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THE JOURNAL OF RADIO RESEARCH & PROGRESS

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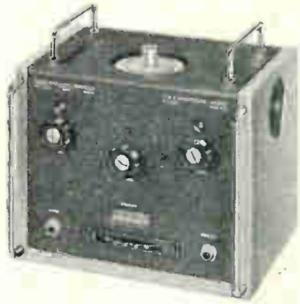

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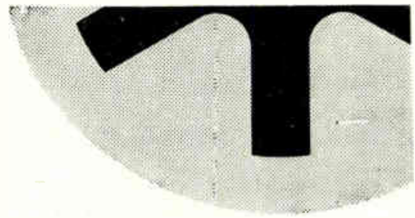

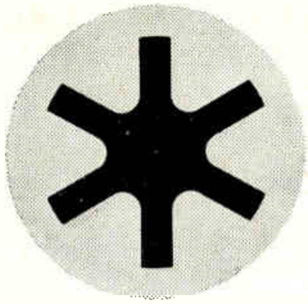

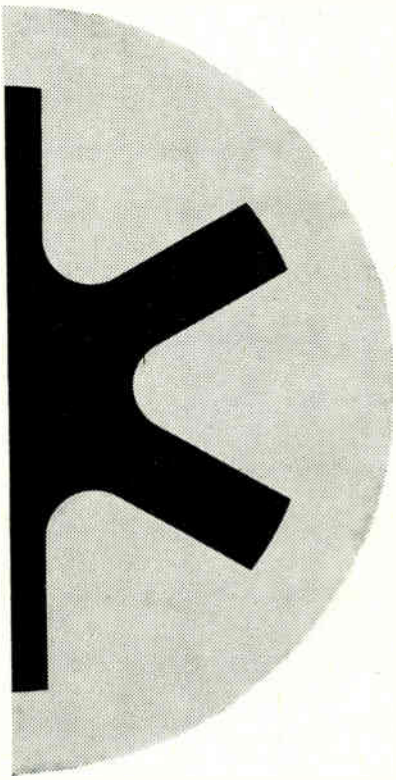

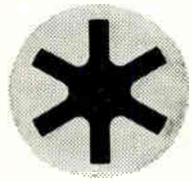

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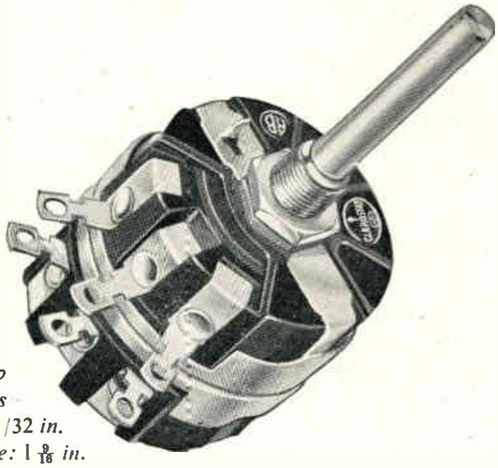

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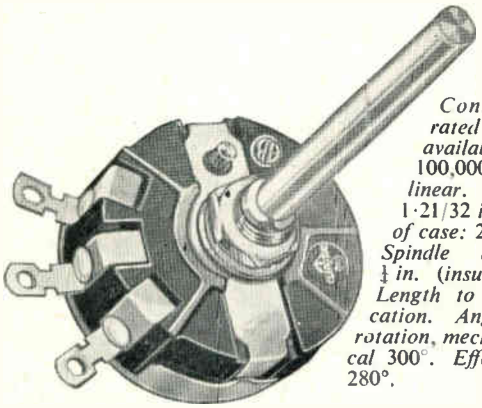

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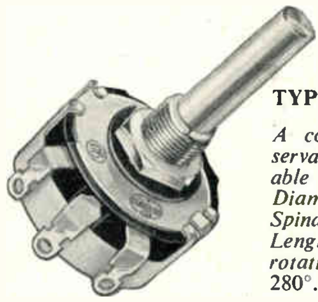

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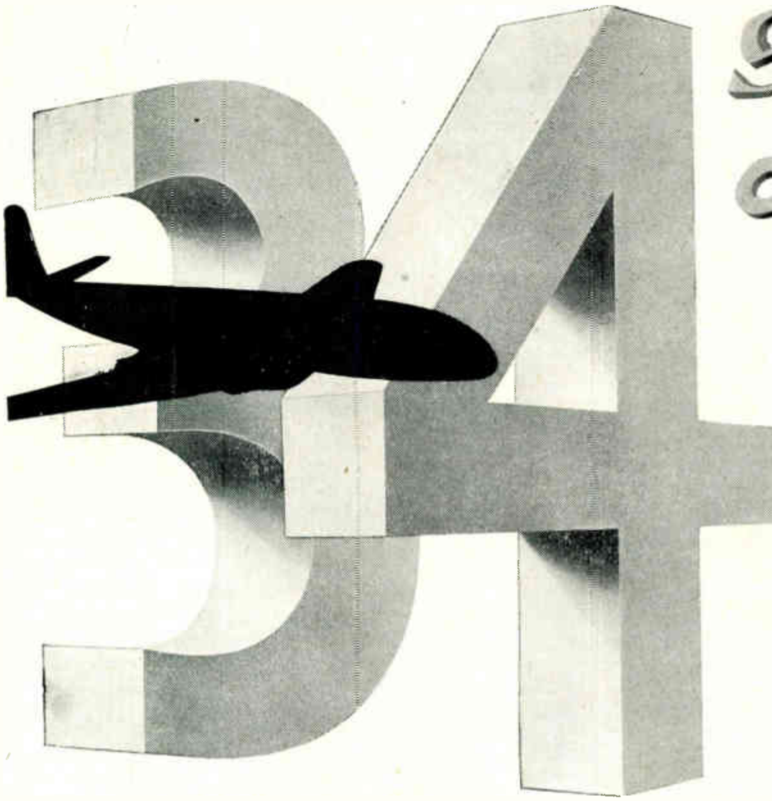

