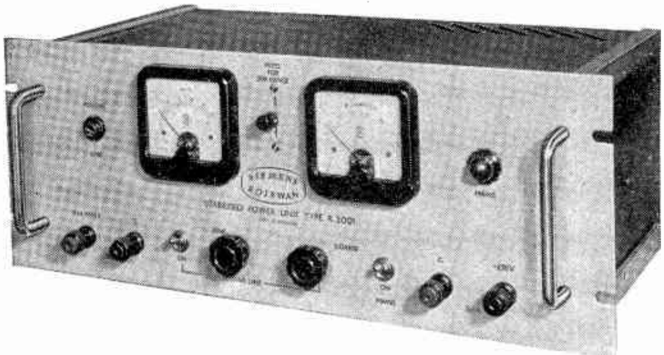
TEXAS INSTRUMENTS LIMITED

TELEPHONE: BEDFORD 68051 DALLAS ROAD BEDFORD CABLES: TEXINLIM BEDFORD

T 31

Stabilised
D.C. Power Supply Units
(Types R.2001, R.2030 and R.2030A)

These stabilised power supply units are examples from a range of Siemens Edison Swan units providing reliable, highly stabilised supplies at relatively low cost.



Type R.2001

Designed to replace H.T. Batteries. Provides a D.C. output continuously variable from 0-100 volts by means of coarse and fine voltage controls. Load 0-50mA.

Types R.2030 and R.2030A

Low voltage supply units designed to replace a 6 volt or 12 volt accumulator. Type R.2030 provides an output adjustable between 0-8 volts. Load 0-4 amps. Type R.2030A gives an output of 0-12 volts. Load 0-3 amps.

SPECIFICATION

UNIT	R.2001	R.2030	R.2030A
D.C. Output	0–100V	0–8V	0–12V
Output Current	50 mA	4 amps	3 amps
Output V change for 10% mains change	0.1V	0.01V	
Output Resistance	2 ohms	0.0025 ohms	
Ripple	1 mV p-p	3 mV p-p	
Correction for 0–4 amp Load change		2–3 milliseconds	
Input Voltage	200–250V 40–100 c.p.s.		

