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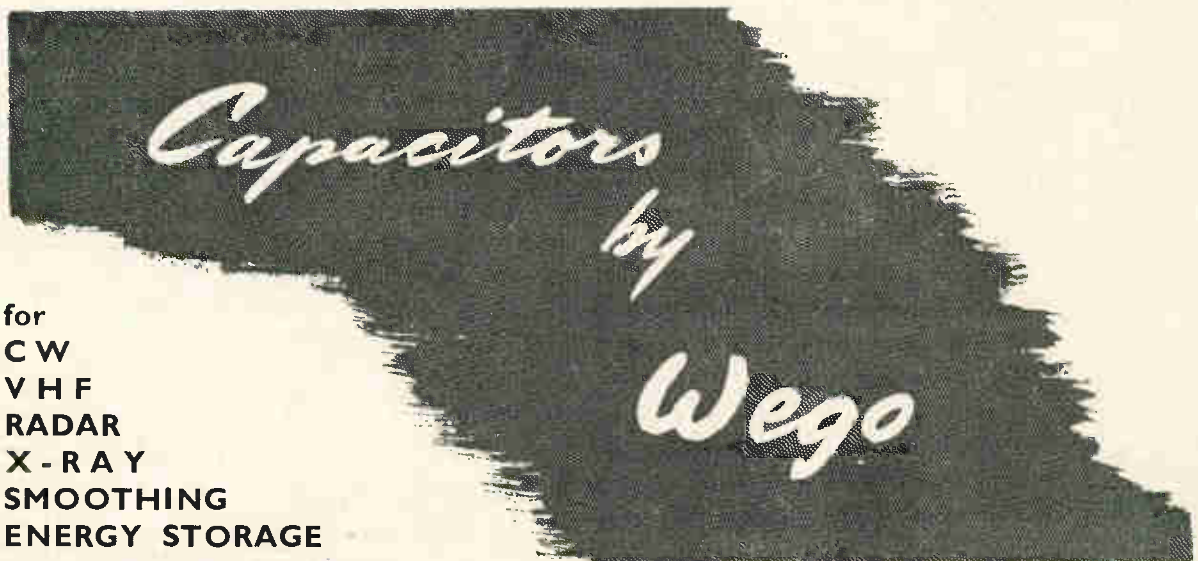

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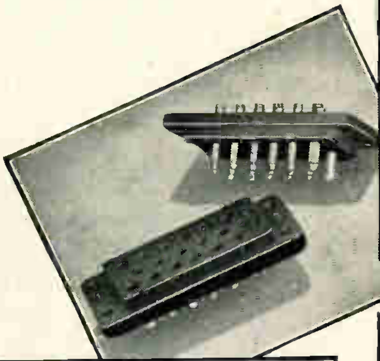

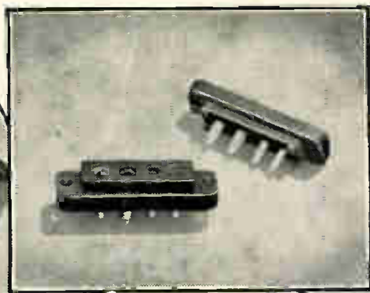

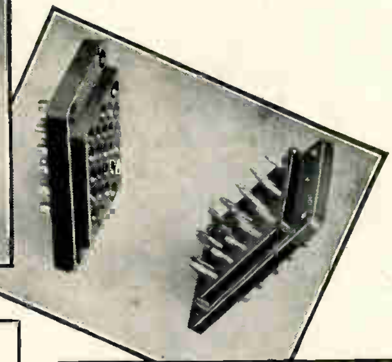

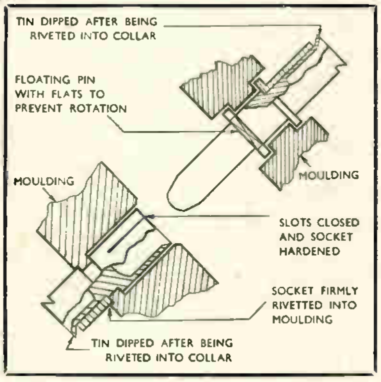

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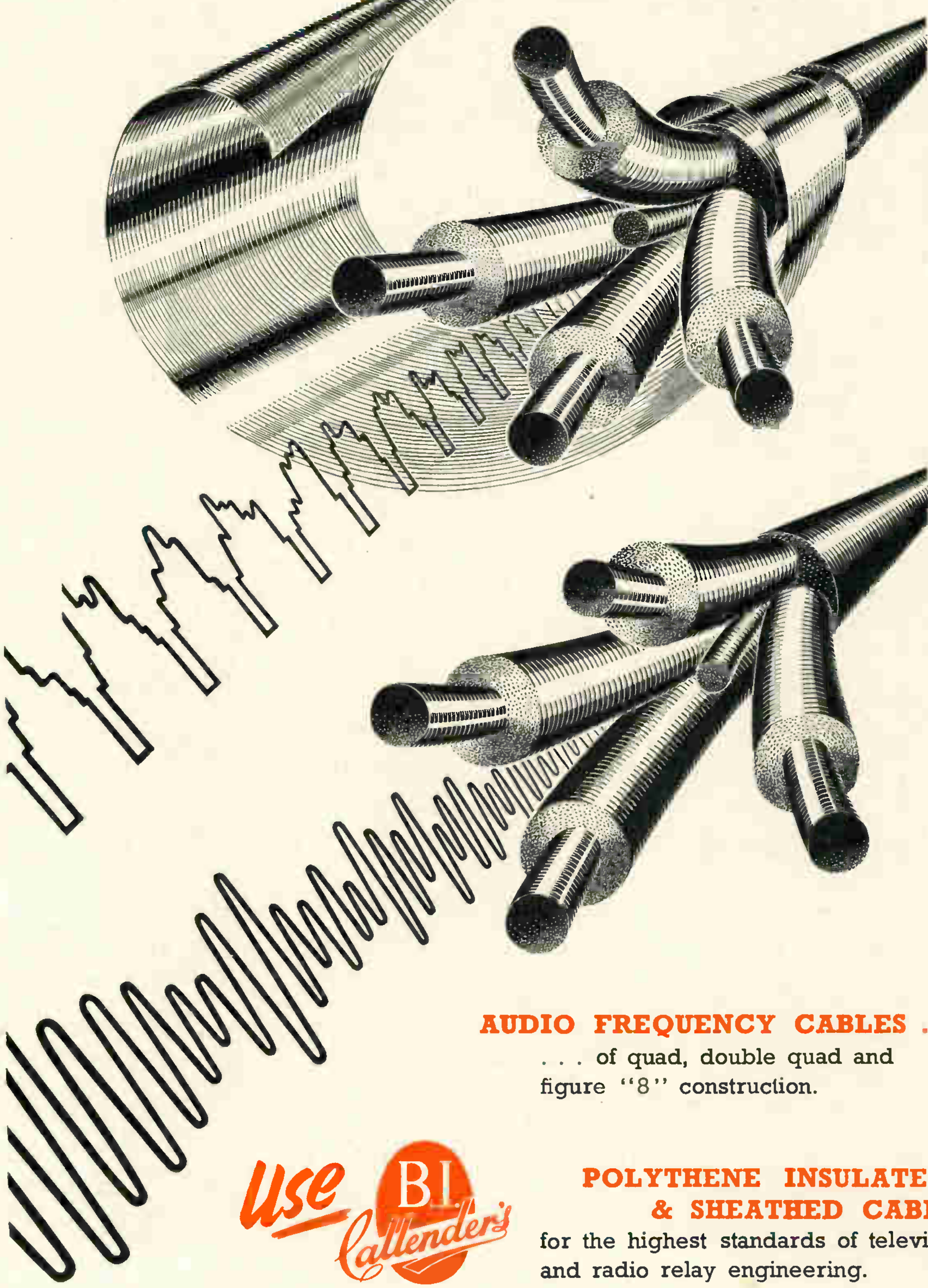

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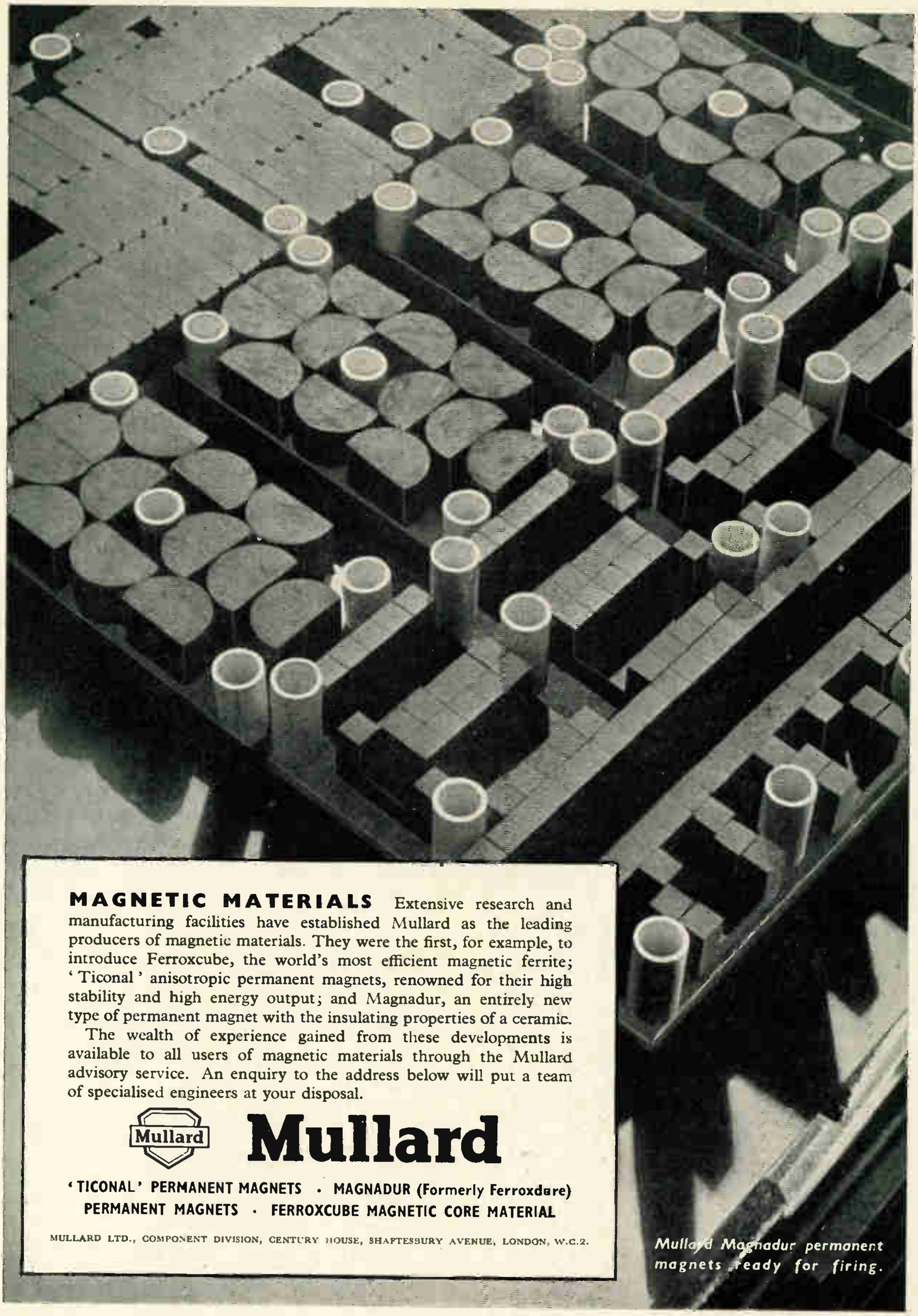

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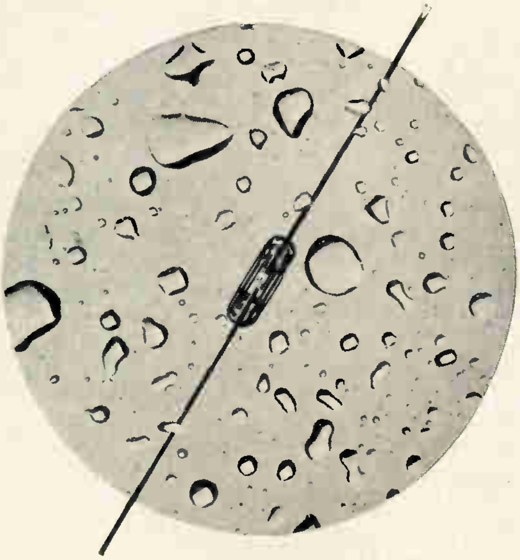

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Electromagnetic Induction from Different Points of View

THE divergency of views that are held on this subject was brought out in a paper recently read before the I.E.E. by Mr. P. Hammond* and in the subsequent discussion. The middle section of the paper was entitled "Electromagnetic Induction" and it is to this that our remarks will be confined.

If Fig. 1(a) represents an electron coming towards us in an unchanging magnetic field, it is equivalent to a current element in the opposite direction and will be associated with a right-handed circular magnetic field which will distort the uniform field in such a way as to cause a downward force on the electron, which is therefore deflected. Fig. 1(b) is much the same except that the electron under consideration is one of the free electrons in the conductor which is moving towards us; the electron will experience a downward force which is equivalent to an e.m.f. in the upward direction. In the Faraday disc armature the circumferential movement of the free electrons causes them to experience a radial force, which, in the absence of any external circuit, will cause a potential difference between the centre and the periphery of the disc. Current will therefore flow in any circuit connected between brushes on the axis and periphery. It is amusing to see the efforts often made to explain this fundamentally-simple generator by means of the variation of magnetic flux through fictitious circuits.

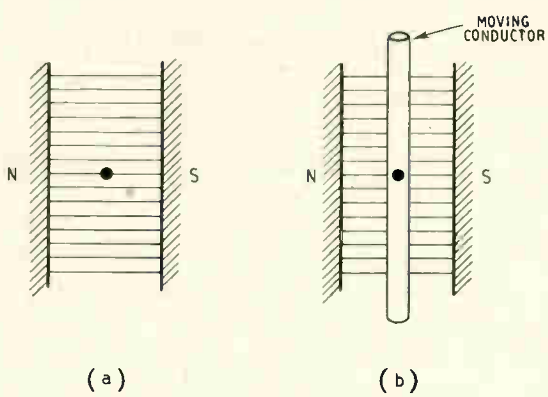


Fig. 1.

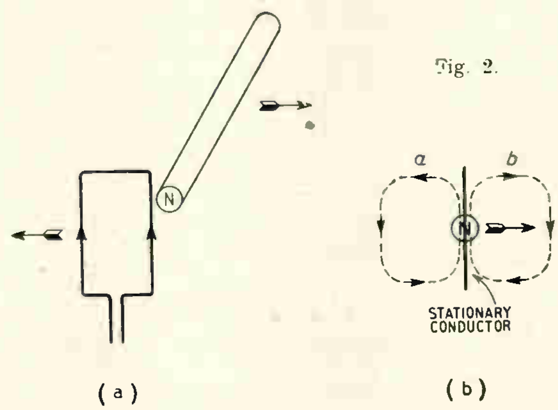


Fig. 2.

Fig 2(a) shows a bar magnet and a conductor moving relatively to each other. To an observer on the magnet the e.m.f. induced in the conductor is due to its movement through an unchanging magnetic field. To the same observer the e.m.f. induced in the rectangle is due to the difference

* A short modern review of fundamental electromagnetic theory. Read 11th Feb., 1954.

between the e.m.fs induced in the two sides because of the difference between the fields through which they are moving, or he might explain it as being due to the decreasing flux through the circuit. This really comes to the same thing, for as the field is everywhere unchanging, the root cause of the e.m.f. is the movement of the conductors in the magnetic field.

But how would this be explained by an observer on the conductor? To him the conductor is at rest and the magnet moving and causing variations in the strength of the magnetic field in all the surrounding space. If we consider the two closed paths in Fig. 2(b), as the pole moves to the right, the flux coming towards us through the loop *a* decreases, whereas that through loop *b* increases, and the e.m.fs induced in the two paths are as shown. The conductor is thus situated in an electric field and, in the case of the rectangle, one side is in a stronger electric field than the other. To one observer the space around the magnet is a simple stationary magnetic field, whereas to the other observer it is a combination of changing magnetic and electric fields.

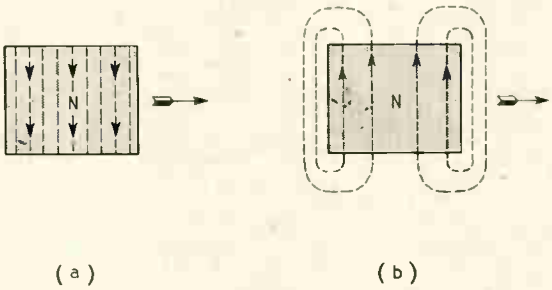


Fig. 3.

One can cover both cases by picturing the lines of force and saying that the induced e.m.f. is due to the relative motion of the conductor and the lines or the cutting of one by the other, but this seems to lack fundamental explanation. To the observer on the magnet nothing is happening inside the magnet, as the field is unchanging, but to the observer on the conductor the magnetic material is moving through its own magnetic field and therefore inducing an e.m.f. within the material as shown in Fig. 3(a), but if one considers two closed paths embracing the leading and trailing edges of the magnet, e.m.fs will be induced in them as shown in Fig. 3(b) due to the changing flux. These two e.m.fs, or the charges that they would produce on the surface, cancel each other, and so give the same result as obtained by the other observer. In all the cases so far considered the e.m.f. is due to the motion of the conductor or of the magnet and can presumably be called motional e.m.f.

Turning from linear motion, which is purely relative, to rotation, which is absolute, let us consider the simple 2-pole dynamo with a smooth armature shown in Fig. 4. Neglecting the armature coil, and assuming the poles to be excited and the armature rotating, the field will be everywhere unchanging, both in the air-gap and in the iron, to the normal stationary observer.

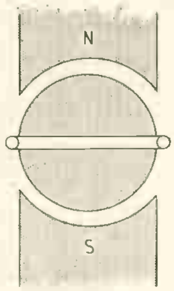


Fig. 4.

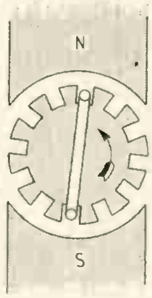


Fig. 5.

If, now, the armature coil is rotated, either alone with the iron armature at rest, or attached to the rotating armature, the coil is moving in a magnetic field which at no point is undergoing any change either in magnitude or direction with respect to time. If, now, the coil is stationary and the polar system is rotated in the opposite direction, so that instead of the coil cutting the lines the lines cut the coil, the result is, of course, the same, and the e.m.f. in both cases can surely be called motional.

In his paper Mr. Hammond deals with this in a very peculiar manner. He divides the field of the stationary poles and the armature into two components B_1 and B_2, the former due to the field magnets and the latter to the surface polarity of the armature. He then says "a purely motional e.m.f. will be induced in the armature coils by their rotation through the field B_1. The component B_2 does, however, experience a time variation *which would be noticed by an observer travelling with the armature coils.*" (The italics are ours). He concludes that "the e.m.f. induced in the coil is thus found to consist of a motional effect due to the movement of the coil through the constant field of the pole magnets, and a transformer effect caused by the variation of the surface polarity of the armature". This is surely quite unjustifiable. Although at any point in the rotating armature fixed with respect to the stationary observer, one iron molecule is being continually replaced by another, the magnetic condition at the point is unchanging, and the surface polarity of the armature, as viewed from the pole, is as stationary as that of the pole itself, and neither B_1 nor B_2 undergo any time variation. His observer travelling with the armature coil—in other words, spinning round with the armature coil—would certainly note a time variation of B_2, but also of B_1. To him they would both, like the poles themselves, appear to be rotating in the opposite direction to that in which he is rotating. The coil is really rotating through a field which to a stationary observer is undergoing no change whatever at any point and the e.m.f. is purely motional.

The conditions are entirely different if the armature is slotted, for the magnetic field at every point in or near the air-gap is changing as the armature rotates, and the induced e.m.f. can be explained in different ways. If one regards the lines of force (or magnetic induction) crossing the air-gap as definite physical entities, sometimes crowded together in a tooth and sometimes widely separated in a slot, then, although a conductor in a slot is in a very weak field, it must cut all the lines emanating from the pole as the slot moves across the pole-face. This it does due to the very high speed at which the lines flash across the slot, a speed many times the peripheral speed of the armature. From this point of view the induced e.m.f. is certainly motional. If, however, we adopt a less materialistic view of the magnetic field, we can explain the induction of the e.m.f. in a different way. The conductors are moving in a very weak field and the motional e.m.f. so induced will be small, but as the armature in Fig. 5 rotates, the flux through the coil, which will be very small in the position shown, will pass through zero and then increase as the teeth on one side of it pass from the N to the S pole and the teeth on the other side from the S to the N pole. The variation of the flux through the coil is thus due to the growth and decay of the flux through the teeth without any flashing across the slots. If the armature were fixed and the poles rotated in the opposite direction, the coil would not be moving, even in a weak field, and the induced e.m.f. would be due entirely to the variation of the magnetic flux through it. Should this be called a motional or a transformer e.m.f.?

G. W. O. H.

External Field of a Widely-Wound Toroid

It has been suggested to us that we might devote an Editorial note to the following question. Fig. 1 shows a toroidal coil wound, say, on a non-magnetic core; the dotted circle is a closed path external to the toroid, and the line integral of H around this path should be zero. If, however, the turns are rather widely spaced, as shown in the figure, and the dotted circle is not much bigger than the toroid, a unit pole being carried round this path would come within the field of the outer conductors as it passed near them, and it is difficult to see how the line integral

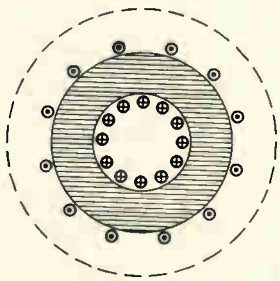
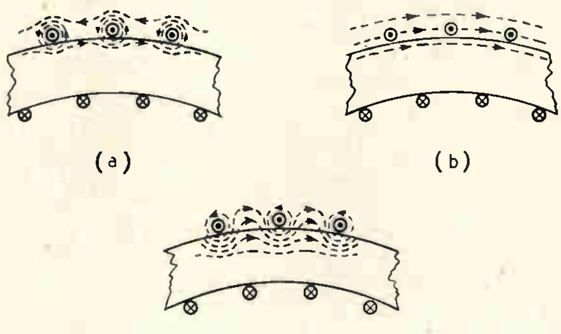


Fig. 1.

could be zero. The answer to the question is seen in Fig. 2, which shows a small piece of the toroid; it is perhaps simpler to assume a cylindrical toroid so that the wires are straight and parallel. Considering the field near the outer surface, Fig. 2(a) shows the field due only to the outer wires, Fig. 2(b) the field due only to the inner wires, and Fig. 2(c) the resultant field. In (a) the field between the wires is very weak, being zero midway between them so that the resultant field there is mainly due to the inner wires. From Fig. 2(c) it is clear that in following the dotted path close to the toroid one passes alternately through fields in opposite directions, thus giving zero line integral. Even if the spacing is not very wide, the same must be true to some extent, because, if, on passing very near the wire, its field can be detected, there must be a reversal of the field in passing from one wire to the next since the line integral must be zero.



(a) (b) (c)

Fig. 2.

G. W. O. H.

CURRENT-NOISE IN COMPOSITION RESISTORS

By D. A. Bell, M.A., Ph.D. and K. Y. Chong, M.Sc.

(Electrical Engineering Department, University of Birmingham)

SUMMARY.—Current-noise appears generally to be a resistance-modulation effect, but in composition resistors the expected relation $\overline{\delta V^2} \propto i^2$ is experimentally found to be valid only for small currents. At large currents there are hysteresis effects in the noise without comparable effects in the d.c. resistance. It is shown that the noise is very sensitive to current concentration, and this could account for the lack of correlation between noise and d.c. resistance with large currents.

1. Dependence of Noise on Steady Current

THE view that current-noise in semi-conductors is a resistance-modulation effect implies that the mean-square fluctuation voltage δv^2 should be proportional to the square of i, the steady current. Since there are also reasons (both from dimensional arguments and from the type of mechanism proposed by Macfarlane in 1947) for supposing that the noise would be proportional to i^{x+1}/f^x the relation between noise and steady current is of theoretical as well as practical interest.

The first set of measurements on carbon-composition resistors showed the noise power to increase less rapidly than as the square of the current, Fig. 1, but closer investigation of the noise with small steady currents, Fig. 2, showed that the linear relationship $\delta v^2 \propto i^2$ was valid for steady currents producing a dissipation of one-tenth or less of the nominal rating of the resistor. (See also Templeton and MacDonald, 1953.)

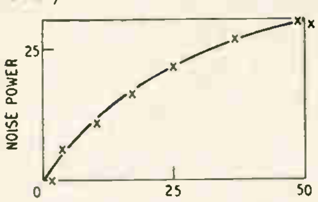


Fig. 1 (above). *Relation between noise and current for large values of current.*

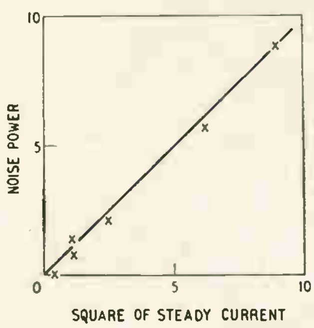


Fig. 2 (right). *With small currents the noise power is nearly proportional to the square of the current.*

Although there are considerable differences in behaviour between different specimens of nominally similar resistor (see below), Fig. 3 is a typical example of a noise characteristic up to a loading of two-thirds of the nominal rating. These characteristics showed differences between successive runs; they could also be changed, often with a substantial reduction of noise, by subjecting the resistors to an overload of six times the nominal dissipation for ten minutes. At the same time the steady voltage versus current characteristic remained constant and ohmic (Fig. 4). From the occurrence of these hysteresis effects it is believed that the departure from the initial proportionality between noise and square of steady current is not due to non-linearity of the essential physical law but to partially-reversible discontinuous changes in the internal structure of the resistor. (An explanation of these changes in terms of thermal instability has been proposed by one of the authors, Chong, 1953).

MS accepted by the Editor, October 1953

2. Importance of Current Density

In a homogeneous material it would be better to express the noise in terms of *current density* and *specific resistivity*, rather than in the customary manner in terms of *current* and *resistance*. Composition resistors are not homogeneous bodies, so that (a) the current-density in the conducting paths within the resistor is far greater than the value obtained by averaging over the whole of the cross-section of the body, and (b) the current-density may vary widely between different points along a given conducting path between the two terminals. Owing

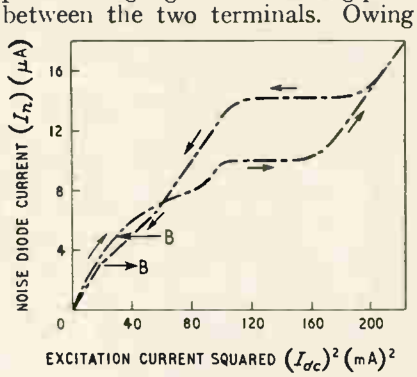


Fig. 3. *Typical example of a noise characteristic.*

to the dependence of the noise on the *square* of the current-density, the effect of the series combination of the unequally-loaded parts will be to make the noise disproportionately dependent on the high-density parts of the path. If the noise originates mainly at comparatively few points where the current-density is high, any thermal or other modification of the structure at these points can greatly change the noise without much affecting the d.c. resistance, in the way found in the experimental results of Figs. 3 and 4.

This view was strikingly confirmed by an experiment designed to show how a redistribution of current paths can drastically change the noise output from a resistor while having only a small effect on the d.c. resistance. Split circular clamps were attached to the outside of the resistor at the ends, Fig. 5, and the 'noisiness' of different systems of connections was compared by adjusting the steady current in each case to give a reference value of noise output, and the results are shown in Table 1. (The identification of terminals is by means of the letters shown in Fig. 5.) It will be seen that the connection of

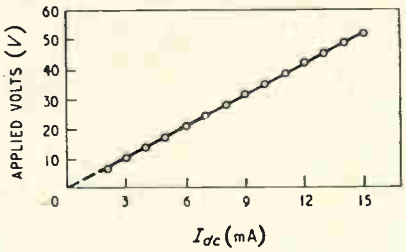


Fig. 4. *Relation between applied volts and current.*

the clamped connections in parallel with the normal terminals makes no significant difference to the resistance but increases the noisiness about 50 times. (The noise is nominally proportional to the square of the d.c., so that reducing the latter seven-fold is equivalent to a 49-fold change of noise in a given resistor.) The further increase of noisiness when clamped terminals alone are used is out of all proportion to the accompanying change of resistance. Since the clamps were making only mechanical contact with the untreated surface of the resistor, it

TABLE 1

Current flow	D.C. for ref. value of noise	Measured resistance (ohms)
A to B	7 mA	3,800
A + C to B + D	1 mA	3,800
C to D	26 μA	5,800
A to B + D	53 μA	3,800
A + C to B	62 μA	3,800

seems likely that the electrical contact was confined to a limited number of points on the surface, thus giving rise locally to high current densities and high noise level.

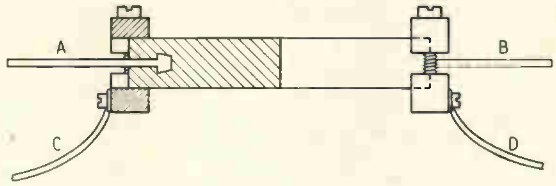


Fig. 5. *A and B are the normal resistor connections; C and D are extra clip contacts.*

3. Variability of Composition Resistors

A difficulty which appears to be common to other types of semi-conductor also (see, for example, the measurements quoted by Torrey and Whitmer on crystal rectifiers) is that different specimens of nominally similar resistors show widely different values of current-noise coefficient. It might be expected that there would be a random variation and, if there were, the dispersion within a group would lead to a straight-line plot on probability graph paper* (taking 'random variation' to mean that the distribution about the mean is Gaussian or follows the so-called 'normal law of errors'). Results for a sample of 20 nominally-similar resistors are plotted in Fig. 6, curve (a), where there appear to be five resistors which are consistent (points within the dotted rectangle), one which is exceptionally good, and 14 which are scattered over rather a wide range of higher values of noise. Earlier results for a sample of six only,

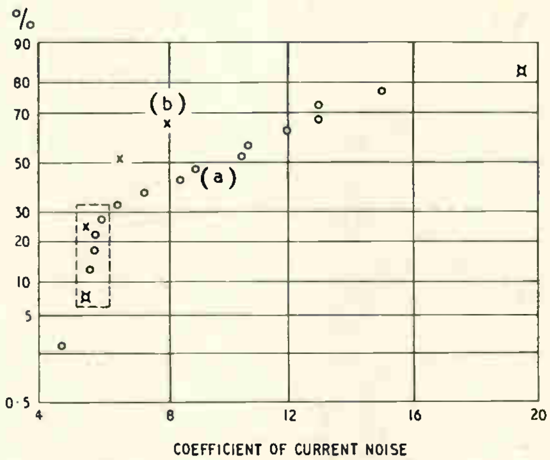


Fig. 6. *Performance of 20 nominally-alike composition resistors, curve (a) and earlier results for six resistors only, curve (b).*

* 'Probability graph paper' as a test for Gaussian distribution was devised by A. F. Dufton (1930) and its use has been reviewed by W. R. Hinton (1949).

curve (b), show a similar trend, and in the graph published by Campbell and Chipman (1949) there is (in the only case for which they record measurements on 20 samples) a tendency for some points to be grouped together near the minimum noise for given conditions, while others spread out to much higher values of noise: this is in their Fig. 4(a), for 1-W metallized resistors of 11,000 Ω. In comparing different types of resistor it is therefore desirable to measure 20 specimens of each, and a suitable representative value to take is the mean value from the closely-spaced group as indicated in our Fig. 6 for group (a). Unfortunately the first set of measurements covered only six specimens of most of the types concerned, and in these instances the lowest individual value of noise has been taken to represent the basic state.

Acceptance of the local-current-density source of noise makes this variation between nominally similar resistors less surprising, for the nominal similarity refers to the overall resistance while the noise, according to the local-density theory, is localized in regions which may be insignificant from the point of view of total resistance.

4. Relation between Noise and Bulk of Composition Resistors

Measurements were made on a range of different types of resistor all nominally of 3·3 kΩ; and the noise coefficient was taken to be the number of microamps of noise-diode current which was equivalent by substitution to the excess noise in the resistor per squared-milliampere of steady current. It was originally supposed that the noise would be inversely proportional to the total volume of the resistor, and that there would probably be a shape factor as well. From a qualitative consideration of the combination of conducting paths in series and in parallel it was thought that for a given value of resistance the noise coefficient might be a function of l^2/Av where l is the length, A the sectional area and v the volume; and it was found empirically that the results for resistors of the same resistance value but different shapes and sizes did fit this formula rather better than several other functions

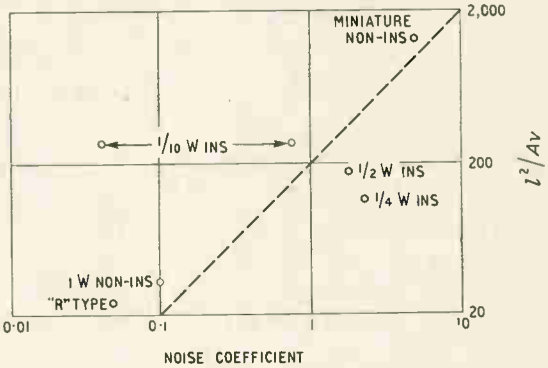


Fig. 7. *Relation between resistor performance and its shape factor. Each point on this diagram is deduced from measurements on a number of samples.*

which were tried. This empirical relationship is illustrated with log-log scales in Fig. 7, where it is clear that there is one type which is very much better than the others and presumably this type has some radical difference in manufacture from the other 1/10th W, which is externally similar but produces as much noise as the average.

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INTERMODULATION DISTORTION IN RECTIFIER MODULATORS

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SUMMARY.—It is shown that non-linear distortion occurs in linear rectifiers, and balanced modulators using linear rectifiers, due to the interference of the signal voltages with the switching function of the carrier. On the assumption of signal voltages low compared with the carrier, a method of calculating the harmonic and intermodulation-distortion products is developed, and a full tabulation is given, for two input signal tones, of all third-order distortion terms based on first- and third-order modulation. An appendix extends this in general form to all orders of modulation and to second- as well as third-order distortion. The values actually tabulated apply to a single rectifier with finite forward and back resistances, and to an ideal ring modulator and ideal balanced shunt and series modulators in which the carrier-generator resistance is zero, and the rectifiers have zero forward and infinite back resistance. For more practical cases, the functions expanded in this paper can be readily fitted into a previous analysis by Belevitch of harmonic distortion in modulator circuits which have finite rectifier and carrier-generator resistances; they thus enable the intermodulation products to be obtained for these circuits. But except in the balanced shunt and series circuits with high carrier-generator resistance, the departure from the performance of ideal modulators is small.

It is shown that the non-linear behaviour described above is not similar to that caused by non-linearity in the rectifier resistances, and cannot be represented by the same equivalent circuits except by an artifice in which different non-linear coefficients are taken in calculating the products of the various orders of modulation, and in which the independent variable of the non-linear equation is the ratio of signal to carrier voltages instead of the signal voltage itself.

LIST OF SYMBOLS

As used by Belevitch:—

F = carrier frequency (c/s)
f = signal frequency (c/s)
v_c = instantaneous carrier voltage
v_s = instantaneous signal voltage

As used by present author in Reference 8:—

p = angular frequency of carrier
$p - q$ = angular frequency of signal
so that q = difference frequency
$E_1 \cos pt$ = carrier voltage
x = ratio of signal/carrier voltage across the rectifier or any circuit containing it which is common to both signal and carrier.
h_1 = coefficient of $\cos pt$ in the Fourier expansion of the modulating or switching function of a single rectifier. This is, in effect, the factor representing the influence of the external circuit on the first-order modulation efficiency. In particular, if $r_f = 0$ and $r_b = \infty$, then $h_1 = 2/\pi$ for any finite value of the external circuit resistance.
r_f = forward resistance of rectifier
r_b = back resistance of rectifier
k = order of modulation; i.e., the order of F in the output components $kF \pm mf_1 \pm nf_2$.

1. Introduction

THIS paper is intended to supplement existing accounts of non-linear distortion in rectifier modulators, and will not attempt to summarize the whole subject. The literature of the subject is not large, however, and is comprised mainly of four contributions in *Wireless Engineer* by Belevitch[1,3,4] and Tucker and Jeynes,[2] together with a few others[5,6,7]. In these papers it is shown that non-linear distortion occurs in modulators such as the ring, shunt and series types (which are balanced rectifier-modulators with a high-level carrier voltage applied longitudinally), due to two main causes:

(a) The non-linearity of the rectifier characteristics when biased either forwards or backwards (i.e., the slope or a.c. resistance is not constant and therefore varies over the cycle of superimposed signal voltage).

(b) The interference with the switching of the rectifiers from forward to backward condition and vice versa—an operation ideally under the sole control of the carrier voltage—caused by the superimposition of the signal voltage.

The first cause is dealt with by Christiansen[6] and is not inherent in the conception of the modulator action. The second cause is inherent in the use of any carrier waveform other than a square-wave; in normal usage, a sinusoidal carrier e.m.f. is provided. Belevitch[1,3,4] has given a very comprehensive and elegant analysis of this effect, but has restricted his consideration to distortion occurring when the applied signal voltage is a single sinusoid.

It is the purpose of the present paper to show how the distortion type (b) can be calculated when the signal comprises two or more tones. As in Belevitch's work, it is assumed that the rectifiers are 'linear', that is to say, they have a constant back resistance r_b and a constant forward resistance r_f, and switch from one to the other at zero applied voltage.* The basis of the

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* The problem of what distortion occurs in practical modulators, in which the rectifiers are not perfect switches, is not so simple to deal with, as all sorts of non-linear laws are encountered. It is clear that the actual law plays a most important part, since Section 4 shows that when square-law

(Footnote continued on next page)

calculation is the analysis recently published by the present author[8] of complex signals applied to a single linear rectifier. Throughout it is assumed that the signal amplitudes are small compared with the carrier.

2. Single Linear Rectifier

It will become clear later that the performance of balanced modulators, however complex, can be described in terms of the equations applying to a single linear rectifier. We will thus examine the intermodulation-distortion of a linear rectifier to which the voltage

$$e = E_1\,[\cos pt + x_1 \cos (p - q_1)\,t + x_2 \cos (p - q_2)\,t] \quad \ldots \quad \ldots \quad (1)$$

is applied; $E_1 \cos pt$ is the carrier, and $x \ll 1$. The analysis will thus cover all ordinary requirements for intermodulation, since as x approaches unity, the distortion will be too large for most applications. Distortion levels around 60 or 70 db below the wanted outputs are commonly specified.