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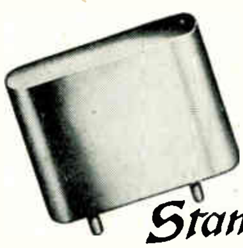

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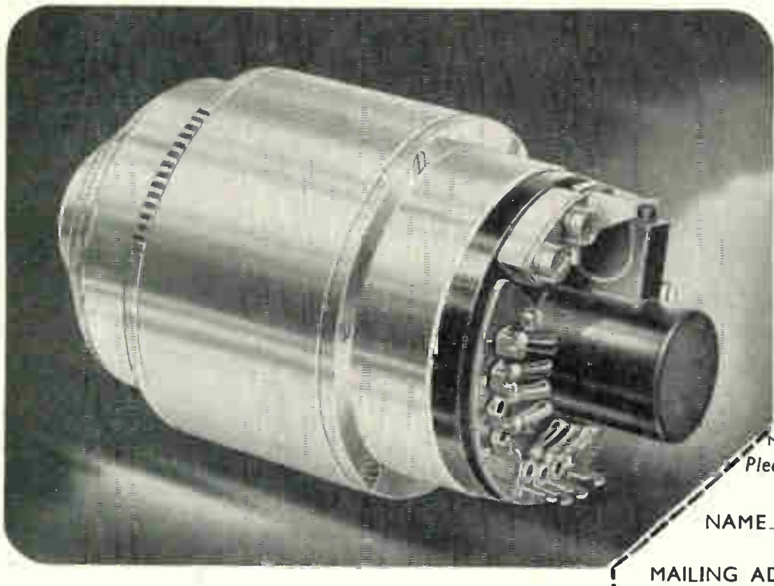

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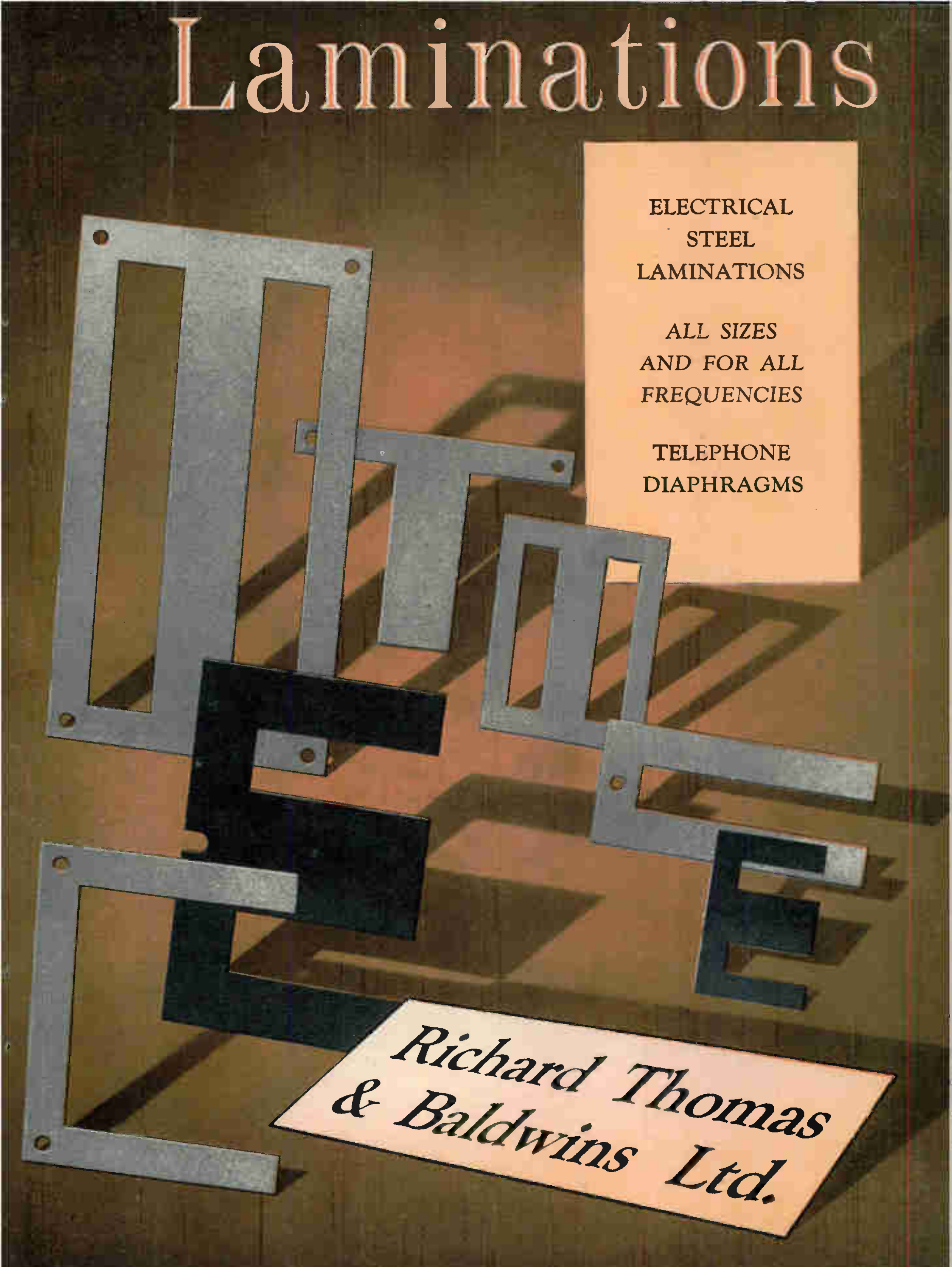
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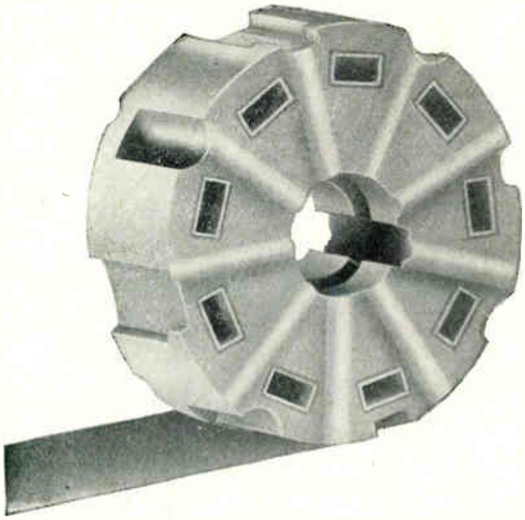