SIEMENS EDISWAN

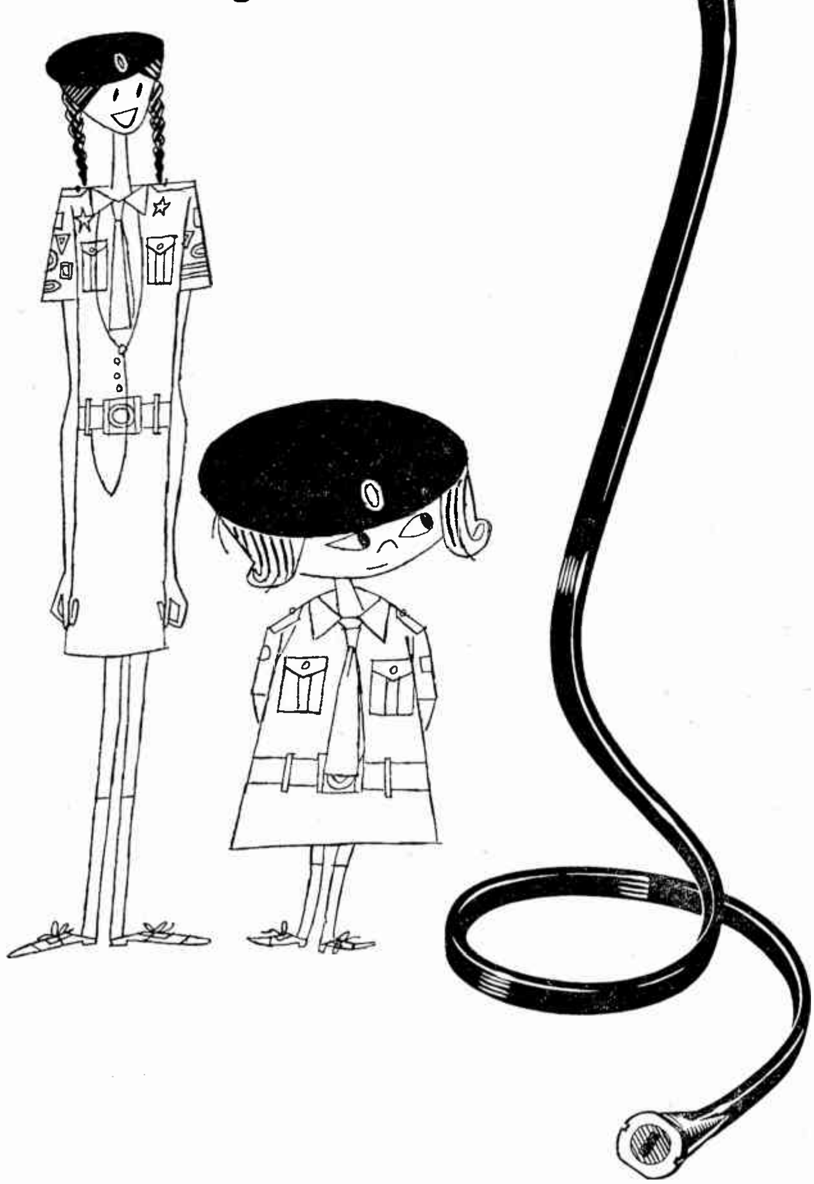
SIEMENS EDISON SWAN LIMITED

AN A.E.I. COMPANY

155 CHARING CROSS ROAD, LONDON, W.C.2 AND BRANCHES
Phone: GERrard 8660. Grams: Sieswan, Westcent, London.

CRC 17/4

The Long and the Short and the 12-feet tall



For Microwaves
Yes—the newest recruit to the WAVEFLEX range of flexible waveguides is a WG 16 guide 12 feet in length, specially produced to meet a customer's requirements.

These guides are produced under exacting scrutiny and tested to rigorous Government specifications. Losses due to attenuation and Standing Wave Ratio are well within the limits set. They are entirely suitable for use in pressurised systems, remain stable when bent or twisted, and are unaffected by extremes of temperature. Power handling characteristics are excellent, and performance covers a broad band. "Type approval" has been given by the Royal Radar Establishment and by the Admiralty.

Waveflex Flexible Waveguides are of three types

Non-twistable
Flexible in the E and H planes.

Twistable
Twistable about the longitudinal axis, as well as being flexible in the E and H planes,
and the new

Premoulded Twisted
Flexible in E and H planes with a built-in longitudinal 'twist' which relieves strain on end flanges.

Standard lengths range from $1\frac{5}{16}$" to 36". All guides can be supplied with any standard type of flange, or special flanges made to your design. The many short (under 6") guides are particularly useful as malalignment units and mechanical decouplers.