The rectifier has a switching function, $\phi(t)$, generated by the applied voltage, and the output is therefore

$$e\,.\,\phi(t) \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

Now $\phi(t)$ is discussed and derived in Reference 8, Equ. (11), and, ignoring the constant term which has no bearing on distortion, is shown to be approximately

$$\phi(t) = h_1 \sum_{n=1}^{\infty} \frac{(-1)^{n-1}}{2n-1} \cos\,[(2n-1)\,(pt - x_1 \sin q_1 t - x_2 \sin q_2 t)] \quad (3)$$

This can be expanded into series form with Bessel functions as coefficients; and if for compactness we put $r = 2n - 1$, so that r takes the values 1, 3, 5, etc., we obtain

$$\frac{\phi(t)}{h_1} = \sum \frac{(-1)^{(r-1)/2}}{r} \text{ multiplied by}$$

$$\begin{aligned}&\cos rpt\,[\{J_0(rx_1) + 2J_2(rx_1)\cos 2q_1 t + \ldots\}\{J_0(rx_2) + 2J_2(rx_2)\cos 2q_2 t + \ldots\}\\ &-\{2J_1(rx_1)\sin q_1 t + 2J_3(rx_1)\sin 3q_1 t + \ldots\}\{2J_1(rx_2)\sin q_2 t + 2J_3(rx_2)\sin 3q_2 t + \ldots\}]\\ &+\sin rpt\,[\{2J_1(rx_1)\sin q_1 t + 2J_3(rx_1)\sin 3q_1 t + \ldots\}\{J_0(rx_2) + 2J_2(rx_2)\cos 2q_2 t + \ldots\}\\ &+\{2J_1(rx_2)\sin q_2 t + 2J_3(rx_2)\sin 3q_2 t + \ldots\}\{J_0(rx_1) + 2J_2(rx_1)\cos 2q_1 t + \ldots\}] \quad \ldots \quad \ldots \quad \ldots \quad (4)\end{aligned}$$

Now, since $x \ll 1$, and since we are not likely to be concerned with large values of r, we can use the following approximations for the Bessel functions:

$$\left.\begin{aligned}J_0(x) &\approx 1\\ J_1(x) &\approx x/2\\ J_2(x) &\approx x^2/8\\ J_3(x) &\approx x^3/48\end{aligned}\right\} \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

From these formulae, any desired or undesired output component may be worked out, easily if rather laboriously.

Before listing products of interest, it is as well to point out that there are traps in interpreting the order of intermodulation and order of modulation. A frequency term such as

$$kF \pm mf_1 \pm nf_2$$

is regarded as of the kth order of modulation, and of $(m + n)$th order of intermodulation, but this does not mean that only the terms in $\cos kpt$ and $\sin kpt$ in Equ. (4) contribute to it. Care must be taken to see that all terms which can contribute are included in the calculation. This point is emphasized in the tables of products below. Note that throughout, all output terms are cosines.

The first-order terms are shown in Table 1.

TABLE 1

Product		Relevant values of r in Equ. (4)	Amplitude (Multiply by $h_1 E_1$)
F, f notation	p, q notation		
$F - f_1$	q_1	1	$x_1/2$
$F - f_2$	q_2	1	$x_2/2$
$F + f_1$	$2p - q_1$	1, 3	$x_1/2$
$F + f_2$	$2p - q_2$	1, 3	$x_2/2$

The third-order intermodulation and harmonic terms based on first-order modulation are shown in the first part of Table 2, and those based on third-order modulation are shown in the second part of Table 2. The latter must not be overlooked in practice, as many of them are of considerable importance and relatively large amplitude.

Third-order distortion terms for higher orders of modulation can be readily calculated from the general second- and third-order expansion of Equ. (2) given in Appendix 1. It will be found that the terms $kF \pm 3f$ have an amplitude

$$x^3\,.\,(-1)^{(k+1)/2}\,.\,h_1 E_1 k/48$$

and the terms $kF \pm f_1 \pm 2f_2$ have an amplitude

rectifiers are used, there is no distortion at all. On the other hand, experience suggests that the distortion magnitudes as calculated for the perfect-switch rectifiers in this paper are not often exceeded in practice, and if an extra margin of say 3–6 db is allowed for effect (a)—it will be seen from Section 9 that this need be allowed only for distortion products of first-order modulation—then designs based on these calculations will not often fail to meet their required performance. At any rate, this will be a much better basis of design than merely to make distortion measurements on one sample modulator and then assume the results obtained will apply to all production modulators of the same nominal design—they will not! Variations from one sample to another may be very large.

$$x_1 x_2^{-} \,.\, (-1)^{(k+1)/2} \,.\, h_1 E_1 \; k/16.$$

The wanted side-tones, $kF \pm f$, have an amplitude approximately

$$x \,.\, (-1)^{(k-1)/2} \,.\, h_1 E_1/2k$$

The distortion on higher orders of modulation is thus worse by a factor approximately k^2 than that on first-order modulation.

In the single linear rectifier all orders of non-linearity are produced.* For instance, there are second-order terms such as

$$h_1E_1 \left[-(x_1^2/8) \cos 2q_1t - (x_2^2/8) \cos 2q_2t + (x_1x_2/4) \cos (q_1 - q_2)t - (x_1x_2/4) \cos (q_1 + q_2)t\right] \quad .. \quad .. \quad (6)$$

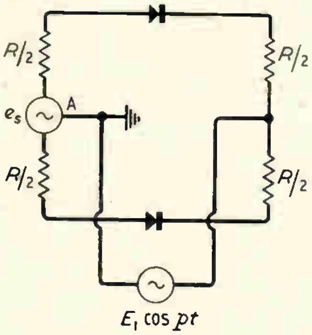

Fig. 1. *Balanced series modulator, where* $e_s = 2E_1 [x_1 \cos (p - q_1) t + x_2 \cos (p - q_2) t]$. *A is the centre point of the balanced signal source.*

but as all terms produced by even-order non-linearity cancel out in perfectly-balanced modulators, they are not important in the present theoretical study. They can, if desired, be readily written down from the expansion given in Appendix 1.

If intermodulation effects for three or more signal tones are required, they can be deduced in the manner discussed in Reference 8, Section 3.3.

3. Balanced Series Modulator

The working of the previous section can be applied directly to a balanced modulator only if the identical conditions are maintained, namely, that the carrier and signal currents flow through the same circuit, so that their ratio is unaffected by the switching of the rectifiers. Fig. 1 shows a balanced series modulator which meets this requirement; a corresponding shunt modulator is easily devised. In this modulator, one rectifier has an applied e.m.f.

$$e_- = E_1 \left[\cos pt + x_1 \cos (p - q_1)\, t + x_2 \cos (p - q_2)\, t\right] \quad .. \quad .. \quad (7)$$

which gives rise to a switching function $\phi_+(t)$, defined by Equ. (3); and the other rectifier has an applied e.m.f.

$$e_- = E_1 \left[\cos pt - x_1 \cos (p - q_1)\, t - x_2 \cos (p - q_2)\, t\right] \quad .. \quad .. \quad (8)$$

giving rise to a switching function $\phi_-(t)$ defined by Equ. (3) but with $-x$ replacing x.

The output voltage is evidently

$$e_+ \,.\, \phi_+ (t) - e_- \,.\, \phi_- (t) \quad .. \quad .. \quad (9)$$

which contains only terms in odd orders of x; i.e., the carrier and all even-order terms are balanced out. The ratio of each of the third-order intermodulation and harmonic terms to the first-order output terms is the same as for the single rectifier. Note that the signal e.m.f. to the whole modulator is $2xE_1 \cos (p - q)\, t$ for each frequency,

TABLE 2

Product		Relevant values of r in Equ. (4)	Amplitude (Multiply by h_1E_1)
F, f notation	p, q notation		
$F - 3f_1$	$2p - 3q_1$	1, 3	$-x_1^3/48$
$F - 3f_2$	$2p - 3q_2$	1, 3	$-x_2^3/48$
$F + 3f_1$	$4p - 3q_1$	3, 5	$-x_1^3/48$
$F + 3f_2$	$4p - 3q_2$	3, 5	$-x_2^3/48$
$F + f_1 + 2f_2$	$4p - q_1 - 2q_2$	3, 5	$-x_1x_2^2/16$
$F + 2f_1 + f_2$	$4p - 2q_1 - q_2$	3, 5	$-x_1^2x_2/16$
$F + f_1 - 2f_2$	$q_1 - 2q_2$	1	$-x_1x_2^2/16$
$F - 2f_1 + f_2$	$2q_1 - q_2$	1	$-x_1^2x_2/16$
$F - f_1 + 2f_2$	$2p + q_1 - 2q_2$	1, 3	$-x_1x_2^2/16$
$F + 2f_1 - f_2$	$2p - 2q_1 + q_2$	1, 3	$-x_1^2x_2/16$
$F - f_1 - 2f_2$	$2p - q_1 - 2q_2$	1, 3	$-x_1x_2^2/16$
$F - 2f_1 - f_2$	$2p - 2q_1 - q_2$	1, 3	$-x_1^2x_2/16$
$3F - 3f_1$	$3q_1$	1	$x_1^3/16$
$3F - 3f_2$	$3q_2$	1	$x_2^3/16$
$3F + 3f_1$	$6p - 3q_1$	5, 7	$x_1^3/16$
$3F + 3f_2$	$6p - 3q_2$	5, 7	$x_2^3/16$
$3F + f_1 + 2f_2$	$6p - q_1 - 2q_2$	5, 7	$3x_1x_2^2/16$
$3F + 2f_1 + f_2$	$6p - 2q_1 - q_2$	5, 7	$3x_1^2x_2/16$
$3F + f_1 - 2f_2$	$2p - q_1 + 2q_2$	1, 3	$3x_1x_2^2/16$
$3F - 2f_1 + f_2$	$2p + 2q_1 - q_2$	1, 3	$3x_1^2x_2/16$
$3F - f_1 + 2f_2$	$4p + q_1 - 2q_2$	3, 5	$3x_1x_2^2/16$
$3F + 2f_1 - f_2$	$4p - 2q_1 + q_2$	3, 5	$3x_1^2x_2/16$
$3F - f_1 - 2f_2$	$q_1 + 2q_2$	1	$3x_1x_2^2/16$
$3F - 2f_1 - f_2$	$2q_1 + q_2$	1	$3x_1^2x_2/16$

* Note that odd orders of non-linearity are always associated with odd orders of modulation, and even orders of non-linearity with even orders of modulation.

so that the ratio of (signal voltage)/(carrier voltage) is $2x$ and not x.

4. Use of Square-Law Rectifiers

If the rectifiers in the previous discussion had been square-law instead of linear they would be represented by an output (v_0) to input (v_1) relationship

$$v_0 = a_1 v_1 + a_2 v_1^2 \qquad \ldots \quad \ldots (10)$$

No distortion products of order higher than the second could then be produced, so that a balanced modulator as shown in Fig. 1, in which all even-order terms are balanced out, would be completely distortionless.

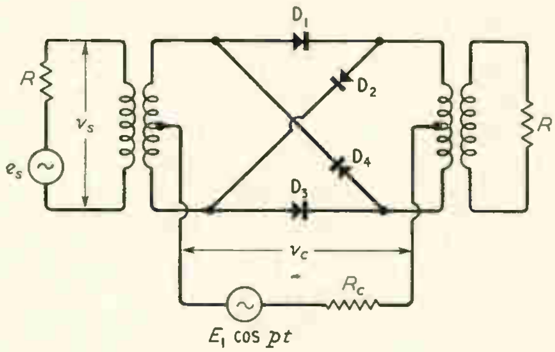

Fig. 2. *Ring modulator, where*

$e_s = 2V_c\,[x_1 \cos (p = q_1)\,t + x_2 \cos (p - q_2)\,t]$;

$v_c = V_c \cos pt$ — *carrier p.d. when* $e_s = 0$;

v_s = *signal p.d. when modulator is not overloading; e.g. around* $pt = 0, \pi$, *etc.*

5. Ring Modulator with Ideal Rectifiers

The ring modulator with rectifiers having $r_f = 0$ and $r_b = \infty$, and a carrier-generator resistance of zero, was dealt with by Belevitch (for a single signal tone) in his first paper[1] and is a particularly simple case. Fig. 2 shows a ring modulator, and putting in the conditions above, it can be seen that so long as the carrier keeps rectifiers D_1 and D_3 conducting and D_2 and D_4 blocking (or vice versa), the output is the signal voltage v_s. As the carrier approaches zero voltage at $pt = \pi/2$, $3\pi/2$, etc., the condition is reached when the signal and carrier currents through one of the blocking rectifiers become equal and opposite, and then that rectifier starts to conduct and unbalances the circuit. For example, if D_1 and D_3 are conducting and the signal current (assumed altering very slowly) is flowing clockwise, then it is D_2 which starts to conduct as the carrier approaches $pt = \pi/2$. As soon as this happens, the signal is short-circuited and the carrier-balance fails, so that the output is then the carrier voltage v_c (i.e., $v_c/2$ is produced across half the primary of the output transformer, giving v_c across the secondary). This condition continues until at $v_c = 0$, D_3 ceases to conduct and D_4 ceases to block, but this change does not alter the output, which is still v_c. Then, as v_c increases negatively, a point is reached when the carrier current in D_1 becomes equal and opposite to the signal current, and then D_1 ceases to block; the output then becomes v_s again, but reversed in phase, and there is no output due to the carrier, since the circuit is once more balanced.

The output voltage can thus be written as

$$v_s\,[\phi_+(t) + \phi_-(t)] + v_c\,[\phi_+(t) - \phi_-(t)] \quad \ldots \quad (11)$$

where $\phi(t)$ is as defined by Equ. (3) with subscripts as used in Equ. (9) and with $h_1 = 2/\pi$.

The way in which this expression represents the circuit action as described above will probably be easier to see with the help of Fig. 3, which shows the waveforms of the switching functions, assuming the signal voltage changes only slowly; i.e., is of relatively low frequency. In these functions, x is the ratio of signal p.d. to carrier p.d. (i.e., v_s/v_c) since this is the ratio of signal current to carrier current through the rectifier which fails, both before and after it switches over.

Equ. (11) can be rewritten as

$$e_+ \,.\, \phi_+(t) - e_- \,.\, \phi_-(t) \qquad \ldots \quad \ldots (12)$$

using the symbolism of Equs. (7) and (8), and this output is identical with that of the balanced series modulator, Equ. (9), except for the value of h_1 which is used in determining $\phi(t)$. This means that although the ring modulator is more efficient, its distortion ratios are identical with those of the previous circuit, provided the (signal voltage)/(carrier voltage) ratio is x instead of $2x$ as used in the previous circuit.

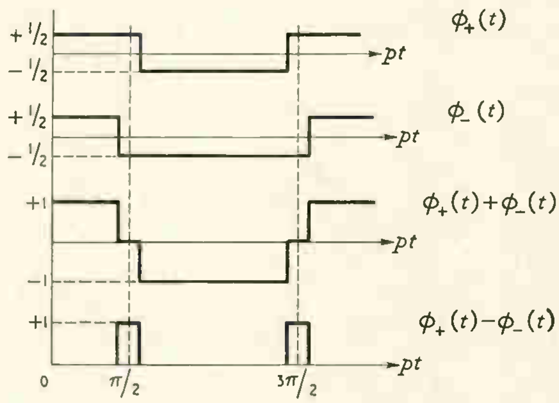

Fig. 3. *Switching functions in ring modulator.*

It should be noted that Belevitch calculated the $F \pm 3f$ and $3F \pm 3f$ products, and the values given here agree exactly with his, provided x is small.

6. Ring Modulator with Non-Ideal Rectifiers

If the rectifiers have finite forward and backward resistances and if the carrier generator resistance is not zero, the analysis of the circuit is greatly complicated. It follows the lines of the previous section, but as (taking the previous example) when D_1 and D_2 are both conducting they do not now completely short-circuit the signal voltage, there is a residual signal voltage opposing the carrier across D_3. Thus D_3 ceases to conduct before v_c reaches zero. Similarly D_4 does not start to conduct until after v_c has become negative. There are thus four points of discontinuity per half-cycle of carrier instead of two, and thus the equation for the output voltage involves two functions like Equ. (12) instead of one, and with coefficients involving r_f, r_b, R and R_c.

Belevitch[3] analyses this case in detail for one signal tone, and his analysis can be extended to two or more signal components by replacing his switching ('signature') functions by the corresponding $\phi(t)$ functions.* But he shows that the difference between the results obtained and those of the ideal modulator (Section 5) is very small in practical cases—generally less than 1 db—and therefore the full analysis is hardly worth while.

7. Shunt and Series Modulators of Conventional Type

The modulator of Fig. 1 was, of course, a special arrangement designed to fulfil a particular requirement of the analysis. Conventional shunt and series modulators do not use a balanced signal source with the carrier returned to its centre point, and they generally have a finite carrier-generator resistance. When they are used in pairs with a common carrier supply, connected in parallel opposition, or when their carrier generator resistance is very small, they do give much the same intermodulation performance as ring modulators. But when used singly, the effect of the carrier-generator resistance is very large; and it is sometimes possible to find a value which gives a minimum distortion 20 or 30 db below that obtained with zero resistance. This was originally described and explained by Tucker and Jeynes[2] and analysed by Belevitch[3,4] Belevitch's analysis can be extended to two or more signal components as previously discussed for the ring modulator.

8. Use of a Square-Wave Carrier Voltage

Since the non-linear effects described above are due to the interference of the signal voltage with the switching action, which is nominally under the control of a sinusoidal carrier e.m.f., it is clear that these effects can be eliminated by using a square-wave carrier provided the signal voltage never exceeds the carrier.

It is not, of course, always convenient or economical to use a square-wave carrier; this applies particularly in carrier telephony, although there are doubtless many special applications where the extra complication would be well justified. But even when the square-wave is used, there is the residual non-linear distortion due to the non-linearity of the rectifier forward and backward resistances. The amount of this distortion is quite unrelated to that due to switching, but measured results given by Christiansen[6] for ring modulators using germanium rectifiers with a square-wave carrier show distortion of the same order as, or slightly higher than, has been calculated for the switching effect. It is thus evident that the use of a square-wave carrier will not necessarily reduce distortion appreciably, and it is necessary in any particular case to investigate the relative magnitudes of the two effects before the trouble and expense of providing a square-wave carrier is decided upon. Furthermore, it is probable that the calculated distortion for the switching effect is somewhat approximate since practical rectifiers do not switch suddenly at zero voltage as has been assumed.

Christiansen shows that the non-linearity of the rectifier forward resistance can be reduced by the use of a resistor in series with each rectifier. Usually an optimum value can be found giving the minimum level of distortion products relative to the wanted components. Negative feedback can also improve this form of distortion.

9. Equivalent Circuits of Ring Modulator with Non-Linear Response

The ring modulator in its linear mode (i.e., all non-linearity of any kind either ignored or made negligible by the use of low signal/carrier voltage ratio) can be represented by a time-invariable lattice followed by a perfect commutator operating at the carrier frequency. The lattice has one pair of arms with a resistance r_f and the other with a resistance r_b; and it is assumed that the rectifiers from which they are derived switch over at zero applied voltage.

Non-linearity of the rectifier forward and backward resistances can be included in this circuit by making the lattice arms voltage-dependent, as shown in Fig. 4, and we may allow for third-order distortion by saying that $r_f(v)$ and $r_b(v)$ are such that the lattice response can be represented by the cubic

$$v_R = a_1 v_S + a_3 v_S^3 \qquad \text{.. .. (13)}$$

* Note that $2\phi(t)$ corresponds to the signature function.

On substituting the input signal expression for v_S, the Fourier expansion of v_R is obtained; and then all the components of v_R are multiplied by the switching function to obtain the series for the output components. This means that all orders of modulation have the same relationships between their first-order sidetones and their third-order intermodulation and harmonic products; actual levels are proportional to $1/k$ (where k is the order of modulation).

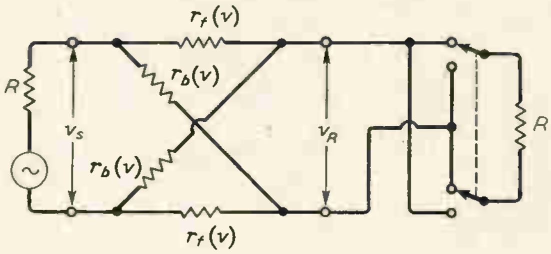

Fig. 4. *Equivalent circuit of ring modulator.*

In the case of third-order non-linear distortion due to the switching-interference effect analysed in this paper, this simple equivalent circuit does not apply, since the relationships within the various orders of modulation are different, and the effect is dependent on the ratio of the signal voltage to the carrier, instead of on its actual magnitude. But since in each order of modulation, the relation between the third-order intermodulation and harmonic terms, and the effect of the third-order distortion on the amplitude of the first-order sidetones, are the same as in Equ. (13), the equivalent circuit of Fig. 4 can be used by means of an artifice. This consists of taking different coefficients in Equ. (13) for each order of modulation and of using x_S in place of v_S. Thus

$$v_R = E_1\,(b_1 x_S + b_3 x_S^3) \quad \ldots \quad \ldots (14)$$

where

$$b_3 = -k^2 b_1/6 \text{ and } x_S = v_S/E_1 \quad \ldots (15)$$

the appropriate value of k being used in calculating the distortion due to the kth order of modulation.

The principle of the equivalent circuit appears to apply to all orders of distortion. Its general application is discussed in Appendix 2.

10. Conclusions

The main conclusions from the work discussed above have been given in the Summary at the head of the paper.

Acknowledgments

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APPENDIX 1

The Complete Output Expression for First-Order Sidetones and Second- and Third-Order Distortion Products from a Single Linear Rectifier with a Large Carrier and Two Small Signal Tones Applied

In the expression below, all terms involving x in higher degree than the third have been neglected. The amplitude of any required frequency component can readily be determined by substituting the appropriate one or two values of r in the relevant output terms.

Output of single linear rectifier with complex signal $E_1\{\cos pt + x_1 \cos(p-q_1)t + x_2 \cos(p-q_2)t\}$ applied

$$= h_1 E_1 \sum_{r=1,3,5\text{ etc.}}^{\infty} (-1)^{(r-1)/2} \text{ of:—}$$

$$\left\{\frac{1}{2r} - x_1^2\left(\frac{1}{4}+\frac{r}{8}\right) - x_2^2\left(\frac{1}{4}+\frac{r}{8}\right)\right\}\cos(r+1)pt$$

$$+\left\{\frac{1}{2r} + x_1^2\left(\frac{1}{4}-\frac{r}{8}\right) + x_2^2\left(\frac{1}{4}-\frac{r}{8}\right)\right\}\cos(r-1)pt$$

$$+\left\{-\frac{x_1}{4} + x_1x_2^2\left(\frac{r}{8}+\frac{r^2}{16}\right) + x_1^3\left(\frac{r}{16}+\frac{r^2}{32}\right)\right\}\cos\{(r+1)p+q_1\}t$$

$$+\left\{-\frac{x_1}{4} + \frac{x_1}{2r} - x_1x_2^2\left(\frac{r}{4}-\frac{r^2}{16}\right) - x_1^3\left(\frac{r}{8}-\frac{r^2}{32}\right)\right\}\cos\{(r-1)p+q_1\}t$$

$$+\left\{x_1\left(\frac{1}{4}+\frac{1}{2r}\right) - x_1x_2^2\left(\frac{r}{4}+\frac{r^2}{16}\right) - x_1^3\left(\frac{r}{8}+\frac{r^2}{32}\right)\right\}\cos\{(r+1)p-q_1\}t$$

$$+\left\{\frac{x_1}{4} + x_1x_2^2\left(\frac{r}{8}-\frac{r^2}{16}\right) + x_1^3\left(\frac{r}{16}-\frac{r^2}{32}\right)\right\}\cos\{(r-1)p-q_1\}t$$

+ terms of the form $\cos\{(r \pm 1)p \pm q_2\}t$ having amplitudes as above but with x_1 replaced by x_2 and vice-versa.

$$+\frac{rx_1^2}{16}\cos\{(r+1)p+2q_1\}t$$

$$+x_1^2\left(\frac{r}{16}-\frac{1}{4}\right)\cos\{(r-1)p+2q_1\}t$$

$$+x_1^2\left(\frac{r}{16}+\frac{1}{4}\right)\cos\{(r+1)p-2q_1\}t$$

$$+\frac{rx_1^2}{16}\cos\{(r-1)p-2q_1\}t$$

+ similar terms of the form $\cos\{(r \pm 1)p \pm 2q_2\}t$ as above

$$-\frac{r^2x_1^3}{96}\cos\{(r+1)p+3q_1\}t$$

$$+\left(\frac{rx_1^3}{16}-\frac{r^2x_1^3}{96}\right)\cos\{(r-1)p+3q_1\}t$$

$$+\left(\frac{rx_1^3}{16}+\frac{r^2x_1^3}{96}\right)\cos\{(r+1)p-3q_1\}t$$

$$+\frac{r^2x_1^3}{96}\cos\{(r-1)p-3q_1\}t$$

+ similar terms of the form $\cos\{(r \pm 1)p \pm 3q_2\}t$

$$+\frac{rx_1x_2}{8}\cos\{(r+1)\,p+q_1+q_2\}t$$
$$+\,x_1x_2\left(\frac{r}{8}-\frac{1}{2}\right)\cos\{(r-1)\,p+q_1+q_2\}t$$
$$+\,x_1x_2\left(\frac{r}{8}+\frac{1}{2}\right)\cos\{(r+1)\,p-q_1-q_2\}t$$
$$+\frac{rx_1x_2}{8}\cos\{(r+1)\,p-q_1-q_2\}t$$
$$-\,x_1x_2\left(\frac{r}{8}+\frac{1}{4}\right)\cos\{(r+1)\,p+q_1-q_2\}t$$
$$+\,x_1x_2\left(-\frac{r}{8}+\frac{1}{4}\right)\cos\{(r-1)\,p+q_1-q_2\}t$$
$$-\,x_1x_2\left(\frac{r}{8}+\frac{1}{4}\right)\cos\{(r+1)\,p-q_1+q_2\}t$$
$$+\,x_1x_2\left(-\frac{r}{8}+\frac{1}{4}\right)\cos\{(r-1)\,p-q_1+q_2\}t$$

$$-\frac{r^2x_1^2x_2}{32}\cos\{(r+1)\,p+2q_1+q_2\}t$$
$$+\,x_1^2x_2\left(\frac{3r}{16}-\frac{r^2}{32}\right)\cos\{(r-1)\,p+2q_1+q_2\}t$$
$$+\,x_1^2x_2\left(\frac{r}{16}+\frac{r^2}{32}\right)\cos\{(r+1)\,p+2q_1-q_2\}t$$
$$+\,x_1^2x_2\left(-\frac{r}{8}+\frac{r^2}{32}\right)\cos\{(r-1)\,p+2q_1-q_2\}t$$
$$-\,x_1^2x_2\left(\frac{r}{8}+\frac{r^2}{32}\right)\cos\{(r+1)\,p-2q_1+q_2\}t$$
$$+\,x_1^2x_2\left(\frac{r}{16}-\frac{r^2}{32}\right)\cos\{(r-1)\,p-2q_1+q_2\}t$$
$$+\,x_1^2x_2\left(\frac{3r}{16}+\frac{r^2}{32}\right)\cos\{(r+1)\,p-2q_1-q_2\}t$$
$$+\frac{r^2x_1^2x_2}{32}\cos\{(r-1)\,p-2q_1-q_2\}t$$

$$-\frac{r^2x_1x_2^2}{32}\cos\{(r+1)\,p+q_1+2q_2\}t$$
$$+\,x_1x_2^2\left(\frac{3r}{16}-\frac{r^2}{32}\right)\cos\{(r-1)\,p+q_1+2q_2\}t$$
$$+\,x_1x_2^2\left(\frac{r}{16}+\frac{r^2}{32}\right)\cos\{(r+1)\,p-q_1+2q_2\}t$$
$$+\,x_1x_2^2\left(-\frac{r}{8}+\frac{r^2}{32}\right)\cos\{(r-1)\,p-q_1+2q_2\}t$$
$$-\,x_1x_2^2\left(\frac{r}{8}+\frac{r^2}{32}\right)\cos\{(r+1)\,p+q_1-2q_2\}t$$
$$+\,x_1x_2^2\left(\frac{r}{16}-\frac{r^2}{32}\right)\cos\{(r-1)\,p+q_1-2q_2\}t$$
$$+\,x_1x_2^2\left(\frac{3r}{16}+\frac{r^2}{32}\right)\cos\{(r+1)\,p-q_1-2q_2\}t$$
$$+\frac{r^2x_1x_2^2}{32}\cos\{(r-1)\,p-q_1-2q_2\}t$$

APPENDIX 2

The General Application of the Equivalent Circuit of Fig. (4) and Section 9

It is shown below that for any order of modulation (k) and for any order of harmonic distortion (m), the equivalent circuit, in which a non-linear lattice precedes a perfect switch (i.e., commutator), holds exactly for switching-interference provided a different set of coefficients is taken for each order of modulation. These coefficients are those of the power series

$$v_R = b_1v_s + b_3v_s^3 + \ldots + b_n\,v_s^n + \ldots \qquad (1)$$

and the different set for each order of modulation is conveniently indicated by a prefix; e.g., ${}_kb_n$ is the coefficient of v_s^n when the kth order of modulation is being considered. The proof is based on the expressions for harmonic distortion in an ideal ring modulator obtained by Belevitch[1], with the correction of a misprint and the addition of correct signs obtained from the paper by Bennett[9] which he quotes. It does not seem possible to extend this proof to intermodulation products without a quite prohibitive amount of labour; but as the circuit is shown in Section 9 to be valid for third-order intermodulation, and is valid for all orders of harmonic distortion, it is not unreasonable to conclude (without further proof) that it is also valid for all orders of intermodulation distortion.

The independent variable is shown in (1) as a voltage, but this is only appropriate if the carrier voltage $E_1 = 1$. In general we should write

$$v_R = E_1\,[b_1x + b_3x^3 + \ldots + b_nx^n + \ldots] \qquad (2)$$

since the non-linearity due to switching interference is dependent on the ratio of signal/carrier voltage, as distinct from the rectifier non-linearity which is, at least mainly, dependent on the signal voltage itself. This is a further point of artificiality about the equivalent circuit, although it gives a means of separating the two effects in practical measurements.

The applied signal is a single tone $xE_1\cos\theta$, where $\theta = (p-q)\,t$ or $2\pi ft$. Let the Fourier series for v_R be

$$E_1\,[c_1\cos\theta + c_3\cos 3\theta + c_5\cos 5\theta + \ldots + c_r\cos r\theta + \ldots] \qquad (3)$$

To relate this to the power series (2), we use the expansion

$$\cos^n\theta = \frac{1}{2^n}\left[\cos n\theta + n\cos(n-2)\,\theta + \frac{n(n-1)}{2!}\cos(n-4)\,\theta + \ldots + \frac{n!}{n!}\cos(n-2n)\,\theta\right] \qquad (4)$$

which is more conveniently written for odd values of n as

$$\cos^n\theta = \frac{1}{2^{n-1}}\left[\cos n\theta + n\cos(n-2)\theta + \frac{n(n-1)}{2!}\cos(n-4)\theta + \ldots\right] \qquad (5)$$

where the series terminates at $\cos\theta$.

From this it is easy to derive the value of c_r in terms of b, thus:

$$c_r = \frac{1}{2^{r-1}}\left[b_rx^r + \frac{1}{2^2}(r+2)\,b_{r+2}x^{r+2} + \frac{1}{2^4}\cdot\frac{(r+4)(r+3)}{2!}\,b_{r+4}x^{r+4} + \ldots\right] \qquad (6)$$

or

$$c_r = \frac{1}{2^{r-1}}\sum_{\gamma=0}^{\infty}\frac{1}{2^{2\gamma}}\cdot\frac{(r+2\gamma)!}{\gamma!\,(r+\gamma)!}\cdot b_{r+2\gamma}\cdot x^{r+2\gamma} \qquad (7)$$

and the values of the b coefficients can be determined by taking special values of the c coefficients, each being that obtained by making the series (3) terminate at it. Thus if the series is made to terminate at the rth term, then from (5) or (6),

$$b_r = 2^{r-1}c_r'/x^r \qquad (8)$$

where c_r' is the special value of c_r.

We are now in a position to test the validity of this for distortion due to switching interference. Belevitch (corrected) shows that the output voltage of an ideal ring modulator, of the kth order of modulation and the mth order of harmonic distortion, is

$$\frac{V_{k,m}}{E_1} = (-1)^{[(k+m)/2]+1} \cdot x^m \frac{\Gamma\left(\frac{k+m-1}{2}\right)}{\pi m!\,\Gamma\left(\frac{k-m+3}{2}\right)} \cdot {}_2F_1 \qquad \ldots \quad (9)$$

where ${}_2F_1$ is the Gaussian hypergeometric function

$$F\left(\frac{k+m-1}{2},\ \frac{m-k-1}{2};\ m+1;\ x^2\right)$$

which can be expanded, if $x<1$, as

$$\begin{aligned} {}_2F_1 = 1 &+ \frac{(k+m-1)(m-k-1)}{4(m+1)} x^2 \\ &+ \frac{(k+m-1)(k+m+1)(m-k-1)(m-k+1)}{16(m+1)(m+2)\,.\,2!} x^4 \\ &+ \ldots \qquad \ldots \quad (10) \end{aligned}$$

Now if we obtain values of b_r as in (8) by using (9) with ${}_2F_1$ put equal to unity, a test of the validity of the equivalent circuit is to see if it then gives the same value of $V_{k,m}$ as obtained from (9) fully expanded. Thus, remembering that the output of the equivalent circuit is v_R followed by a perfect commutator [i.e., the output of kth order of modulation and mth order of harmonic distortion is $2E_1\, c_m\, (-1)^{(k-1)/2}/\pi k$] we say that

$${}_kc_m' = \frac{k}{2}(-1)^{(m-1)/2} \cdot \frac{\Gamma\left(\frac{k+m-1}{2}\right)}{m!\,\Gamma\left(\frac{k-m+3}{2}\right)} \cdot x^m \quad (11)$$

So that*

$${}_kb_m = 2^{m-2}\,(-1)^{(m-1)/2} \frac{\Gamma\left(\frac{k+m-1}{2}\right)}{m!\,\Gamma\left(\frac{k-m+3}{2}\right)} \cdot k \quad (12)$$

Using this, the complete value of ${}_kc_m$ is, from (7).

$${}_kc_m = \frac{k}{2^{m-1}} \sum_{\gamma=0}^{\infty} \frac{1}{2^{2\gamma}} \cdot \frac{(m+2\gamma)!}{\gamma!(m+\gamma)!} \cdot 2^{m-2+2\gamma}\,(-1)^{[(m-1)/2]+\gamma} \cdot \frac{\Gamma\left(\frac{k+m-1+2\gamma}{2}\right)}{(m+2\gamma)!\,\Gamma\left(\frac{k-m+3-2\gamma}{2}\right)} x^{m+2\gamma} \quad \ldots \quad (13)$$

$$= (-1)^{(m-1)/2} \cdot \frac{k}{2} \cdot x^m \sum_{\gamma=0}^{\infty} (-1)^{\gamma} \cdot \frac{\Gamma\left(\frac{k+m-1+2\gamma}{2}\right)}{\Gamma\left(\frac{k-m+3-2\gamma}{2}\right)} \cdot \frac{x^{2\gamma}}{\gamma!\,(m+\gamma)!} \quad \ldots \quad (14)$$

and noting that

$$\frac{\sum_{\gamma=0}^{\infty} \Gamma\left(\frac{m-k-1+2\gamma}{2}\right)}{\Gamma\left(\frac{k-m+3}{2}\right)\Gamma\left(\frac{m-k-1}{2}\right)} = \sum_{\gamma=0}^{\infty} \frac{(-1)^{\gamma}}{\Gamma\left(\frac{k-m+3-2\gamma}{2}\right)} \quad \ldots \quad (15)$$

and introducing a factor $m!\,\Gamma\left(\frac{k+m-1}{2}\right)$ in both numerator and denominator, (14) becomes

$${}_kc_m = \frac{k}{2}(-1)^{(m-1)/2} \cdot x^m \cdot \frac{\Gamma\left(\frac{k+m-1}{2}\right)}{m!\,\Gamma\left(\frac{k-m+3}{2}\right)} \sum_{\gamma=0}^{\infty} \frac{\Gamma\left(\frac{k+m-1+2\gamma}{2}\right)\Gamma\left(\frac{m-k-1+2\gamma}{2}\right) \cdot m!}{\Gamma\left(\frac{k+m-1}{2}\right)\Gamma\left(\frac{m-k-1}{2}\right) \cdot (m+\gamma)!\,\gamma!} x^{2\gamma} \quad \ldots \quad (16)$$

and on multiplying this by $2E_1\,(-1)^{(k-1)/2}/\pi k$ to obtain the actual voltage from the complete equivalent circuit including the commutator, we find that this is exactly the same as $V_{k,m}$ given by (9). [Note that the expression following the summation sign in (16) is the generalized form of the hypergeometric function ${}_2F_1$.]

This proves that the equivalent circuit of Fig. 4, used as described, is entirely self-consistent in respect of harmonic-distortion.

For numerical evaluation, it should be remembered that $\Gamma(\frac{1}{2}) = \sqrt{\pi}$, $\Gamma(-\frac{1}{2}) = -2\sqrt{\pi}$, and that for any variable z,

$$\Gamma(z+1) = z\Gamma(z) = z(z-1)\Gamma(z-1) \text{ etc.}$$

Any intermodulation product can now be rapidly determined, since it is easy to prove from the power series by induction that products, $kF \pm mf_1 \pm nf_2$, of the $(m+n)$th order of intermodulation, have an amplitude given to the first order of accuracy (still assuming $x \leqslant 1$) by

$$\frac{2(-1)^{(k-1)/2} E_1}{\pi k} \cdot {}_kb_{m+n} \cdot \frac{(m+n)!}{2^{m+n-1}\, m!\, n!} x_1^m x_2^n \quad \ldots \quad (17)$$

where ${}_kb_{m+n}$ is determined from Equ. (12).

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* Note that ${}_kb_1$ is unity, whatever the value of k, and that ${}_kb_3 = -k^2/6$. In general, ${}_kb_m$ is a polynomial in k of the $(m-1)$th degree.

AUTOMATIC TUNING FOR PRIMARY RADAR

By S. Ratcliffe, B.Sc.