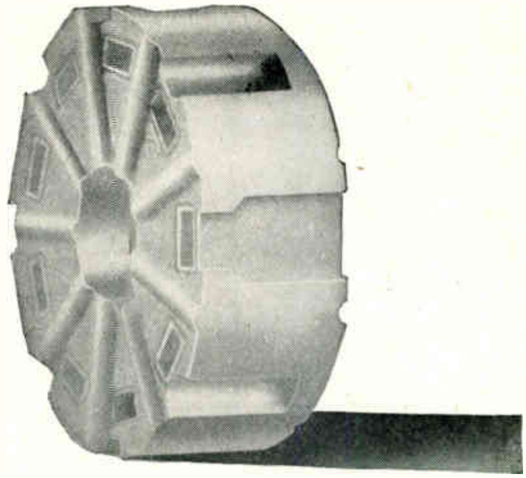

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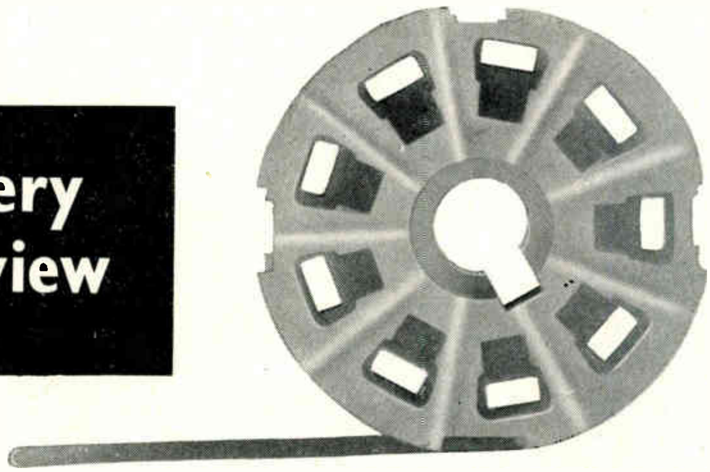

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The Journal of Radio Research and Progress

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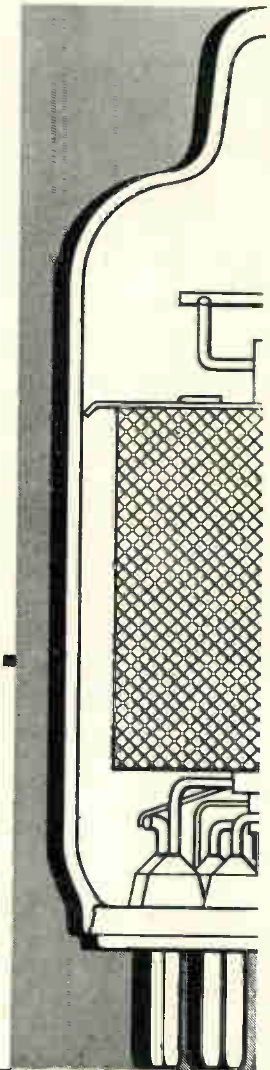
EF86



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WIRELESS ENGINEER

Vol. 30 NOVEMBER 1953 No. 11

The Quantum

IN a recent number of the Vienna *Elektrotechnik und Maschinenbau* there was an interesting article by Professor J. Fuchs entitled 'Der Sinn des Wirkungsquantums', in which he traced the development of the idea and discussed its place in physical concepts and formulae. According to Fuchs the formula associated with the name of Max Planck was originally put forward by him merely as a link between two radiation formulae due to Wien and Rayleigh-Jeans respectively, and it was only when measurements by Rubens in 1900 gave results in strict agreement with his formula that Planck was faced with the problem of giving it a physical meaning—a problem that has worried physicists ever since.