WAVEFLEX

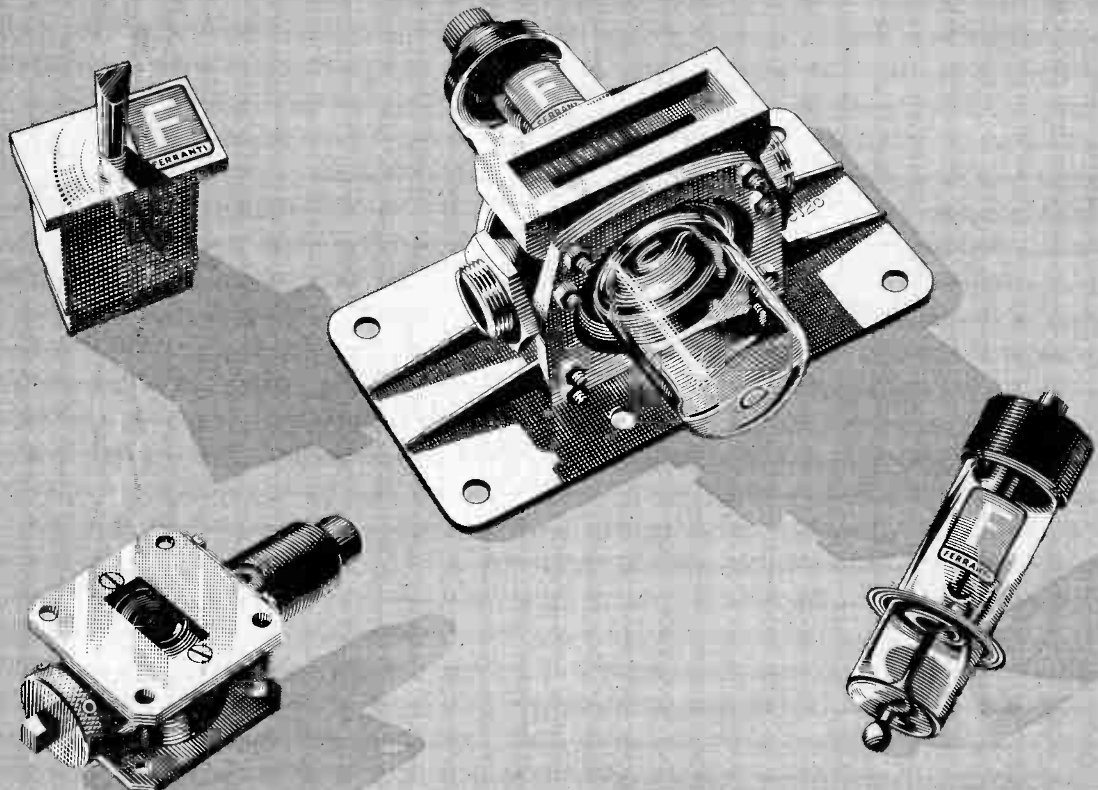
FLEXIBLE AND TWISTABLE WAVEGUIDES

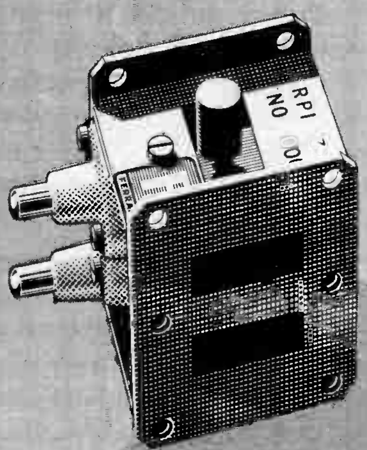
Wherever wave transmission involves directional change or vibration, and movement will be encountered, WAVEFLEX Flexible Waveguides offer the complete solution. Present production embraces WG 10, WG 10a, WG 15 and WG 16 but planned extensions to other bands are already under way. Enquiries are particularly invited for the twistable, the premoulded twisted and the new twistable WG 22 Waveguides. Write now for full technical details.

GABRIEL MANUFACTURING CO. LTD.

Newton Road, Torquay, Devon. *Telephone: Kingskerswell 3333*

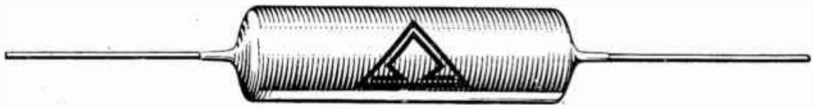
MEMBER OF THE TECALEMIT GROUP OF COMPANIES





RESISTA

high stability carbon layer and wire wound resistors



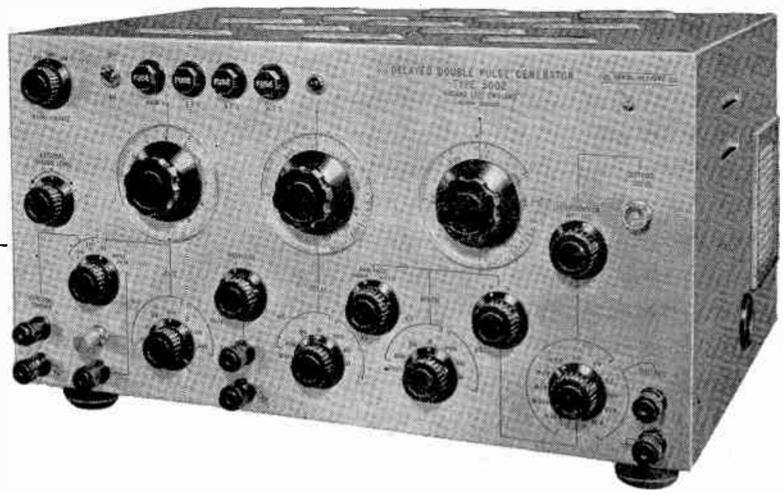
- ★ *High stability to ±0.5% stability and tolerance.*
- ★ *Wire wound to ±0.02% tolerance.*
- ★ *Any non-preferred values.*
- ★ *Very high and low OHMIC values.*
- ★ *Precision and quality at low prices.*

This new comprehensive range of resistors is proving extremely popular with manufacturers of high-quality electrical equipment. They can be used with complete confidence in all types of installation and reflect a considerable saving in production costs. The RESISTA range includes a wide variety of special types including sub-miniatures. They represent the highest precision and quality at the most competitive prices.