(Continued from p. 131, May issue)

Complete A.F.C. Loop

For the purposes of the following discussion, radar a.f.c. systems will be divided into four categories:—

(a) Systems in which the change in oscillator frequency is controlled by some non-linear circuit whose behaviour is determined by the sign of the tuning error indicated by the discriminator, and not by its magnitude (once the error exceeds some low value).

(b) Systems in which the change in oscillator frequency is proportional to the magnitude of the tuning error, as well as its sign, and in which the change in oscillator frequency occurs during the interval between transmitter pulses. It will be assumed that the change in oscillator frequency during the transmitter pulse itself is negligible.

(c) Systems similar to those in (b) but in which the frequency correction is actually applied during the pulse, and in which some 'clamp' circuit is then used to hold the tuning fixed at this point until the transmitter next fires.

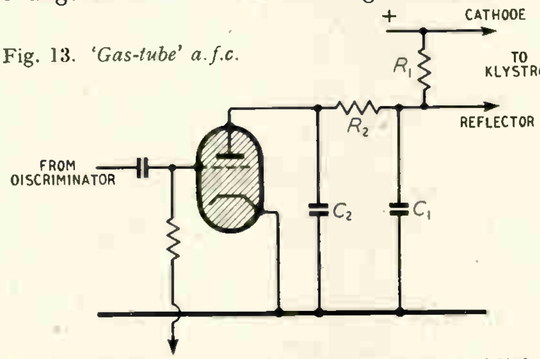
(d) Hybrid systems which use some combination of the three methods listed above.

For the moment, it will be assumed that such tuning errors as may occur do not exceed the limits of the linear portion of the discriminator curve.

The a.f.c. systems mentioned under (a) above, contain some form of circuit having a small number of stable states, usually two. It is arranged that the oscillator frequency drifts in a direction determined by the state of this circuit and at a rate determined by the circuit constants.

A simple example of this type of circuit is the 'gas-tube' a.f.c.,[7] illustrated in Fig. 13. The oscillator frequency is controlled by variation of the reflector voltage. When in the neighbourhood of the desired frequency, the reflector, which is connected to a capacitor C_1 charging through a resistance R_1, is allowed to drift steadily more positive until the desired frequency is passed. A positive output pulse is then received from the discriminator which causes a gas triode to fire. This gas triode then discharges a capacitor C_2 which is connected through R_2 to the larger capacitor C_1, feeding the reflector. Having discharged C_2, the gas triode is extinguished; C_2 then starts to recharge to the reflector potential, lowering the potential across C_1 in the process. This lowering of the reflector potential normally changes the sign of the tuning error, so that the gas triode will remain extinguished until the charging of the reflector network carries the oscillator frequency back through the desired value. Thus, in operation, the gas triode fires at such a rate as to keep the mean frequency of the klystron at the desired value; any change in the transmitter frequency being accompanied by a change in the rate at which the gas triode fires.

Fig. 13. *'Gas-tube' a.f.c.*



A detailed analysis of this circuit is published elsewhere,[7] and for purposes of the present discussion a simplification of the problem will suffice. Suppose that the sensitivity of the discriminator is infinite; that is, that an infinitesimal error will suffice to produce a voltage large compared with the grid base of the gas triode. Suppose also that the reflector voltage drifts up or down at the same rate (p Mc/s per pulse, say) according to the sign of the tuning error. Then the excursions of the oscillator from the desired value will have a peak amplitude of $\pm p$ Mc/s where p is the rate of follow of the a.f.c. system in Mc/s per pulse.

In practice, there will be a threshold level of error below which the signal from the discriminator will be too small to trigger the gas triode, it is impossible for the circuit designer to maintain p at the desired value over the working range of the reflector voltages, and further variations arise from tolerances on the gas-triode characteristics. A typical design may well have a minimum rate of follow (Mc/s per pulse) which is no higher than 0·05 of the maximum tuning error, so that an a.f.c. of this type must be very slow or inaccurate or both.

Another application of this type of control is the 'bang-bang' control system used with thermally-tuned oscillators. Its performance is not appreciably better than the figure quoted above for the 'gas-tube' system.

No general analysis of systems of this type will be attempted here although McColl[11] has indicated one possible approach. In general, an *ad hoc* argument will suffice to reveal the salient characteristics of any given system and, in fact, the ease with which the performance can be analysed is one of the major attractions of this type of circuit.

One further example[12] of a control system of this type may be cited. This uses mechanical tuning of the oscillator. The tuning shaft is coupled through a differential gear to two 'stepping' motors, similar to those used to drive a uniselector, the two mechanisms being arranged to drive the tuner shaft in opposite directions. The two motor windings are controlled by relays in the anode circuits of a pair of valves fed in push-pull from the discriminator output. If the tuning error, in either direction, exceeds some threshold value, one of the relays will close, and the tuning shaft will move by one step in the required direction. The process will then repeat until the error has fallen below the threshold value. This control system can thus be regarded as having three stable states, corresponding to positive, negative, and negligible error. The stability condition for this system is that a single step of the motor must never cause the tuning to change by twice the threshold level, otherwise the desired central state will not be achieved. The rate of follow of this system is then fixed by the speed of the motor in steps per second.

It will be seen that there are inherent limitations on the speed and locking accuracy of control systems of this type, and a better performance is possible if full use is made of the information available from the discriminator as to the relative magnitude of the tuning error.

The control systems most commonly used in radar a.f.c. fall into category (b) of the classification given above. The pulses from the discriminator are 'stretched' in some way, and the resulting error voltage, possibly after some further 'smoothing', controls the oscillator frequency. In a reflector a.f.c. system, for example, the reflector is often connected to the anode of a Miller integrator, into the grid of which is fed a current proportional to the error voltage. The frequency, therefore, drifts in a suitable direction at a rate proportional to the tuning error existing at the time; or, more correctly, at a rate proportional to the error voltage output from the discriminator or smoothing circuit.

The above account of the operation could, perhaps with some change in nomenclature, equally well have referred to any of the other servo-mechanisms which are sometimes associated with a radar, a lock-following strobe for example. It is not, however, commonly necessary or possible for such systems to move their output shaft, or its equivalent, any significant distance between one pulse and the next, so that servos are commonly designed by negative-feedback theory; i.e., as though error information were continuously available from the discriminator, the 'undesirable' modulation of this voltage at the p.r.f. being removed by suitable filtering.[13]

If we are to attempt an a.f.c. system which is to deal with scanner pulling, we require an approach to the design problem which takes account of the limitation on the rate at which information is available from the discriminator, since the transmitter frequency may change appreciably between one pulse and the next. In a reflector a.f.c. system, for example, this may be the only important limitation on the response speed of the system. A system limited in this way by the supply of error information forms what McColl[11] terms a 'sampling' servo-mechanism, and its analysis is more complicated than that of one with continuous feedback.

If the tuning error in an a.f.c. system is measured relative to the cross-over point on the discriminator characteristic, there are two types of error that can arise. The first of these is a static error that will persist even if the transmitter frequency remains constant for a long period. When the steady-state is reached, the fact that the error voltage must be derived from pulses at finite intervals of time is not important, and the steady-state error e_s will be given by normal negative-feedback theory as

$$e_s = E/(1 + G) \quad \quad (3)$$

where E is the error that would exist in the absence of a.f.c. feedback, and G can be defined as the ratio of the change in local-oscillator frequency, due to the operation of the a.f.c., to the tuning error which produces the a.f.c. effect. There is usually no difficulty in providing sufficient gain to reduce this static error to negligible proportions, and it will be neglected in discussing the more serious dynamic errors which arise when the transmitter frequency is subject to variation.

The problem of determining the dynamic response and stability conditions for a sampling servo is discussed in some detail in the Appendix to this article. A brief non-mathematical examination of the problem is given here.

The a.f.c. is given a piece of information about the tuning error each time the transmitter fires, and must arrange to take some suitable action after each pulse with a view to reducing the error when the transmitter next fires. In other words, the problem is to extrapolate from some given data. The a.f.c. will base its extrapolation on the most recent pulse and possibly on some earlier ones, the number of pulses used for this purpose being fixed by the number of independent elements (time-lags) in the feedback loop which are capable of 'remembering' a voltage from one pulse to the next.

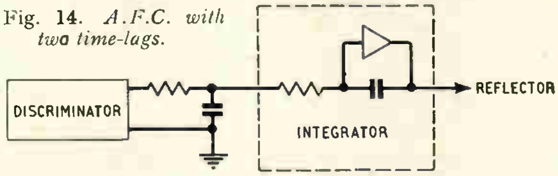
This statement perhaps requires some justification. Consider, for example, the circuit shown in Fig. 14 which possesses two 'time-lags'—both capacitance-resistance networks. The voltage across the first capacitor will depend on all the error voltages which have ever appeared across the discriminator output, so that it may appear that the a.f.c. will base its behaviour at any instant on all the previous error pulses and not merely on the two most recent. However, the response at any instant is completely determined by the voltages on the two capacitors, and these two voltages can be determined from a knowledge of the last two errors. Thus, although previous errors may have been responsible for the building-up of charges on these two capacitors, the behaviour of the a.f.c. is completely determined if the last two errors are known. Hence the statement that the behaviour depends only on these two errors.

If the a.f.c. is to follow a smooth variation in the transmitter frequency which takes place over a period of many pulses, the more of these pulses taken into account in extrapolating, the greater will be the possible accuracy with which the extrapolation can be performed. It has been shown, however, when discussing scanner pulling, that in practice the a.f.c. is required to deal with a more or less random fluctuation in frequency from one pulse to the next, and under these circumstances the a.f.c. will be slowed down by any time-lags which retain information about pulses other than the most recent, just when the most rapid reaction is required.

If the a.f.c. is to extrapolate with reasonable success, it is essential that it should have an appropriate rate of response to an error. It is shown in the Appendix that a single time-lag extrapolating a.f.c., given optimum design, has a response time-constant which may be as great as $R/2$ pulse intervals, where R is the ratio of the maximum to the minimum possible gain round the a.f.c. loop. Thus a good transient response can only be guaranteed if R is kept low. Similar results hold for systems having more than one time-lag. Although it is possible to devise circuits which to some extent compensate for tolerances on gain when following a smoothly-varying transmitter frequency, no circuit has been devised which has the same effect when faced with a rapid and random variation in frequency.

Pending detailed discussion of the problem of reducing the tolerances contributing to R, it can be taken that a value of 6–8 represents the minimum that is obtainable in any type of control system.

Fig. 14. *A.F.C. with two time-lags.*



The third type of a.f.c. system defined above was invented by C. Baron, of R.R.E., in an attempt to avoid the fundamental limitations of the extrapolating systems. In this system, often termed an 'instantaneous' a.f.c., the feedback loop is given sufficiently short time-constants to enable the device to operate during the transmitter pulse as a true negative-feedback system, and a 'clamp' circuit is then used to hold the reflector potential constant during the interval between pulses. An example of a practical circuit of this type is given later, and the Appendix discusses the theory of this system. Except possibly for very short pulses, the design does not present any major difficulty; analysis shows that, although the performance is still determined by the tolerances, it is less sensitive to them than is the extrapolating a.f.c. The classification of linear a.f.c. systems as either 'extrapolating' or 'instantaneous' is convenient, but the boundary line is not necessarily clear. Given a sufficiently wide pulse, an a.f.c. designed as an extrapolating system may, in practice, make an appreciable correction to the frequency during the pulse, so that the system becomes some sort of hybrid.

Throughout the preceding discussion it has been assumed that the tuning error lies within the peaks of the discriminator curve. An a.f.c. is capable of 'holding-in' so long as the tuning error, after the application of the correction, lies within the discriminator peaks. In the absence of the a.f.c. correction, the tuning error may be more than 10 times as great as the discriminator peak separation, and for an uncorrected tuning error of this magnitude the discriminator output will have fallen virtually to zero, so that the a.f.c. will be unable to deal with this error once it has developed. It is, therefore, necessary to distinguish between the a.f.c. 'hold-in' range, which is usually determined by the electronic-tuning range of the klystron, and the 'pull-in' range which gives the limits of initial tuning error with which the a.f.c. can cope. This latter is usually about twice the discriminator peak separation, though it may be greater if the discriminator has been designed to have large skirts to its response curve. Unless some additional steps are taken, therefore, the a.f.c. cannot be relied upon to deal with frequency drifts in excess of the pull-in range. It is, therefore, customary to provide some means of searching over the electronic-tuning range of the oscillator for the wanted signal, both on switching-on and on any later occasion, this search usually being stopped automatically when signals are received from the discriminator.

Earlier reflector a.f.c. technique provided preset adjustments of the mechanical tuning and of the potential applied to the reflector in the absence of a voltage from the discriminator. The search circuit has the added advantage that it reduces the accuracy with which these presets need be adjusted. It must be pointed out that it is not necessary to use electronic tuning for this search, since by means of the mechanical tuner one can search a much wider band of frequencies, if this should be necessary. For the moment, it will be assumed that both the search and the a.f.c. loop use electronic tuning, discussion of the other possibilities being left till later.

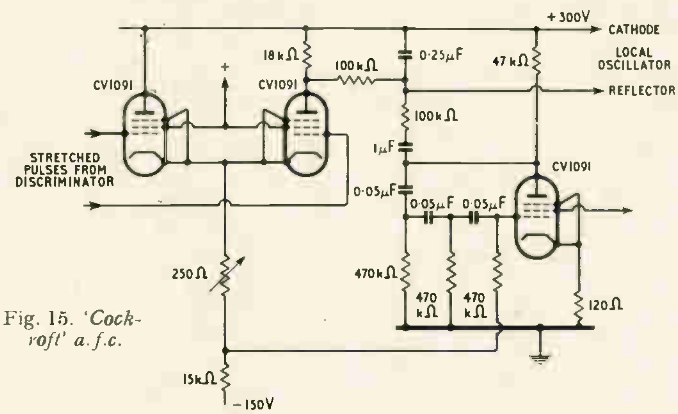


Fig. 15. *'Cockroft' a.f.c.*

The search of the electronic-tuning range is usually performed by applying a 2–6-c/s sawtooth or sine wave to the reflector. Two types of circuit have been used for this purpose.

The first of these is exemplified by the 'Cockroft' circuit[14] shown in outline in Fig. 15. This searches by applying a fraction of the output of a 4-c/s phase-shift oscillator to the reflector. The two discriminator diodes develop negative outputs, and the stretched pulses are applied to the grids of a long-tailed pair. The difference voltage at one of the anodes is then used to control the reflector, the frequency response of the detector outputs and of the reflector voltage being shown in Fig. 9. In addition to the difference voltage available at this anode, a signal also appears across the common cathode load. This signal being proportional to the sum of the two inputs, the frequency response to this point is similar to that of a normal band-pass coupling. Thus a negative signal appears at this cathode whenever signals are received down the a.f.c. chain. A d.c. coupling connects this cathode to the grid of the phase-shift oscillator, so that the latter is biased-off when i.f. signals are received down the a.f.c. amplifier.

The sequence of operations on switching on the equipment is as follows: before the transmitter commences firing, the phase-shift oscillator will be sweeping the klystron frequency through its electronic-tuning range. When the transmitter begins to fire, and the oscillator frequency approaches the correct tuning point, a tuning-correction voltage will be received from the anodes of the long-tailed pair, while a negative voltage appears at the cathode. This latter voltage will bias-off the phase-shift oscillator while the reflector will be maintained at approximately the correct potential by the error voltage at the anode. The phase-shift oscillator, when biased-off, will tend to apply a voltage surge to the reflector, and if the klystron is to stay on frequency, the main a.f.c. loop must be able to cancel this surge. This can be achieved if the frequency of the phase-shift oscillator is sufficiently low, so that the anode will not rise too quickly to h.t. potential.

The second circuit avoids the complicated situation which arises when the Cockroft a.f.c. circuit approaches the correct frequency during the search. Examples of this circuit are to be found in a number of systems developed both in this country and in the U.S.A. (In the "Microwave Receivers" volume of the Radiation Laboratory series an arrangement of this type is repeatedly but quite inaccurately referred to as the 'diode phantastron' circuit.) In the usual form of this circuit, shown in Fig. 16, the reflector is coupled to the anode of a Miller integrator which forms part of the normal a.f.c. loop. The bias conditions are so adjusted that the anode of the integrator 'bottoms' if no signal is received from the discriminator and pulse-stretching circuits, and transitron feedback is provided between screen and suppressor of this valve. Thus, in the absence of a signal from the discriminator, this integrator will act as a free-running sawtooth generator. The transitron operating conditions are so chosen that the amplitude of the sawtooth is somewhat greater than that required to cover the electronic-tuning range of the klystron.

When the transmitter is switched on, a signal will be received at some point in the search cycle. When a negative output is received from the pulse-stretcher, the Miller run-down of the anode will be slowed down and finally arrested at a suitable anode potential. The valve will thereafter function as an integrator in the normal manner. A residual tuning error must exist in order to hold the reflector at a suitable potential but, by making the steady-state gain sufficiently high, this error can be kept down to negligible proportions.

The transitron circuit is not highly satisfactory in a Service equipment, since it makes use of various valve characteristics which are not controlled. The difficulties are increased when it is attempted to maintain only a small amplitude of oscillation about a mean potential suitable for a particular klystron. It is, therefore, usual to provide two preset adjustments of the transitron operating conditions. It is, however, possible to replace this somewhat crude transitron oscillator by a more 'designable' circuit in which the limits of the anode excursion are more clearly defined.

In many equipments the frequency range over which the search is conducted is comparable with the i.f. It is then found that signals are received from the a.f.c. mixer, not only when the oscillator is approximately on tune, but also when the difference between the l.o. and transmitter frequency is any submultiple of the desired i.f. This effect is due to harmonics of the difference frequency generated in the mixer. The amplitude of these harmonics falls off as their order increases, so that it is unusual for trouble to be experienced from harmonics other than the second and third.

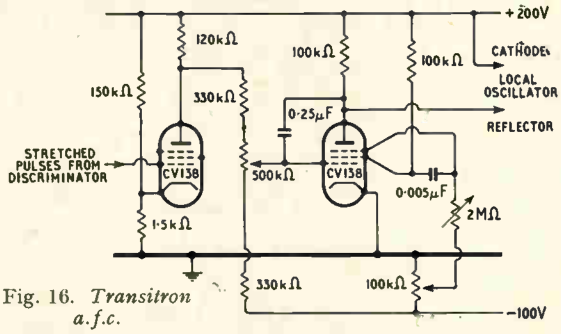


Fig. 16. *Transitron a.f.c.*

It is essential to arrange the search so that the a.f.c. locks only to the wanted signal, and not to the weaker harmonic responses. Even when the input levels are correctly chosen, the amplitude of these harmonics differs from the wanted signal by only 15–20 db (perhaps 25 db if a balanced a.f.c. mixer is used), so that if the relative amplitude of the signals is the only means of distinguishing between them, the margin of safety is quite small. For this reason it is not unusual to provide a preset adjustment of the signal level at which locking is to occur. Modern practice, however, is to make use of the improved tolerances on mixer crystals and better waveguide techniques to remove the necessity for this adjustment.

Reference has been made to the effect on the performance of the tolerances on the gain round the feedback loop of an extrapolating a.f.c. system. A list is given below of the more significant of the many factors contributing to variations in the gain. The importance of some of these factors depends on whether the l.o. or the transmitter drive to the a.f.c. mixer is the greater of the two, since, to a first approximation, the output will be directly proportional to the smaller of the two signals and independent of the larger.

TABLE 2

Typical tolerances on:	*Contribution to R*
(a) Transmitter power reaching a.f.c. mixer	1·8
(b) Local-oscillator drive	
(c) Gain of a.f.c. circuits	2
(d) Slope of electronic-tuning characteristic (at centre of mode)	1·5
(e) Variation in slope over mode	2·5

The tolerances on l.o. and transmitter drive are likely to be much the same but, other things being equal, the final value of R (ratio maximum/minimum possible loop gain) will be lower if the local-oscillator drive is the smaller of the two. If the l.o. is providing the signal controlling the mixer output, the fall in the l.o. output as the side of the mode is approached will tend to compensate for the increase in slope of the electronic tuning characteristic of the klystron.

If the transmitter drive is taken as the lesser of the two mixer inputs, a probable final value of R is 14 while if the l.o. is the smaller, the reduction in gain at the edges of the mode will reduce R by about 1·3; i.e., to 11. In a wideband system these figures can readily be exceeded because of the necessarily greater variation in the drive to the mixers and in the loading of the oscillator cavity, the Q of which determines the slope of the electronic-tuning characteristic.

It is possible to provide an a.g.c. system round the a.f.c. loop but this a.g.c. cannot take account of the slope of the electronic-tuning characteristic, and its application to the above system is not justifiable.

Wide-Range A.F.C.

The frequency range over which a reflector a.f.c. can hold in is limited by the electronic-tuning range of the klystron. In some higher-frequency equipments this range is barely adequate to deal with probable variations in the frequency of the magnetron and the klystron.

There is a limit to the electronic-tuning range that can be achieved before the efficiency of the klystron falls too low and its noise output becomes too high, and it is, therefore, often desirable to adopt some technique which does not confine the operation of the a.f.c. system to the limits of the electronic-tuning range, especially as war-time experience showed the difficulty of producing a reliable electronic a.f.c. for airborne radars. There is, therefore, much to be said for applying the policy, already adopted in the waveguide system, of designing circuits to operate over the widest possible frequency range without adjustment.[15]

Various attempts have therefore been made at producing an a.f.c. system having the wider tuning range obtainable by mechanical means. A purely mechanical a.f.c. is not capable of dealing with scanner pulling, and similar short-term fluctuations in transmitter frequency, and most proposals involve some combination of electronic and mechanical tuning.

One arrangement consists essentially of a tuning motor which searches a wide band of frequencies until the wanted signal is located. The motor is then switched off, a reflector a.f.c. is allowed to take over, and is expected to deal with all subsequent variations. This system cannot be said to have overcome in any very satisfactory manner the limitations of a purely electronic a.f.c., and it is now more usual to use both forms

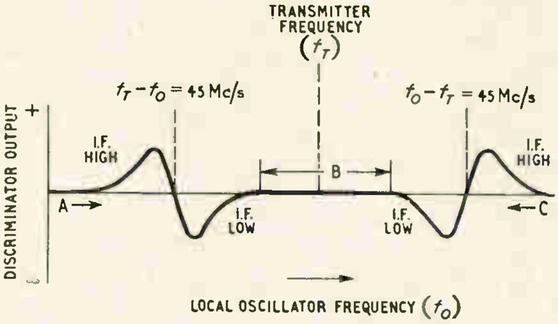


Fig. 17. *Output from i.f. discriminator.*

of correction simultaneously. If an attempt is made to connect two normal servos in parallel, both using the discriminator output to control the frequency, it will be found that the design problems become insuperable, since each servo receives an input only by virtue of the defects of the other, and it becomes impossible to determine either the steady-state position of the mechanical tuner or the condition that will ever be reached.

This problem can be avoided if some technique is adopted which divides the responsibility between the two servos in a less ambiguous manner. The first suggestion on these lines appears to have come from G. H. Nibbe, of Radiation Laboratories, M.I.T., who proposed to attempt to supplement the thermal tuning of a local oscillator with the electronic-tuning characteristic of the reflector.

Nibbe suggested that the design problem could be tackled by employing a reflector a.f.c. to hold the frequency at the desired value while the thermal tuner was controlled by a circuit which attempted to maintain the local-oscillator power output at its maximum value. Fig. 11 shows that if the mechanical tuning is varied while the frequency is kept constant by means of the reflector, the power output will fall off as the reflector voltage departs from the value corresponding to the centre of the electronic-tuning range. Thus the combined action of these two control systems will maintain the oscillator on tune with both the mean reflector potential and the mechanical tuning near their optimum positions. This technique has been successfully adopted for the control of a motor-tuned oscillator.[16]

In a wide-range a.f.c., many of the problems discussed earlier are accentuated and some new ones are introduced. For example, the tolerances on the local-oscillator output and on the slope of the electronic-tuning characteristic must both increase with the tuning range. This means a reduction in the response speed of the system, and the increased tolerances on the l.o. output may add to the difficulties with harmonics. In practice, however, it is probable that the theoretical loss in performance of a wide-range system is considerably less than the losses in performance of a narrow-band system due to errors in the adjustment of the preset controls. All that has happened, on changing over to a wide-range system, is that much of the responsibility for the performance of the equipment has been shifted back from the maintenance mechanic or equipment user to the designer.

Some new problems arise when designing a wide-range search system. First, there are, as shown in Fig. 17, two l.o. frequencies which give rise to a beat at the desired i.f., but the sense of the feedback voltage obtained from a conventional i.f. discriminator can only be correct for one of them, thus complicating the search problem. Another difficulty is that the process of searching the entire tuning range may take considerably longer than in a narrow-band system, times of the order of 5–10 secs being not unusual. This delay is not important on first switching on but, particularly in high-power systems, there may be momentary interruptions in the transmitter output due to arcing, etc. If the search oscillator begins to run when these interruptions occur, and the system takes 10 secs or so to locate the transmitter frequency after the transmitter output is restored, the effect of these disturbances on the performance of the equipment will be greatly exaggerated. It is therefore necessary to design the control circuits so as to minimize the time wasted by such interruptions.

To tune a klystron over a wide range of frequencies, it is necessary to vary the physical dimensions of the cavity by some means. The first determined attempts at solving the wide-range a.f.c. problems were made in the U.S.A. and resulted in the development of a number of thermally-tuned local oscillators and some a.f.c. circuits for use with them. Improved techniques have now completely displaced thermally-tuned valves and the problems of thermal tuning will not be discussed further.

Reference has already been made to the difficulties of producing a purely mechanical servo-mechanism to provide the a.f.c. correction. In the present state of the art, it is not easy to see how these difficulties can be overcome to make possible a good a.f.c. performance. Work has therefore been devoted mainly to the possibility of using a mechanical tuner to conduct the initial search, and to handle long-term drifts, while a reflector a.f.c. system deals with scanner pulling, power supply 'flutter', microphony, and any other short-term fluctuations.

Two possible techniques have been described briefly above. The second of these uses the discriminator output to operate a normal reflector a.f.c. system (either 'instantaneous' or 'extrapolating'), while a separate circuit adjusts the mechanical tuning to keep the power output of the local oscillator (measured in terms of the direct current in a crystal) at or near its maximum value. This ensures that the mechanical tuning is set to approximately the correct mean frequency; i.e., that the mean operating point is in the centre of the 'reflector mode'.

Since the power output of the oscillator will vary with frequency it is desirable to maintain the crystal current at its maximum value rather than at some standard figure. This can be done by allowing the mechanical tuning to 'hunt' continuously about the optimum, a circuit being used to reverse the motor direction whenever the crystal current begins to fall perceptibly. In the absence of a signal from the discriminator, the motor is arranged to sweep to and fro over the tuning range until the desired frequency is located.

It will be appreciated that in normal operation the local-oscillator frequency is not varying to any significant extent during this motion of the mechanical tuner, since the reflector a.f.c. is simultaneously adjusting the reflector voltage so as to maintain the i.f. at the desired value. The change in crystal current to which the motor control reacts is, in fact, directly due to the change in reflector voltage rather than to the change in mechanical tuning.

Readers who are aware of the considerable complications which arise in circuits for use with the thermally-tuned valves will notice that there is little difficulty in searching with this reflector-cum-mechanical system, due to the separation of the two functions of searching (performed by

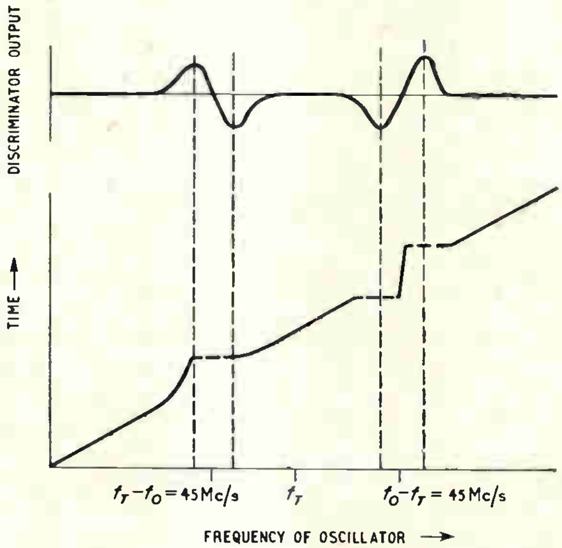


Fig. 18. *Variation of oscillator frequency during search.*

the motor) and locking (performed by the reflector system). During the search period the crystal-current circuit is rendered inoperative, and the motor is made to run continuously through the tuning range. If the end of the range is reached the direction of travel is reversed, either by means of a limit switch or a cam. Fig. 18 shows the manner in which the frequency of the oscillator will vary as the motor searches over the band. When the i.f. differs widely from the correct value, no output will be received from the discriminator controlling the reflector voltage, and the oscillator frequency will be substantially the resonant frequency of the cavity. If the 'wrong' channel is reached first, the reflector voltage will begin to vary so as to oppose the change in frequency. When the limit of its control is reached, the frequency will jump discontinuously to the other side of the discriminator curve, when the cavity will again take complete control of the frequency. As the 'right' channel is reached, the frequency of the oscillator will be pulled-in to the value giving the desired i.f., and it will be held at this point over a range of settings of the mechanical tuner. The search is now at an end, and the motor control circuit may be allowed to take control of the motion of the tuning. This transition may be initiated by a valve which is biased-off when signals are received in both halves of the discriminator, since this means that the i.f. is in the cross-over region, and this only occurs when the wanted channel has been reached. Thus it is necessary to provide a circuit similar to that used to stop the phase-shift search oscillator in the Cockroft circuit of Fig. 15. It is necessary to provide sufficient loop gain to ensure that in the neighbourhood of the 'wrong' channel, the limit of the klystron's electronic-tuning range is reached before the frequency comes sufficiently near to the peaks of the discriminator to operate the search-stopping mechanism.

The absence of any wrong-channel problem makes possible the use of much simpler circuits than are required to obtain even an indifferent performance from a thermally-tuned valve. With a typical motor control circuit, and an 'extrapolating' a.f.c., seven valves are required after the discriminator. This is a saving of two valves over the Nibbe-Durand circuit[7] which gives a greatly inferior a.f.c. performance.

To maintain oscillations over a wide tuning range, it is necessary to gang a potentiometer to the tuning mechanism, in order to maintain the correct mean reflector voltage as the frequency is varied.

The crucial point in the design of the mechanical system and the motor control circuit is the choice of the speed at which the mechanical tuner is driven by the motor. There are four main factors:—

(a) The life of the mechanism will decrease with increasing motor speed.

(b) The control circuit is operated by the derivative of the crystal current. The faster the drive the lower the gain required in the control circuit, and therefore the less the danger of trouble from microphony, hum on the supply lines, etc.

(c) The rate of mechanical tuning in Mc/s per pulse should not be too high, otherwise the reflector a.f.c. may lag noticeably with a subsequent loss in tuning accuracy. This limitation is usually only serious at low p.r.fs.

The drive may be either continuous, or discontinuous. The advantage of a discontinuous drive is that a high rate of change of current is produced for a small amount of wear in the tuning mechanism, but there does not appear to exist a satisfactory motor which can produce a discontinuous motion in both directions. The discontinuous drive may, of course, be obtained by other means, for example, by switching the power supplied to the motor, or by using a Geneva mechanism. The former involves some circuit complications, and it is very doubtful whether a motor of normal design would not give a longer life when running continuously. The Geneva mechanism involves some mechanical complication, but there is plenty of evidence from their use in film projection, etc., that a design is possible which has a perfectly satisfactory life.

In the design of the motor control circuits, it is necessary to take account of the shape of the waveform which the crystal-current amplifier is to handle during normal operation of the circuit. For a discontinuous drive and a low drive/rest ratio the nature of the waveform is fairly obvious. The waveform is not so obvious when a continuous drive is used, since it is a function of the acceleration characteristics of the motor. A useful empirical approach is to construct the drive mechanism and to complete the reflector a.f.c. loop, but to arrange to reverse the motor by a key which is operated by hand whenever the crystal current is observed to drop to the limit within which it is intended that the circuit should maintain the current. The frequency with which the key is depressed should be measured and the amplifier is designed on the assumption that the input is a sine wave of this frequency. It is advisable to cut off the frequency response of the amplifier at 2 to 4 times this frequency, but the low-frequency response of the amplifier should be maintained down to at least one-tenth of the fundamental frequency, so that the only short time-constant is that feeding the trigger circuit at the end of the amplifier. This differentiating time-constant is probably best made roughly equal to a fifth of the time interval needed for a complete cycle of the motor operation.

Two other circuits[17,18] which have been tried under laboratory conditions use the discriminator output to control both the motor and the reflector a.f.c. The reflector a.f.c. system shown in Fig. 19 has a threshold tuning error, below which the error voltage is too small to overcome the bias on either of the diodes. Thus the reflector a.f.c. will reduce the tuning error only to this threshold value, which is not greater than about 10% of the receiver bandwidth, so that the presence of the residual error need not affect the performance of the equipment. The sign of the residual error is the same as that of the tuning error which the reflector a.f.c. is being called on to correct, so that if a tuning motor is controlled by a 'bang-bang' circuit actuated by the error voltage from the discriminator, this residual error can provide sufficient information to enable the motor to take over from the reflector a.f.c. as much of the task of keeping on tune as mechanical limitations permit.

In the second experimental system the reflector correction is applied through an a.c. coupling, so that, in the steady-state condition, the entire correction is applied by the mechanical tuner, the reflector correction being reserved to deal with transients.

The transient response, for small errors, is identical with that of a normal extrapolating a.f.c. provided that the reflector coupling time-constant is long compared with the response time of the a.f.c. loop. Since this latter should not exceed about five pulses, whereas the coupling time-constant should be greater than the response time of the mechanical servo system, this condition is easily met.

A formal analysis of the stability conditions of a complete a.f.c. of this type, taking due account of the modification of the behaviour of the a.f.c. due to the finite p.r.f., offers no theoretical difficulty but is extremely laborious. It has already been shown, however, that the reflector a.f.c. will have a response time very much less than that of the mechanical system, so that the analysis may be simplified without much loss of accuracy, by discussing the response of the mechanical system on the assumption that the response of the reflector a.f.c. is instantaneous. The virtual impossibility of allowing for the effect of 'stiction' and backlash sets a limit to value of these calculations.

These last two circuits have the advantage that they can be used with a local oscillator having an irregular or discontinuous 'mode'. A disadvantage is that the reflector-voltage tracking circuit must work very well if the local-oscillator drive is not to vary considerably over the mechanical-tuning range due to variation of the position of the operating point on the reflector 'mode'. The 'crystal-current' circuit will always settle down with the operating point in the centre of the mode; the reflector-voltage tracking need only be good enough to maintain some sort of oscillation during the search. Further research is also necessary to determine whether the 'bang-bang' circuit used with the second scheme is capable of maintaining the tuner in the correct position in the presence of appreciable scanner pulling, which will greatly confuse the information reaching the trigger circuit.

Present experience, admittedly somewhat limited, suggests that the 'crystal-current' scheme is capable of giving perfectly satisfactory results with any oscillator having a reasonable 'mode'. Freedom from discontinuities over the electronic-tuning range is also necessary for the satisfactory functioning of the reflector a.f.c. This limitation is therefore inherent in any a.f.c., and the use of this type of system does little more than emphasize it. It is believed that this circuit will offer a satisfactory solution to almost any wide-range a.f.c. problem. The other two schemes have not been developed to a state where they can be recommended with any confidence, but they are mentioned here for the sake of completeness.

Some Practical A.F.C. Circuits

Previous writers[12] have commented on the remarkable diversity of circuits which have been used for radar a.f.c. No attempt will be made here to catalogue these circuits or discuss their genesis. A few outlines will be given of some recent designs which avoid the more serious defects of earlier circuits, illustrate minor design problems, and which may help readers to judge the amount of complexity involved in the techniques under discussion.

Fig. 19 shows a simple electronic a.f.c. with a 'transitron' type search. In the absence of a signal from the discriminator, V_4 and V_5, which replace the conventional transitron, will apply a saw-tooth sweep to the reflector. The detailed mechanism is as follows: V_4 is so biased that, in the absence of an input signal, its anode will run down till V_5 begins to conduct. The screen current of V_4 will then begin to rise, and the feedback to the grid of V_4 through the blocking-oscillator transformer T_2 and V_5 will cause V_4 to swing into heavy screen and grid current. The screen of V_4 will fall nearly to earth potential, and the cathode of V_5 will

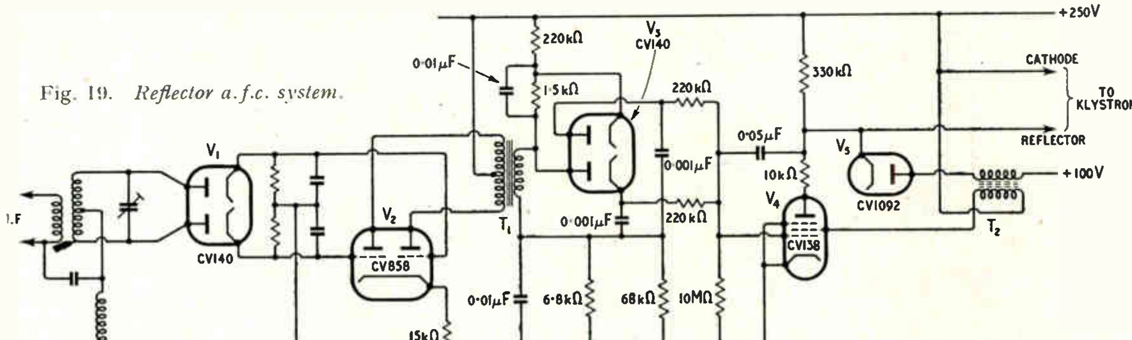


Fig. 19. *Reflector a.f.c. system.*

therefore rise by an amount nearly equal to the h.t. voltage multiplied by the turns ratio of T_2. This is chosen to suit the sweep voltage required by the klystron. Soon after the peak of the swing has been reached, V_5 will cut off and the heavy negative-feedback through the Miller capacitor will take over. V_4 and V_5 thus generate a saw-tooth sweep for the reflector.

When signals are received, the discriminator diodes V_1 will charge their load capacitors during the transmitter pulses. The resulting video signals, somewhat lengthened by the discharge time-constant of the diode loads, are applied to the grids of a long-tailed pair, V_2. The current pulses in the two anodes are subtracted from each other in the pulse transformer T_1, which has a 2 : 1 step down from each anode so that a peak current of about 20 mA is available from the secondary. The voltage gain of V_2 is very low, because of the low anode-load impedance, so that the use of triodes in this stage presents no difficulty.