Planck found that the energy of the electric oscillation occurring within an atom could only assume certain values, which were exact multiples of a certain minimum, and that the amount of energy radiated when the oscillation jumped from one mode to another was therefore equal to this minimum; the energy was therefore radiated in quanta or photons. The amount of energy in each quantum was found to be strictly proportional to the frequency of the radiation, so that $W = hf$; if W is in joules and f in cycles per second, Planck's constant $h = 6{\cdot}625 \times 10^{-34}$ joule-second. One can form a mental picture of energy or of power, which is energy divided by time, but the product of energy and time is difficult to conceive and suggests that h is not itself a fundamental quantity, but is derived from other really fundamental quantities, such as the charge of an electron. Many leading physicists have wrestled with this problem, but at a conference in Edinburgh in 1949 Bohr and Rosenfeld concluded that it was now clear that the problem of the elementary particle could not be solved within the framework of quantum dynamics. This was nearly forty years after Sommerfeld had stated that the existence of the atom must be regarded as the result of the existence of the quantum. De Broglie has also stated that, although Planck's celebrated hypothesis has helped us to understand the relationship between many phenomena, we still do not understand the quantum itself.

In 1915 Sommerfeld investigated the mechanics of the atom in an attempt to explain the fine structure of the hydrogen spectrum, and introduced as the limiting impulse moment of the electrons in their elliptical path a term $p_0 = e^2/c$. Now this has the same dimensions as Planck's constant, and putting $e = 4{\cdot}8 \times 10^{-10}$ e.s. unit, $c = 3 \times 10^{10}$ cm/sec and $h = 6{\cdot}625 \times 10^{-27}$ erg-sec Sommerfeld found that the ratio

$$\frac{h/2\pi}{e^2/c} = 137{\cdot}04$$

This figure often occurs in quantum formulae and is known as Sommerfeld's fine structure constant. It was shown in 1939 that Planck's constant can be expressed as

$$h = \frac{e^2}{2} \cdot 137 \cdot \left(\frac{\mu_0}{\epsilon_0}\right)^{\frac{1}{2}}$$

where $h = 6{\cdot}625 \times 10^{-34}$ coulomb-weber, $e = 1{\cdot}602 \times 10^{-19}$ coulomb, $\mu_0 = 1{\cdot}2566 \times 10^{-6}$ H/m and $\epsilon_0 = 8{\cdot}8543 \times 10^{-12}$ F/m. A coulomb-weber is, of course, the same as a joule-sec but the formula was expressed in this way to emphasize the fact that it contains only electric and magnetic quantities; viz., coulombs, webers, and the permeability and permittivity of space. It can also be expressed in purely mechanical terms; if m is the equivalent mass of the electron, v its velocity and r the radius of its path in the atom, then $m.v.r$ is the impulse moment, which has the same dimensions as h, and one finds the formula $m.v.r = n.h/2\pi$, where $n = 1, 2, 3$, etc. Since h, in whatever formula it occurs, can always be replaced by a combination of definite fundamental quantities, such as the electronic charge e, Fuchs concludes that it is not a fundamental quantity but a derived constant.

G. W. O. H.

DETECTION OF PULSE SIGNALS IN NOISE

By D. G. Tucker, D.Sc., A.M.I.E.E., and J. W. R. Griffiths, B.Sc., A.M.I.E.E.

(H.M. Underwater Detection Establishment)

SUMMARY.—The following paper is concerned with the problem of detecting, mainly by means of a visual display, a pulse signal which is accompanied by noise. It is assumed that the noise has a uniform and continuous power spectrum over a bandwidth which at least includes the spectrum of the pulse. Interest centres around the conditions where the pulse can just be detected, and it is shown that for these conditions there are several methods of preparing the signal for display which give better detection than merely extracting the envelope of the wave by plain rectification. Methods discussed include the coherent detector, modified rectifier circuits, phase-modulation display, pulse-to-pulse integration, pulse-to-pulse correlation, and optimum filtration.

Introduction

THERE have recently been many papers published on the subject of detection of signals in noise. Some of the more recent [1–8] are listed in the references at the end of this paper, and references to numerous earlier papers may be found in them. Nevertheless, the present authors feel they have some new matter to record and hope that their different point of view on some of the possibilities already published may help in clarifying the subject. It should be explained that the other authors who have written on this matter appear to have experience mainly of radar systems—or at any rate of high-frequency radio systems—where pulse lengths are measured in microseconds and repetition periods in milliseconds. The present authors' experience is that of asdic systems, where pulse lengths are measured in milliseconds and repetition periods in seconds; they think that this 1000/1 ratio of time scales has forced them to a rather different approach to pulse-to-pulse integration problems. However, they have had detailed and critical discussions with radar experts, and believe no fundamental differences arise.