G. A. STANLEY PALMER LIMITED
Maxwell House, Arundel Street, London, W.C.2
Telephone: Temple Bar 3721

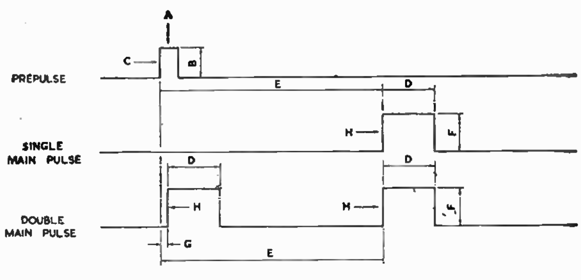
10,000,000 to 1

continuously variable RANGE of Square Pulse Width, Delay and Rate



- **SINGLE OR DOUBLE PULSES of either polarity.**
- **FAST RISE TIME—10 to 12 mμsec. independent of control settings over a range of 20mV. to 2V.**
- **AMPLITUDE—Accurately calibrated within 2%—20mV to 50V.**
- **NEGLIGIBLE RINGING, OVERSHOOT, JITTER and NO SAG.**
- **INTERNAL RATE GENERATOR 0·1c/s to 1Mc/s.**
- **EXTERNAL TRIGGERING up to 2·5Mc/s from any wave form.**
- **SINGLE PULSE or PAIR OF PULSES by push button.**

ALL THESE FEATURES make the NAGARD 5002 the most versatile PULSE GENERATOR available today



A. 0·2μSEC.
B. 20VOLTS +ve OR −ve
C. RISETIME 0·035μSEC
D., E. 0·2μSEC. TO 2SEC. CONTINUOUSLY VARIABLE ±5% ±0·02μSEC.
F. 20mV TO 50V ±2% IN 11 STEPS VARIABLE +ve OR −ve
G. APPROX. 0·04μSEC
H. RISETIME 10–12mμSEC. 20mV TO 2V. 5–50V DEPENDENT ON LOAD

REPETITION RATE CONTINUOUSLY VARIABLE 0·1c/s TO 1Mc/s ±5% BY INTERNAL GENERATOR, OR TO 2·5Mc/s BY EXTERNAL TRIGGER SOURCE.

The diagram on the left represents its output

You can now visualise some of its many applications, for example:—

TRANSISTOR CIRCUITRY
The testing of response characteristics, particularly in computer applications.

COMPUTERS
And all devices depending on gating circuits can be tested for response times.

TRANSMISSION
Characteristics of lines, networks, delays, etc.

WIDE RANGE ACCURATE TRIGGER DELAY
For C.R.O.s without this feature

SERVICING
Oscilloscopes and fault-tracing in a variety of electronic devices.