To complete the process of converting the error pulses to d.c. V_3 is provided. The twin diodes are connected back-to-back with a volt or so of bias on each. In the absence of an error pulse of amplitude exceeding this bias, both diodes remain non-conducting, and the anode of V_4 can only drift very slowly. In the presence of an appreciable tuning error, one section of V_3 will conduct, charging its load capacitor to the peak output voltage of T_2. This voltage will be integrated by V_4, producing a suitable change in the reflector voltage of the klystron and hence a compensating shift in its frequency. The discharge time constants of the load capacitances and resistances of V_3 should not exceed the time interval between two successive pulses, so that the complete loop will behave substantially as the 'single time-lag' system discussed in the Appendix, which also explains the factors influencing the choice of the speed of response of the feedback loop to a tuning error. In practice this speed, when chosen, is adjusted by the choice of the Miller feedback capacitance round V_4.

This arrangement differs in a number of important particulars from many published circuits of this type. First, there are no inter-stage d.c. couplings, so that it is not necessary to provide presets to take up component tolerances. In addition, the blocking-oscillator type of feedback avoids the drawbacks of the transitron, and the symmetry of the video portion of the discriminator eliminates many of the troubles which otherwise arise from the capacitance unbalance in the diode loads. The tolerances on the reflector characteristics of the klystron, a CV2116, are such that no adjustment of reflector voltage is needed even when the klystron is replaced.

Fig. 20 shows the circuit used by C. Baron in his pioneer work on the 'instantaneous' a.f.c. Three i.f. stages V_1, V_2 and V_3 feed a Foster-Seeley

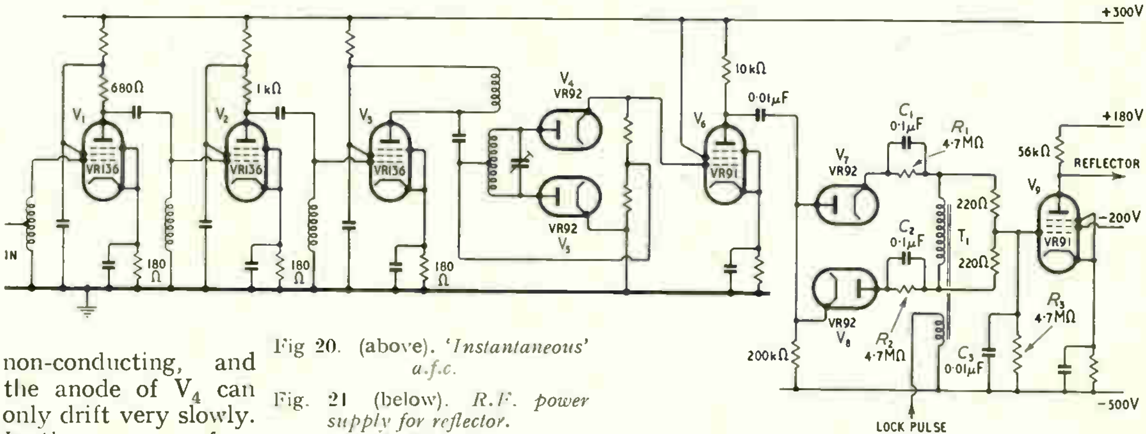
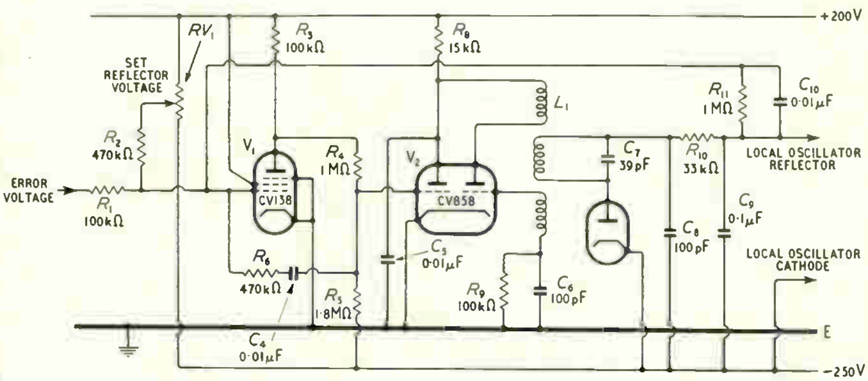


Fig 20. (above). '*Instantaneous' a.f.c.*

Fig. 21 (below). *R.F. power supply for reflector.*



discriminator having a peak separation of 22 Mc/s. The discriminator output is amplified by the video stage V_6 and fed into the 'memory' capacitor C_3 through a 'clamp' circuit consisting of a pair of diodes V_7 and V_8 which are switched on when a 'lock' pulse is applied to the transformer. The current taken by the two diodes produces a more or less steady bias across the capacitors C_1 and C_2, so that the diodes will be non-conducting in the intervals between locking pulses. Thus C_3 will remain charged when the locking pulse ends. (The time constants C_1R_1, C_2R_2 and C_3R_3 are very long compared with the reciprocal of the p.r.f.)

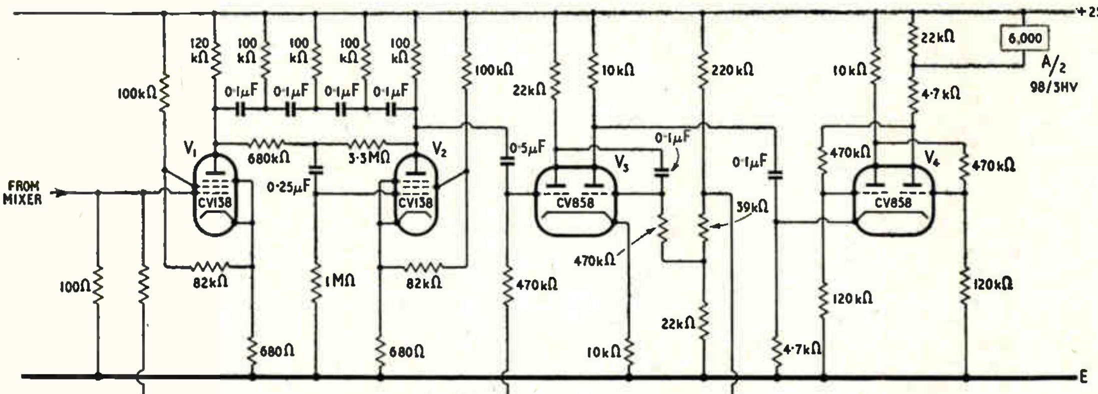


Fig. 22. *Motor control circuit.*

After further amplification the voltage across C_3 is applied to the reflector. A mathematical analysis of the stability conditions for this circuit is given in the Appendix. It is not proposed to enter into a lengthy discussion of the circuit problems peculiar to the 'instantaneous' a.f.c. system, such as that of devising a suitable clamp circuit. The Appendix shows that, except possibly for very short pulses, the design does not present any major difficulty. It is also shown that this system is less sensitive to circuit tolerances than the extrapolating type of a.f.c., since the instantaneous scheme can always be designed to cope with circuit tolerances, provided that the a.f.c. circuit bandwidth is increased sufficiently.

The circuit of Fig. 20 is open to a number of minor criticisms, but it is included here because examples of this type of circuit are rare, and because this example is of some historic interest.

A recurring problem in the design of reflector a.f.c. systems is set by the necessity of providing a d.c. coupling between the last a.f.c. stage and the reflector of the klystron. In most early equipments, as in the examples given above, the klystron was operated with its cavity at a positive potential with respect to earth. The cathode could then be directly connected to the anode of a control valve. This arrangement usually complicates the power supply problems, and has the disadvantage that the metal body of the klystron and the mechanical tuner are at a dangerous potential. There are also insulation problems, in the coupling between the klystron output and the mixers, for example. These objections become even more cogent if a high-voltage klystron is used.

These difficulties can be avoided if an r.f. power supply is used for the reflector. A suitable circuit is shown in Fig. 21. One section of a double triode (V_2) maintains oscillation at a frequency of about 500 kc/s in L_1. The voltage across a secondary winding is rectified by the diode and applied between cathode and reflector of the klystron. This voltage can be varied by means of the grid of the other triode section of V_2, which modulates the oscillator by varying the anode voltage. The grid of V_2 is d.c. coupled to the amplifier stage V_1. The error voltage, consisting of stretched pulses from the discriminator, is fed into the 'see-saw' input circuit of V_1, the other arm being taken to the reflector feed. Thus the steady-state gain from V_1 to the reflector is fixed by the ratio of R_{11} to R_1. V_1 is necessary mainly to provide a phase-reversal to allow negative-feedback to be applied round the loop. The rate of response of the reflector voltage to an error is controlled by the integrating capacitor C_{10}. The risk of phase-shift oscillations round this local-feedback loop is avoided by making the filter time-constant (C_9R_{10}) long compared with the other time-lags in the system (e.g., R_8C_5 and R_9C_6), and by restricting the a.c. loop gain by further feedback (R_6C_4) round V_1. To maintain the correct d.c. conditions at the grid of V_1, it is necessary to feed in some d.c. from the positive h.t. line. If klystron tolerances make it necessary, an adjustment (RV_1) can be provided to set the mean reflector voltage to suit the individual klystron.

By virtue of the heavy feedback round the loop, this arrangement is quite stable and linear, and the reflector voltage can be controlled over a much wider range than is ever necessary. Although this technique requires an extra valve in the a.f.c. circuit, the simplification of the power supplies usually proves a more than adequate recompense.

in one of the anode circuits. Briefly, the mechanism of division is as follows:—

Negative pulses applied to the cathode of V_4 produce little effect since whichever section of V_4 happens to be conducting will merely be driven a little further into grid current; positive pulses applied to the cathode will cause a rise in the

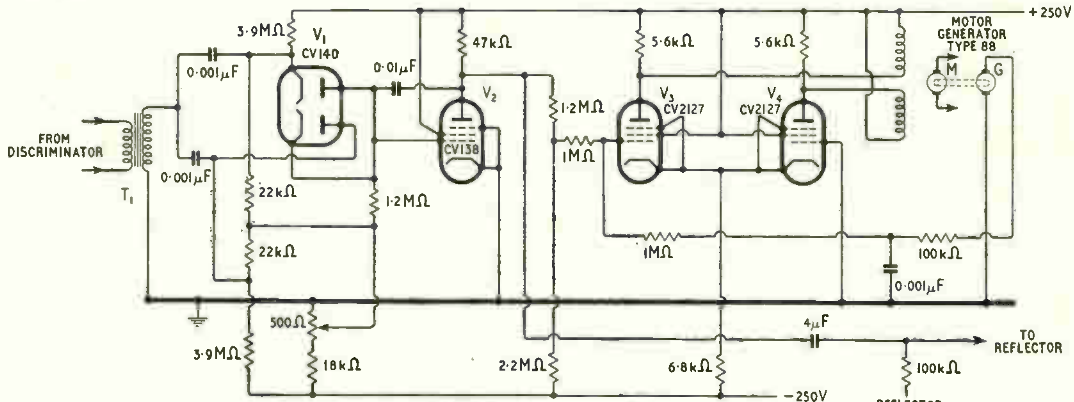


Fig. 23. *Reflector-mechanical a.f.c.*

The circuit illustrated in Fig. 21 was designed for use with a low-voltage klystron, but the technique can be even more profitably applied to high-voltage valves. If it is not convenient to take feedback directly from the reflector to the grid of V_1, it is possible to provide duplicate output windings and rectifiers, one of which develops a voltage between the cathode and reflector of the klystron, while the output of the other is earthed at one end and is used to provide the feedback. Such an arrangement overcomes the main objection to the use of high-voltage oscillators in conjunction with a.f.c. systems.

Fig. 22 shows an experimental circuit used for controlling the tuning motor of a reflector-cum-mechanical a.f.c. using the crystal-current scheme. The direct current in the a.f.c. mixer (obtained from the input to the a.f.c. i.f. amplifier, after filtering, develops a voltage across a 100-ohm resistor in the grid of an amplifier stage V_1. The resultant voltage, after further amplification by V_2, is differentiated and applied to the grid of a trigger circuit V_3, which trips whenever the mixer current falls sufficiently. To prevent this circuit being tripped by hum, microphony, or video pulses from the a.f.c. mixer, the amplifier stage V_2 is arranged to have a frequency response which commences to fall above about 10 c/s. This fall is brought about by the phase-shifting feedback network between the anodes of V_1 and V_2. The trigger circuit V_3, when tripped by a signal from V_2, produces a pulse of about 0·05 sec duration which is differentiated and fed into the cathode of the 'divide-by-two' circuit based on V_4. This is a d.c. coupled multivibrator with a relay

anode potential of the conducting triode, applying a positive swing to the grid of the non-conducting section. This grid voltage will be much greater than that at the cathode because of the gain of the conducting triode; the net result being to turn-off the formerly conducting section and vice versa, further reversal during the pulse being prevented by the inductive surge developed across the relay winding at the first reversal.

When V_4, and therefore the relay A/2, changes its state, the direction of running of the mechanical tuner motor is reversed. (The motor wiring is omitted from Fig. 22.) The motor will thus 'hunt' to-and-fro about the position where the crystal current is a maximum. The local oscillator used with the circuit shown was an X-band valve, the 723A/B. The tuner was driven by a two-way 'stepper' motor running at a speed of about 2 or 3 steps per second, each step producing a frequency shift of about 2 Mc/s. The 723A/B is not mechanically very suitable for use in this way, but was used merely to try out the principles. The search mechanism consists of a relay in the reflector control circuit which operates when the reflector loop pulls in. When this relay is operated, the above circuit is allowed to control the motor, but when the relay is open, the tuner will be driven to-and-fro over its range under the control of a toggle mechanism which reverses the motor when the limits are reached. Normally, when the correct frequency is reached during the search, the a.f.c. will lock in, the relay will operate and the motor control circuit can take over.

Fig. 23 shows an experimental version of a wide-range a.f.c. using an a.c. coupling to the reflector of the klystron. The latter was a high-voltage X-band valve, the CV129, whose tuning mechanism was driven by a Velodyne through a suitable gearbox. The gear ratio was chosen so that, at full speed, the tuner was driven at about 10 Mc/s per second. This is more than adequate to deal with any thermal drift. The CV129 has an electronic tuning range of only about 3 Mc/s, since this valve was not intended for use with an a.f.c., but the assembly of CV129, gearbox and Velodyne happened to be available in the laboratory.

An a.f.c. mixer, i.f. amplifier, and discriminator of conventional design, not shown in Fig. 23, supply error pulses to the transformer T_1. These error pulses pass through the diodes V_1, and are added to the output of V_2. The bias on the diodes prevents them conducting during the intervals between the pulses, but otherwise has little influence on the operation of the circuit. The output of V_2 is a.c. coupled to the reflector of the CV129, and also controls the input to the long-tailed pair (V_3, V_4) and hence the field current of the motor.

Before connecting the a.c. coupling to the reflector of the CV129, the mechanical a.f.c. was first tested alone. It was found that there was a marked tendency to instability, largely, it was believed, because of the considerable 'stiction' on the tuner shaft of the CV129, which provided the mechanical load on the Velodyne. This 'stiction' is abnormally high because the tuning mechanism of this valve is far from ideal for use with a motor drive, but high stiction is likely to be a feature of almost any such mechanism, because of the infinitesimal movement required.

On connecting up the reflector feedback, this was found to exert a marked stabilizing action on the system, which then behaved in a perfectly satisfactory manner. It appears quite likely that a simpler type of servo motor could have been used. The search system is again omitted from the diagram, but that described in the previous example is equally applicable here.

(*To be concluded*)

References will appear at the end of the last instalment.

NEW BOOKS

The Electromagnetic Field and its Engineering Aspects

By G. W. Carter, M.A., M.I.E.E. Pp. 360 + xvi. Longmans, Green & Co., 6 and 7 Clifford Street, London, W.1. Price 35s.

The author is Professor of Electrical Engineering in the University of Leeds, and the book is dedicated to his father, the late F. W. Carter, M.A., D.Sc., F.R.S., for many years Consulting Engineer to the B.T-H. Co. The object of the book is stated to be to give a thorough-going non-mathematical treatment of the theory of electromagnetism, a secure grasp of which has been made more than ever necessary by recent developments in engineering. Its first aim is to explain how electromagnetic apparatus behaves. One must not be misled by the term non-mathematical, for, as the author confesses, the treatment goes deeper than is usual in non-mathematical books, and the final chapter is devoted to 'the Principle of Relativity'. The rationalized m.k.s. system is used throughout. A very full list of symbols is given at the beginning and one receives a shock to note that, contrary to B.S.I. recommendations and general usage, the symbol for e.m.f. is $\mathcal{E}$ and that for electric force is E; this is not a misprint but is used throughout the book.

Chapter 1 consists of a brief historical introduction and a note on units. Chapter 2 deals with the electric field of charges in free space, and Gauss' theorem. In the formula $D = \epsilon E$, ϵ is called the 'electric constant' and when applied to free space it is called the 'primary electric constant'. Chapter 3 is entitled 'Conductors, Insulators and Capacitance', and concludes with a section on electrostatic screening. Chapter 4 deals with energy and the mechanical forces in the electric field, concluding with the electrostatic voltmeter and quadrant electrometer. In Chapter 5 we leave electrostatics and turn to electric-circuit theory and Ohm's law; the electrolytic tank method of plotting electric fields is described. In Chapter 6 we turn to magnetism, current loops and magnetic dipoles, magnetic potential and circuits, and Biot-Savart's law. What is generally called the absolute permeability of free space is called the primary magnetic constant and represented by the symbol η_0. The symbol μ is used for the ratio η/η_0 just as $\kappa = \epsilon/\epsilon_0$ is used for permittivity. Chapter 7 deals with the magnetic effects of iron. Different types of iron and steel are discussed and also various alloys and powdered cores; although ferrites do not appear to be mentioned, references are given to works on the subject, and it is explained that the account given is very incomplete. Chapter 8 is entitled 'electromagnetic induction', and deals with flux threading, flux cutting, slotted d.c. armatures, self and mutual inductance, etc. The energy and mechanical forces in the magnetic field are discussed in Chapter 9; the theory of the flux meter is given; hysteresis, eddy currents and Maxwell stresses are discussed. Chapter 10 is devoted to vector potential and its application to the calculation of self and mutual inductance. Chapter 11 deals with skin effect in a very thorough manner. Chapters 12 and 13 are devoted to Maxwell's equations, electromagnetic waves in space and in coaxial cables, waveguides, oscillating dipoles, the Poynting and Slepian vectors; one section is entitled 'the transition from low to high-frequency engineering'. The final chapter gives a brief discussion of the principle of relativity and its effect in electrical engineering.

A number of mainly numerical examples is given at the end of each chapter and the solutions are given at the end of the book.

In some cases the methods adopted by the author seem to be unnecessarily complicated, due to "the author's strong preference for the direct application of Neumann's law $\mathcal{E} = -d\Phi/dt$", which leads him to apply it where its application necessitates the picturing of fictitious circuits. A striking example of this is the

Faraday disc rotating in a uniform field parallel to the axis, so that every free electron experiences a radial force; this e.m.f. causes a p.d. between the axis and the periphery, and it seems quite unnecessary to explain it as being due to increasing flux linkage in a fictitious circuit. The author returns to it again in the chapter on relativity, and we are told that "the relativistic effect arises from the fact that the disc material is moving across the circuit line PQ". An interesting variant of the ordinary Faraday disc would be obtained by using for the external circuit a cylindrical can a little larger than the disc, with brushes all round the rim rubbing on the edge of the disc, and with the spindle prolonged to make contact with the centre of the bottom of the can. This would provide some further exercise for the Neumann law enthusiasts.

These remarks are not intended as an adverse criticism of the book, but rather the reverse, for they show that it is thought-provoking and attempts "to tie up a lot of loose ends which are commonly left untied". The aim of the book is stated to be "to give the reader a comprehensive view of the theory of electromagnetism, with constant reference to its engineering applications", and this it certainly does. It is a book that can be recommended to every serious student of electrical engineering. G. W. O. H.

Transistoren

By Dr. M. J. O. Strutt. Pp. 166 with 121 illustrations. S. Hirzel Verlag, Zürich. Price 21 Swiss francs.

The author is Professor and Director of the Institute for Advanced Electrotechnics at Zürich. This volume is No. 18 of a series of monographs in German on electric communications published under the direction of Professor Feldtkeller, of Stuttgart. Work on transistors has been going on at Zürich since 1948, and this book is based on a course of lectures on the subject given during the session 1952–53. It takes into account, although perhaps not completely, the published literature on the subject up to the middle of 1953; references are given to 148 published papers. In the preface the author says that it is now certain that in wide fields the valve will be replaced by the transistor, as its advantages in the saving of power and space are obvious. The difficulties of standard production are, especially in the U.S.A., being overcome.

The book is divided into ten chapters, of which the first, entitled Introduction, is divided into seven sections with the following headings: (1) What is a transistor? (2) Characteristics of the point transistor, (3) Compared with those of a passive 4-pole, (4) Equivalent diagram of a point transistor, (5) Power gain, (6) Ways of connecting it as a 4-pole, (7) Stability. This opening chapter occupies 19 pages, and deals with all these points in a very thorough manner. There is an unfortunate muddle on pages 4 and 5 in the numbering of the equations; in five places the references to Section 2 should be to Section 3, but this is a minor detail and it is soon spotted.

The second chapter deals with the electronics of the material, slow moving electrons, electrons in atoms, the wave nature of the electron and then electrons in solid material, with special reference to semi-conductors. The third chapter is entitled Contact Electronics and deals with contact between different metals, between metals and semi-conductors, and between different semi-conductors, and concludes with a section on rectification at the contact. Having devoted three chapters to the electronics of the materials, Chapter IV deals with their use in point transistors, the current multiplication at the collector, the formation of the material at the surface, time of transit, capacitance, inductance and noise. Chapter V deals with surface or junction transistors of various types and phototransistors. Duality and Analogues, to which we devoted the Editorial in March 1952, form the subject of Chapter VI. Two chapters are devoted to the use of transistors in the early stages and in the late stages of amplifiers, the noise-to-signal ratio, bandwidth, distortion and comparison with valves. Chapter IX deals with the production of oscillations with back coupling and negative resistance, and the final chapter with methods of measuring the important properties of semi-conductors and of transistors. Although most of the material has already been published in some form or other, some chapters contain hitherto unpublished developments of this earlier work.

The book contains a very clear exposition and a very thorough discussion of almost all that is known or, at least, was known a year ago, of the properties and applications of this minute but revolutionary element of electronic equipment, but some readers may find it rather compressed; the Principles of Transistor Circuits, by R. F. Shea, recently published, contains about three and a half times as many pages, and yet the reviewer in the May *Wireless World* thought that some of the chapters were "disappointingly scanty".

G. W. O. H.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for April 1954

Date 1954 April	Frequency deviation from nominal: parts in 10^8		Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
	MSF 60 kc/s 1429-1530 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1	−1·1	+2	NM
2	−1·1	+2	+53·2
3	−1·1	+1	NM
4	−1·1	+1	NM
5	NM	NM	+47·5
6	−1·2	+2	+45·2
7	−1·2	+2	+42·9
8	−1·2	+2	+42·5
9	−1·1	+2	+38·6
10	−1·1	+2	NM
11	−1·1	+1	NM
12	−1·1	+2	+34·3
13	−1·1	+2	+33·4
14	−1·1	+2	+31·8
15	−1·3	+1	+30·6
16	−1·3	+1	NM
17	−1·3	+2	NM
18	−1·2	+1	NM
19	−1·2	+2	NM
20	−1·2	+2	+23·6
21	−1·2	+2	+22·1
22	−1·2	+3	+20·4
23	−1·2	+3	+19·3
24	−1·3	+3	NM
25	−1·3	+3	NM
26	−1·3	+1	+15·3
27	−1·2	+4	+14·2
28	−1·2	+3	+13·3
29	−1·1	+4	+12·0
30	−1·2	+4	+10·3

The values are based on astronomical data available on 1st May 1954.

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

NM = Not Measured.

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.

ACOUSTICS AND AUDIO FREQUENCIES

534.121.1 1631

Concerning Combined and Degenerate Vibrations of Plates.—M. D. Waller. (*Acustica*, 1953, Vol. 3, No. 6, pp. 370–374.)

534.2-13 1632

Combined Translational and Relaxational Dispersion of Sound in Gases.—M. Greenspan. (*J. acoust. Soc. Amer.*, Jan. 1954, Vol. 26, No. 1, pp. 70–73.)

534.2-14 1633

Wave Propagation in a Randomly Inhomogeneous Medium: Part 2.—D. Mintzer. (*J. acoust. Soc. Amer.*, Nov. 1953, Vol. 25, No. 6, pp. 1107–1111.) The region of validity of the single-scattering approximation used in part 1 (931 of April) is found by considering the next higher approximation. Some results are given for the case where the sound wavelength is large compared with the size of the inhomogeneity.

534.213.4 : 534.833.1 1634

Transmission of Sound through a Stretched Membrane.—U. Ingard. (*J. acoust. Soc. Amer.*, Jan. 1954, Vol. 26, No. 1, pp. 99–101.) Analysis is given for a system comprising a membrane stretched across a tube. Conditions at resonance and antiresonance are studied by means of equivalent electrical circuits. Measurements made with a tube of diameter 3 in. and a fundamental frequency of 300 c/s gave results in agreement with the theory.

534.213.4-13 1635

The Attenuation of Sound in Small Tubes.—G. T. Kemp & A. W. Nolle. (*J. acoust. Soc. Amer.*, Nov. 1953, Vol. 25, No. 6, pp. 1083–1086.) Report of measurements over the frequency range 3·8–20 kc/s, using air-filled tubes of internal radius 0·238 and 0·0292 cm respectively. The results are compared with values calculated from the Kirchhoff theory.

534.231 1636

Mean Force on a Sphere in a Spherical Sound Field: Part 1 (Theoretical).—T. F. W. Embleton. (*J. acoust. Soc. Amer.*, Jan. 1954, Vol. 26, No. 1, pp. 40–45.) Theory developed by King (*Proc. roy. Soc. A*, 1934, Vol. 147, pp. 212–240) is presented in a form suitable for calculating the radiation force on a sphere in the field of a progressive spherical wave. The analysis is valid for all values of the ratio of wavelength to sphere radius. For large distances from the source the force is a repulsion varying as the inverse square of the distance. As the source is approached the force becomes an attraction, at a point depending on the frequency and on the sphere radius.

534.231 : 534.6 1637

Mean Force on a Sphere in a Spherical Sound Field: Part 2 (Experimental).—T. F. W. Embleton. (*J. acoust. Soc. Amer.*, Jan. 1954, Vol. 26, No. 1, pp. 46–50.) Measurements were made of the deflection of hollow glass spheres, of radius 0·228–3·029 cm, suspended by fine glass fibres in the field of a progressive sound wave, using frequencies between 300 and 7 000 c/s. Results give values of the force on the sphere in good agreement with values derived from the theory in part 1 (1636 above).

534.232 : 534.321.9 : 538.652 1638

Investigations on Magnetostrictive Oscillators at Frequencies between 80 and 500 kc/s.—G. Steinkamp. (*Acustica*, 1953, Vol. 3, No. 6, pp. 399–404. In German.) Measurements of the electromechanical and the mechano-acoustic efficiencies of five oscillators were made, from which the electroacoustic efficiency was deduced. Hysteresis and eddy-current losses were also determined. The losses due to internal magnetization processes are discussed in detail.

534.321.9 : 534.22.094.1 1639

Modulation Method of Measurement of Ultrasonic Dispersion.—V. A. Zverev. (*C. R. Acad. Sci. U.R.S.S.*, 1st Aug. 1953, Vol. 91, No. 4, pp. 791–794. In Russian.) The method was used in the experimental determination of dispersion in a long Ni wire, using a 1-Mc/s ultrasonic wave modulated at 100 kc/s. Fractional differences of velocity down to $\sim 2{\cdot}4 \times 10^{-3}$ were readily observed.

534.321.9 : 534.23 1640

Scattering of Underwater Plane Ultrasonic Waves by Liquid Cylindrical Obstacles.—S. J. Bezuszka. (*J. acoust. Soc. Amer.*, Nov. 1953, Vol. 25, No. 6, pp. 1090–1095.) Report of a theoretical investigation.

534.321.9-14 **1641**

Ultrasonic Transmission through Porous Bodies in Liquids.—G. Schmid & H. Knapp. (*Z. angew. Phys.*, Dec. 1953, Vol. 5, No. 12, pp. 463–472.) Report of measurements at 350 kc/s, using clay diaphragms in a N/1000 KCl solution.

534.373 : 534.213 **1642**

Boundary-Layer Absorption in a Spherical Resonator.—I. D. Campbell. (*Acustica*, 1953, Vol. 3, No. 6, pp. 395–398.) Calculation of the effective wall admittances representing viscous and thermal losses at the boundary leads to a solution of the decay of symmetrical modes of oscillation of a gas within a closed spherical resonator. A numerical example is calculated which permits assessment of the relative amounts of the absorption due to viscosity and thermal conductivity.

534.373 : 534.414 **1643**

A Study of the Factors influencing the Damping of an Acoustical Cavity Resonator.—R. F. Lambert. (*J. acoust. Soc. Amer.*, Nov. 1953, Vol. 25, No. 6, pp. 1068–1083.) Absorption measurements were made using a transverse-particle-velocity pickup device [1057 of 1950 (Hartig & Lambert)]. The results indicate that mechanical losses due to wall vibrations are experimentally significant. Empirical formulae are derived for the relaxation time of air molecules.

534.414 **1644**

On the Theory and Design of Acoustic Resonators.—U. Ingard. (*J. acoust. Soc. Amer.*, Nov. 1953, Vol. 25, No. 6, pp. 1037–1061.) Absorption and scattering from resonators in a free field and in walls are discussed. The effect on the resonance frequency of different aperture geometries is investigated. Design charts are presented for obtaining maximum resonance absorption, taking account of viscosity, heat conduction and radiation.

534.414 : 534.231 **1645**

The Near Field of a Helmholtz Resonator Exposed to a Plane Wave.—U. Ingard. (*J. acoust. Soc. Amer.*, Nov. 1953, Vol. 25, No. 6, pp. 1062–1067.) Calculated values for the near field of a spherical resonator, assuming a uniform velocity distribution in the aperture, are in good agreement with measured values. The case of a resonator at the end of a tube is also discussed.

534.614-14 **1646**

Measurement of the Velocity of Sound in the Ocean.—R. K. Brown. (*J. acoust. Soc. Amer.*, Jan. 1954, Vol. 26, No. 1, pp. 64–67.) An arrangement providing a continuous record of sound velocity as a function of depth is described. A phase-comparison principle is used; the operating frequency is 500 kc/s.

534.614-14 **1647**

Velocity of Sound in Glycerol.—F. A. A. Fergusson, E. W. Guptill & A. D. MacDonald. (*J. acoust. Soc. Amer.*, Jan. 1954, Vol. 26, No. 1, pp. 67–69.) A method for measuring the velocity of sound in liquids is based on the principle of a method used to determine critical angle in optics. The liquid is placed in a thin-walled cell which is rotated in a tank of a second liquid for which the velocity of sound is known. The value obtained for glycerol at 25°C is $1\,964 \pm 10$ m/s at a frequency of 7·5 Mc/s. Values for glycerol-water mixtures are also reported.

534.614-14 **1648**

A Method of Measuring the Velocity of Sound in a Few Grams of a Liquid.—J. H. Janssen. (*Acustica*, 1953, Vol. 3, No. 6, pp. 391–394.) A slight modification of the e.m. sound generator of St. Clair (2455 of 1941) permits the determination of the resonance frequency of a bar clamped at its centre, with and without a liquid column on top of it. The velocity of sound in the liquid can then be calculated.

534.614-14 : 534.321.9 **1649**

The Measurement of the Velocity of Sound in Liquids.—G. M. Graham. (*J. acoust. Soc. Amer.*, Nov. 1953, Vol. 25, No. 6, pp. 1124–1127.) A technique suitable for use in the frequency range 0·5–5 Mc/s is described. Measurements on distilled water are reported.

534.64 **1650**

The Acoustic Impedance of a Porous Layer at Oblique Incidence.—J. S. Pyett. (*Acustica*, 1953, Vol. 3, No. 6, pp. 375–382.) The specific normal impedance of a layer of anisotropic porous material for a plane wave incident in any direction is calculated in terms of the two propagation parameters of the material, on the assumption that one of the principal axes of the material is normal to the surface. Experiments on the same lines as those of Shaw (302 and 303 of February) gave results for rockwool in agreement with calculations.

534.76 **1651**

Basic Principles of Stereophonic Sound.—W. B. Snow. (*J. Soc. Mot. Pict. Telev. Engrs*, Nov. 1953, Vol. 61, No. 5, pp. 567–587. Discussion, pp. 587–589.) A review of published theory, with examples of practical applications. 68 references.

534.79 **1652**

A Technique and a Scale for Loudness Measurement.—W. R. Garner. (*J. acoust. Soc. Amer.*, Jan. 1954, Vol. 26, No. 1, pp. 73–88.) A true ratio scale of loudness is obtained by interrelating two scales derived from judgments made on two different principles, namely (*a*) equisection judgments, in which the observer adjusts a series of tones between two fixed limits to give a series of equal loudness intervals, and (*b*) fractionation judgments, in which the observer is instructed to adjust a tone to obtain a specified fraction of the initial loudness. Results of experiments indicate the validity of the assumptions made.

534.79 **1653**

Theory of Loudness Level and Loudness Sensation.—G. Quietzsch. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1953, Vol. 5, Nos. 9/10, pp. 169–177.) A summarized account of recent work on loudness level and loudness sensation with particular reference to the relation between the phon and sone scales.

534.833 **1654**

Electronic Sound Absorber.—H. F. Olson & E. G. May. (*J. acoust. Soc. Amer.*, Nov. 1953, Vol. 25, No. 6, pp. 1130–1136.) A microphone, amplifier and loudspeaker are arranged in a feedback system so as to reduce the sound pressure near the microphone. Reductions of 10–25 db can be achieved over a three-octave range at low audio frequencies. A suitable microphone is the electronic type previously described [3413 of 1947 (Olson)].

534.833 **1655**

Measurement of Sound Absorption Coefficients at the Research Centre for Cinematography in Rome.—G. Parolini. (*Ann. Télécommun.*, Dec. 1953, Vol. 8, No. 12, pp. 391–394.) Measurements were made with warble tones in a reverberation chamber. Results for 31 types of sound-absorbent materials of Italian manufacture, and for costumes of various periods, are tabulated.

534.834 : 534.861.1 **1656**

Acoustical Measurements on Components of Air-Conditioning Installations for Broadcasting and Television Studios.—G. Venzke. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1953, Vol. 5, Nos. 11/12, pp. 224–231.)

534.84 1657

An Empirical Acoustic Criterion.—T. Somerville. (*Acustica*, 1953, Vol. 3, No. 6, pp. 365–369.) A parameter R has been evolved which expresses the divergence of the reverberation-time/frequency characteristic from the mean reverberation time T_m. A parameter D has also been evolved which is a measure of the general decay irregularity for the studio under test. When $(D + 0{\cdot}7\ R)$ is plotted against T_m, the points representing studios and halls subjectively classified as having good acoustic properties are all found to lie within a particular area.

534.85 1658

Reduction of Background Noise in the Reproduction of Gramophone Disk Records.—M. J. de Cadenet. (*Rev. Son*, Dec. 1953, No. 9, pp. 311–322.) A review of basic circuits for use particularly with modern microgroove records.

621.395.61 1659

An Acoustic Lens as a Directional Microphone.—M. A. Clark. (*J. acoust. Soc. Amer.*, Nov. 1953, Vol. 25, No. 6, pp. 1152–1153.) Discussion of the radiation pattern of a microphone in which the transducer is located at or near the focus of a lens, e.g. of slant-plate type, the two being joined by a conical horn.

621.395.623.7 1660

A Physical Approach to the Generalized Loudspeaker Problem.—O. K. Mawardi. (*J. acoust. Soc. Amer.*, Jan. 1954, Vol. 26, No. 1, pp. 1–14.) The problem formulated by Foldy & Primakoff (264 of 1946) is studied and an expression is derived relating the sound pressure at a point in the field to the electrical signal at the loudspeaker terminals, a transfer function being defined by analogy with the methods of network analysis. For the simple case of a circular loudspeaker mounted in an infinite baffle, the integral equation obtained is solved exactly.

621.395.625 (083.74) 1661

I.R.E. Standards on Sound Recording and Reproducing: Methods for Determining Flutter Content, 1953.—(*Proc. Inst. Radio Engrs*, March 1954, Vol. 42, No. 3, pp. 537–541.) Standard 53 I.R.E. 19S2.

621.395.625.3 1662

Accurate Measurement of the Speed of Transport of Magnetic Tapes.—F. Gallet. (*Rev. Son*, Oct. 1953, No. 7, pp. 252–255.) The mains voltage is recorded on a length of the tape under test, and the magnetic-spectrum method is applied to locate the positions of the alternations. The length of tape occupied by 100 alternations gives the speed that would be obtained with a constant mains frequency of 50 c/s. An alternative method for continuous measurements is mentioned.

621.395.625.6 : 537.228.3 1663

Electro-Optic Sound-on-Film Modulator.—R. O'B. Carpenter. (*J. acoust. Soc. Amer.*, Nov. 1953, Vol. 25, No. 6, pp. 1145–1148.) Description of a variable-density sound-recording system using a crystal modulator of the type described previously [2209 of 1953 (Dressler & Chesnes)].

621.395.92 1664

Recent Advances in Hearing Aids.—L. G. Hector, H. A. Pearson, N. J. Dean & R. W. Carlisle. (*J. acoust. Soc. Amer.*, Nov. 1953, Vol. 25, No. 6, pp. 1189–1194.) Three models are described; the output powers are about 1, 5 and 40 mW respectively.

AERIALS AND TRANSMISSION LINES

621.372.2 1665

Experimental Study of the Transmission of Centimetre Waves along Wire Waveguides.—P. Chavance & B. Chiron. (*Ann. Télécommun.*, Nov. 1953, Vol. 8, No. 11, pp. 367–378.) Measurements over the frequency range 3–10 kMc/s are reported; six different polyethylene-covered Cu wires were used. Various types of launching device are described. Results indicate that the greater the reduction of phase velocity, the greater is the concentration of energy round the guide; the degree of concentration increases with the thickness of the dielectric coating. Discontinuities give rise to both reflection and radiation. Rain caused attenuations of 3–7 db/100 m at 3·15 kMc/s; a film of soot produced an attenuation of only 1·5 db/100 m. A power of 250 kW was transmitted without complications. Possible applications are discussed.