Many authors[2,3] have restricted themselves to systems using envelope detection (i.e., plain rectifier circuits), while others[6,7] approach the subject from the point of view of information theory and detection probabilities. Coherent detection[4,5,7]—where the detector circuit is controlled by a pure local oscillation and not by the noisy input signal, thus inferring knowledge of the signal frequency and phase—has been shown to give improved results, and there is no doubt that pulse-to-pulse integration[1,3] (or superposition or correlation) gives improved detection. But in all previous work detection is exclusively on the basis of amplitude-modulated waveforms and the phase-modulation is not utilized. The present paper attempts to review all these various aspects and put them in reasonable perspective.

Throughout this paper it is assumed that there is no frequency-shift of the received pulse relative to the transmission or to the background due, e.g., to Doppler effects. This restriction affects the discussion very little, but it is worth pointing out that where such frequency-shifts do occur they can usually be exploited to make the practical realization of improved detection somewhat simpler.

Terminology, Definitions and Criteria

It seems advisable to discuss and define a few terms and criteria that are used later, as confusion has been found to arise over them.

By *visual display* is meant some way of presenting the received information that depends on vision for its final appreciation. In this paper—as in most—the meaning will be limited to only a very few different forms of visual display. Most papers consider only a cathode-ray display, and often only the so-called A-scan, which is merely an

MS accepted by the Editor, February 1953

oscillogram of the envelope or 'detected' waveform; i.e., signal amplitude on the Y-plates, time-base (repetition frequency = signal-pulse repetition frequency) on the X-plates. (In echo-ranging, the position of the pulse on the X-axis indicates the range of the reflection which produces it.) Intensity-modulated cathode-ray displays are more commonly used, however, in which the signal amplitude modulates the spot brightness instead of deflecting the spot. This method is evidently suitable for presenting information in two dimensions, as in a P.P.I. (plan-position-indicator) display. Another form of cathode-ray display which has not been much considered is the Periodmeter[9] or phase-modulation display; in this, the time between successive zero-crossings (or every second crossing) of the received waveform is measured and displayed on the Y-deflection (with a very large zero-offset) against the time-base on the X-plates as for the A-scan. Since the phase-modulation is smaller when the signal is present than when it is absent, the presence of the signal pulse can be detected by a smaller fluctuation over a period of about the pulse-duration. This method of display, as shown later, may have advantages over the A-scan.

In addition to these cathode-ray displays, the authors have had much experience with the *chemical recorder*[10] which has been hitherto exclusively used in asdic and echo-sounder work. This machine, which uses iodized paper travelling at right angles to the time-base deflection, provides an intensity-modulated display, since a larger signal amplitude releases more iodine under the stylus point and makes a darker mark on the paper. The display is, however, only unidimensional, since the paper movement is intended only to record the next trace adjacent to, but not on top of, the previous one. This display has a perfect 'memory', as the record is permanent.

Generally speaking, the signal received in a pulse system is repetitive to a considerable extent, i.e., the signal pulse occurs at nearly the same place in the range (or X) deflection on each successive transmission. If the received pulse is produced by reflection in a uniform and constant medium from an object whose position is fixed relative to the transmitter, the pulses occur at exactly the same place; if the object or transmitter is moving, then successive received pulses do not exactly coincide in range. In any repetitive system improved detection can be obtained by using the increasing information obtained as more repetitive signals are examined. *Pulse-to-pulse integration* occurs when successive signals are added together so that the wanted pulse coincides and adds on a voltage basis every time, but the noise, being random, does not coincide and so adds only on a power basis. *Pulse-to-pulse correlation* is obtained when the successive signals, instead of being added together, are compared, either visually or by some instrument, so that a continued repetition of a small pulse in the same or similar place quickly becomes evident even if it is not detectable on a single transmission.

It should be noted that confusion is possible over the words *detector* and *detection*. The context must be relied on to make it clear whether the words have their everyday connotation (i.e., detection = being able to observe) or the limited meaning originating from the use of the name 'detector' for a rectifier circuit which demodulates a modulated radio wave.

Finally, the term *signal/noise ratio* needs some discussion. In the case where the signal and noise are simultaneously present there is no confusion possible; the ratio concerned is that of the signal component to the noise component of the total waveform. Confusion can arise, however, in pulse systems, where the term can be extended to include criteria of detectability. In this paper, as in previous ones by one of the authors,[5] we shall use the following criteria of signal/noise ratio:—

(a) For the input waveform to the receiver, we use a signal/noise ratio (R_1) which is defined as the ratio of signal to noise components while both are present together. (This is the ordinary definition, as above.)

(b) For intensity-modulated displays, we use a criterion of detectability (R_A) defined as the ratio of the total output when the signal pulse is present to that when it is absent.

(c) For A-scan displays, and possibly for any case where the pulse is long compared with the reciprocal of the noise bandwidth or where the detectable value of R_1 is made very low ($\ll 1$) by pulse-to-pulse integration, we use a criterion of detectability (R_B) defined as the ratio of the change in d.c. output component when the signal pulse is added, to the noise voltage (r.m.s. fluctuation about the mean value) either in the presence or in the absence of the signal pulse.