To get the best results use it with a NAGARD OSCILLOSCOPE

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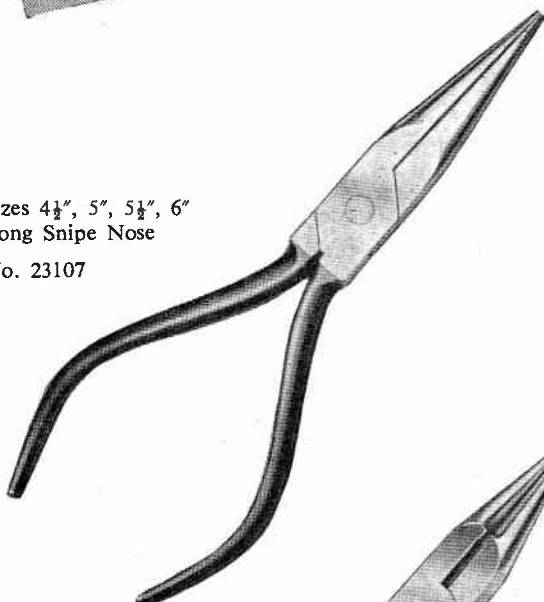
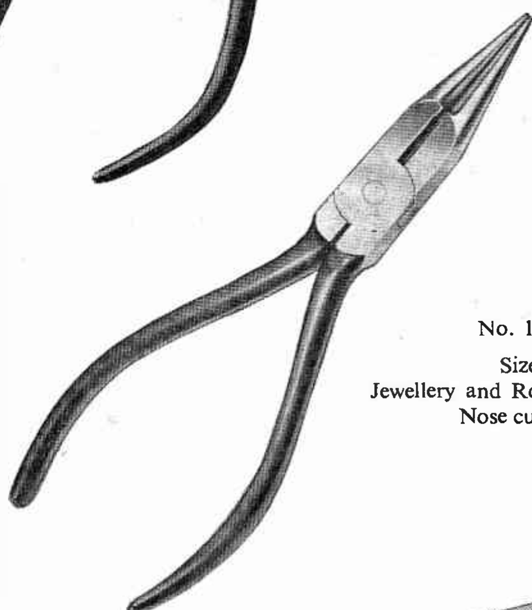
Wilkinsons Special Purpose Tools for the Radio, Electrical and Electronic Engineer



Sizes 4½", 5", 5½", 6"
Long Snipe Nose
No. 23107



No. 11064
Size 5½"
Jewellery and Rosary
Nose cutting

Wilkinsons Tools Ltd.
(A member of the Staveley Coal & Iron Co. Ltd. Group)
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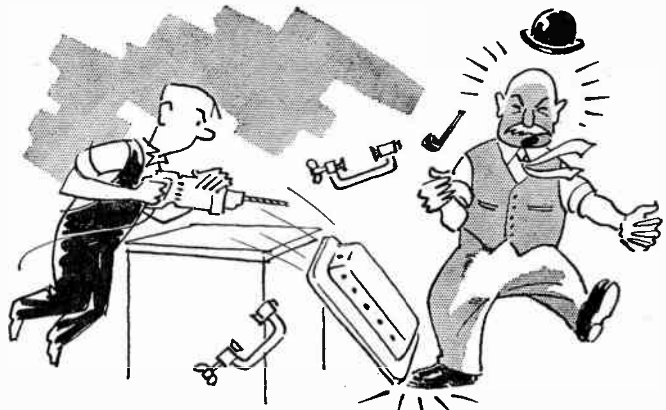
X66

Write today for our Special Purpose Tools Catalogue



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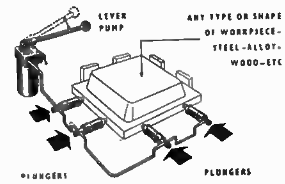
NEWTON TRADE PRODUCT MARK
Hydraulic
CLAMP
SYSTEM

Cut work clamping costs with the Newton Hydraulic Clamp System. Plungers can be grouped in numbers to suit any particular job. Holding pressure of from 800 lbs. to 4 tons. Maximum pressure with smallest plungers. Easily adaptable to any fixture design.



Write for fully illustrated technical booklet. 243/J2.

POWER JACKS LIMITED
VALETTA RD., ACTON, LONDON, W.3
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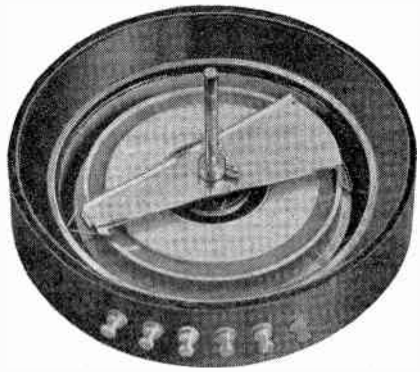

AK-pot (up to 250 kΩ)
with top-plate removed showing

BUILT-IN CLUTCH-ACTUATOR

Wiper brushes on track only when energised. Illustration shows size 31 Radar-type. Write for details and data on other pots, servo-motors and electromagnetic clutches.