621.372.2 1666

Attenuation and Power-Handling Capability of Helical Radio-Frequency Lines.—J. H. Bryant & E. J. White. (*Trans. Inst. Radio Engrs*, Nov. 1953, Vol. MTT-1, No. 2, pp. 33–36.) A method is given for calculating the cold insertion loss and the power-handling capacity of a helical line with or without an outer coaxial cylindrical conductor.

621.372.2 + 535.41] : 621.3.012 1667

On a Property of a Family of Equiangular Spirals and its Application to Some Problems of Wave Propagation.—K. Landecker. (*J. appl. Phys.*, Jan. 1954, Vol. 25, No. 1, pp. 41–48.) The locus of all end points of the normals drawn from a point in a plane to the members of a family of equiangular spirals is a circle. The standing-wave pattern in a uniformly attenuating medium may, in simple cases, be represented by a vector diagram using a symmetrical pair of spirals. Construction of the locus circle enables the voltage and current maxima and minima to be determined. Transmission-line and light-wave interference examples are given, and the application in conjunction with a Smith chart is indicated.

621.372.2 : 621.315.212 : 621.317.733 1668

Some Steady-State Characteristics of Short Irregular Lines.—A. Rosen. (*Electronic Engng*, March 1954, Vol. 26, No. 313, pp. 90–96.) Analysis is based on representation of the distribution of impedance irregularities by a Fourier series. Measurements of impedance and of attenuation coefficient are discussed. A description is given of a bridge with variable-capacitance ratio arms for testing factory lengths of coaxial cable. See also 1295 of May.

621.372.21.029.6 1669

Low-Loss Microwave Line.—F. J. Tischer. (*Arch. elekt. Übertragung*, Dec. 1953, Vol. 7, No. 12, pp. 592–596.) Analysis of wave propagation along a line consisting of parallel strip conductors and a dielectric cross-bar. The transverse magnetic field parallel to the conductors is zero, hence no induced electric currents flow along the strips. Sections can therefore be coupled directly. The attenuation/wavelength characteristics of the line are compared with those of rectangular and circular waveguides.

621.372.8 1670

Propagation of Microwaves through a Cylindrical Metallic Guide Filled Coaxially with Two Different Dielectrics: Part 2.—S. K. Chatterjee. (*J. Indian Inst. Sci.*, Section B, July 1953, Vol. 35, No. 3, pp. 103–117.) The field components and propagation characteristics for the TM mode are derived. The hollow waveguide is treated as a special case. Part 1: 328 of February.

621.372.8 1671

Propagation of Microwaves through a Cylindrical Metallic Guide Filled Coaxially with Two Different Dielectrics: Part 3.—S. K. Chatterjee. (*J. Indian Inst. Sci.*, Section B, Oct. 1953, Vol. 35, No. 4, pp. 149–169.) The propagation characteristics for the TE_{01} and TM_{11} modes are derived. The results indicate that the phase velocity for a given mode can be adjusted to a pre-assigned value by a suitable choice of the dielectric constants and radii of the two media. In the case of the TE_{01} mode most of the energy is located in the medium having the higher dielectric constant. Part 2: 1670 above.

621.372.8 **1672**

Annular Resonant Slots in Dielectric-Filled Circular Waveguide.—M. Cohn. (*Trans. Inst. Radio Engrs*, Nov. 1953, Vol. MTT-1, No. 2, pp. 39–44.) The variation of susceptance with frequency over the range 8·2–9·6 kMc/s was investigated experimentally for irises with annular slots of various dimensions, in polystyrene-filled waveguides operating in the TE_{11} mode. Relations between the resonance frequency and Q value and the slot dimensions are derived. Results are shown in graphs.

621.372.8 **1673**

Graphical Analysis of Measurements on Multiport Waveguide Junctions.—S. Stein. (*Proc. Inst. Radio Engrs*, March 1954, Vol. 42, No. 3, p. 599.) Extension of previous work [3191 of 1953 (Storer et al.)].

621.372.8 **1674**

On the Propagation Constant in Gentle Circular Bends in Rectangular Wave Guides—Matrix Theory.—A. T. de Hoop. (*J. appl. Phys.*, Jan. 1954, Vol. 25, No. 1, p. 136.) Correction to paper noted in 968 of April.

621.396.67.012.12 : 517.864 **1675**

Solution of the Integral Equation of a Linear Aerial.—L. D. Bakhrakh. (*C. R. Acad. Sci. U.R.S.S.*, 1st Oct. 1953, Vol. 92, No. 4, pp. 755–758. In Russian.) The radiation-pattern function is related to the aerial current by a Fredholm-type equation of the first kind. The conditions are derived for the solutions to be in the form of (*a*) a finite sum of Mathieu harmonics and (*b*) a convergent series of Mathieu functions. The solution can be extended to the case of a plane aerial.

621.396.67.029.62 : 621.397 **1676**

Band III Television Aerials.—F. R. W. Strafford. (*Wireless World*, April 1954, Vol. 60, No. 4, pp. 181–184.) Requirements of aerials for receiving the proposed British transmissions in band III are discussed on the basis of the known performance of band-I aerials.

621.396.673.029.51 **1677**

Design and Adjustment of the Aerial at the National Long-Wave Station [at Allouis].—R. Chaste. (*Ann. Radioélect.*, Oct. 1953, Vol. 8, No. 34, pp. 301–312.) Design details are given of the folded $\lambda/4$ vertical aerial operating on 164 kc/s with bandwidth 20 kc/s [886 of March (Gaillard)]. Preliminary measurements were made on a model, scale 1 : 170.

621.396.674.3 **1678**

The Dipole and its Radiation.—A. Spork. (*Elektrotech. u. Maschinenb.*, 15th Dec. 1953, Vol. 70, No. 24, pp. 545–554.) Theory introductory to a series of papers on aerials is presented.

621.396.674.33 **1679**

The Symmetrical Biconical Aerial with Arbitrary Cone Angle.—L. Robin & A. Pereira-Gomes. (*Ann. Télécommun.*, Dec. 1953, Vol. 8, No. 12, pp. 382–390.) Schelkunoff's method (1049 of 1942) is extended to the general case, with the semi-angle of the cone, ψ, having any value in the range $0 < \psi < \pi/2$. Numerical computations are made for $\psi = 15°$, $30°$, $45°$, $60°$ and $75°$. The corresponding normalized impedances at the end and at the centre of the dipole are then calculated, taking into account the first two complementary internal waves and the first three external waves, in addition to the principal waves. Results are shown graphically.

621.396.676.012.12 **1680**

Electrically Small Antennas and the Low-Frequency Aircraft Antenna Problem.—J. T. Bolljahn & R. F. Reese. (*Trans. Inst. Radio Engrs*, Oct. 1953, Vol. AP-1, No. 2, pp. 46–54.) Quasi-static measurement techniques suitable for the determination of the radiation pattern and receiving sensitivity of electrically small aerials of both the electric- and magnetic-dipole types are described. These involve measurements on models in an electrostatic cage or an electrolyte tank. The theory, applications and limitations of these methods are discussed.

621.396.677 **1681**

A Simplified Calculation for Dolph-Tchebycheff Arrays.—G. J. van der Maas. (*J. appl. Phys.*, Jan. 1954, Vol. 25, No. 1, pp. 121–124.) A fuller account of work previously reported (634 of March).

621.396.677.3 **1682**

A Four-Terminal Network Theory for the Multi-Elements-Beam-Array with Reflector. (On an Adjustment Method of the Reflector and the Tailwire).—K. Nagai, R. Sato & G. Sato. (*Technol. Rep. Tohoku Univ.*, 1953, Vol. 17, No. 2, pp. 131–147.) The aerial array is treated as a four-terminal network with input terminals at the feed points of the energized element and output terminals at the junction of the reflector and stub (tailwire), the stub being considered as a load. From the equations obtained, the array characteristics and the optimum stub length are calculated. Fair agreement with experimental values is obtained.

621.396.677.3.012.12 **1683**

Radiation Pattern Synthesis.—E. A. Laport. (*RCA Rev.*, Dec. 1953, Vol. 14, No. 4, pp. 533–545.) A computational method is described for the synthesis of radiating arrays having zero radiation in specified directions. An extension of this method includes the solutions for radiation patterns from uniform current sheets of any specified aperture. Examples of radiation pattern shaping by using various primary arrays together with various secondary arrays are given.

621.396.677.7 **1684**

Passive Radiating Systems in Waveguides.—M. L. Levin. (*C. R. Acad. Sci. U.R.S.S.*, 1st Aug. 1953, Vol. 91, No. 4, pp. 807–810. In Russian.) A generalization of results obtained earlier for the characteristics of radiation from a resonant slot in a waveguide. The generalization includes the case of an arbitrary non-resonant radiating system consisting of passive metal aerials and slots in waveguide walls.

621.396.677.85 **1685**

Propagation in an Artificial Dielectric.—R. Fortet. (*Ann. Télécommun.*, Nov. 1953, Vol. 8, No. 11, pp. 361–366.) Periodic structures of the type used for microwave lenses are considered, comprising 'free zones', which are dielectrically homogeneous, alternating with regions in which the dielectric material is partitioned by metal strips into 'restricted zones'. The simple case studied by Brillouin (1584 of 1949) is examined and a more generalized theory is developed.

AUTOMATIC COMPUTERS

681.142 : [551.510.3 : 535.325 **1686**

An Analogue Computer for the Solution of the Radio Refractive-Index Equation.—Johnson. (See 1760.)

681.142 : 621.385.832 **1687**

Automatic Beam Current Stabilization for Williams Tube Memories.—R. J. Klein. (*Trans. Inst. Radio Engrs*, Dec. 1953, Vol. EC-2, No. 4, pp. 8–11.) One of the storage points in the memory tube is used as a test point. At regular intervals this is cleared and recharged, and the output is sampled, the beam current being adjusted to keep the sampled output at a constant level. The whole operation takes 40 μs to complete. Circuit details are given.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.011.21 **1688**

A Note on calculating the Total Impedance of Impedors connected in Parallel.—W. Boesch. (*Frequenz*, Nov. 1953, Vol. 7, No. 11, pp. 331–335.) Three methods, one involving use of a desk calculating machine, and another use of a linear nomogram, are compared and contrasted from the point of view of speed and accuracy.

621.3.018.783 : 621.376.3 **1689**

Nonlinear Distortion of a Random Signal. Application to Frequency Modulation.—P. J. M. Clavier. (*Câbles & Transm.*, Oct. 1953, Vol. 7, No. 4, pp. 293–300.) Analysis is based on the instantaneous statistical properties of a random signal, the energy spectrum for a given bandwidth being determined from the appropriate correlation functions. Application of the method is illustrated in the case of f.m. when propagation time is considered. Since the calculation takes account of intermodulation products over a complete frequency band the method has advantages over that based on intermodulation of discrete frequencies.

621.314.22.018.75 : 538.221 : 621.318.1 **1690**

Pulse Permeability and Losses in Magnetic Materials subjected to Rectangular D.C. Pulses.—Einsele. (See 1826.)

621.318.435.3 : 621.375.3 **1691**

Transductor with High-Impedance Control-Winding Termination.—W. Schmidt. (*Arch. elekt. Übertragung*, Dec. 1953, Vol. 7, No. 12, pp. 574–578.) The dynamic behaviour of this transductor is represented by equations similar to those for a transductor with a low-impedance control-winding termination (2579 of 1953).

621.372 : 512.831 **1692**

Matrix Analysis of Linear Time-Varying Circuits.—L. A. Pipes. (*Trans. Inst. Radio Engrs*, Dec. 1953, No. PGCT-2, pp. 91–105.)

621.372 + 621.3.018.78] (083.74) **1693**

I.R.E. Standards on Circuits: Definitions of Terms in the Field of Linear Varying Parameter and Nonlinear Circuits, 1953.—(*Proc. Inst. Radio Engrs*, March 1954, Vol. 42, No. 3, pp. 554–555.) Standard 53 I.R.E. 4S1.

621.372.029.63 .64 : 621.315.212 **1694**

A Coaxial Magic-T.—T. Morita & L. S. Sheingold. (*Trans. Inst. Radio Engrs*, Nov. 1953, Vol. MTT-1, No. 2, pp. 17–23.) The basic design consists of a coaxial-line T junction including a shielded loop whose axis is located at the centre of the T; the operation is similar to that of the hybrid coil or the waveguide magic T. Two modifications are described, one using direct coupling between the loop and the loads. Standing-wave characteristics of experimental models for operation at wavelengths of about 10 cm are given; a s.w.r. < 2 can be obtained over a frequency band of about 10%. Application to phase measurements is described briefly.

621.372.221 : 621.385.029.63/.64 **1695**

Filter-Helix Traveling-Wave Tube: Part 1—The Filter Helix, a New Circuit Element for Traveling-Wave Amplifiers and Oscillators.—Dodds & Peter. (See 1963.)

621.372.412 **1696**

The Properties and Manufacture of Piezoelectric Quartz Crystals.—H. L. Downing. (*J. Brit. Instn Radio Engrs*, March 1954, Vol. 14, No. 3, pp. 130–138.) Reprint. See 667 of March.

621.372.412 : 534.133 **1697**

Thickness-Shear and Flexural Vibrations of Contoured Crystal Plates.—R. D. Mindlin & M. Forray. (*J. appl. Phys.*, Jan. 1954, Vol. 25, No. 1, pp. 12–20.) Equations of motion expressing approximately and separately the thickness-shear and flexural modes are derived from data on the coupled modes for a plate of uniform thickness [1861 of 1951 and 2156 of 1952 (Mindlin)]. These equations are applied to the case of (*a*) a plate of double-wedge shape, and (*b*) a plate with bevelled edges. Results support the finding that thickness-shear motion is localized at the centre, flexural motion at the edges.

621.372.412 : 534.133 **1698**

Thickness-Shear Vibrations of Piezoelectric Crystal Plates with Incomplete Electrodes.—R. D. Mindlin & H. Deresiewicz. (*J. appl. Phys.*, Jan. 1954, Vol. 25, No. 1, pp. 21–24.) The method developed in 1697 above (Mindlin & Forray) is applied to determine the influence of electrode length on thickness-shear frequencies. For a partially coated crystal these frequencies depend on the coupling between the coated and uncoated portions considered as separate systems.

621.372.412 : 534.133 **1699**

Suppression of Overtones of Thickness-Shear and Flexural Vibrations of Crystal Plates.—R. D. Mindlin & H. Deresiewicz. (*J. appl. Phys.*, Jan. 1954, Vol. 25, No. 1, pp. 25–27.) Analysis showing how electrodes coated on the surfaces of thickness-shear plates may be shaped so that only one resonance mode is excited.

621.372.413 **1700**

Derivation of Equivalence Theorem of a Weakly Coupled Electromagnetic Cavity-Resonator System.—E. Ledinegg & P. Urban. (*Arch. elekt. Übertragung*, Dec. 1953, Vol. 7, No. 12, pp. 561–568.) A general proof of the equivalence between a system of weakly coupled cavities and a system of coupled resonant LCR circuits is derived from Maxwell's equations by means of a first-order perturbation calculation. The frequency dependence of the system is deduced from the equation of state. The case of two coupled cavity resonators is considered in an example.

621.372.413 **1701**

Calculation of the Resonant Properties of Electrical Cavities.—S. Bertram. (*Proc. Inst. Radio Engrs*, March 1954, Vol. 42, No. 3, pp. 579–585.) The method described is based on measurements made on a model in an electrolyte tank. Calculated values of resonance wavelength and Q are in good agreement with directly measured values.

621.372.414 : 517.5 **1702**

Characteristic Functions of Real Resonators.—V. M. Vakhnin. (*C. R. Acad. Sci. U.R.S.S.*, 1st Aug. 1953, Vol. 91, No. 4, pp. 779–782. In Russian.) Orthogonal characteristic functions for a piecewise-inhomogeneous, one-dimensional system are considered in relation to resonators with energy losses at the boundaries. The analysis provides an explanation of previously published experimental results.

621.372.5 1703

An Iterative Method for Network Synthesis.—R. E. Scott & R. L. Blanchard. (*Trans. Inst. Radio Engrs*, Dec. 1953, No. PGCT-2, pp. 19–29.) The numerical method described is based on the complex critical frequencies of the insertion-loss function. Applications to networks with up to four poles and four zeros and to a crystal filter are illustrated.

621.372.5 1704

Synthesis of Transfer Functions with Poles Restricted to the Negative Real Axis.—A. D. Fialkow & I. Gerst: L. Weinberg. (*J. appl. Phys.*, Dec. 1953, Vol. 24, No. 12, pp. 1525–1527.) Comment on 1929 of 1953 and author's reply.

621.372.5 : 621.396 1705

Radio-Frequency Phase-Difference Networks: a New Approach to Polyphase Selectivity.—Cifuentes & Villard. (See 1910.)

621.372.5.018.78 1706

Quasi-Distortionless Filter Functions.—J. L. Stewart. (*Trans. Inst. Radio Engrs*, Dec. 1953, No. PGCT-2, pp. 39–54.) The quasi-distortionless filter network function is specified by expressing the input and output functions as time power series and equating their coefficients. Such filter functions are derived and discussed.

621.372.54 1707

The Direct Method of Filter and Delay Line Synthesis.—M. J. E. Golay. (*Proc. Inst. Radio Engrs*, March 1954, Vol. 42, No. 3, pp. 585–588.) Extension of previous work (1497 of 1946). "The direct method consists in postulating an indefinitely recurring network in which each recurring element is coupled to all elements of the network, and in determining the values of these elements and of their indefinitely extended couplings which will yield the exact phase characteristics desired within the exact pass band desired." Examples are given of the application of the method to the design of ideal band-pass filters.

621.372.54 1708

The Synthesis of Reactance Networks consisting only of Rejector Circuits in Series, by the Method of Partial Networks.—T. O'Callaghan. (*Frequenz*, Oct. 1953, Vol. 7, No. 10, pp. 299–306.) The modification required in a filter network with given resonance and anti-resonance frequencies in order to introduce a further prescribed resonance or antiresonance frequency is determined. Numerical examples illustrate the method, which is also applied to the reduction of the number of resonance frequencies and to a mechanical oscillator.

621.372.54 1709

The Historical Development of Filter Theories.—W. Klein. (*Frequenz*, Nov. 1953, Vol. 7, No. 11, pp. 326–331.)

621.372.54 1710

Elementary Ladder-Type Filters.—J. Oswald. (*Câbles & Transm.*, Oct. 1953, Vol. 7, No. 4, pp. 325–358.) A system of classification of conventional ladder-type filters based on image parameters is described. An elementary filter is defined as a network with two prescribed image impedances and minimum attenuation. These are generally composed of conventional half-sections, but in certain cases, particularly for band-pass filters, synthesis involves sections with three branched impedances which cannot be resolved into matched half-sections. Formulae with examples of networks for low-pass, high-pass, band-pass and band-stop filters are given in an appendix.

621.372.54 : 621.314.7 1711

***RC* Active Filters.**—J. G. Linvill. (*Proc. Inst. Radio Engrs*, March 1954, Vol. 42, No. 3, pp. 555–564.) A type of filter is described in which the active element is a negative-impedance converter using transistors (2582 of 1953). The number of capacitors required is only equal to the total number of reactance elements in the corresponding *LC* filter. Theory is given, and tests on low-pass, high-pass and band-pass filters are reported.

621.372.543.2 1712

Impedance Transformations in Band-Pass Filters.—T. J. O'Donnell & E. M. Williams. (*Elect. Engng, N.Y.*, Dec. 1953, Vol. 72, No. 12, p. 1105.) Digest of paper to be published in *Trans. Amer. Inst. elect. Engrs*, 1953, Vol. 72. The design of band-pass impedance-transforming filters is derived from the equivalence between an asymmetrical band-pass filter and a symmetrical section plus an ideal transformer.

621.372.543.2 : 621.375.221 1713

Design of a Simple Band-Pass Amplifier with Approximate Ideal Frequency Characteristics.—W. E. Bradley. (*Trans. Inst. Radio Engrs*, Dec. 1953, No. PGCT-2, pp. 30–38.) The application of the 'pole and zero' method of synthesis is illustrated.

621.372.543.2 : 621.376.3 1714

F.M. Transient Response of Band-Pass Circuits.—R. E. McCoy. (*Proc. Inst. Radio Engrs*, March 1954, Vol. 42, No. 3, pp. 574–579.) A general formula is derived for the instantaneous deviation of output frequency resulting from a step change of input frequency. Calculations are presented for large frequency shifts in a single-tuned circuit and for small frequency shifts in amplifiers with various numbers of tuned circuits. For small input steps near the middle of the pass band the transient response is the same for f.m. as for a.m. Stagger tuning gives rise to overshoot, but the extent of this is small if the amount of stagger is no more than enough to provide maximally flat response in the steady state.

621.372.543.29 1715

One Design Method of Band-Pass Filter for V.H.F.—K. Nagai, R. Sato & N. Saito. (*Technol. Rep. Tohoku Univ.*, 1953, Vol. 17, No. 2, pp. 120–130.) Design formulae for band-pass filters constructed of two-wire and coaxial cables are derived and numerical examples are given. The theoretical and experimental results are in fair agreement.

621.373 : 513.83 1716

'Hereditary' Dynamic Systems with Discontinuities ['à déferlement'].—T. Vogel. (*Ann. Télécommun.*, Nov. 1953, Vol. 8, No. 11, pp. 354–360.) The characteristics distinguishing nonlinear from linear oscillations are examined. A class of dynamic systems termed 'discontinuous' ('à déferlement') is defined; cases are discussed for which Poincaré's topological methods can be used to determine the periodic solutions and the stability. Such systems are included in the larger class of 'hereditary' systems defined by Volterra (i.e. systems whose operation is affected by their previous history). See also 630 of 1952.

621.373.4 1717

Single-Stage Phase-Shift Oscillator.—W. Bacon. (*Wireless Engr*, April 1954, Vol. 31, No. 4, pp. 100–104.) A method of design is developed in which the phase-shifting network contains an arbitrary number of *RC* or *CR* cells and the value of the first series impedance is half that of the others. Simple formulae are derived for the network impedance and oscillation frequency.

621.373.5 1718

Transistor Oscillators.—H. E. Hollmann. (*Arch. elekt. Übertragung*, Dec. 1953, Vol. 7, No. 12, pp. 585–591.) The self-oscillation conditions and frequency-deviation characteristics of a three-point transistor-oscillator are derived by analogy with the retarding-field oscillator, and a comparison is made with measured characteristics of point-contact and junction-type transistor oscillators. Various circuits are examined from the point of view of suitability of application of the two types of transistor. The discussion is limited to the frequency range below α cut off.

621.375.132.012 1719

The Audio-Frequency Negative-Feedback Amplifier as a Filter Circuit.—F. Steiner & I. Riess. (*Öst. Z. Telegr. Teleph. Funk Fernsehtech.*, Nov./Dec. 1953, Vol. 7, Nos. 11/12, pp. 141–147.) A single-stage *RC* feedback amplifier is represented by an equivalent damped parallel *LC* circuit and a locus diagram for amplification is constructed. The method is applied to a two-stage amplifier, an expression analogous to form factor in filter theory being derived in terms of feedback ratio. This is useful for determining the effect of feedback on the frequency characteristic in a.f. amplifier design.

621.375.2 1720

Bandpass Amplifiers.—Z. Jelonek & R. S. Sidorowicz. (*Wireless Engr*, April 1954, Vol. 31, No. 4, pp. 84–99.) A design method is presented based on the required overall gain, half-power bandwidth and centre frequency of the pass band. The amplifier considered is composed of a number of identical groups each comprising one or more amplifying stages and having a maximum-flatness frequency response. Conditions for maximum possible gain are found. Regeneration due to anode-grid capacitances is discussed; the amount of distortion introduced into the frequency response is related to a regeneration coefficient α. When α is sufficiently large, oscillations occur. Stability conditions for single-tuned and stagger-tuned amplifiers are derived.

621.375.221 1721

Distributed Amplifier Theory.—D. V. Payne. (*Proc. Inst. Radio Engrs*, March 1954, Vol. 42, No. 3, pp. 596–598.) Discussion on 2612 of 1953.

621.375.222.029.3 1722

A New Transformerless Amplifier Circuit.—K. Onder. (*J. audio Engng Soc.*, Oct. 1953, Vol. 1, No. 4, pp. 282–286.) Four valves in a bridge circuit are driven in appropriate phase and amplitude so as to deliver equal power to a common load. No d.c. flows through the load; negative feedback is easily applied. A 9-W and an 18-W amplifier are described; response characteristics are flat over the audio range at all levels of operation. The 18-W unit weighs 4 lb 3 oz, including power supply and pre-amplifier.

621.375.3 1723

Magnetic Amplifier performs Analytical Operations.—L. A. Finzi & R. A. Mathias. (*Elect. Engng, N.Y.*, Dec. 1953, Vol. 72, No. 12, p. 1097.) Digest of paper to be published in *Trans. Amer. Inst. elect. Engrs*, 1953, Vol. 72.

621.395.665 1724

Audio Automatic Volume Control Systems.—F. W. Roberts & R. C. Curtis. (*J. audio Engng Soc.*, Oct. 1953, Vol. 1, No. 4, pp. 310–316.) Applications of a.v.c. are noted and various circuits are reviewed. Methods of eliminating audible 'thump' occurring on application of a direct control voltage, of deriving the control voltage, and of reducing time delay are discussed. Measurement techniques are outlined.

621.395.665.1 1725

The New Compressor-Expander System of the French Post Office.—M. Lagarde, R. Blondé & R. Derosier. (*Câbles & Transm.*, Oct. 1953, Vol. 7, No. 4, pp. 301–308.) The new unit has improved operating characteristics and is less than one-quarter the size of the former type [306 of 1951 (Lagarde et al.)]. Two Si diodes replace the Cu_2O rectifiers in the potentiometer system, and two Ge diodes the double-diode valve in the detector unit. Other valves are replaced by miniature types. Performance tests are reported.

GENERAL PHYSICS

530.112 1726

Proposal for a New Aether Drift Experiment.— H. L. Furth. (*Nature, Lond.*, 9th Jan. 1954, Vol. 173, No. 4393, pp. 80–81.) An experiment using standing microwaves is outlined.

535.1 : 537.228 1727

Deductions regarding the Constitution of Light Energy.—J. Stark. (*Z. Phys.*, 17th Nov. 1953, Vol. 136, No. 2, pp. 221–223.) A discussion based on the results of experiments previously reported (1292 of 1953). The field associated with a particle of light, as defined by Planck, is nonoscillatory, the frequency associated with the light being due to the circulation of e.m. energy in the particle.

535.37 : 537.228 1728

The Mechanism of Electroluminescence: Part 2—Applications to the Experimental Facts.—D. Curie. (*J. Phys. Radium*, Dec. 1953, Vol. 14, No. 12, pp. 672–686.) Theory previously developed (1373 of May) is used to interpret observed phenomena, including temperature and frequency effects, fluctuations of brightness corresponding to field alternations, initial rise of brightness, and surface effects. Good crystallization, abundance of luminescence centres and donor levels, and absence of other defects are conditions favouring the production of electroluminescence. Various theories of the phenomenon are compared in an appendix.

535.41 + 621.372.2] : 621.3.012 1729

On a Property of a Family of Equiangular Spirals and its Application to Some Problems of Wave Propagation.—Landecker. (See 1667.)

535.42 : 538.566 1730

The Diffraction of Waves in passing through an Irregular Refracting Medium.—J. A. Fejer. (*Proc. roy. Soc. A*, 22nd Dec. 1953, Vol. 220, No. 1143, pp. 455–471.) "A relation is derived between the angular power spectrum of waves emerging from a thin diffracting screen random in two dimensions and the auto-correlation function describing the irregularities of the field as it emerges from the diffracting screen. The special case of an isotropic screen characterized by an auto-correlation function which depends only on distance and not on direction is discussed for normal and oblique incidence. Multiple scattering of waves caused by volume irregularities of the dielectric constant is considered. The angular power spectrum and the auto-correlation function describing the irregularities of the diffraction field caused by multiple scattering in a thick slab are determined. The results are compared with those obtained by Hewish [1284 of 1952] for a thin phase-changing screen. The results are discussed in terms of problems arising in the study of radio-wave propagation."

537.213 1731
A Note on Uniqueness Proofs for Boundary-Value Problems in Potential Theory and Steady Heat Conduction.—M. E. Rayner. (*Quart. J. Mech. appl. Math.*, Dec. 1953, Vol. 6, Part 4, pp. 385–390.) By allowing the boundary to tend to infinity in two directions, and considering the problem in a semi-infinite region, it is possible to prove uniqueness by classical methods for a very large class of boundary conditions.

537.311.5 : 517.9 1732
Variational Methods for Problems in Resistance.—J. F. Carlson & T. J. Hendrickson. (*J. appl. Phys.*, Dec. 1953, Vol. 24, No. 12, pp. 1462–1465.) Schwinger's variational methods are used to calculate upper and lower limits for the resistance of a cylinder with a specified transverse distribution of potential at one end.

537.311.62 : 517.392 1733
The Evaluation of the Surface Impedance in the Theory of the Anomalous Skin Effect in Metals.—A. N. Gordon & E. H. Sondheimer. (*Appl. sci. Res.*, 1953, Vol. B3, Nos. 4/5, pp. 297–304.) A method is described for obtaining series expansions for the integrals which represent the surface impedance.

537.52 1734
Note on the Quantitative Development of the [gas] Discharge Characteristic.—E. Pfender. (*Z. angew. Phys.*, Dec. 1953, Vol. 5, No. 12, pp. 450–453.)

537.52 : 551.594.223 1735
Investigation of Ball Lightning by means of Models.—H. Nauer. (*Z. angew. Phys.*, Dec. 1953, Vol. 5, No. 12, pp. 441–450.) Experiments with gas discharges are described.

538.2 1736
U.S.S.R. Research on Magnetism.—S. Rosenblum. (*Nuovo Cim.*, 1953, Supplement to Vol. 10, No. 4, pp. 441–458. In French.) A brief historical survey with comprehensive bibliography.

538.214 1737
Effect of the Surface on the Magnetic Properties of an Electron Gas.—F. S. Ham. (*Phys. Rev.*, 1st Dec. 1953, Vol. 92, No. 5, pp. 1113–1119.) The energy levels of free electrons confined in a finite cylindrical box with a uniform axial magnetic field are computed using the WKB approximation; the form of approximation appropriate to various boundary conditions is discussed. The results are used to calculate the susceptibility of the system.

538.3 1738
On Reciprocity Theorems in Electromagnetic Theory.—T. H. Crowley. (*J. appl. Phys.*, Jan. 1954, Vol. 25, No. 1, pp. 119–120.) Five different types of sources of e.m. fields are considered, and the corresponding reciprocity theorems stated.

538.3 1739
A Scalar Representation of Electromagnetic Fields.—H. S. Green & E. Wolf. (*Proc. phys. Soc.*, 1st Dec. 1953, Vol. 66, No. 408A, pp. 1129–1137.) It is shown that in a region which is free from currents and charges any e.m. field may be rigorously derived from a single, generally complex, scalar wave function. The momentum density and energy density defined in terms of this function differ from those given by the usual expressions, but the differences disappear on integration over any arbitrary macroscopic domain.

538.5 1740
Induction of Currents by Moving Charges.—Ya. N. Fel'd. (*C. R. Acad. Sci. U.R.S.S.*, 21st Nov. 1953, Vol. 93, No. 3, pp. 447–450. In Russian.) An analysis is made of the field due to a charge moving between a pair of parallel disk electrodes, connected externally by a conductor. A general expression for the current flowing through the conductor is derived in terms of the electrode radius, r_0, and separation, a, and the velocity of the charge, u. A periodic discharge could be obtained with particular values of r_0, a and u.

538.561 : 662.2 1741
Electromagnetic Waves emitted on Detonation of Explosives.—H. Kolsky. (*Nature, Lond.*, 9th Jan. 1954, Vol. 173, No. 4393, p. 77.) Small electrical disturbances, detected incidentally while using explosives to produce sharp stress pulses, were later investigated separately. The effect was displayed on a high-speed c.r.o. The potential of a wire probe reached a maximum after about 50 μs and then decayed to zero. The effect is probably due to the production of positive and negative ions having different mobilities.

538.566 1742
Vector Modifications inherent in the Doppler Effect for Waves Propagated in a Dielectric Medium.—M. Risco. (*J. Phys. Radium*, Dec. 1953, Vol. 14, No. 12, pp. 657–662.) The propagation of plane e.m. waves in a moving medium is characterized by four vectors, as compared with two for a stationary medium, because the directions of the electric and magnetic fields no longer coincide with the respective inductions. Relations between these vectors are derived and their physical significance is discussed.

538.566 1743
The Electromagnetic Transmission Characteristics of the Two-Dimensional Lattice Medium.—H. S. Bennett. (*J. appl. Phys.*, June 1953, Vol. 24, No. 6, pp. 785–810.) A theoretical investigation is made of systems with obstacles arranged to form a regular lattice in planes perpendicular to the direction of propagation, the obstacles having elliptical cross-section. An examination is made of the influence of obstacle size, shape and spacing, and of the wave polarization on the transmission characteristics. No restriction is introduced as regards the size of the obstacle in relation to λ. The theory is relevant to problems of microwave lenses and transmission circuits of travelling-wave valves.

538.566 : 535.42 1744
Microwave Diffraction Measurements in a Parallel-Plate Region.—R. V. Row. (*J. appl. Phys.*, Dec. 1953, Vol. 24, No. 12, pp. 1448–1452.) Diffraction of a plane or a cylindrical wave by wedges, covered with 0·003-in. silver foil, was investigated experimentally by a probe method. The wedge, the 3·185-cm-λ line source and the probe were located between two large parallel duralumin sheets spaced ½-in. apart. The measured values of field strength are compared with values derived from previously published theory.

538.566 : 538.311 1745
Radiation from a Line Source Adjacent to a Conducting Half Plane.—J. R. Wait. (*J. appl. Phys.*, Dec. 1953, Vol. 24, No. 12, pp. 1528–1529.) Expressions for the field are obtained in terms of tabulated functions by using an alternative method of analysis to that of Harrington (2967 of 1953).

538.566.2 1746
The Fields of a Line Source of Current over a Stratified Conductor.—J. R. Wait. (*Appl. sci. Res.*, 1953, Vol. B3, Nos. 4/5, pp. 279–292.) Expressions for the electric and

the magnetic fields are derived, assuming the wire to be infinitely long and parallel to the conductor. At large distances the results can be generalized to apply to any source of horizontally polarized waves situated on the surface of a half-space consisting of an arbitrary number of layers.

538.569.4.029.64 1747

Absorption of Microwaves in Gases.—Krishnaji & P. Swarup. (*J. appl. Phys.*, Dec. 1953, Vol. 24, No. 12, p. 1525.) Absorption coefficients in the 3-cm waveband were determined for ammonia, ethyl alcohol and acetaldehyde at a pressure of 745 mm Hg, and for methyl iodide and acetone at 20 cm Hg and 25 cm Hg, respectively.

539.23 : 537.533.8/.9 1748

The Decomposition of Thin Films on Bombardment with Slow Electrons.—D. A. Wright & J. Woods. (*Proc. phys. Soc.*, 1st Dec. 1953, Vol. 66, No. 408B, pp. 1073–1086.) The threshold energies of the bombarding electrons were determined for the onset of electron absorption, decomposition and secondary emission. The films investigated included BaO, $BaCl_2$, $BaSO_4$ and the alkali halides. The experiments were carried out using diode or triode valve systems, with an oxide-coated cathode operated at 400°C as the source of the bombarding electrons.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5 : 621.396.96 1749

Radio Astronomy: Part 3—Radiolocation of Meteors.—W. Dieminger. (*Arch. elekt. Übertragung*, Dec. 1953, Vol. 7, No. 12, pp. 555–560.) A review of measurement techniques and results, with a note on local ionization effects of meteors. Part 2: 1407 of May (Siedentopf).

523.5 : 621.396.96 1750

The Effect of Radar Wavelength on Meteor Echo Rate.—V. R. Eshleman. (*Trans. Inst. Radio Engrs*, Oct. 1953, Vol. AP-1, No. 2, pp. 37–42.) Theory is given which includes the effects of high electron density, the linear rate of trail formation, and the initial column radius on the meteor echo power and hence on the number of echoes received. The theories of Lovell & Clegg (2782 of 1948) and others are discussed critically. A comparison with experimental results is made.

523.7 : 550.38 1751

A Note on Solar-Terrestrial Relationships.—N. C. Gerson. (*J. atmos. terr. Phys.*, Jan. 1954, Vol. 4, No. 6, pp. 341–342.) Corpuscular emission from the invisible solar disk and deflection of the particles into certain trajectories could account for the lack of correlation between particular solar and geophysical phenomena.

523.746 1752

Identification of Sunspots.—H. Alfvén. (*Tellus*, Nov. 1953, Vol. 5, No. 4, pp. 423–445.) Previous work (3104 of 1948) indicates a statistical correlation between spots on opposite hemispheres during successive sunspot cycles. Analysis of sunspots between latitude 4°S and 4°N appears to support this view. A number of sunspots are identified as being produced by the same original disturbance in the solar core; the mechanism is explained on the basis of the magneto-hydrodynamic theory of sunspots.

523.854 : 621.396.822 1753

Position and Identification of a Bright Extended Radio Source in Gemini.—J. E. Baldwin & D. W. Dewhirst. (*Nature, Lond.*, 23rd Jan. 1954, Vol. 173, No. 4395, pp. 164–165.) More accurate observations are reported on a previously detected source [121 of 1952 (Ryle et al.)].

523.854 : 621.396.822 1754

The Radio Brightness Distributions over Four Discrete Sources of Cosmic Noise.—B. Y. Mills. (*Aust. J. Phys.*, Dec. 1953, Vol. 6, No. 4, pp. 452–470.) The experimentally determined isophotes, obtained from measurements with a variable-aerial-spacing radio interferometer at 101 Mc/s, are compared with the optical features of the nebulae.

523.854 : 621.396.822 1755

Galactic Radiation at Radio Frequencies: Part 5—The Sea Interferometer.—J. B. Bolton & O. B. Slee. (*Aust. J. Phys.*, Dec. 1953, Vol. 6, No. 4, pp. 420–433.) The factors governing the interference pattern of a Lloyd's-mirror type of interferometer at frequencies between 40 and 400 Mc/s are discussed and three systems for reducing the effect of background noise are described. Part 4: 1905 of 1952 (Bolton & Westfold).

523.854 : 621.396.822 : 550.510.535 1756

Galactic Radiation at Radio Frequencies: Part 6—Low-Altitude Scintillations of the Discrete Sources.—J. G. Bolton, O. B. Slee & G. J. Stanley. (*Aust. J. Phys.*, Dec. 1953, Vol. 6, No. 4, pp. 434–451.) Scintillations of four discrete sources of altitudes from 0° to 10° were observed at frequencies in the range 40 to 300 Mc/s. A strong correlation is established between the occurrence of scintillations and sporadic E. Part 5: 1755 above (Bolton & Slee).