The criterion R_B has been used by Burgess[2] and Smith,[5] though not with the same symbol. It is probably the most generally useful criterion, as it is more easily interpreted for numerical calculation; only two alternative interpretations are possible, namely, whether the noise voltage is taken in the presence or absence of signal. R_A is harder to interpret, as it is not quite certain how to define total output; the obvious way is to take the r.m.s. of the d.c. and a.c. components, but as the authors feel that the intensity of the display may well more nearly correspond to 'average peak' values owing to a sort of peak-charging effect, they suggest that to take the d.c. component plus the r.m.s. of the a.c. component might accord better with practical conditions.

Spencer[3] (following other authors) has used as his criterion the ratio of peak pulse amplitude to mean noise amplitude. This, however, is not considered further in this paper.

Single-Pulse Detection

Any discussion of the detection of a single pulse, with no repetition, is to some extent unrealistic, as it is not a usual practical problem in the sort of system (e.g., echo-ranging) we are mainly concerned with. But it provides a convenient way of comparing the performance of different types of circuit, and the effect of repetition and pulse-to-pulse integration and correlation can be added quite separately. Under this head, therefore, we shall consider various types of detector circuit in relation to the recognition of a pulse signal against a noise background.

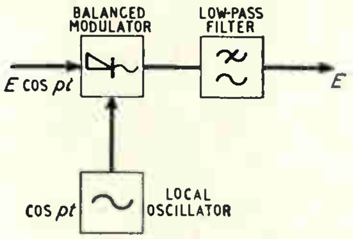


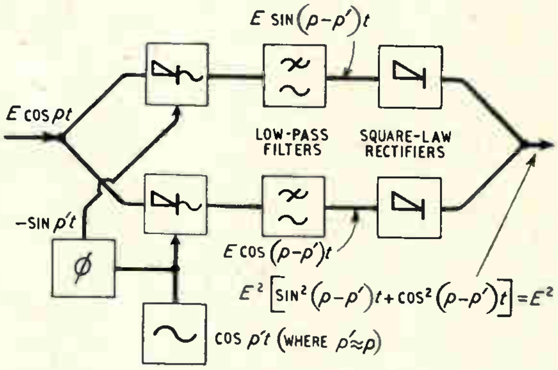
Fig. 1 (left). *The coherent detector.*

Fig. 2 (right). *Two-path detector system.*

Linear and Coherent Detectors

The most usual way of obtaining a signal for display is to rectify the received signal so that the output signal is the envelope waveform. The circuit which does this is usually called the 'linear' detector, presumably because the output signal is a linear function of the input envelope; this very relationship, of course, makes it an essentially non-linear circuit. By contrast, the coherent detector does not extract the envelope function, but is strictly linear. The coherent detector may be defined as one in which the switching of the rectifier is controlled entirely by a pure local oscillation of the same frequency and phase as the carrier of the wanted pulse (see Fig. 1). This is done in practice by using a balanced modulator with a switching oscillation of amplitude very large compared with that of the signal. The homodyne and synchrodyne circuits are an approximation to the coherent detector, but as their local oscillation is derived from the input signal (purified as far as the synchronizing circuit allows) it is preferred to restrict the term 'coherent detector' to the ideal case with a local signal quite free from noise.

It may happen that the frequency and phase of the incoming signal may not be known sufficiently accurately for coherent detection as described above, and the signal/noise ratio may be too poor for synchrodyne action. In such cases, some of the advantages of the coherent detector may be obtained by branching the input into a two-path circuit using two modulators switched by two supplies from the local oscillator in phase quadrature. The system was fully described in 1947 by N. F. Barber[13] and also by one of the present authors.[14] It is shown schematically in Fig. 2. The local frequency need not, in this system, be of exact frequency and phase. The two paths can be recombined after square-law rectification or re-modulation. The main advantage of this type of circuit is that, as in the coherent detector, it permits filtration at the envelope frequency at a part of the system where purely linear relationships



still apply. In the so-called 'linear' detector, intermodulation takes place between the signal and noise, and filtration at the envelope frequency (i.e., at the output) is much less effective in improving signal/noise performance.[4,5] But, as in the two-path scheme rectifiers are ultimately used, it has not the properties of the coherent detector which are discussed in the following paragraphs.

The relative merits of the 'linear' and coherent detectors have been discussed quite fully in two recent papers,[4,5] and therefore only the main features of their performance will be discussed here. The criteria used are R_A and R_B (as already defined) with the alternative interpretations indicated thus:

$$R_{A1} = \frac{\text{(d.c. output due to signal + noise) + (r.m.s. output of l.f. noise when signal present)}}{\text{(d.c. + r.m.s. value of l.f. output, when signal absent)}} \quad \ldots \quad (1)$$

$$R_{A2} = \frac{\text{r.m.s. of d.c. + l.f. components when signal present}}{\text{r.m.s. of d.c. + l.f. components when signal absent}} \quad \ldots \quad (2)$$

$$R_{B1} = \frac{\text{change in d.c. on application of signal}}{\text{l.f. noise when signal is present}} \quad \ldots \quad (3)$$

$$R_{B2} = \frac{\text{change in d.c. on application of signal}}{\text{l.f. noise when signal is absent}} \quad \ldots \quad (4)$$

Values of these criteria are tabulated in Table 1. Owing to the complexity of the full expressions, only simplified expressions are given here on the assumptions that either $R_1 \ll 1$ or $R_1 \gg 1$. The full expressions are given in the Appendix.