KYNMORE ENGINEERING COMPANY LTD.
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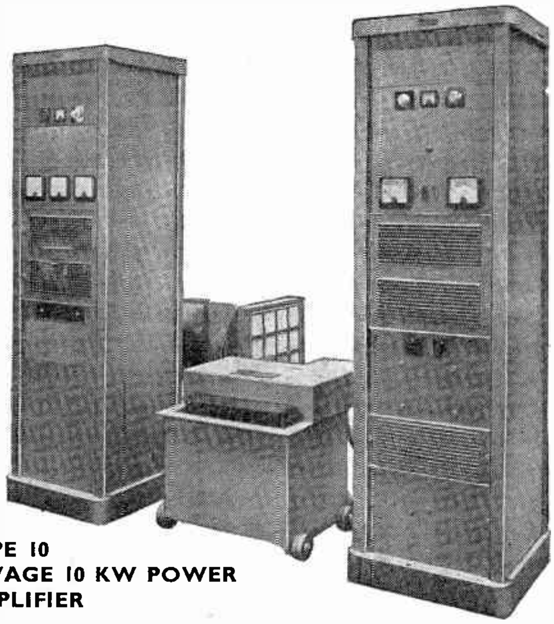
Telephone: TRAfalgar 2371 (3 lines)

Variable high power Amplifiers



TYPE 10
SAVAGE 10 KW POWER AMPLIFIER

The Savage 10 KW Amplifier is designed to meet the high power drive requirements of large vibrators. It has an output of 10 KW (continuous sine wave rating) over the frequency band 40 c/s. to 10 Kc/s. The output has eight secondary sections of 41 1/4 V. each which may be cross connected to give a range of output voltages from 41 1/4 V. to 330 V. Accessibility of components and ease of installation are features of this very compact 10 KW amplifier.

SAVAGE AMPLIFIERS
are suitable for driving 60 cycle American equipment; 400-2500 cycles for aircraft equipment; ultrasonic power supplies for cleaning, drilling, etc.

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W. BRYAN SAVAGE LTD

designers and manufacturers of amplifiers and vibrators for modern industry

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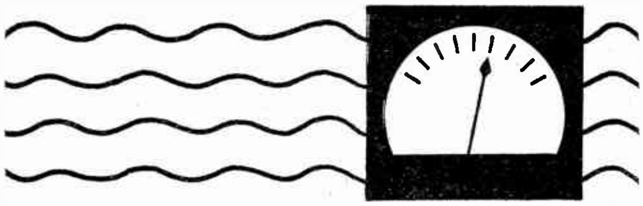
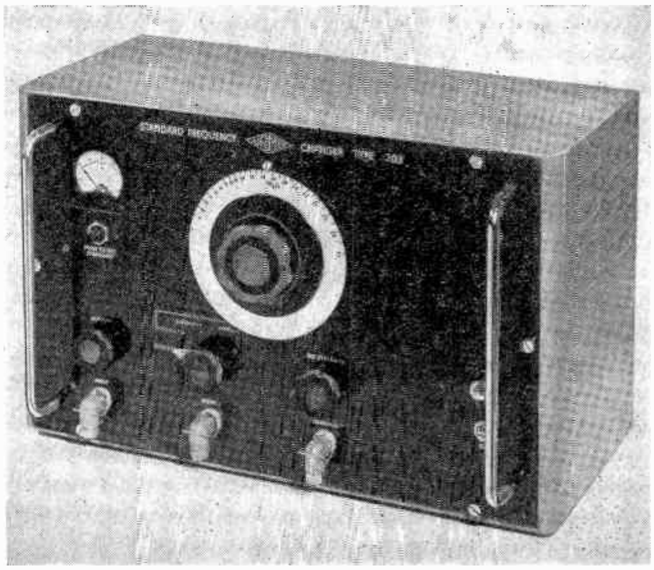


Converting & Measuring Frequency



This AIRMEC instrument extends the range of your Frequency Measuring Equipment . . .

The Airmec Standard Frequency Changer Type 203 accepts unmodulated signals within the range 30 c/s to 31 Mc/s. Signals below 1.1 Mc/s are passed through the instrument with some amplification, whilst signals between 1.1 and 31 Mc/s are converted to a frequency between 30 c/s and 1.1 Mc/s for operating an electronic counter. The frequency of the output signal is entirely dependent upon the frequency of the input signal and the 1 Mc/s standard signal. No frequency errors are introduced in the Frequency Changer.