550.385 : 551.55 : 551.510.535 1757

Note on Geomagnetic Disturbance as an Atmospheric Phenomenon.—E. H. Vestine. (*J. geophys. Res.*, Dec. 1953, Vol. 58, No. 4, pp. 539–541.) Possible relations between ionospheric winds and atmospheric electric currents associated with magnetic storms are briefly discussed.

550.385 : 551.556 1758

The Immediate Source of the Field of Magnetic Storms.—E. H. Vestine. (*J. geophys. Res.*, Dec. 1953, Vol. 58, No. 4, pp. 560–562.) Results of a comparison of mean values of initial phase of magnetic storms at Huancayo, Cheltenham (Md), Honolulu and San Juan (Porto Rico), indicate that major immediate sources of magnetic-field changes during storms in all phases are within or near the atmospheric region. See also 1757 above.

551.510.3 : 535.325 1759

A Statistical Survey of Atmospheric Index-of-Refraction Variation.—C. M. Crain, A. W. Straiton & C. E. von Rosenberg. (*Trans. Inst. Radio Engrs*, Oct. 1953, Vol. AP-1, No. 2, pp. 43–46.) Survey of microwave measurements made at heights of 2 000–25 000 ft over the Pacific Ocean, California and Ohio, using an airborne refractometer described by Crain & Deam (3493 of 1952).

551.510.3 : 535.325] : 681.142 1760

An Analogue Computer for the Solution of the Radio Refractive-Index Equation.—W. E. Johnson. (*J. Res. nat. Bur. Stand.*, Dec. 1953, Vol. 51, No. 6, pp. 335–342.) A computer in use at the C.R.P.L. comprises basic computation circuits incorporated in a bridge circuit. A 10-turn potentiometer is modified so as to give an approximately exponential variation of the resistance with the shaft rotation, corresponding to the curve of the saturated-water-vapour term in the equation.

551.510.535 **1761**
Irregularities in the Ionosphere.—R. Roy & J. K. D. Verma. (*J. geophys. Res.*, Dec. 1953, Vol. 58, No. 4, pp. 473–485.) Electron clouds in both the E and the F regions have been detected at vertical incidence by means of the equipment described by Banerjee & Roy (1677 of 1953). Photographs of typical echo records are shown; the power and persistence of the echoes are discussed.

551.510.535 **1762**
The Problem of Locating Inhomogeneities in the Ionosphere.—A. A. Gorozhankina. (*C. R. Acad. Sci. U.R.S.S.*, 21st Nov. 1953, Vol. 93, No. 3, pp. 459–461. In Russian.) Vertical soundings were made using two frequencies, $f_1 = 4{\cdot}3$ Mc/s and $f_2 = 6{\cdot}4$ Mc/s, chosen so that f_1 was reflected at the F_1 layer or the lower part of the F_2 layer and f_2 near the middle of the F_2 layer. The correlation factors ρ between pairs of the four parameters defined by $k = (\Delta\overline{R}/\Delta t)/\overline{R}$, derived from the amplitudes R of the ordinary and the extraordinary f_1 and f_2 signals recorded over periods t of a few minutes, are calculated and their significance is discussed. A high degree of correlation ($\rho > 0{\cdot}9$) between the four parameters indicates that the inhomogeneous region lies in the path common to the f_1 and f_2 signals.

551.510.535 **1763**
Recombination and Diffusion and Spread Echoes from the Ionosphere.—T. L. Eckersley. (*Proc. phys. Soc.*, 1st Dec. 1953, Vol. 66, No. 408B, pp. 1025–1038.) The inclusion of a diffusion term in the differential equation of layer formation, to account for ionization due to the influx of charged particles, leads to a solution in terms of a doubly periodic elliptic function. This solution indicates the possibility of a periodic variation of ionic density within the F layer such as to account for spread echoes which have been observed. Layers produced by ultraviolet radiation are many wavelengths thick, those due to particle ionization are only two or three wavelengths thick and are embedded in the background of ionization. The existence of auroral curtains provides further evidence of ionization by particles.

551.510.535 **1764**
Origin of the E Layer of the Ionosphere.—E. Bauer & Ta-You Wu. (*Phys. Rev.*, 1st Dec. 1953, Vol. 92, No. 5, pp. 1101–1105.) The electron densities for the E and F_1 layers resulting from photo-ionization of molecular and atomic oxygen are calculated on the basis of the work by Moses & Wu on the distribution of O_2 and of O in and above the oxygen dissociation layer (129 of 1952 and 102 of 1953). While the results are in reasonable agreement with observations for the E layer, it is not possible to account for the observed value of recombination coefficient for the F_1 layer by any theory based on processes involving only oxygen atoms and molecules.

551.510.535 **1765**
Method of Determining the True Height of the Ionospheric Layers: Part 2—Application of the Exact Value of the Refractive Index (Ordinary-Ray Case).—É. Argence & M. Mayot. (*J. geophys. Res.*, Dec. 1953, Vol. 58, No. 4, pp. 493–496. In French.) Values obtained using the exact value of the refractive index (Appleton-Hartree) are somewhat higher than those obtained with the approximate formulae used in part 1 (3290 of 1953).

551.510.535 **1766**
Recombination in the F Layers.—A. L. Stewart. (*Nature, Lond.*, 23rd Jan. 1954, Vol. 173, No. 4395, p. 165.) The mechanism proposed by Banerji (1058 of April) is critically discussed.

551.510.535 : 551.594.12 **1767**
Further Discussion of Kelso's Paper on a Method for Determination of the Distribution of Electron Density in the Ionosphere.—L. Kraus. (*J. geophys. Res.*, Dec. 1953, Vol. 58, No. 4, pp. 551–553.) Comment on 1997 of 1953 (Manning).

551.510.535 : 551.594.5 : 550.384 **1768**
Correlation of Magnetic, Auroral, and Ionospheric Variations at Saskatoon.—J. H. Meek. (*J. geophys. Res.*, Dec. 1953, Vol. 58, No. 4, pp. 445–456.) Analysis of measurements made between December 1951 and April 1952. A connection exists between the maximum elevation of auroral light above the northern horizon and the maximum amplitude of variation of H. Some types of sporadic-E reflecting layers appear more frequently during geomagnetic disturbances.

551.594.223 : 537.52 **1769**
Investigation of Ball Lightning by means of Models.—H. Nauer. (*Z. angew. Phys.*, Dec. 1953, Vol. 5, No. 12, pp. 441–450.) Experiments with gas discharges are described.

551.594.5 : 551.510.535 : 551.55 **1770**
On the Production of Glow Discharges in the Ionosphere by Winds.—O. R. Wulf. (*J. geophys. Res.*, Dec. 1953, Vol. 58, No. 4, pp. 531–538.) A source of excitation of the airglow and aurora may be found in the potential difference generated by zonal ionospheric winds cutting the earth's magnetic field. A return path of current flow or glow discharge in the higher ionosphere appears plausible. See also 3790 of 1945.

551.594.5 + 550.385] : 621.396.96 **1771**
Radio Echoes observed during Aurorae and Geomagnetic Storms using 35 and 74 Mc/s Waves simultaneously.—L. Harang & B. Landmark. (*J. atmos. terr. Phys.*, Jan. 1954, Vol. 4, No. 6, pp. 322–338.) Observations made at Kjeller in 1952 and at Tromsoe during early 1953 are reported. Echoes are only received when the aerial is directed towards the North; the echo amplitude has a maximum value when the aerial is directed horizontally, and decreases rapidly with increased elevation of the aerial. The results indicate that there is no correlation between the position of visible aurorae and echo range but that there is a correlation between the occurrence of high E_s-layer critical frequency and the reception of echoes. During geomagnetic storms echoes on 74 Mc/s are observed only during the more severe phases. Scattering from the ground or sea reflected back via an intense E_s layer is advanced in explanation of the observations.

LOCATION AND AIDS TO NAVIGATION

621.396.932 + 621.396.969.3 **1772**
Radio in the Service of the Seaman.—C. V. Robinson. (*Proc. Instn elect. Engrs*, Part I, Jan. 1954, Vol. 101, No. 127, pp. 27–28.) Chairman's address, I.E.E. Southern Centre. Communications and navigational aids are reviewed.

621.396.96 : 551.578 **1773**
Polarization of Radar Echoes from Meteorological Precipitation.—I. M. Hunter. (*Nature, Lond.*, 23rd Jan. 1954, Vol. 173, No. 4395, pp. 165–166.) A 'circularly polarized' radar for operation on a wavelength of 3·2 cm has been constructed, in which the polarization parameters are accurately known and controlled (voltage ellipticity ratio 0·99 on transmission and 0·95 on reception) and spurious signals due to departure from circularity are eliminated by means of a grating in the

waveguide which separates the incoming signals into two orthogonal components. Error signals can be reduced to a level of -50 db for point targets. Measurements on various types of precipitation are tabulated and discussed.

621.396.968 : 551.578.1 **1774**

Rain Clutter Measurements with C.W. Radar Systems Operating in the 8-mm Wavelength Band.—D. G. Kiely. (*Proc. Instn elect. Engrs*, Part III, March 1954, Vol. 101, No. 70, pp. 101–108.) Measurements were made using (*a*) a single aerial for transmission and reception, with polarization duplexing, and (*b*) separate aerials for transmission and reception, without duplexing. The amplitude of the clutter was given in terms of decibels below valve power level and was plotted against rainfall rate. The clutter was about 35 db greater in case (*a*) than in case (*b*). The maximum useful separation between the centres of the transmission and reception aerials is about 2 ft 6 in. No definite result was obtained for reduction of clutter by use of the orthogonal polarization technique, but the reduction reported by other observers was generally confirmed.

621.396.969.13 **1775**

Height Indicator.—P. H. Leidier. (*Ann. Radioélect.*, Oct. 1953, Vol. 8, No. 34, pp. 313–317.) The principle of operation of the height indicator Type IS 330 is described. It is associated with 10-cm radar equipment Type ER221 measuring azimuth and range. The c.r.o. presentation shows range as abscissa and height relative to the station as ordinate. Sources of error are considered and the accuracy of the system is determined. A block diagram of the unit is given.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 : 535.37 **1776**

Effect of Illumination Intensity, Field Strength and Temperature on the Dark Current and Photocurrent of Phosphors.—H. Gobrecht, D. Hahn & H. J. Kösel. (*Z. Phys.*, 8th Dec. 1953, Vol. 136, No. 3, pp. 285–292.) An experimental investigation of various phosphors is reported. The photocurrent increases linearly with the illumination intensity, at low intensities, and exponentially with electric field strength. It increases with temperature without exhibiting saturation in phosphors with a bimolecular luminescence mechanism, but reaches saturation early, or remains constant, in phosphors with a monomolecular mechanism. Results for both photocurrent and dark current are shown graphically.

535.215 : 546.682.231 **1777**

Photoconductivity of Indium Selenide.—D. E. Bode & H. Levinstein. (*J. opt. Soc. Amer.*, Dec. 1953, Vol. 43, No. 12, pp. 1209–1210.) Results are given of measurements of the spectral response of evaporated films at 20°C and -78°C.

535.215 : 546.817.221 **1778**

Photoelectromagnetic and Photoconductive Effects in Lead Sulphide Single Crystals.—T. S. Moss. (*Proc. phys. Soc.*, 1st Dec. 1953, Vol. 66, No. 408B, pp. 993–1002.) Bulk photoeffects have been observed. The quantum efficiency is approximately equal to unity between 0·9 and 2·9 μ, and falls rapidly at 3 μ. The optical activation energy is 0·41 eV. Carrier lifetimes, which are roughly proportional to the square of the resistivity, lie between 6×10^{-10} and 9×10^{-6} sec.

535.215.1 : 537.311.33 **1779**

Sensitization of the Internal Photoeffect in Semiconductors by Chlorophyll and Similar Pigments.—E. K. Putseiko & A. N. Terenin. (*C. R. Acad. Sci. U.R.S.S.*, 21st June 1953, Vol. 90, No. 6, pp. 1005–1008. In Russian.) The spectral response of ZnO is altered by adsorption of chlorophyll so that in addition to the 3 600–3 700 Å maximum, two additional maxima occur in the visible region, near 4 300 Å and between 6 600 and 6 700 Å. Similar results were obtained with other pigments. Other semiconductors investigated include PbO, ZnS, CdS, CdI_2 and PbI_2. The chlorophyll precipitates themselves did not show any photoeffects.

535.37 **1780**

Brightness Waves and Transitory Phenomena in the Quenching of Luminescence by Alternating Electric Fields.—G. Destriau. (*J. appl. Phys.*, Jan. 1954, Vol. 25, No. 1, pp. 67–71.) An experimental study was made of the variations of brightness of sulphide phosphors excited by means of a half-wave X-ray generator and subjected to a field alternating at the same frequency (50 c/s) as that of the X-ray generator. Secondary brightness maxima were observed lagging a quarter of a cycle behind the normal X-ray-excited maxima. The amplitude of the secondary maxima varied rapidly from cycle to cycle immediately after application of the field.

535.37 **1781**

Variations in the Decay of Phosphorescence with Frequency of Applied Electric Field.—K. W. Olson. (*Phys. Rev.*, 1st Dec. 1953, Vol. 92, No. 5, p. 1323.) Results of measurements on (Zn, Cd)S-Cu phosphors at 80, 800 and 5 000 c/s are reported.

535.37 **1782**

Mechanism of Impurity Poisoning in the Luminescence of Zinc Sulfide Phosphors with Manganese Activator.—R. H. Bube, S. Larach & R. E. Shrader. (*Phys. Rev.*, 1st Dec. 1953, Vol. 92, No. 5, pp. 1135–1139.) Poisoning effects of Fe, Co and Ni were studied. The relative importance of three possible poisoning mechanisms considered depends on the activator and poison proportions and on the energy of the exciting radiation.

535.37 **1783**

Electroluminescence in Thin Films of ZnS:Mn.—R. E. Halsted & L. R. Koller. (*Phys. Rev.*, 15th Jan. 1954, Vol. 93, No. 2, pp. 349–350.) No luminescence is produced by application of a unidirectional field below breakdown strength, apart from flashes on application and removal of the field. With alternating fields, the brightness varies at a frequency twice that of the applied field. The dependence of the effect on temperature and frequency is investigated.

535.37 **1784**

The Penetration of Electrons into Luminescent Material.—W. Ehrenberg & J. Franks. (*Proc. phys. Soc.*, 1st Dec. 1953, Vol. 66, No. 408B, pp. 1057–1066.) The penetration of electrons of energies between 10 and 40 keV into single crystals of Tl-activated iodides of K, Rb and Cs, and of tungstates of Ca and Cd, and a luminescent plastic, was measured on microphotographs of the luminous region. The diameter of the electron beam was $<0{\cdot}75\,\mu$.

535.37 **1785**

The Manganese Emission in ABF_3 Compounds.—H. A. Klasens, P. Zalm & F. O. Huysman. (*Philips Res. Rep.*, Dec. 1953, Vol. 8, No. 6, pp. 441–451.) Investigation of the fluorescence under cathode-ray excitation of perovskite-type compounds activated with Mn. The A metals used were Na, K, Rb and Cs, the B metals Mg, Zn, Cd, Ca and Sr. A tentative explanation is given of the shift from orange to green occurring with increase of the Mn-F distance.

535.37 1786

Electroluminescent Zinc Sulfide Phosphors.—H. H. Homer, R. M. Rulon & K. H. Butler. (*J. electrochem. Soc.*, Dec. 1953, Vol. 100, No. 12, pp. 566–571.) The preparation of phosphors giving blue, green and yellow electroluminescence is described. The activation of the green and blue phosphors is produced by Cu together with Pb, that of the yellow phosphors by Cu with Pb and Mn. A controlled amount of chloride is included in all the phosphors.

535.37 : 537.311.33 1787

Electroluminescence of Insulated Particles.—L. Burns. (*J. electrochem. Soc.*, Dec. 1953, Vol. 100, No. 12, pp. 572–579.) An examination is made of possible mechanisms whereby luminescence could be produced in a region on or near the surface of a semiconductor particle embedded in a dielectric and subjected to an electric field. The localization of the field, the presence of electrons in the conduction band, the colour shift with frequency in some phosphors, and the efficiency of the phosphor are discussed.

536.587 : 548.55 : 546.289 1788

Temperature Regulator for Germanium Metallurgy.—G. Lehmann & C. Meuleau. (*Onde élect.*, Dec. 1953, Vol. 33, No. 321, pp. 678–683.) The importance of temperature control in the production of single crystals for Ge diodes and triodes is emphasized. In the system described the supply to an induction furnace is controlled by means of a thermometric resistance bridge and a servo motor. The temperature stability achieved is within $\pm\frac{1}{6}$°C at 930°C, for mains voltage variations up to 10%.

537.226 1789

Origin of Ferroelectricity in Barium Titanate and other Perovskite-Type Crystals.—H. D. Megaw. (*Acta cryst., Camb.*, 10th Nov. 1952, Vol. 5, Part 6, pp. 739–749.)

537.226.2 : 546.23 1790

The Dielectric Constant of Amorphous Selenium at Wavelengths of 1 cm and 3 cm.—Y. Klinger & E. W. Saker. (*Proc. phys. Soc.*, 1st Dec. 1953, Vol. 66, No. 408B, p. 1117.) A cylindrical-cavity resonator operating in the H_{01} mode was used for measurements. The weighted mean value obtained is $6{\cdot}37_6$.

537.228.1 : 548.0 1791

Dynamic Determination of Elastic and Piezoelectric Coefficients.—R. Bechmann. (*Telefunken Ztg*, Dec. 1953, Vol. 26, No. 102, pp. 353–365.) The determination of the coefficients of piezoelectric crystals from measurements of resonance frequency, crystal dimensions, density and dynamic capacitance is discussed in detail; crystals of various structures are considered.

537.311.31 1792

The Propagation of Electrons in a Strained Metallic Lattice.—S. C. Hunter & F. R. N. Nabarro. (*Proc. roy. Soc. A*, 22nd Dec. 1953, Vol. 220, No. 1143, pp. 542–561.) The problem is studied using a perturbation technique in which the perturbing potential is proportional to the elastic strain rather than to the displacement. For the approximation of nearly free electrons, the resulting potential depends only on the Fermi energy of the electrons and not on their interaction with the ionic lattice. In a higher approximation this interaction is taken into account. The method is used to estimate the resistivity produced by dislocations of edge and screw types in Na and Cu.

537.311.31 1793

Electrical Conductivity of Metals at High Current Density.—E. S. Borovik. (*C. R. Acad. Sci. U.R.S.S.*, 1st Aug. 1953, Vol. 91, No. 4, pp. 771–774. In Russian.) Measurements were made using both steady currents at very low temperatures and current pulses of densities of the order of 10^6 A/cm² at 20·4° and 78°K. Results indicate that Pt, W and Cu obey Ohm's law at densities up to 5–8 × 10^6 A/cm²; large deviations were found for Bi at 0·5–1 × 10^6 A/cm².

537.311.31 1794

The Extra-Resistivity due to Vacancies in Copper, Silver and Gold.—P. Jongenburger. (*Appl. sci. Res.*, 1953, Vol. B3, Nos. 4/5, pp. 237–248.)

537.311.33 1795

Some Problems in the Diffusion of Minority Carriers in a Semiconductor.—S. Visvanathan & J. F. Battey. (*J. appl. Phys.*, Jan. 1954, Vol. 25, No. 1, pp. 99–102.) "The exact solutions to the problems of the diffusion of minority carriers involved in the measurement of surface recombination velocity in a semiconductor with a sample geometry bounded by two infinite planes are presented. The reduction of the exact solutions to simple forms used in the analysis of experimental data is shown."

537.311.33 1796

Contribution to the Theory of Conduction in Semiconductors.—W. H. Isay. (*Ann. Phys., Lpz.*, 15th Dec. 1953, Vol. 13, Nos. 6/8, pp. 327–348.) Given the conductivity rise and decay curves of a CdS crystal, and assuming a three-energy-band model, the various parameters of the associated system of differential equations can be calculated.

537.311.33 1797

Effect of Traps on Carrier Injection in Semiconductors.—H. Y. Fan. (*Phys. Rev.*, 15th Dec. 1953, Vol. 92, No. 6, pp. 1424–1428.) Two effects are considered, (*a*) that on photoconductivity, and (*b*) that on the spread of excess carrier concentration under an applied field. Trapped minority carriers, by causing an increase in majority carrier concentration, give rise to increased photoconductivity which may vary nonlinearly with light intensity and have a very long time constant. An expression is derived for the distribution of excess minority carriers as a function of time and space, neglecting diffusion. Results are in agreement with observations.

537.311.33 1798

New Semiconducting Compounds: Part 2.—H. Welker. (*Z. Naturf.*, April 1953, Vol. 8a, No. 4, pp. 248–251.) Measurements of conductivity as a function of temperature are reported for specimens of the crystalline compounds InSb, GaSb and AlSb; the widths of the respective forbidden energy bands are determined. Characteristics are shown of the rectifier effect in AlSb, GaSb, GaAs and InP and of the transistor effect in InP. Part 1: 1454 of May.

537.311.33 : 539.165 : 621.314.6 1799

The Electron-Voltaic Effect in *p-n* Junctions induced by Beta-Particle Bombardment.—P. Rappaport. (*Phys. Rev.*, 1st Jan. 1954, Vol. 93, No. 1, pp. 246–247.) Current/voltage characteristics are shown for Si-alloy wafer-type units bombarded from a 50-millicurie Sr^{90}-Y^{90} source. One of the units has been used as power supply for an a.f. transistor oscillator. The efficiency of conversion of the radioactive power is about 0·4%. Ge units were also investigated.

537.311.33 : 546.23 1800

Frequency Dependence of the Electrical Conductivity of Polycrystalline Selenium.—H. Gobrecht & H. Hamisch. (*Z. Phys.*, 17th Nov. 1953, Vol. 136, No. 2, pp. 234–247.) The conductivity of Se layers deposited on glass by evaporation was measured at frequencies up to 14 Mc/s and at temperatures between −25° and +50°C. The results are shown graphically and are discussed in relation to the Maxwell-Wagner theory for a capacitor with two different dielectrics.

537.311.33 : 546.23 : 548.55 **1801**
Electrical Properties of Selenium: Part 3 — Microcrystalline Selenium Metal Doped.—H. W. Henkels & J. Maczuk. (*J. appl. Phys.*, Jan. 1954, Vol. 25, No. 1, pp. 1–11.) Measurements were made on two crystalline forms of pure Se and on samples with small amounts of various metals added. Thermoelectric power and the d.c. and 200-Mc/s resistivities are studied as functions of temperature and a theoretical energy-band model for Se is developed. Part 2: 1335 of 1952 (Henkels).

537.311.33 : [546.28 + 546.289 **1802**
Mobility of Impurity Ions in Germanium and Silicon.—J. C. Severiens & C. S. Fuller. (*Phys. Rev.*, 1st Dec. 1953, Vol. 92, No. 5, pp. 1322–1323.) The diffusion constants of Li in Si and Ge, and of Cu in Ge, as calculated from the experimentally determined mobilities, are in agreement with previously published results.

537.311.33 : [546.28 + 546.289 **1803**
Direct Measurement of the Dielectric Constants of Silicon and Germanium.—W. C. Dunlap, Jr, & R. L. Watters. (*Phys. Rev.*, 15th Dec. 1953, Vol. 92, No. 6, pp. 1396–1397.) Single-crystal Si and Ge samples were doped with gold to produce material of sufficiently high resistivity at the working temperature of 77°K, and were formed into small parallel-plate capacitors. The permittivity of Si is $11{\cdot}7 \pm 0{\cdot}2$ at 1 Mc/s and remains constant within $\pm$ 1% over the range 500 c/s–30 Mc/s. The corresponding value for Ge is $15{\cdot}8 \pm 0{\cdot}2$; its much greater apparent variation with frequency is ascribed to sample inhomogeneity.

537.311.33 : [546.28 + 546.289] : 532.2 **1804**
Shapes of Floating Liquid Zones between Solid Rods.—P. H. Keck, M. Green & M. L. Polk. (*J. appl. Phys.*, Dec. 1953, Vol. 24, No. 12, pp. 1479–1481.) The method described by Keck & Golay (*Phys. Rev.*, 15th March 1953, Vol. 89, No. 6, p. 1297) for crystallizing Si from a melt involves holding the molten Si between two vertically aligned solid rods of Si. The shape and stability of the molten Si are investigated theoretically.

537.311.33 : 546.28 **1805**
Gold as a Donor in Silicon.—E. A. Taft & F. H. Horn. (*Phys. Rev.*, 1st Jan. 1954, Vol. 93, No. 1, p. 64.) The experimentally determined resistivity/temperature characteristic indicates that Au produces a donor level 0·33 eV above the occupied band in Si.

537.311.33 : 546.289 **1806**
Lattice-Scattering Mobility in Germanium.—F. J. Morin. (*Phys. Rev.*, 1st Jan. 1954, Vol. 93, No. 1, pp. 62–63.) The temperature dependence of the lattice-scattering mobility of holes and electrons was determined from conductivity measurements. The variation with temperature of the ratio Hall-mobility/conductivity-mobility for holes suggests that the valence band is composed of multiple surfaces of minimum energy.

537.311.33 : 546.289 **1807**
Preparation of *p-n* Junctions by Surface Melting.—K. Lehovec & E. Belmont. (*J. appl. Phys.*, Dec. 1953, Vol. 24, No. 12, pp. 1482–1484.) A r.f.-heating method of preparing single or multiple *p-n* junctions is described. The influence of (*a*) heater temperature, (*b*) forced cooling, and (*c*) thickness of the slice of the single-crystal semiconductor on the position of the solid/liquid interface is calculated from a one-dimensional model.

537.311.33 : 546.289 : 535.215 **1808**
Infrared Photoconductivity due to Neutral Impurities in Germanium.—E. Burstein, J. W. Davisson, E. E. Bell, W. J. Turner & H. G. Lipson. (*Phys. Rev.*, 1st Jan. 1954, Vol. 93, No. 1, pp. 65–68.) Experimental determination at liquid-He temperatures of the photoconductivity in the wavelength range 1–38 μ of *n*-type and *p*-type Ge containing various donor and acceptor impurities.

537.311.33 : 546.41-31 **1809**
Calcium Oxide, an Amphoteric Semiconductor.—K. Hauffe & G. Tränckler. (*Z. Phys.*, 17th Nov. 1953, Vol. 136, No. 2, pp. 166–178.) Conductivity measurements on CaO at 600°C indicate that at oxygen pressures $>10^{-2}$ Torr CaO is a *p*-type semiconductor, at pressures $<10^{-2}$ Torr an *n*-type semiconductor. Addition of Li_2O results in *p*-type conduction even in a high vacuum, the conductivity increasing with increasing O_2 pressure and decreasing in high vacuum. Addition of Y_2O_3 gives opposite results, and is beneficial in alkaline-earth oxide cathodes.

537.311.33 : 546.47-31 **1810**
Conductivity and Hall Effect of ZnO at Low Temperatures.—S. E. Harrison. (*Phys. Rev.*, 1st Jan. 1954, Vol. 93, No. 1, pp. 52–62.) Results of measurements on single-crystal and polycrystalline specimens are used in assessing the adequacy of various energy-level models of ZnO.

537.311.33 : 546.482.21 **1811**
Demonstration of Traps in CdS Single Crystals.—W. Muscheid. (*Ann. Phys., Lpz.*, 15th Dec. 1953, Vol. 13, Nos. 6/8, pp. 322–326.) Results of an experimental investigation of the effect of illumination and rate of heating or cooling on the dark current indicate that traps are produced by incorporating oxygen in the CdS crystals.

537.311.33 : 546.482.21 : 535.215 **1812**
The Effect of Oxygen on the Conductivity of CdS Single Crystals.—W. Muscheid. (*Ann. Phys., Lpz.*, 15th Dec. 1953, Vol. 13, Nos. 6/8, pp. 305–321.) Measurements are reported of the dark conductivity and photoconductivity at various pressures of air and oxygen, in the temperature range 300°–700°K.

537.311.33 : 546.482.21 : 538.632 **1813**
Hall-Effect Measurements on CdS Crystals. — H. Diedrich. (*Ann. Phys., Lpz.*, 15th Dec. 1953, Vol. 13, Nos. 6/8, pp. 349–352.) Results of measurements in the absorption region at various illumination intensities give excess-electron mobilities of 20–400 cm/s per V/cm. The mobility depends on the conduction-electron concentration.

537.311.33 : 546.482.21.03 **1814**
Optical and Electrical Properties of Cadmium-Sulphide Single Crystals.—H. Gobrecht & A. Bartschat. (*Z. Phys.*, 17th Nov. 1953, Vol. 136, No. 2, pp. 224–233.)

537.311.33 : 546.561-31 **1815**
Electrical Conductivity of Cuprous Oxide.—A. I. Andrievski, V. I. Voloshchenko & M. T. Mischchenko. (*C. R. Acad. Sci. U.R.S.S.*, 1st June 1953, Vol. 90, No. 4, pp. 521–523. In Russian.) Results of an experimental investigation indicate that the conductivity of the Cu_2O surface layer is proportional to the number of crystalline grains per unit area and is independent of the method of production.

537.311.33 : 546.682.86 **1816**
A Note on the Semiconducting Compound InSb.—F. A. Cunnell, E. W. Saker & J. T. Edmond. (*Proc. phys. Soc.*, 1st Dec. 1953, Vol. 66, No. 408B, pp. 1115–1116.) Measurements were made on an ingot purified by zone

melting. The variations of the conductivity and of the Hall constant with distance along the ingot from the pure end, and with temperature, are shown graphically. The results are discussed briefly.

537.311.33 + 538.63 : 546.762.21 **1817**

Electrical and Galvanomagnetic Properties of Sulphides of Chromium.—N. P. Grazhdankina & I. G. Fakidov. (*C. R. Acad. Sci. U.R.S.S.*, 21st Nov. 1953, Vol. 93, No. 3, pp. 429–430. In Russian.) Classification according to resistivity places these sulphides intermediate between metals and semiconductors. Measurements were made of (*a*) the resistance at very low temperatures, (*b*) the Hall effect, and (*c*) the magnetoresistance effect. The results of (*a*) and (*c*) are shown graphically.

537.311.33 : 546.817.221 **1818**

Interpretation of Hall Effect and Resistivity Data in PbS and Similar Binary Compound Semiconductors.—W. W. Scanlon. (*Phys. Rev.*, 15th Dec. 1953, Vol. 92, No. 6, pp. 1573–1575.) Discrepancies in published data on the width of the forbidden energy gap for PbS are accounted for by the occurrence of thermal changes in composition during experimental runs. Measurements confined to temperatures below 500°K, for which such changes do not take place, give a figure of $0 \cdot 37 \pm 0 \cdot 01$ eV, in agreement with optical data.

537.311.33 : 621.314.634 **1819**

Mode of Operation of CdSe Intermediate Layers in Selenium Rectifiers.—A. Hoffmann & F. Rose. (*Z. Phys.*, 17th Nov. 1953, Vol. 136, No. 2, pp. 152–165.) An experimental investigation. Results indicate that the blocking action in commercial Se rectifiers with counter electrodes containing Cd occurs at the CdSe/Se boundary, CdSe and Se being *n*-type and *p*-type semiconductors respectively. The CdSe layer has not a sufficiently high resistivity to act as an 'isolating' hole layer with multipoint contact rectification.

537.312.8 **1820**

Effect of Composition on Change of Electrical Resistance of Ni-Mn Alloys in a Magnetic Field.—I. Ya. Dekhtyar. (*C. R. Acad. Sci. U.R.S.S.*, 1st Dec. 1953, Vol. 93, No. 4, pp. 637–639. In Russian.) The variation of the fractional change of resistance of Ni-Mn alloys with variation of the percentage of Mn in the alloy was determined experimentally at several values of the magnetic field strength. The influence of Cu or Sn impurity atoms was also investigated. The results are presented graphically.

538.213 : 621.3.042.15 **1821**

Effective Permeabilities of Regular Configurations of Spherical Ferromagnetic Particles.—I. Lucas. (*Frequenz*, Oct. 1953, Vol. 7, No. 10, pp. 289–295.) Estimation of effective permeability is based on an approximation to the magnetic-field distribution near the points of contact of the particles. For the closest packing, the effective permeability, μ_w, is related to the bulk permeability of the material, μ, by the formula $\mu_w = \sqrt{2}\,\pi \log \mu$. Agreement with measurements on sendust (Fe-Si-Al-alloy) and carbonyl-iron powder cores is satisfactory.

538.221 **1822**

An Amplitude- and Temperature-Dependent After Effect in α-Fe at −70°C.—G. Sorger. (*Z. angew. Phys.*, Nov. 1953, Vol. 5, No. 11, pp. 406–413.) An anomaly in the complex permeability observed at about −70°C in Fe and Fe-Si alloys with α-type lattice was investigated. The after-effect decreases with an increase of applied field strength. The activation energy is 0·3 eV/atom. See also 1631 of 1952 (Feldtkeller et al.).

538.221 **1823**

Developments in Sintered Magnetic Materials.—J. L. Salpeter. (*Proc. Inst. Radio Engrs*, March 1954, Vol. 42, No. 3, pp. 514–526.) Reprint. See 3334 of 1953.

538.221 : 538.561.029.6 **1824**

Ferromagnetic Resonance using Waves from a Mass Emitter.—K. A. Volkova. (*C. R. Acad. Sci. U.R.S.S.*, 1st April 1953, Vol. 89, No. 4, pp. 655–658. In Russian.) Measurements were made of the heat dissipated in Fe and in Ni specimens subjected to a steady magnetic field of 100–13 500 oersted and a superimposed h.f. field of mean frequency variable between 21·4 and 125 kMc/s. The latter field was produced by the wide-band mass emitter described by Glagoleva-Arkad'eva (2106 of 1943). The experimental results are shown graphically. Ferromagnetic dispersion is briefly discussed.

538.221 : 538.632 **1825**

Spontaneous Hall Effect in Ferromagnetics.—J. Smit & J. Volger. (*Phys. Rev.*, 15th Dec. 1953, Vol. 92, No. 6, pp. 1576–1577.) Results of measurements on several specimens of Ni and Ni alloys are tabulated and discussed.

538.221 : 621.318.1 : 621.314.22.018.75 **1826**

Pulse Permeability and Losses in Magnetic Materials subjected to Rectangular D.C. Pulses.—T. Einsele. (*Frequenz*, Oct. 1953, Vol. 7, No. 10, pp. 281–289.) Measurements of maximum pulse permeabilities and total losses under steady operating conditions are reported for cores of siferrit 2000 Tr 7, permenorm 3601 and mumetal. When simplifying assumptions are made, the maximum pulse power that can be dealt with by a transformer of specified core material, for given pulse duration and repetition rate, can be calculated.

538.221 : 621.318.124 + 621.318.134 **1827**

Ferromagnetic Ferrites.—Y. Lescroël. (*Câbles & Transm.*, Oct. 1953, Vol. 7, No. 4, pp. 273–292.) Characteristic properties of ferrites are examined with reference to three materials developed since 1950, namely (*a*) fermalite, a mixed Mn-Zn ferrite, of which three principal types are manufactured; Type 1002 for coils operating at frequencies up to 1 Mc/s; Type 2001 for wide-band transformers, and reaction coils; Type 3001 for h.f. power and pulse transformers; (*b*) fernilite, a Ni-Zn ferrite for applications at frequencies up to 100 Mc/s; (*c*) fercolite, a Co ferrite used for light-weight permanent magnets.

538.221 : 621.318.124 + 621.318.134 **1828**

Magnetic Resonance in Ferrimagnetics. — R. K. Wangsness. (*Phys. Rev.*, 1st Jan. 1954, Vol. 93, No. 1, pp. 68–71.) The general expression for the magnetic resonance frequency of the two-sublattice model of a ferrimagnetic crystal has essentially the same form as that originally obtained for the ferromagnetic case provided that the product of the molecular field coefficient and the net magnetization is large compared to the applied and anisotropy fields.

538.221 : 621.318.134 **1829**

Effect of Cross-Section Area and Compression upon the Relaxation in Permeability for Toroidal Samples of Ferrites.—R. E. Alley, Jr, & F. J. Schnettler. (*J. appl. Phys.*, Dec. 1953, Vol. 24, No. 12, pp. 1524–1525.) The real and the imaginary components of permeability of MnZn and NiZn ferrites were determined experimentally over the frequency range 25 kc/s–2 Mc/s. Only the MnZn showed a change in the relaxation frequency with change of cross-section. The two ferrites also responded differently to compression. The results are shown graphically.

538.221 : 621.318.134 **1830**

Magnetization in Nickel Ferrite-Aluminates and Nickel Ferrite-Gallates.—L. R. Maxwell & S. J. Pickart. (*Phys. Rev.*, 1st Dec. 1953, Vol. 92, No. 5, pp. 1120–1126.) Measurements were made on materials prepared by substituting trivalent Al and Ga for part or all of the trivalent Fe in $NiO.Fe_2O_3$. Graphs show the variation with composition of the unit cell size and of the saturation magnetization extrapolated to 0°K.

538.221 : 621.318.134 **1831**

Ionic Distribution deduced from the g-Factor of a Ferrimagnetic Spinel: Ti^{4+} in Fourfold Co-ordination.—E. W. Gorter. (*Nature, Lond.*, 16th Jan. 1954, Vol. 173, No. 4394, pp. 123–124.)

538.221 : 621.318.134 **1832**

Ferromagnetic Resonance of Nickel-Zinc Ferrites.—E. I. Kondorski & N. A. Smol'kov. (*C. R. Acad. Sci. U.R.S.S.*, 11th Nov. 1953, Vol. 93, No. 2, pp. 237–240. In Russian.) The relaxation times and g-factors of three ferrites were calculated from results of an experimental determination of the real and imaginary components of the magnetic permeability and the dielectric constant at wavelengths of 3·2 and 8·6 cm with the sample placed in a steady magnetic field. The experimental method is described. Results are tabulated and, for one ferrite, shown graphically.

538.222/.224 **1833**

Resonance Absorption in Metals at Centimetre Wavelengths.—S. G. Salikhov. (*C. R. Acad. Sci. U.R.S.S.*, 11th Nov. 1953, Vol. 93, No. 2, pp. 241–244.) Measurements are reported on 28 pure metals at temperatures of 90° and 290°K using a frequency of 9·378 kMc/s. The table of results gives the atomic susceptibility, paramagnetic absorption, g-factor, line width and line intensity. Some results are also shown graphically.