The full range of values of these criteria are shown graphically in Figs. 3-6.

It will be seen that, with the exception only of R_{B2} at high values of R_1, the coherent detector gives better detection than the 'linear' detector. It is thus fairly safe to generalize by concluding that the coherent detector is better for pulse signals in noise under the usual threshold conditions, where interest centres in the region where the pulse can just be detected.

It is difficult to illustrate these results by practical examples since, when single pulses are concerned, experimental work cannot ignore the statistical aspect of detection in the way the theoretical comparisons have done. But Fig. 6 shows some typical A-scan displays for both types of detector with the same input signal. It will be seen that the coherent detector gives an improvement, and that this is of the order given by the appropriate criteria. Figs. 7 and 8 show some chemical recorder traces, and although in this case pulse-to-pulse correlation is obtained, this does not alter the fact that for the values of R_1 concerned, the agreement between practice and the theoretical results for R_{A1} is good.

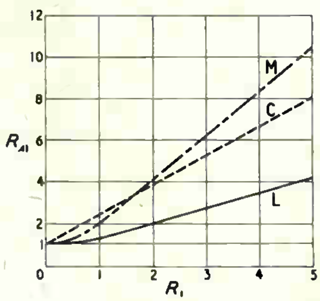


Fig. 3 (above). *Criterion R_{A1} plotted against the input signal/noise ratio R_1; curve L for the 'linear' detector, curve C for the coherent detector and curve M for the modified 'linear' detector.*

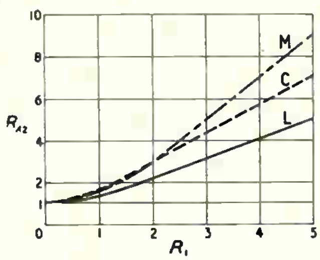


Fig. 4 (left). *Criterion R_{A2}.*

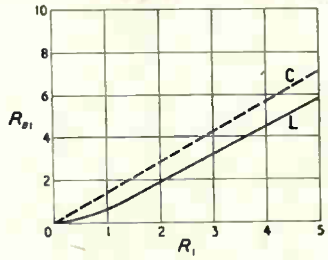


Fig. 5 (above). *Criterion R_{B1}.*

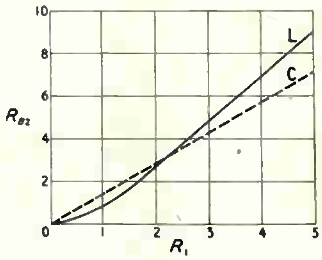


Fig. 6 (left). *Criterion R_{B2}.*

Modified Rectifier Detectors

Suggestions are often made that modification of the normal detector circuits can bring about an improvement in the output signal/noise ratio. Two such circuits, namely a diode detector modified so as to remove the d.c. due to the noise and a diode detector biased so that it only conducts when the input exceeds a predetermined level, are examples. The latter scheme is rather similar to the behaviour of an intensity-modulated cathode-ray display.

These are all forms of contrast expansion and are very similar to the square-law detector. The higher peaks are accentuated relative to the lower peaks and, with signal/noise ratios greater than

TABLE 1

Detector		R_{A1}	R_{A2}	R_{B1}	R_{B2}
'Linear'	$R_1 \ll 1$	$1 + R_1^2/2$	$1 + R_1^2/2$	R_1^2	R_1^2
	$R_1 \gg 1$	$\frac{1}{2}[\sqrt{2}R_1 + 1 + 1/(2\sqrt{2}R_1)]$	R_1	$\sqrt{2}R_1 - \sqrt{(\pi/2)}$	$2R_1 - 1$
Coherent	$R_1 \ll 1$	$1 + \sqrt{2}R_1$	$1 + R_1^2$	$\sqrt{2}R_1$	$\sqrt{2}R_1$
	$R_1 \gg 1$	$1 + \sqrt{2}R_1$	$\sqrt{2}R_1$	$\sqrt{2}R_1$	$\sqrt{2}R_1$