SPECIFICATION

Frequency Ranges:	30 c/s—31 Mc/s
Input Signal Amplitude:	150 mV to 3V RMS
Input Impedance:	1000 ohms in parallel with 40 pF
Output Signal Amplitude:	1.5 to 2.5V RMS
Output Impedance:	200 ohms
Standard 1 Mc/s Input:	300 mV to 5V RMS required.

IMMEDIATE DELIVERY

Airmec FREQUENCY MEASUREMENT

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Telephone: High Wycombe 2060

ELECTRONIC & RADIO ENGINEER

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Press date for the July 1959 issue is first post 23rd June 1959

SITUATIONS VACANT

ATOMIC ENERGY ESTABLISHMENT, WINFRITH

EXPERIMENTAL staff are required in the Techniques Laboratory of the High Temperature Reactor Division, with experience in one or more of the following:
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BP has a vacancy for a Technical Assistant in the Exploration Division of its Research Centre at Sunbury-on-Thames. There will be 3-4 months of field work per year in the United Kingdom. Age under 27. Several years' experience is required in the building, testing and use of electronic equipment in the audio and sub-audio (low frequency) range. Desirable, but not essential, qualifications are H.N.C. in Light Current Engineering or City & Guilds Telecommunications. Applicants must have no National Service obligations. Salary according to age and experience. Non-contributory Pension Scheme. Removal and settling-in allowances payable in certain cases. Luncheon Club. Write, giving full details, quoting reference H.4645, to Box 6065, c/o 191 Gresham House, E.C.2. [1329

BP has a vacancy for an Electronic Engineer for work in the Instrument Group at its Research Station at Sunbury-on-Thames. Applications are invited from men with a University Degree in Physics or Electrical Engineering or equivalent. Some experience in the design and development of electronic instruments preferable. Work will include development of instruments for the control and testing of internal combustion engines, for process control, and for chemical and physical measurements. Age under 35. Salary according to age, qualifications and experience. Non-contributory Pension Fund. Assisted House Purchase Scheme. Removal expenses and settling-in allowance payable in certain cases. Luncheon Club. Write, giving full details, quoting H.3303A, to Box 6055, c/o 191 Gresham House, E.C.2. [1327

CHIEF Technical Officer required by Posts and Telegraphs Department, Nigeria Federal Government, on contract for tour of 12/24 months in first instance. Salary scale (including Inducement Addition) £1,350 rising to £1,674 a year. Outfit allowance £60. Gratuity at rate of £150 a year. Free passages for officer and wife. Assistance towards children's passages and grant up to £150 annually towards maintenance in U.K. Liberal leave on full salary. Established Civil Servant's Pension rights may be preserved. Candidates should have about five years' practical experience in radio systems including V.H.F. and V.F.T. multiplex systems. Possession of appropriate C. and G. Certificate an advantage. Write to the Crown Agents, 4 Millbank, London, S.W.1. State age, name in block letters, full qualifications and experience and quote M2C/42585/EO. [1330

SITUATIONS VACANT

ELECTRO-ENCEPHALOGRAPHY RECORDIST, GRADE I

APPLICATIONS are invited from experienced technicians. The department also serves other hospitals in the district and a new research unit for neuro-surgery which opens shortly. Whitley Council salary scales and conditions apply. Apply, giving qualifications, experience, and quote two referees to Medical Superintendent, Parkside Hospital, Macclesfield. [1242

TENDERS

GHANA BROADCASTING SYSTEM
External Broadcasting Project

TENDERS are invited by the Crown Agents for Overseas Governments and Administrations acting for and on behalf of the Ghana Government for the supply of four 100 kW Short Wave Broadcasting Transmitters, Auxiliary Equipment and Aerial Arrays together with the necessary buildings to accommodate the equipment for installation near Accra. Tenderers must be prepared to supply staff for installation, operation and maintenance of equipment and training of local staff.
COPIES of the specifications, drawings, form of tender and conditions of contract, may be obtained from the Crown Agents, 4 Millbank, London, S.W.1, by application quoting the reference Contract EE2 Ghana 1815 on payment of a deposit of £10. The deposit will be returned on receipt of a bona fide tender.
TENDER documents will be available for issue as from the 29 May, 1959, and completed tenders must be delivered to the Crown Agents' office not later than 1 p.m. on Monday, 31 August, 1959.
THE Ghana Government does not undertake to accept the lowest or any tender nor be responsible for, or pay for, expenses or losses which may be incurred by any tenderer in the preparation of his tender. [1326