538.567 : 539.234 : 546.57 **1834**

The Electrical Properties of Thin Evaporated Silver Films at 3 000 Mc/s.—F. J. Tischer. (*Z. angew. Phys.*, Nov. 1953, Vol. 5, No. 11, pp. 413–415.) The films discussed are of a type used in microwave attenuators and terminating resistors. Radiation penetration depth and reflection and transmission coefficients are investigated, taking into account the effect of the mica base and the displacement current in the layer. The theoretical and experimental values of the reflection and transmission coefficients are compared and the material constants are calculated.

538.653.2 : [546.72 + 546.74 **1835**

Effect of Plastic Deformation on the Magnetization Curves of Iron and Nickel in High-Strength Magnetic Fields.—V. V. Parfenov. (*C. R. Acad. Sci. U.R.S.S.*, 21st Nov. 1953, Vol. 93, No. 3, pp. 435–438. In Russian.) Report of experimental investigation using field strengths up to 10 000 oersted.

547.476.3 **1836**

Dielectric Properties of Some Double Tartrates.—F. Jona & R. Pepinsky. (*Phys. Rev.*, 15th Dec. 1953, Vol. 92, No. 6, p. 1577.) No dielectric anomaly is found for the Na-Rb salt in the temperature range 4·2°K–333°K but a dielectric anomaly appears in the $NaNH_4$ salt at 109°K.

548.0 : 539.378.3 **1837**

Some Predicted Effects of Temperature Gradients on Diffusion in Crystals.—A. D. LeClaire: J. A. Brinkman: W. Shockley. (*Phys. Rev.*, 15th Jan. 1954, Vol. 93, No. 2, pp. 344–346.) Discussion on 1122 of April and author's reply.

549.514.51 : 539.31 **1838**

Anelasticity of Quartz.—R. K. Cook & R. G. Breckenridge. (*Phys. Rev.*, 15th Dec. 1953, Vol. 92, No. 6, pp. 1419–1423.) Measurements were based on the method of Cook & Weissler (650 of 1951). Anelasticity, i.e. internal friction, has a maximum value at a temperature between room temperature and the quartz inversion temperature of 573°C, the value being greatly increased when foreign atoms are present in the lattice. Activation energies and relaxation time constants were deduced for 4 quartz bars studied.

621.315.61 **1839**

Progress in Dielectrics.—S. Whitehead. (*Elect. Rev., Lond.*, 11th Dec. 1953, Vol. 153, No. 24, pp. 1309–1313.) A survey based on a paper presented at the British Association.

621.315.61 : 538.569.3 **1840**

Absorption of Millimeter Waves in Dielectric Solids.—W. L. Brooks, J. H. Greig, C. Pine, W. G. Zoellner & J. H. Rohrbaugh. (*J. opt. Soc. Amer.*, Dec. 1953, Vol. 43, No. 12, pp. 1191–1194.) The real and imaginary parts of the complex refractive index of polystyrene, polymethyl methacrylate and a high-melting-point paraffin were determined at wavelengths between 1·79 and 4·18 mm from measurements of the magnitudes and phase angles of the transmission coefficients, using harmonics of a 12·5 mm-λ Type-3J31 magnetron.

621.315.612.6.011.5.029.42 **1841**

The Dielectric Relaxation of Glass and the Pseudocapacity of Metal-to-Glass Interfaces, measured at extremely Low Frequencies.—J. Volger, J. M. Stevels & C. van Amerongen. (*Philips Res. Rep.*, Dec. 1953, Vol. 8, No. 6, pp. 452–470.) The dispersion of the dielectric constant was determined experimentally over the frequency range 0·1–20 c/s, at various temperatures. From theoretical considerations, the main relaxation time is inversely proportional to the transition probability of Na^+ ions jumping between adjacent interstices. The very high frequency- and temperature-dependent electrode capacitances are discussed.

669-426.2 **1842**

The Production of Fine Wires by Electrolytic Polishing.—H. R. Haines & B. W. Mott. (*J. sci. Instrum.*, Dec. 1953, Vol. 30, No. 12, pp. 459–460.) A method suitable for reducing the diameter of brittle or easily oxidized metal wires is described; it has been used successfully for Th, U, Zr and nichrome.

MATHEMATICS

517.51/.52 **1843**

On the Summation of Infinite Series in Closed Form.—A. D. Wheelon. (*J. appl. Phys.*, Jan. 1954, Vol. 25, No. 1, pp. 113–118.)

517.942.9 **1844**

The Meaning of the Vector Laplacian.—P. Moon & D. E. Spencer. (*J. Franklin Inst.*, Dec. 1953, Vol. 256, No. 6, pp. 551–558.) The distinction between the vector and scalar Laplace operators is discussed. A general equation is developed for the vector Laplace operator in any orthogonal curvilinear coordinate system; use of this equation enables electrodynamic problems to be formulated simply by means of the vector Helmholtz equation.

517.949.8 **1845**

On Mildly Nonlinear Partial Difference Equations of Elliptic Type.—L. Bers. (*J. Res. nat. Bur. Stand.*, Nov. 1953, Vol. 51, No. 5, pp. 229–236.) Justification is given for the use of the finite-difference method for the numerical treatment of these equations.

519.271.3 : 620.113.2 1846

The Efficiency of Sequential Sampling for Attributes: Part 2—Practical Applications.—H. C. Hamaker. (*Philips Res. Rep.*, Dec. 1953, Vol. 8, No. 6, pp. 427–433.) The simplified equations valid under Poisson conditions are applied to the design of a slide rule from which the operating characteristic can be read off when two fundamental parameters are known. Part 1: 2715 of 1953.

519.272.119 1847

Error due to Finite Integration Time-Limits in the Determination of Autocorrelation Functions.—P. Blassel. (*Ann. Télécommun.*, Dec. 1953, Vol. 8, No. 12, pp. 406–414.)

MEASUREMENTS AND TEST GEAR

529.786 1848

Annual Fluctuations in Quartz Clock Error and Frequency Drift.—H. J. M. Abraham. (*Nature, Lond.*, 9th Jan. 1954, Vol. 173, No. 4393, pp. 73–74.) Report of an investigation of the performance of several quartz clocks in the southern hemisphere.

538.652.08 1849

A Method of Measuring Magnetostriction.—A. W. Cochardt. (*J. appl. Phys.*, Jan. 1954, Vol. 25, No. 1, pp. 91–95.) The absolute magnitude of magnetostriction is determined from torsion tests on unmagnetized and magnetized wires. The error is about 3% at stresses > 1 000 p.s.i.

621.317.3 : 621-526 1850

Measurement of L.F. Response of Servo-Mechanism Components.—E. J. P. Long. (*Engineer, Lond.*, 4th Dec. 1953, Vol. 196, No. 5106, pp. 722–725.) A description is given of apparatus for producing a sinusoidal signal in the frequency range 0·5–20 c/s, together with suitable equipment for measuring attenuation and phase shift at these frequencies. The method consists in operating a magslip so that one complete revolution of the rotor corresponds to one cycle of the required low frequency, and using the magslip as a mechanical modulator for a h.f. carrier. The display end of the apparatus consists of a two-channel oscilloscope.

621.317.3 : 621.385.029.6 1851

Rapid Method of testing Magnetrons in the Nonoperating State.—J. W. Dodds. (*Le Vide*, Nov. 1953, Vol. 8, No. 48, pp. 1429–1431.) The resonance frequency and the effect of loading are determined using an arrangement in which the magnetron terminates a line fed by a frequency-modulated klystron. The signal is picked up by a standing-wave detector, and the resonance condition is indicated by the shape of the curve displayed on a c.r.o. Under-coupling and over-coupling are easily identifiable. Relevant theory is given.

621.317.311 1852

The Measurement of Very Small Direct Currents.—M. W. Jervis. (*Electronic Engng*, March 1954, Vol. 26, No. 313, pp. 100–105.) A review of methods used for measuring currents in the range 10^{-16}–10^{-6}A, with particular reference to thermionic and capacitor modulator electrometers. 32 references.

621.317.329 : 621.373.413 1853

Simple Method of Measurement of the Distribution of the Electromagnetic Field in a Resonant Cavity.—A. Septier. (*C. R. Acad. Sci., Paris*, 8th Feb. 1954, Vol. 238, No. 6, pp. 658–660.) The determination of the field is based on observed variations of resonance frequency with displacement of a suitably shaped obstacle inside the cavity. Frequency variations of 1 part in 10^5 can be detected.

621.317.335.3 : 546.171.1 1854

Dispersion in Ammonia in 3 cm Region.—Krishnaji & P. Swarup. (*Z. Phys.*, 8th Dec. 1953, Vol. 136, No. 3, pp. 374–378. In English.) The variation of electric susceptibility with pressure was determined experimentally using a standing-wave technique; the results are tabulated and shown graphically. The value of the susceptibility at atmospheric pressure is $5{\cdot}3 \times 10^{-3}$.

621.317.342 1855

Measurements of Phase Angles.—A. van Weel. (*Philips Res. Rep.*, Dec. 1953, Vol. 8, No. 6, pp. 471–475.) Discussion of an improved system based on the method described previously (3659 of 1953).

621.317.382 : 538.632 : 537.311.33 1856

Application of the Hall Effect in a Semiconductor to the Measurement of Power in an Electromagnetic Field.—H. M. Barlow. (*Nature, Lond.*, 2nd Jan. 1954, Vol. 173, No. 4392, pp. 41–42.)

621.317.443.029.4/.5 1857

A Radio-Frequency Permeameter.—P. H. Haas. (*J. Res. nat. Bur. Stand.*, Nov. 1953, Vol. 51, No. 5, pp. 221–228.) Theory and description of a permeameter for frequencies up to 20 Mc/s. The measurement of permeability and losses is made by inserting the ferromagnetic material in the form of a toroidal core into the short-circuited secondary of a transformer, the primary of which is connected to a r.f. bridge or Q meter.

621.317.7 : 537.54 : 621.396.822.029.64 1858

A Microwave Wide-Band Noise-Generator in Coaxial-Line Form.—W. Friz. (*Fernmeldetech. Z.*, Dec. 1953, Vol. 6, No. 12, pp. 583–587.) A description is given of the characteristics and construction of a gas-discharge noise generator for the 6·5–20-cm wavelength range, with the discharge path inside a coaxial line. The advantages of this type over the diode type are (*a*) nearly constant internal resistance, and (*b*) constant power distribution over a wide frequency range. The upper and lower frequency limits are discussed.

621.317.733 1859

Resistance Bridge Sensitivity and Output Formulas.—P. M. Andress. (*Elect. Engng, N.Y.*, Dec. 1953, Vol. 72, No. 12, p. 1095.) Digest of paper to be published in *Trans. Amer. Inst. elect. Engrs*, 1953, Vol. 72.

621.317.733 : 621.314.2 1860

Bridges with Coupled Inductive Ratio Arms as Precision Instruments for the Comparison of Laboratory Standards of Resistance or Capacitance.—C. W. Oatley & J. G. Yates. (*Proc. Instn elect. Engrs*, Part III, March 1954, Vol. 101, No. 70, pp. 91–100.) An estimate is made of errors likely to be caused by deviation of the transformer from the ideal, in respect of imperfect coupling, winding resistance, stray capacitance and eddy currents in the core. Measurements on six transformers designed to give low errors are reported. For ratios up to 100 : 1 there is no difficulty in constructing transformers whose effective ratios are equal to their turns ratios within 1 part in 10^4; even at 1000 : 1 the error need not exceed 1 or 2 parts in 10^4.

621.317.733 : 621.372.2 : 621.315.212 1861

Some Steady-State Characteristics of Short Irregular Lines.—Rosen. (See 1668.)

621.317.755 1862

A Portable High-Speed Cathode-Ray Oscillograph.—S. Waring & B. Murphy. (*J. sci. Instrum.*, Dec. 1953, Vol. 30, No. 12, pp. 469–471.) The instrument described uses a sealed-off c.r. tube and an asymmetrical thyratron timebase which can be operated to give either a single sweep or a repeated sweep at frequencies up to 500 c/s.

621.317.761 **1863**
A Stroboscopic Frequency Meter.—C. W. McLeish & D. H. Rumble. (*Proc. Inst. Radio Engrs*, March 1954, Vol. 42, No. 3, pp. 594–596.) The instrument described is capable of giving a direct indication to the nearest kilocycle in the range 3–30 Mc/s, and is adapted for building in with the signal source (e.g. receiver local oscillator).

621.317.784.029.64 **1864**
Automatic Milliwattmeter for Electromagnetic Radiation.—J. C. van den Bosch & F. Bruin. (*Appl. sci. Res.*, 1953, Vol. B3, Nos. 4/5, pp. 260–264.) An instrument for the 1–10-cm waveband, with two ranges giving respectively 1 and 10 mW full scale deflection, comprises a Wheatstone bridge, one arm of which is formed by a thermistor enclosed in a waveguide; the bridge is balanced by varying the current through the thermistor by means of the valve circuit described.

621.317.789 : 621.373.2.029.65/.66 **1865**
Boltzmann Interferometer.—J. L. Farrands & J. Brown. (*Wireless Engr*, April 1954, Vol. 31, No. 4, pp. 81–83.) Use of the interferometer for investigating the power spectrum of spark-type oscillators is described. The spectrum is given by the Fourier transform of the interference pattern.

621.373.029.3 : 621.376.222 **1866**
A Low-Distortion Pulse Modulator.—F. Brunner. (*Öst. Z. Telegr. Teleph. Funk Fernsehtech.*, Nov./Dec. 1953, Vol. 7, Nos. 11/12, pp. 147–149.) Description of a circuit designed for investigating transients in electro-acoustic systems. A signal tone is modulated by a square-wave voltage the frequency of which is continuously variable between 4 c/s and 100 kc/s. The modulator circuit comprises two hexodes so arranged that the modulating voltage itself is eliminated in the output.

621.373.1.029.42 : 621-526 **1867**
Forcing Function Generator using Conductive Plastic.—L. W. Norman. (*Elect. Engng, N.Y.*, Dec. 1953, Vol. 72, No. 12, p. 1112.) Digest of paper to be published in *Trans. Amer. Inst. elect. Engrs*, 1953, Vol. 72. A potentiometer made of conductive plastic is used to obtain a signal of frequency between 0·05 and 60 c/s and of controllable waveform, for testing servomechanisms.

621.397.2.001.4 **1868**
The Application of Pulse Technique in the Testing of Television Transmission Circuits.—H. Röschlau. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1953, Vol. 5, Nos. 9/10, pp. 187–190.) The pulse-echo meter used by the Federal German Post Office provides either direct-voltage square-wave pulses of $12{\cdot}5 \times 10^{-8}$ sec duration, or alternately positive and negative square-wave pulses of duration $4{\cdot}7 \times 10^{-8}$ sec. It can be used either for the detection of cable irregularities or, with the addition of a special filter, as mismatch indicator. The double-reflection test set operates with a sine-squared test pulse having a repetition rate equal to the television-picture line frequency.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

535.824 : 621.397.6 **1869**
Television Optics.—Zschau. (See 1932.)

621.317.083.7 **1870**
Telemetering Systems.—W. Bösch. (*Microtecnic*, 1952, Vol. 6, No. 6, pp. 303–311 & 1953, Vol. 7, Nos. 1–3 & 5, pp. 19–24, 85–91, 142–148 & 256–260.) A comprehensive review.

621.384.611 **1871**
Acceleration of Partially Stripped Heavy Ions.—R. S. Livingston. (*Nature, Lond.*, 9th Jan. 1954, Vol. 173, No. 4393, pp. 54–57.) Description of the 63-in. cyclotron at Oak Ridge (Tennessee) and of nuclear research carried out with this machine; the hot-cathode source used to produce partially stripped ions is a unique feature.

621.384.611 **1872**
Characteristics of a Proposed Double-Mode Cyclotron.—M. J. Jakobson & F. H. Schmidt. (*Phys. Rev.*, 15th Jan. 1954, Vol. 93, No. 2, pp. 303–305.) Operation of a two-dee fixed-frequency cyclotron with the dee voltages either in phase opposition or in the same phase is discussed. The system would be useful for accelerating ions heavier than α particles.

621.384.612 **1873**
The Profile of the [pole] Pieces of Strong-focusing Cosmotrons with Small Diameter.—G. Sasson. (*C. R. Acad. Sci., Paris*, 22nd Feb. 1954, Vol. 238, No. 8, pp. 885–888.)

621.384.622 **1874**
Improvement of the Performance of a Linear Accelerator by bunching the Electrons before Injection.—M. Papoular. (*C. R. Acad. Sci., Paris*, 15th Feb. 1954, Vol. 238, No. 7, pp. 789–791.) Analysis indicates that considerable increase of efficiency is attainable by this method.

621.384.622.1 **1875**
The Magnitude of the Divergence caused by the Accelerating Gaps in Linear Ion Accelerators.—M. Y. Bernard. (*C. R. Acad. Sci., Paris*, 8th Feb. 1954, Vol. 238, No. 6, pp. 675–677.)

621.385.833 **1876**
A New Microscope Principle.—J. M. Cowley. (*Proc. phys. Soc.*, 1st Dec. 1953, Vol. 66, No. 408B, pp. 1096–1100.) A high-resolution image may be derived from a large number of 'dark field' images of normal resolution obtained by varying the angle of incidence of the electron beam in a standard electron microscope. The practical limitations of this theoretical result are discussed.

621.385.833 **1877**
Calculation of the Axial Potential of Electron Lenses constituted by Two Coaxial Cylinders of Different Diameters.—P. Ehinger. (*C. R. Acad. Sci., Paris*, 22nd Feb. 1954, Vol. 238, No. 8, pp. 879–881.) The axial position of the equipotential surface whose mean curvature is zero is introduced as reference parameter.

621.387.4 : 519.2 **1878**
Stochastic Theory of Electronic Counters.—F. Pollaczek. (*C. R. Acad. Sci., Paris*, 15th Feb. 1954, Vol. 238, No. 7, pp. 766–768.)

621.387.424 **1879**
The Behaviour of Counters with External Cathode and Pure Methyl-Alcohol Filling, subjected to Gamma Radiation.—D. Blanc. (*C. R. Acad. Sci., Paris*, 8th Feb. 1954, Vol. 238, No. 6, pp. 673–675.) Report of an experimental investigation to determine the value of pressure for best performance.

621.387.424 1880

The Delay in the Build-up of Halogen-Quenched Counters.—D. van Zoonen. (*Appl. sci. Res.*, 1953, Vol. B3, Nos. 4/5, pp. 377–389.) Continuation of the experiments of van Zoonen & Prast (1463 of 1953).

621.389 : 620.16 1881

Symposium on 'Vibration Methods of Testing'.—(*J. Brit. Instn Radio Engrs*, March 1954, Vol. 14, No. 3, pp. 93–126.) Text of, and discussion on, four papers, as listed below, dealing with the application of electronics to the measurement of mechanical vibrations.

Vibration Generators, Ancillary Equipment and Applications (pp. 94–100).—H. Moore.

Electronic Stroboscopes (pp. 101–105).—F. M. Savage.

Resistance Strain Gauges and Vibration Measurement (pp. 106–114).—P. Jackson.

Electronic Aids to Vibration Measurement (pp. 115–124).—R. K. Vinycomb.

621.397.9 : 612.1 1882

Particle Counting by Television Techniques.—L. E. Flory & W. S. Pike. (*RCA Rev.*, Dec. 1953, Vol. 14, No. 4, pp. 546–556.) Description, with circuit diagram, of a closed-circuit television system designed for counting blood cells.

621.385 1883

Electron Optics. [Book Review]—Klemperer. (See 1976.)

PROPAGATION OF WAVES

621.396.11 1884

Radio Communication by Scattering from Meteoric Ionization.—V. R. Eshleman & L. A. Manning. (*Proc. Inst. Radio Engrs*, March 1954, Vol. 42, No. 3, pp. 530–536.) The amplitude and duration of forward scattered echoes from individual meteor trails, and the probability of detecting randomly oriented trails over an oblique propagation path are examined. The results indicate that for a frequency of about 15 Mc/s the trail ionization would give a practically continuous signal over a path of about 1 000 km. For v.h.f. signals, scattering from the meteor trails is at least an important contributory factor in the propagation; further observations are required in order to evaluate this factor.

621.396.11 : 551.510.52 1885

The Troposphere as a Medium for the Propagation of Radio Waves: Part 2.—H. Bremmer. (*Philips tech. Rev.*, Dec. 1953, Vol. 15, No. 6, pp. 175–181.) Continuation of survey noted in 1548 of May. A discussion of scattering is presented, based on theory and observations.

621.396.11 : 551.510.535 1886

Real and Complex Wave Polarization in the Ionosphere.—J. C. W. Scott. (*J. geophys. Res.*, Dec. 1953, Vol. 58, No. 4, pp. 437–443.) "The relation between the polarization ellipse in the wave front and the complex polarization at vertical incidence in a slowly-varying horizontally-stratified ionosphere is reviewed. Charts and a table are given showing the sense, orientation and eccentricity of the polarization ellipse under all conditions of plasma frequency, collisional frequency, wave frequency, and magnetic field intensity and direction."

621.396.11 : 551.510.535 1887

On the Coupled Wave Equations of Magneto-ionic Theory.—J. M. Kelso. (*J. geophys. Res.*, Dec. 1953, Vol. 58, No. 4, pp. 431–436.) The coupled wave equations describing the propagation of a radio wave incident vertically on a horizontally stratified ionosphere [1884 of 1942 (Försterling)] and the basic defining relations are presented so as to show explicitly the confusing effects resulting from the different forms used by other authors. The varions notations are tabulated.

621.396.11 : 551.510.535 1888

High-Frequency Scatter-Sounding Experiments at the National Bureau of Standards.—R. Silberstein. (*Science*, 25th Dec. 1953, Vol. 118, No. 3078, pp. 759–763.) Four basic types of presentation used in investigating skip distance by means of back-scattered echoes are discussed, namely (*a*) intensity/range at fixed frequency, (*b*) range/time at fixed frequency, (*c*) p.p.i., and (*d*) sweep-frequency records. Experimental results confirm that skip-distance determination by back-scatter methods is not always possible. The technique calls for skilled personnel familiar with the regular behaviour of the ionosphere in the region.

621.396.11 : 551.510.535 1889

Maximum Usable Frequencies and Lowest Usable Frequencies for the Path Washington to Resolute Bay.—G. H. Hanson, H. V. Serson & W. Campbell. (*J. geophys. Res.*, Dec. 1953, Vol. 58, No. 4, pp. 487–491.) WWV transmissions were monitored continuously at Resolute Bay, 4 009 km away, for a period of one year; values of the m.u.f.'s and l.u.f.'s were determined from the records. The predicted m.u.f.'s were found to be too low during the night and, for some months, too high during the day. The l.u.f.'s were lower than expected for propagation by one reflection from the F_2 layer.

621.396.11.029.55 1890

Measurements of Angle of Arrival in the Short-Wave Range.—K. Vogt. (*Fernmeldetech. Z.*, Nov. 1953, Vol. 6, No. 11, pp. 537–539.) The procedure and results of measurements reported since 1934 are reviewed. The need for further investigation of the dependence of angle of arrival on sunspot cycle and direction of propagation is evident.

621.396.11.029.6 1891

Metre, Decimetre and Centimetre Waves far beyond the Horizon.—E. Roessler. (*Frequenz*, Nov. 1953, Vol. 7, No. 11, pp. 313–319.) A review of recent literature on long-distance u.h.f. propagation, and of the consequences for the planning of radio transmissions and links.

621.396.11.029.62 1892

A Simple Method for estimating U.S.W. Propagation within Line of Sight.—E. Prokott. (*Telefunken Ztg*, Dec. 1953, Vol. 26, No. 102, pp. 346–352.) The distribution of illumination intensity over a relief map illuminated by a lamp placed at the top of a model of the transmitter aerial mast is in fair agreement with measured line-of-sight u.s.w. field-strength distribution. A correction factor for diffraction is given.

621.396.11.029.62 1893

Report of Propagation Tests on 45 and 66·6 Mc/s in the Central Mediterranean.—J. Roux. (*Ann. Radioélect.*, Oct. 1953, Vol. 8, No. 34, pp. 318–330.) During summer 1952 and early 1953 continuous records were made of field strength, signal/noise ratio, interference, and meteorological data for two over-sea transmission paths of 245 and 285 km respectively, proposed for radio relay links between Europe and North Africa. Results of an analysis of the records are presented, these include details of interfering signals, the most remarkable of which were from Berlin, from Monte Cavo, near Rome, and television transmissions from Paris.

621.396.11.029.64/.65 1894

Atmospheric Attenuation of Millimetre and Centimetre Waves.—A. Perlat & J. Voge. (*Ann. Télécommun.*, Dec. 1953, Vol. 8, No. 12, pp. 395–405.) Tables and curves showing the attenuation per kilometre in the 10-mm–10-cm frequency range due to various forms of precipitation, fogs and cloud are assembled from the literature. Meteorological statistics for France and North Africa are also presented. From these, the maximum attenuation to be expected over a given path during a particular percentage of time can be estimated for each meteorological factor.

RECEPTION

621.376.33 **1895**

High-Linearity Demodulation of Frequency-Modulated Oscillations.—H. Meinke. (*Fernmeldetech. Z.*, Dec. 1953, Vol. 6, No. 12, pp. 571–577.) The conditions are derived for high linearity of a demodulator consisting of four reactances, one effective resistance and one rectifier. The characteristics of several possible circuits are considered briefly; one circuit is considered in detail.

621.396.621 **1896**

Analysis of a Limiter as a Variable-Gain Device.—L. R. Kahn. (*Elect. Engng, N.Y.*, Dec. 1953, Vol. 72, No. 12, pp. 1106–1109.) The limiter is treated as a device whose gain is varied so that the amplitude-modulation component of the input signal is eliminated. The method is demonstrated by considering the side-band spectrum of the phase-modulation component of a two-tone signal and the mechanism of the common-limiter diversity system.

621.396.621 **1897**

Post-War Development of Broadcast Receivers [in Western Germany].—E. Klotz & G. Schaffstein. (*Telefunken Ztg*, Dec. 1953, Vol. 26, No. 102, pp. 372–382.) A survey with particular reference to the development of v.h.f. f.m. receivers.

621.396.621.54 : 621.314.7 **1898**

The Transistor as a Mixer.—J. Zawels. (*Proc. Inst. Radio Engrs*, March 1954, Vol. 42, No. 3, pp. 542–548.) Operation is considered for signal frequencies up to v.h.f., three frequency ranges being distinguished, namely (*a*) the lowest range, in which the conversion performance depends primarily on the α value of the transistor; (*b*) a middle range in which the base resistance is the dominant factor; (*c*) the high-frequency range where the capacitance shunting the reverse-biased emitter determines the performance. Methods for calculating the conversion gain are indicated, and some measurements are reported. At the low frequencies a junction transistor can give as good a performance as a crystal-diode mixer followed by a junction transistor amplifier. Point-contact types can give conversion gain in the v.h.f. band, but their noise factor is relatively high.

621.396.82 : 621.376.3 **1899**

Reception of an F.M. Signal in the Presence of a Stronger Signal in the same Frequency Band, and other Associated Results.—R. M. Wilmotte. (*Proc. Instn elect. Engrs*, Part III, March 1954, Vol. 101, No. 70, pp. 69–75.) Analysis based on the amplitude variation of the vector resultant of two f.m. signals of different frequencies is used to derive methods for (*a*) receiving a weak signal in the presence of a stronger one, (*b*) reducing the interfering effect of a weak signal on a stronger one, and (*c*) transmitting a number of weak coded messages with an f.m. signal and in the same band. The bandwidths required by various f.m. multiplex systems are compared.

621.396.822 **1900**

On the Detection of Sure Signals in Noise.—R. C. Davis. (*J. appl. Phys.*, Jan. 1954, Vol. 25, No. 1, pp. 76–82.) Several criteria for an optimum pre-detection filter are compared. When the input noise is Gaussian, the optimum filter is identical with the linear filter which maximizes the output signal/noise ratio. The stability of the optimum filter is discussed in some detail, the discussion applying in the practical case only to finite-memory filters. The determination of an optimum signal shape of given energy content and duration is considered.

621.396.828 **1901**

Review of the Pilot Radio-Interference-Suppression Campaign in Iserlohn.—O. Schmidt. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1953, Vol. 5, Nos. 9/10, pp. 191–193.) Details of the organization, measurements and equipment required in dealing with some 600 industrial and domestic sources of interference are given. Difficulties in design and supply at reasonable cost of efficient mains filter units were encountered.

621.396.621(083.72) **1902**

British Standard Glossary of Terms for the Electrical Characteristics of Radio Receivers (B.S. 2065 : 1954). [Book Review]—Publishers: British Standards Institution, London, 1954, 6s. (*Wireless World*, April 1954, Vol. 60, No. 4, p. 188.)

STATIONS AND COMMUNICATION SYSTEMS

621.376.3 : 517.942.922 **1903**

The Representation of a Frequency-Modulated Oscillation by means of Bessel Functions.—V. Poledna. (*Frequenz*, Nov. 1953, Vol. 7, No. 11, pp. 336–338.)

621.376.3 : 621.396.82 **1904**

Reception of an F.M. Signal in the Presence of a Stronger Signal in the same Frequency Band, and other Associated Results.—Wilmotte. (See 1899.)

621.376.5 **1905**

Gaussian Pulses.—J. P. Vasseur. (*Ann. Radioélect.*, Oct. 1953, Vol. 8, No. 34, pp. 286–300.) The advantage of using Gaussian-shaped pulses in p.m. systems is noted. Pulses of this shape may be derived from a narrow pulse by means of a *RC* filter chain. A band-pass filter of this type is described. The a.m. and f.m. distortion which may occur in transmitting these pulses is calculated.

621.39.001.11 **1906**

Bandwidth, Holding Time and Signal/Noise Ratio for Various Forms of Communication in Relation to Shannon's Theory.—K. O. Schmidt. (*Fernmeldetech. Z.*, Dec. 1953, Vol. 6, No. 12, pp. 555–563 & Jan. 1954, Vol. 7, No. 1, pp. 33–43.)

621.39.001.11 **1907**

The Application of Information Theory to Data-Transmission Systems, and the Possible Use of Binary Coding to increase Channel Capacity.—J. F. Coales. (*Proc. Instn elect. Engrs*, Part III, March 1954, Vol. 101, No. 70, p. 76). Discussion on 253 of January.

621.39.001.11 **1908**

Exact Interpolation of Band-Limited Functions.—A. Kohlenberg. (*J. appl. Phys.*, Dec. 1953, Vol. 24, No. 12, pp. 1432–1436.) A general formula is derived for the spectrum of a multiply periodic, a.m. sequence of pulses. The result is used to show that a function which lies in a frequency band (W_0, $W_0 + W$) is completely determined by its values at a set of points of density $2W$, the points consisting of two similar groups with spacing $1/W$ shifted with respect to each other. This verifies a supposition commonly accepted in communication theory.

621.395.724 : 621.395.44 1909

Standardized Equipment Type 51L.—J. Malézieux, R. Sueur & M. Lebedinsky. (*Câbles & Transm.*, Oct. 1953, Vol. 7, No. 4, pp. 313–324.) Description of new French Post Office equipment standardized on the rack principle, particularly for carrier-current repeater stations.

621.396 : 621.372.5 1910

Radio-Frequency Phase-Difference Networks: a New Approach to Polyphase Selectivity.—M. G. Cifuentes & O. G. Villard, Jr. (*Proc. Inst. Radio Engrs*, March 1954, Vol. 42, No. 3, pp. 588–593.) The possibilities of polyphase methods [313 of 1951 (Macdiarmid & Tucker)] for obtaining selectivity are extended by using r.f. instead of a.f. phase-difference networks. The design of these networks is discussed. A polyphase selective system suitable for s.s.b. transmission or reception is described.

621.396.1 1911

Theoretical Investigation of the Appropriate Spatial Distribution of Frequency Channels for Uniform Coverage of Large Plane Areas.—P. Thiessen. (*Frequenz*, Nov. 1953, Vol. 7, No. 11, pp. 319–325.) Simplifying assumptions are made, so that the location of transmitting stations, when the number of available frequency channels is limited, can be considered solely from a geometrical point of view. Two types of grid-net are considered, rectangular and triangular, and geometrical patterns are reviewed for which adjacent-channel and co-channel interference shall be a minimum. Under ideal conditions the least number of channels required is 9, and the desirable number for adequate coverage without interference is 12 or 13.

621.396.5 : 621.318.57 1912

A Differential-Feedback Suppressor with Independent Two-Way Threshold-Level Adjustment.—F. Rumpel. (*Fernmeldetech. Z.*, Nov. 1953, Vol. 6, No. 11, pp. 540–543.) A description of the voice-operated switching system used at Frankfurt for the transatlantic radiotelephone service. A comparator circuit controls the switching operation during pauses and break-in working. Sensitivity in each direction can be adjusted independently to prevent interruptions due to noise at the receiving end.

621.396.61/.62 : 621.373.421.13 1913

A Frequency-Generating System for V.H.F. Communication Equipment.—G. J. Camfield. (*Proc. Instn elect. Engrs*, Part III, March 1954, Vol. 101, No. 70, pp. 85–90.) Technique developed in connection with mobile communication systems is described. 2 000 channel frequencies are provided, using only 32 crystals, comprising three groups of ten plus one group of two. Selection is accomplished by positioning four switches, the operation thereafter being automatic. Compared with the use of a separate crystal to control each frequency, the system offers the advantages of flexibility, quartz economy, simpler maintenance, and the possibility of spacing the channels more closely.

621.396.65 : 621.396.93 1914

Some Aspects of the Design of V.H.F. Mobile Radio Systems.—E. P. Fairbairn. (*Proc. Instn elect. Engrs*, Part III, March 1954, Vol. 101, No. 70, pp. 53–64. Discussion, pp. 64–68.) Present practice in mobile radio systems is outlined. Methods of operation of complete simplex, two-frequency simplex ('dusimplex') and duplex systems are described. The merits of a.m. and f.m. are compared, and the possibility of making better use of available frequency bands is discussed. Valve problems are mentioned briefly.

621.396.66 : 621.372.56 1915

A Programme Fading Circuit.—R. C. Whitehead. (*B.B.C. Quart.*, Winter 1953/1954, Vol. 8, No. 4, pp. 252–256.) Attenuators comprising temperature-sensitive resistors are used; the resistance values are controlled by variation of a locally generated supply. Methods of segregating the control current from the signal current are considered; frequency filtering is recommended for this purpose.

621.396.66 : 621.396.712 1916

Monitoring and Two-Channel Operation of the RIAS Berlin Medium-Wave Transmitter.—O. v. Broecker. (*Telefunken Ztg*, Dec. 1953, Vol. 26, No. 102, pp. 342–345.) The common control equipment for the 100-kW and 200-kW transmitters is described. The operation of these transmitters either in parallel (see 1943 below) on 989 kc/s, or separately on 683 kc/s and 989 kc/s respectively, is also described.

621.396.7 : 621.396.65.029.55 1917

The Tangier Radio Relay System.—C. G. Dietsch. (*RCA Rev.*, Dec. 1953, Vol. 14, No. 4, pp. 557–568; *Trans. Inst. Radio Engrs*, Jan. 1954, Vol. CS-2, No. 1, pp. 65–68.) An illustrated descriptive account. The station relays telegraph, telephone and radiophoto transmissions between New York and several African, Asian and European stations. Factors considered in the choice of the geographical location are discussed. Double- and triple-diversity receivers and 1–15-kW transmitters with rhombic aerials are used, with operating frequencies in the 4–22-Mc/s range.

621.396.71.029.51 1918

Radio Equipment of the Long-Wave Telegraphy Station at Mainflingen near Aschaffenburg.—E. Meinel. (*Fernmeldetech. Z.*, Nov. 1953, Vol. 6, No. 11, pp. 528–537.) Illustrated descriptions are given of two 60-kW and four 50-kW transmitters operating on frequencies between 46 and 125 kc/s in the European telegraphic service. Specifications are given for the 100–200-m aerial masts which carry four T aerials and two triangular fan-type aerials, the latter equipped for diplex operation.

621.396.932 + 621.396.969.3 1919

Radio in the Service of the Seaman.—C. V. Robinson. (*Proc. Instn elect. Engrs*, Part I, Jan. 1954, Vol. 101, No. 127, pp. 27–28.) Chairman's address, I.E.E. Southern Centre. Communications and navigational aids are reviewed.

621.396.97 + 621.397.5 1920

Inaugural Address [of I.E.E. President].—H. Bishop. (*Proc. Instn elect. Engrs*, Part I, Jan. 1954, Vol. 101, No. 127, pp. 1–10.) A review of the development of broadcasting and television, with special reference to the services of the B.B.C., and to the need for international co-operation. See also *Nature, Lond.*, 6th Feb. 1954, Vol. 173, No. 4397, pp. 248–249.

SUBSIDIARY APPARATUS

621-526 1921

Step- to Frequency-Response Transforms for Linear Servo Systems.—L. C. Ludbrook. (*Electronic Engng*, Jan.–March 1954, Vol. 26, Nos. 311–313, pp. 27–30, 51–55 & 122–126.) Five known methods for finding frequency response from a given graph of linear mode step response are discussed. Theory and practical computing routines are given for a new method based on straight-line-segment approximation to the given step-response graph. Approximate laws relating cut-off frequency and frequency and amplitude of maximum response to the shape and time scale of the step response are derived. Discrepancies between experimental results and those derived from the theory are ascribed mainly to non-linearity of practical systems.

621.3.013.783 **1922**

Electromagnetic Shielding with Transparent Coated Glass.—E. I. Hawthorne. (*Proc. Inst. Radio Engrs*, March 1954, Vol. 42, No. 3, pp. 548–553.) The shielding effectiveness of multilayer coated plane glass structures is investigated by determining their transmission factor for plane e.m. waves. Three types of glass are considered, at frequencies up to 10 kMc/s. Transmission-line theory is used in the analysis, the coatings being represented by shunting resistances. Theoretical and experimental results are compared. Attenuations of 40 db and more can be obtained with tolerable loss of transparency.

621.311.6 : 537.311.33 : 539.165 **1923**

The Electron-Voltaic Effect in *p-n* Junctions induced by Beta-Particle Bombardment.—Rappaport. (See 1799.)

621.311.6 : 621.317.3 **1924**

Supply for Precision Measurement.—(*Elect. J.*, 11th Dec. 1953, Vol. 151, No. 24, pp. 1927–1929.) A low-impedance generator giving a highly stable comprehensive supply over the frequency range 40–2 500 c/s is based on a Wien-bridge oscillator with stabilized d.c. supplies.

621.316.722.1 **1925**

Voltage Stabilizers for Microwave Oscillators.—F. A. Benson & G. V. G. Lusher. (*Electronic Engng*, March 1954, Vol. 26, No. 313, pp. 106–110.) Analysis indicates that the effect of variation of valve heater voltage in commonly used stabilizers [3555 of 1949 (Benson)] is not serious in the case of series-parallel arrangements but may be serious where a single series valve is used.

TELEVISION AND PHOTOTELEGRAPHY

621.397.2.001.4 **1926**

The Application of Pulse Technique in the Testing of Television Transmission Circuits.—Röschlau. (See 1868.)

621.397.311.2 : 621.372.55 **1927**

Aperture Compensation for Television Cameras.—R. C. Dennison. (*RCA Rev.*, Dec. 1953, Vol. 14, No. 4, pp. 569–585.) Aperture distortion, caused by the finite size of the scanning aperture, can be compensated by using a dispersionless transversal filter [55 of 1941 (Kallmann)] designed to provide a given frequency and phase characteristic. Design equations for fixed- and variable-boost aperture compensators are developed.

621.397.5 : 621.396.1 **1928**

Television Coverage.—J. A. Saxton. (*Wireless World*, April 1954, Vol. 60, No. 4, pp. 173–176.) The influence of the terrain on the service areas of v.h.f. and u.h.f. transmitters is examined. Analysis of field-strength measurements for an area such as that around London indicates that the median field strength varies with distance according to a law of the same form as that for a smooth spherical earth, but the absolute measured values fall progressively below the theoretical values as the frequency increases. The significance of these results for television transmission in bands I, III, IV and V is discussed.

621.397.5 + 621.396.97 **1929**

Inaugural Address [of I.E.E. President].—Bishop. (See 1920.)

621.397.5 : 771.53 **1930**

Special Films for Recording Television Transmissions.—W. Behrendt. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1953, Vol. 5, Nos. 11/12, pp. 235–237.) Details are given of an orthochromatic film whose gradation can be varied over a wide range in processing.

621.397.5 : 778.5 **1931**

Some Fundamental Aspects of Telerecording.—C. B. B. Wood. (*J. Telev. Soc.*, Oct./Dec. 1953, Vol. 7, No. 4, pp. 143–151.) The advantages and disadvantages of waveform and picture recording are examined, and practical systems of the latter class are discussed.

621.397.6 : 535.824 **1932**

Television Optics.—H. Zschau. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1953, Vol. 5, Nos. 11/12, pp. 217–223.) A survey of the applications of optics in television and of television in optics. 40 references.

621.397.61 : 621.372.553 **1933**

Design of Phase Equalizers for Television-Transmission Systems.—E. Menzer & H. Voelkel. (*Fernmeldetech. Z.*, Dec. 1953, Vol. 6, No. 12, pp. 578–582.) The characteristics of an asymmetrical bridged-T section are given; application to the correction of a given phase distortion is discussed. A calculation is made for a phase equalizer for a 5-Mc/s low-pass filter, and the theoretical performance is compared with experimentally determined phase/frequency characteristics. A variable phase equalizer is briefly described.

621.397.611 : 778.5 **1934**

Scanning Film Negatives for Television.—E. Legler. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1953, Vol. 5, Nos. 11/12, pp. 232–234.) Gamma correction and the production of correct black-level values are discussed and a suitable circuit is illustrated. The interference ratio obtained with negatives is smaller than with positives.

621.397.611.2 **1935**

Television Pick-up Tube for both Light and X-ray Pictures.—L. Heijne, P. Schagen & H. Bruining. (*Nature, Lond.*, 30th Jan. 1954, Vol. 173, No. 4396, p. 220.) Brief note of a camera tube using a photoconductive layer of lead oxide evaporated on to a pyrex window.

621.397.611.2 : 621.397.82 **1936**

Interference Ratio and Types of Interference in Picture Scanners used in Germany at Present.—W. Dillenburger. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1953, Vol. 5, Nos. 11/12, pp. 209–216.) A discussion of the effects on picture quality of shot noise, mains hum, microphony and the granular structure of the mosaic target or fluorescent screen. The supericonoscope, superorthicon and a flying-spot scanner are considered.

621.397.62 **1937**

The Importance of the D.C. Component.—D. C. Birkinshaw. (*J. Telev. Soc.*, Oct./Dec. 1953, Vol. 7, No. 4, pp. 176–177.) Discussion on 552 of February.

621.397.62 **1938**

Two-Band Television Receivers.—G. H. Russell. (*Wireless World*, April 1954, Vol. 60, No. 4, pp. 189–192.) The choice of a suitable i.f. for British receivers covering bands I and III is discussed. The best frequency appears to be 35·25 Mc/s.

621.397.62 : 621.396.662 **1939**

12-Channel Television Tuner.—(*Wireless World*, April 1954, Vol. 60, No. 4, pp. 162–164.) The tuner covers the v.h.f. bands I and III, with five switch positions for band I and seven for band III. It comprises a cascode signal-frequency amplifier and a frequency changer, and provides an output at i.f. A simplified circuit diagram is shown.

621.397.62 : 621.396.662.029.63 **1940**

A Capacitive-Tuned Ultra-High-Frequency Television Tuner.—E. M. Hinsdale, Jr, & I. D. Baumel. (*RCA Rev.*, Dec. 1953, Vol. 14, No. 4, pp. 461–481.) A tuned circuit is described consisting of a split-stator capacitor and the metal box enclosing it. Two of these circuits are used in the preselector and one in the oscillator of a 470–890-Mc/s tuner. The construction, circuit details and performance figures are given.

621.397.621.2 : 621.397.335 **1941**

Flywheel Scanning and Synchronizing Circuits.—H. Fairhurst. (*J. Telev. Soc.*, Oct./Dec. 1953, Vol. 7, No. 4, pp. 152–159.) Various systems used for flywheel synchronization are examined. Methods for eliminating effects due to variations of temperature and of mains frequency are discussed.

TRANSMISSION

621.376.32 **1942**

Investigation of a Reactance-Valve Circuit for the Frequency Modulation of Oscillators.—H. Behling. (*Telefunken Ztg*, Dec. 1953, Vol. 26, No. 102, pp. 367–370.) The circuit given is sufficiently linear and stable for use in the u.s.w. band. The modulation characteristic is derived for an arrangement using two Type EF802 valves. A comparison is made between calculated and experimentally determined characteristics for an operating frequency of about 7·5 Mc/s.

621.396.61 : 621.396.712 **1943**

Parallel Connection of Two Independent High-Power Sections in the RIAS Berlin 300-kW Medium-Wave Broadcasting Transmitter.—K. Müller. (*Telefunken Ztg*, Dec. 1953, Vol. 26, No. 102, pp. 335–341.) The bridge-type network used in coupling the output of the 100-kW and 200-kW transmitters to the common aerial is described and design considerations are discussed. Zero phase difference between the two transmitters is obtained by inserting a phase-shifting network in the line between the master oscillator and the 100-kW transmitter.

621.396.61 : 621.396.8 **1944**

Subjective and Objective Testing of the Transmission Quality of Medium-Wave Transmitters.—K. H. Baer & H. Lauer. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1953, Vol. 5, Nos. 9/10, pp. 178–186.) Details of subjective tests of music transmission on four medium-power transmitters are given and the results compared with objective distortion measurements. Three transmitters gave very satisfactory performance. Making allowance for distortion in the programme circuits, distortion in the transmitter should be less than 1% between 120 and 1 000 c/s to give good musical quality but may be slightly higher outside this range.

VALVES AND THERMIONICS

537.533 : 621.385.029.6 **1945**

Space Charge Waves in Inhomogeneous Electron Beams.—G. Kent. (*J. appl. Phys.*, Jan. 1954, Vol. 25, No. 1, pp. 32–41.) A sheet beam confined between two plane electrodes by a large applied magnetic field provides the simplified model required for analysis. It is assumed that space charge is not neutralized, hence the beam is inhomogeneous in velocity and charge density. For small inhomogeneity and continuous variation of all quantities as functions of the inhomogeneity, growing waves are not possible. The divergence between this conclusion and those of Haeff (1825 of 1949) is discussed.

537.533.8 **1946**

The Angular Distribution of the Secondary Electrons of Soot.—J. L. H. Jonker. (*Philips Res. Rep.*, Dec. 1953, Vol. 8, No. 6, pp. 434–440.) Experimental determination using equipment previously described (1769 of 1952). The results for soot and nickel are compared.

621.314.63 + 621.314.7 **1947**

Physical Mechanism of [crystal] Rectifiers and Transistors.—E. Spenke. (*Z. angew. Phys.*, Dec. 1953, Vol. 5, No. 12, pp. 472–480.) A survey of known types, based on electronic-semiconductor theory.

621.314.632 : 546.289 : 537.312.6 **1948**

Thermal Effects in Point-Contact Rectifiers.—H. L. Armstrong. (*J. appl. Phys.*, Jan. 1954, Vol. 25, No. 1, p. 136.) Correction to paper abstracted in 1242 of April.

621.314.7 **1949**

Irradiation of Transistors.—C. D. Florida, F. R. Holt & J. H. Stephen. (*Nature, Lond.*, 27th Feb. 1954, Vol. 173, No. 4400, pp. 397–398.) Measurements have been made on point-contact transistors with collector current limited by an external resistance. When the transistor was given a dose of about 10^{14} neutrons/cm^2, the decay of collector current after cessation of emitter current was speeded up, with little or no deterioration of the other parameters. The improvement appears to be permanent, and is attributed to the production of recombination centres such as lattice defects in the bulk material.

621.314.7 **1950**

Joining Solutions at the Pinch-Off Point in "Field-Effect" Transistor.—R. C. Prim & W. Shockley. (*Trans. Inst. Radio Engrs*, Dec. 1953, No. PGED-4, pp. 1–14.) In a field-effect transistor with the drain connection biased beyond 'pinch-off' in respect to the gate, the potential distribution is difficult to determine by analytic methods. An approximate solution is obtained which is valid over the whole length of the channel, and which completes one aspect of theory developed in 879 of 1953 (Shockley).

621.314.7 **1951**

Water Vapor and the "Channel" Effect in *n-p-n* Junction Transistors.—R. H. Kingston. (*Phys. Rev.*, 15th Jan. 1954, Vol. 93, No. 2, pp. 346–347.) Experiments are reported which indicate that the conductance of the channels described by Brown (166 of January) depends on the water-vapour pressure. The mechanism of the effect is discussed.

621.314.7 **1952**

Behavior of Germanium-Junction Transistors at Elevated Temperatures and Power-Transistor Design.—L. D. Armstrong & D. A. Jenny. (*Proc. Inst. Radio Engrs*, March 1954, Vol. 42, No. 3, pp. 527–530.) Limitations on the operation of transistors at high temperatures, resulting from the increased thermal production of hole-electron pairs, are discussed in relation to *n-p-n* and *p-n-p* types for a dissipation of about 1 W. Methods of cooling are described [see also 1247 of April (Giacoletto)] whereby satisfactory α values are maintained at collector currents >100 mA.

621.314.7 **1953**

Calculation of the Surface Recombination Current in the Junction Transistor prepared by Fusion.—J. Laplume. (*C. R. Acad. Sci., Paris*, 8th March 1954, Vol. 238, No. 10, pp. 1107–1109.) The geometry of this type of transistor is such that volume recombination is negligible in comparison with recombination at the junction surfaces. An expression for the recombination current is derived on the assumption of a Laplacian concentration of injected charge carriers in the base.

621.314.7 **1954**

Evaluation of the Current Gain in the Fused-Junction Transistor.—J. Laplume. (*C. R. Acad. Sci., Paris*, 22nd March 1954, Vol. 238, No. 12, pp. 1300–1301.) The current gain α depends on the recombination current, a formula for which was derived previously (1953 above), and on the emitter current I_e. An approximate expression is derived for I_e based on the geometry of the junctions and on certain assumptions regarding the distribution of concentration of charge carriers. It is concluded that (*a*) α increases with collector radius, (*b*) for constant collector radius α exhibits a maximum when the emitter is slightly smaller than the collector, (*c*) when the emitter is larger than the collector α decreases rapidly. An optimum value of α is derived.

621.314.7 : 546.289 **1955**

A P-N-P Triode Alloy Junction Transistor for Radio-Frequency Amplification.—C. W. Mueller & J. I. Pankove. (*RCA Rev.*, Dec. 1953, Vol. 14, No. 4, pp. 586–598; *Proc. Inst. Radio Engrs*, Feb. 1954, Vol. 42, No. 2, pp. 386–391.) A transistor having low resistance and low capacitance is obtained by using a thick wafer of low-resistivity Ge and reducing this thickness, by drilling, to about 0·0005 in. at the region where the junctions are formed. A gain of up to 39 db at 455 kc/s can be obtained using a neutralized circuit, and up to 12 db at 10 Mc/s without neutralization. Characteristics are compared with those of the Type TA-153 transistor described by Law et al. (876 of 1953).

621.383 **1956**

Development of Photoemissive Receivers for the Far Ultraviolet Region. Photocells and Electron Multipliers.—V. Schwetzoff. (*Rev. gén. Élect.*, Feb. 1954, Vol. 63, No. 2, pp. 71–96.) A detailed study of methods of preparation of photocathodes and report of results for photocells and electron multipliers for operation at wavelengths below 2 000 Å. Prolonged heat treatment produced cathodes with very good characteristics. Ag-O-Cs and Sb-Cs photocathodes and Cu-Be electrodes for electron multipliers were specially studied. 199 references.

621.383.27 : 621.311.6 **1957**

Constancy of Photomultiplier Gain.—R. Wilson. (*J. sci. Instrum.*, Dec. 1953, Vol. 30, No. 12, pp. 472–474.) An examination is made of features of the supply circuit which affect the gain of the photomultiplier, both focused and unfocused types being considered. Supply-circuit modifications are suggested by which the gain can be stabilized.

621.383.27 : 621.396.822 **1958**

Origin of Large-Amplitude Pulses in Photomultiplier Background Noise.—Y. Koechlin, I. Pelchowitch & A. Rogozinski. (*C. R. Acad. Sci., Paris*, 8th Feb. 1954, Vol. 238, No. 6, pp. 660–662.) Measurements on an E.M.I. Type-6260 photomultiplier at low temperature indicate that large-amplitude background pulses are to a great extent attributable to scintillations of the glass, caused by various radiations.

621.383.5 **1959**

Modification of the Spectral Response Curve of a Photocell due to Illumination Fatigue.—G. Blet. (*C. R. Acad. Sci., Paris*, 1st Feb. 1954, Vol. 238, No. 5, pp. 578–579.) Measurements made before and after a fatiguing illumination (1251 of April) indicate that the response is diminished throughout the spectrum without any particular manifestation in the neighbourhood of the illumination wavelength, and that the degree of fatigue is independent of this wavelength; the effect is most marked at the long-wave end of the spectrum.

621.385.029.63/.64 **1960**

Some Remarks on Thermal Effect in an Electron Beam on the Gain of Traveling-Wave Tubes.—K. Udagawa & M. Sumi. (*J. appl. Phys.*, Jan. 1954, Vol. 25, No. 1, pp. 135–136.) Comment on 2580 of 1951 (Parzen & Goldstein).

621.385.029.63/.64 **1961**

A Wideband Power Mixer Tube.—H. R. Johnson. (*Trans. Inst. Radio Engrs*, Dec. 1953, No. PGED-4, pp. 15–27.) The mixer valve consists of a short input helix, a drift tube and an output helix arranged successively along an electron beam and surrounded by a magnetic field. The input helix and drift tube are maintained at low potential with respect to the cathode, the output helix is operated at higher potential. The valve thus combines low modulating-voltage requirements with high power output over a wide frequency band. Expressions are developed for carrier input power, sideband output power and modulating power. Results for experimental valves operating at about 3 kMc/s and having Pierce-type guns with L cathodes are reported.

621.385.029.63/.64 **1962**

High Power Traveling-Wave Tube Gain and Saturation Characteristics as a Function of Attenuator Configuration and Resistivity.—J. J. Caldwell, Jr. (*Trans. Inst. Radio Engrs*, Dec. 1953, No. PGED-4, pp. 28–32.) Preliminary results of tests on three valves with different distributed attenuator configurations show that knowledge of loss alone is insufficient to determine attenuator performance and that the effect of loss on circuit impedance and phase velocity must also be known. For a given configuration, the attenuator with the higher resistivity gave the better performance.

621.385.029.63/.64 : 621.372.221 **1963**

Filter-Helix Traveling-Wave Tube: Part 1 — The Filter Helix, a New Circuit Element for Traveling-Wave Amplifiers and Oscillators.—W. J. Dodds & R. W. Peter. (*RCA Rev.*, Dec. 1953, Vol. 14, No. 4, pp. 502–532.) A filter helix consists essentially of a helical transmission line with periodic inhomogeneities which cause reflections. It is basically a special type of a one-dimensional lattice. Various possible constructions are discussed, and three types are distinguished. See also 1524 of 1953 (Dodds et al.).

621.385.029.64/.65 **1964**

Wave Propagation along a Magnetically-Focused Cylindrical Electron Beam.—W. W. Rigrod & J. A. Lewis. (*Bell Syst. tech. J.*, March 1954, Vol. 33, No. 2, pp. 399–416.) Theory given by Fletcher (1811 of 1950) for a beam in which the electrons move in straight lines (corresponding to infinitely strong magnetic field) is extended to the case where the electrons follow spiral paths (Brillouin flow, corresponding to finite magnetic field). The field equation is solved for the cases where the beam is surrounded (*a*) by a helix, and (*b*) by a drift tube. The gain constant for the cylindrical beam with Brillouin flow is greater than that of a similar beam with rectilinear flow but less than that of a thin hollow cylindrical beam with rectilinear flow, for the same radius, current and voltage.

621.385.029.64/.65 : 513.647.1 : 621.317.336 **1965**

Note on Helix Impedance Measurements using an Electron Beam.—D. A. Watkins & A. E. Siegman. (*J. appl. Phys.*, Jan. 1954, Vol. 25, No. 1, p. 133.) Addendum to 3753 of 1953.

621.385.032.216 **1966**

A Hollow Beam Cathode.—G. E. Mueller. (*Trans. Inst. Radio Engrs*, Dec. 1953, No. PGED-4, pp. 33–36.) Tests on hollow cathodes of various volumes and apertures show that hollow electron beams with current densities of 25 A/cm² can be produced at a temperature of 1 100°C. The emission comes from the coating on the aperture edge, the large values of current being due to continual replenishment of the active Ba at this edge from the reservoir within the cavity. Cathode current is nearly proportional to anode potential. This departure from the 3/2 power law can be explained on the basis of the cathode geometry.

621.385.032.216 **1967**

A Study of the Evaporation Products of Alkaline-Earth Oxides.—I. Pelchowitch. (*Philips Res. Rep.*, Feb. 1954, Vol. 9, No. 1, pp. 42–79.) Report of a comprehensive investigation of the evaporation characteristics of oxide-cathode materials; a mass-spectrometer method was used. BaO evaporates mainly in the form of the oxide when coated on Pt or Ni or coated in admixture with SrO and CaO on Pt. At high temperatures the BaO/Pt system gives off Ba_2O_2 ions. With the SrO/Pt and CaO/Pt systems the main evaporation product is the free element, accompanied by the oxide and the singly ionized metal ions. In the BaO/Ta system the main product is free Ba, accompanied by the oxide and by singly ionized Ba ions. The evaporation-rate/temperature curves exhibited critical points, whose existence was confirmed by resistance measurements on the oxides. The influence of the different base metals was demonstrated.

621.385.032.216 **1968**

The Properties of Oxide Cathodes regarded as Mixed Semiconductors.—J. Ortusi. (*Ann. Radioélect.*, Jan. 1954, Vol. 9, No. 35, pp. 3–36.) After activation, the BaO layer is a mixed semiconductor, i.e. with both ionic and electronic conduction. The ionic conduction, which is due to the mobility of the oxygen ion, reacts on the electronic conduction, and modifies the emission characteristics. The electronic conduction is of n type, based on two groups of donors, the first consisting of Ba atoms, the second of F centres.

621.385.032.216 : [546.42-31 + 546.431-31 **1969**

The Thermionic Emission from Thin Films of Barium and Strontium Oxide.—J. Woods & D. A. Wright. (*Brit. J. appl. Phys.*, Feb. 1954, Vol. 5, No. 2, pp. 74–76.) An experimental investigation is reported. The four principal results are: (*a*) emission from an evaporated film of BaO is a maximum at a layer thickness of 10^{-5} cm and, for an activated film, is approximately equal to that from a sprayed BaO cathode; (*b*) SrO behaves in a similar way, with a lower level of emission; (*c*) emission from BaO sprayed on SrO is similar to that from BaO alone; (*d*) SrO evaporated on BaO gives greater emission than either BaO or SrO alone, the d.c. emission for a SrO layer of thickness 10^{-5} cm being 9 mA/cm² at 550°C.

621.385.1 **1970**

Valve Noise produced by Electrode Movement.—P. A. Handley & P. Welch. (*Proc. Inst. Radio Engrs*, March 1954, Vol. 42, No. 3, pp. 565–573.) The effects of faulty electrode positioning and resonant vibrations of electrodes are distinguished. Expressions are derived for the resonance frequencies of the electrodes in terms of the constructional details; use of the results to improve valve design is discussed. Methods of measuring valve noise are indicated.

621.385.2/.3 **1971**

Self-Heating Thermionic Tubes.—E. G. Hopkins. (*Proc. Instn elect. Engrs*, Part III, March 1954, Vol. 101, No. 70, pp. 77–83.) Three experimental types of valve with mutually bombarding oxide cathodes (3209 of 1950) are described, namely (*a*) a diode in which the resistance is controlled by varying the spacing between the cathodes, the valve thus constituting a variable resistor, (*b*) a triode having a grid midway between the cathodes and functioning as a variable resistor or power oscillator, and (*c*) a diode in which the heated cathode area depends on the current, resulting in stabilization of the voltage between the cathodes. All three types operate direct from 240-V 50-c/s mains.

621.385.2 **1972**

Transit-Time Effects in the Diode under Retarding-Field Conditions.—F. W. Gundlach. (*Philips Res. Rep.*, Dec. 1953, Vol. 8, No. 6, pp. 419–426. In German.) Transit-time effects are taken into account in the calculation of the convection current at the anode. A calculation of the admittance shows the large contribution by electrons which do not reach the anode. Errors in the results of Knol & Diemer (290 of 1953) are noted.

621.385.2 : 537.525.92 **1973**

Approximate Solutions of the Space-Charge Problem for Some Unusual Electrode Geometries.—H. F. Ivey. (*J. appl. Phys.*, Dec. 1953, Vol. 24, No. 12, pp. 1466–1472.) The expressions for the space-charge-limited current for coaxial cylindrical or concentric spherical electrodes are usually given as functions of R/r. By applying the technique suggested by Matricon & Trouvé (1514 of 1951), expressions are obtained for R'/r, where R' is the radius of the equivalent cylindrical or spherical collector and r the radius of the emitter, for 27 different electrode arrangements. The results are tabulated.

621.385.3 **1974**

Development of a New Premium Twin Triode.—H. E. Stumman & J. W. Ritcey. (*RCA Rev.*, Dec. 1953, Vol. 14, No. 4, pp. 482–491.) Description of the construction and manufacturing control of the Type 6101 valve, which is electrically and mechanically interchangeable with the Type 6J6 valve, but is specially designed for reliability.

621.385.832 **1975**

Viewing Storage Tube with Halftone Display.—M. Knoll, P. Rudnick & H. Hook. (*RCA Rev.*, Dec. 1953, Vol. 14, No. 4, pp. 492–501.) The construction and operation is described of a c.r. storage tube of the transmission-control type [3760 of 1953 (Smith & Brown)]. Writing speeds up to 3×10^6 dots/sec and an image persistence of ~10 min for half-tone pictures have been obtained using a pulse restoring method.

621.385 **1976**

Electron Optics. [Book Review]—O. Klemperer. Publishers: Cambridge University Press, 1953, 471 pp., 50s. (*Nature, Lond.*, 30th Jan. 1954, Vol. 173, No. 4396, p. 182.) An authoritative work, for those engaged in design and research in the field of electron optics as well as for physicists working in other fields.

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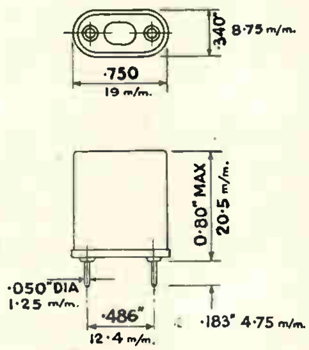

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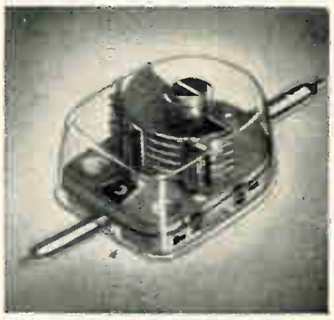

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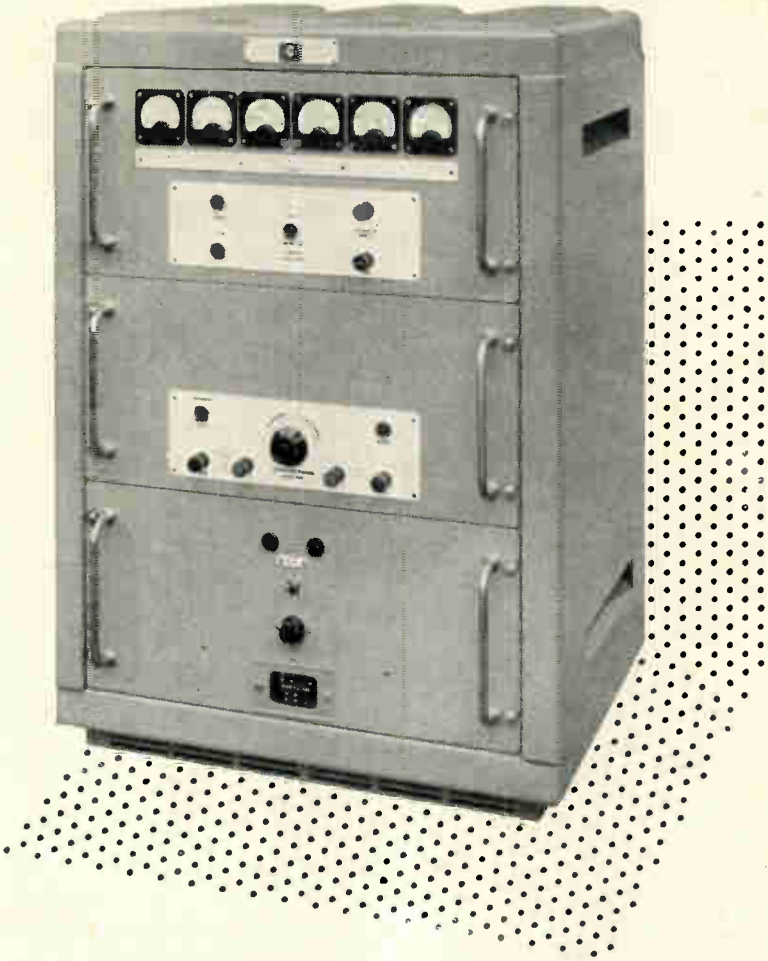

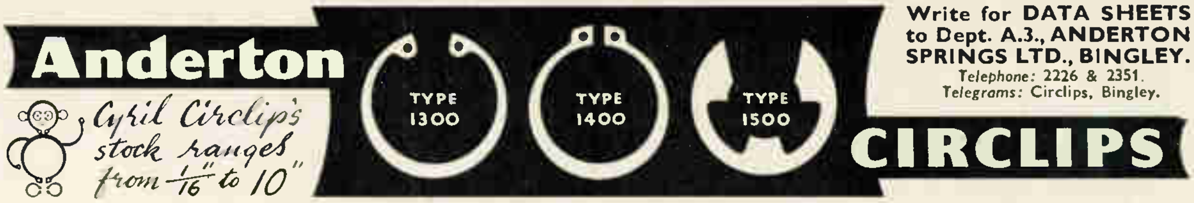

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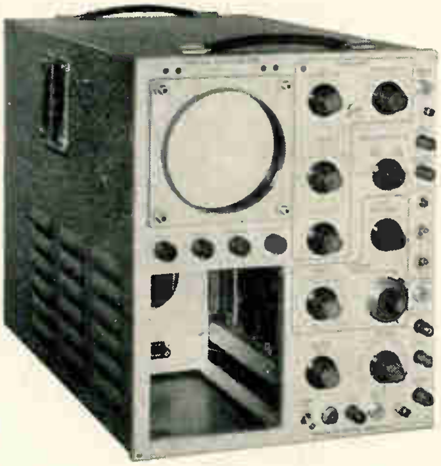

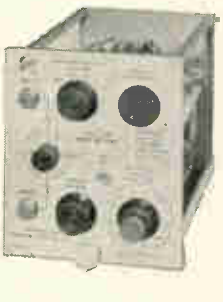

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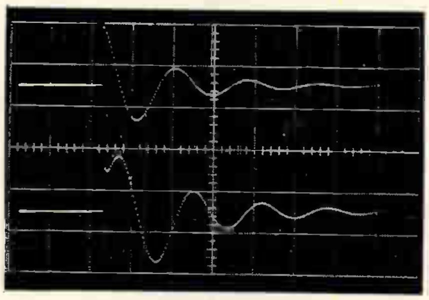

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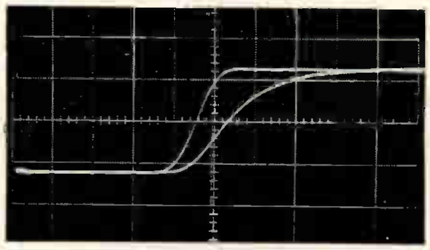

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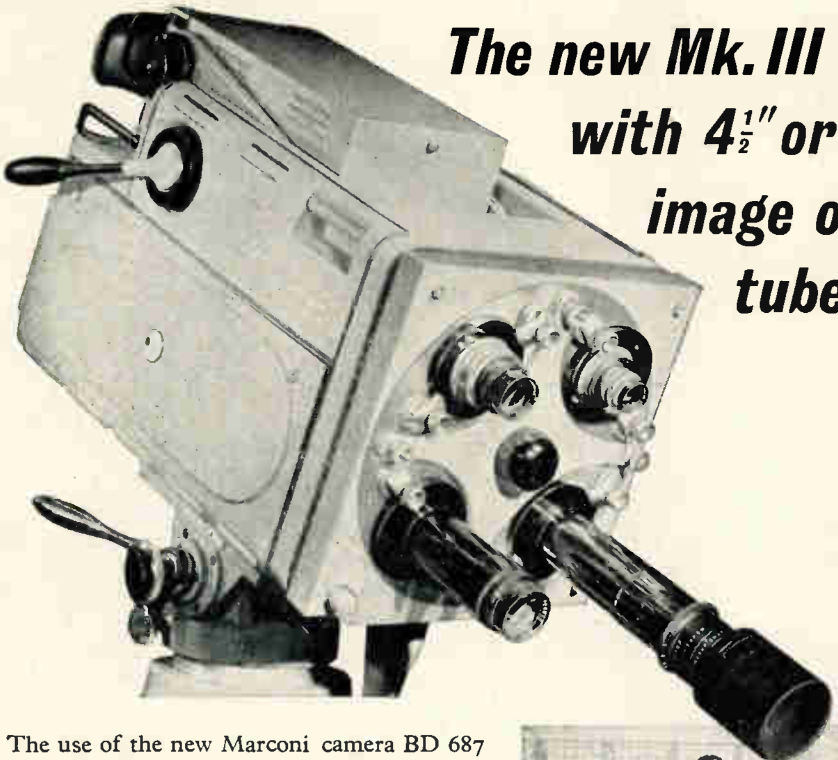

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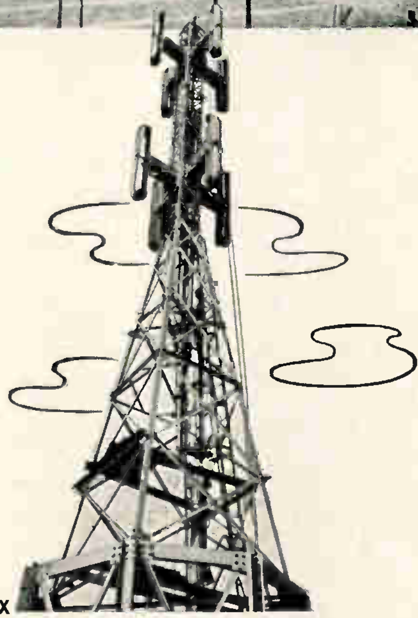

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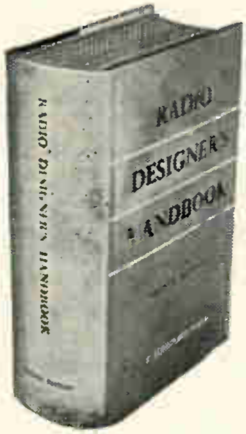

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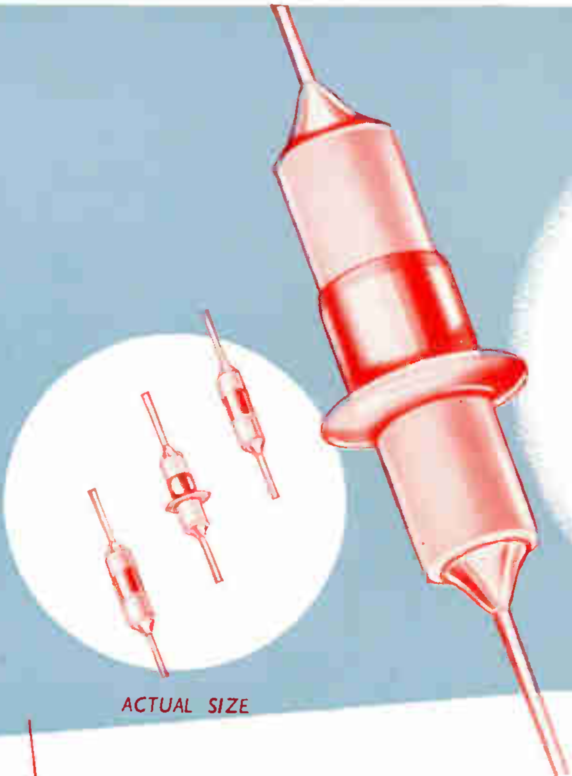

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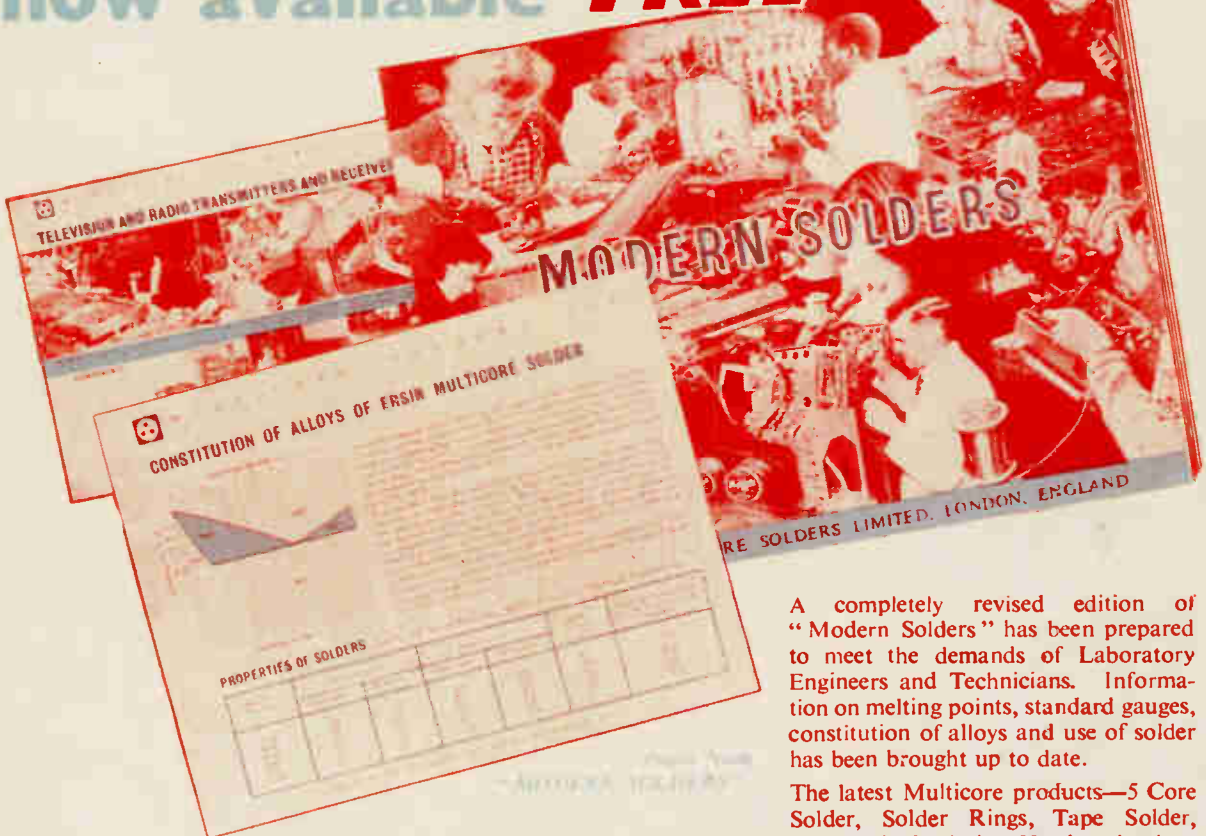

Ersin Multicore is the only solder in the world supplied containing 3 or 5 cores of Ersin Flux, a high-grade rosin which has been subjected to a complex chemical process to increase its fluxing action without impairing in any way the well-known non-corrosive properties of the original rosin.

The core construction ensures flux continuity which, with the correct soldering technique, avoids dry or H.R. joints. There are no wasted lengths of solder without flux.

Unless otherwise specified, Ersin Multicore Solder is supplied as standard to manufacturers in 6 alloys, 9 gauges and 2 flux percentages on nominal 7 lb. reels. Approved by A.I.D., G.P.O., A.R.B., D.T.D.599, R.C.S.1,000 and all U.S.A. Federal specifications. Widely used by U.S.A. manufacturers where it is specially imported from England.

A completely revised edition of "Modern Solders" has been prepared to meet the demands of Laboratory Engineers and Technicians. Information on melting points, standard gauges, constitution of alloys and use of solder has been brought up to date.

The latest Multicore products—5 Core Solder, Solder Rings, Tape Solder, Automatic Soldering Head and others —are described with their specific uses. "Modern Solders" is fully illustrated and contains many useful graphs and tables.

Available only to manufacturers and laboratories using solder. Please apply on your official paper for a copy of this useful reference book. Other interested users of solder are invited to write for complimentary copies of "Hints on Soldering" and "Multicore Solders Technical Summary".

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