RADAR MATERIALS

An expanding research establishment situated in pleasant surroundings in South Northants requires:

Physicists **Chemists**
Research Engineers **Technicians**

to join a group working on the development and assessment of special materials for both transmission and absorption of microwave radiation.

Investigations are being carried out involving the development of silicone, polyester, epoxy and other resins, natural and synthetic elastomers, and ferrites and ceramics with unique magnetic and dielectric properties over the frequency range 500—50,000 Mc/s.

Candidates are required for theoretical studies on the propagation of electromagnetic waves through solid media, and on the physical performance of suitable materials; for the development of electronic measuring techniques; and for investigations on the processing technology of materials to achieve the required properties.

Recent and prospective graduates, graduates with industrial or university-post-graduate experience, and versatile technicians are needed. Experience in this particular field is not, however, essential. Starting salaries are based on qualifications and experience, and prospects for advancement lie either in research and development, or in supervising the introduction of projects into pilot and full-scale production.

Apply to Box No. 3565, c/o "Electronic & Radio Engineer," quoting as reference RADAR/12.

TUITION

THE COLLEGE OF AERONAUTICS
One Year Post-Graduate Course in
Control Engineering 1959-60

APPLICATIONS are invited for the one year course in Control Engineering which begins on 5 October, 1959. Prospective students must be of graduate status in an appropriate branch of science or engineering and, preferably, have had practical experience in industry or an establishment.
THE purpose of the course is to give instruction in the techniques employed in control engineering as applied in a wide range of industries. Students wishing to do so may specialise in the applications of control engineering to aeronautics.
FURTHER information and forms of application may be obtained from The Warden, The College of Aeronautics, Cranfield, Bletchley, Bucks. [1323

BOROUGH POLYTECHNIC
BOROUGH ROAD, LONDON, S.E.1
TRANSISTOR SUMMER SCHOOL

A full-time day course of lectures on fundamentals of transistor theory and applications will be held from 13th-17th July, 1959. In addition to the basic lectures, there will be reviews of the general field of devices and applications by specialists from industry.
THE course is intended to give engineers with some knowledge of electronic measurements and circuits an introduction to transistors and allied devices.
FEE for the course £10.
FURTHER particulars and forms of application from the Secretary. [1331

PATENTS

PATENT No. 797599 "Valve pin connector" for sale or licence. Apply: Chatwin & Company, Chartered Patent Agents, 253 Gray's Inn Road, London, W.C.1. [1332

PATENT No. 773720 "Broad band tuned circuit amplifier" for sale or licence. Apply: Chatwin & Company, Chartered Patent Agents, 253 Gray's Inn Road, London, W.C.1. [1333

PATENT No. 797136 "Grounded grid amplifier filament lead-in and impedance matching structure" for sale or licence. Apply: Chatwin & Company, Chartered Patent Agents, 253 Gray's Inn Road, London, W.C.1. [1334

PATENT No. 726871 for "Improvements in Electric Time-Switches for Signalling Circuits." Owners desire to meet all demands for the utilisation of this Patent and invite inquiries from manufacturers and others in Great Britain prepared to assist in its commercial exploitation. Address Pollak, Mercer & Tench, Audrey House, Ely Place, London, E.C.1. [1321

PATENT No. 732604 for "Improvements in or relating to Goniometers and like devices." Owners desire to meet all demands for the utilisation of this Patent and invite inquiries from manufacturers and others in Great Britain prepared to assist in its commercial exploitation. Address Pollak, Mercer & Tench, Audrey House, Ely Place, London, E.C.1. [1322

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REPRESENTATIVES wanted for correspondence abroad, high earnings, written instructions. Write Fortuna Publishing Co. Vienna 1/8, Postfach 49. [1324

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