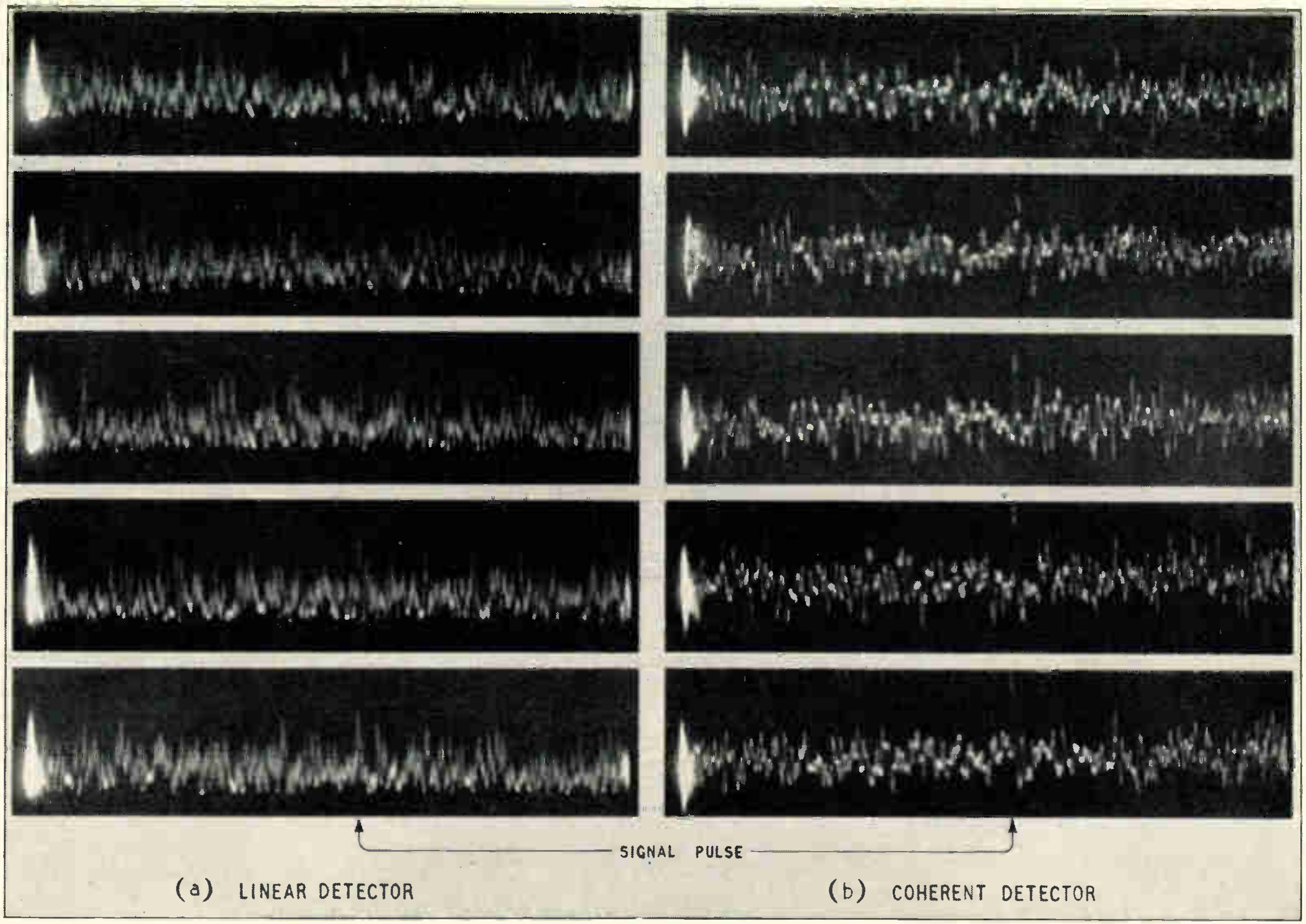


Fig. 7. *Photographs of single traces on cathode-ray A-scan display; pulse duration = 2 ms; white-noise background of 500-c/s bandwidth; input signal/noise ratio (i.e., R_1) = 3 for both types of detector. For each type of detector 5 single scans are shown, chosen at random.*

unity, the signal can be made large compared with the average background. However, the peaks of the noise are also accentuated and consequently if there is any increase in detectability it cannot be very large. If the signal/noise ratio were such as to make detection possible on every single transmission then the contrast circuit would make the signal stand out even more, but since there is already 100% detectability no further information is extracted.

There might be a small improvement generally, by removing the d.c., since this amounts to post-detector filtration (discussed later) and would reduce the noise energy without reducing that of the signal provided the signal were repetitive; i.e., it implies that there is more than one sample of information. Even assuming this to be the case little part is played by the d.c. in the detection on such displays as the A-scan.

It has been calculated using the criterion R_A that removing the d.c. from the output would increase the output signal/noise ratio—see Figs. 3 and 4 which compare the linear detector with and without the d.c. removed. No second curve is drawn for R_B since it is unaffected—the definition of R_B does not involve the d.c. due to the noise.

Hence it can be seen that extreme care should be taken in applying these signal/noise criteria since they do not take into account the statistical nature of the noise accompanying the signal. It should be pointed out that these difficulties apply mainly to the case where the receiver bandwidth is restricted to that necessary to pass the signal so that the signal has a similar character to the noise. It might well be found that 'contrast' circuits give a real improvement in detection when the signal pulse is long compared with the noise pulses.

Phase-Modulation Display

It is well known that random, or 'white', noise has not only a random amplitude distribution but also a random phase distribution. The signal pulse, however, is a portion of carrier wave which should be free of phase-modulation. Thus phase-modulation is a property by which the signal and noise can be distinguished, as an alternative and possibly even in addition to the envelope-amplitude relationships. When the signal is present, the phase-modulation is less than when it is absent, and a periodmeter display is able to indicate the presence of the signal by a reduction in the scatter of the points displayed. (The periodmeter displays the duration of each successive cycle of the incoming wave, measured between zero-crossings of the wave. These latter are determined by amplifying the input signal wave and then limiting it very heavily so that only a phase-modulated square-wave is left.)

In a previous paper[8] it was shown that if the frequency-band corresponding to the input band is filtered out from the output of a balanced limiter, then the signal/noise ratio (R_1; i.e., the ratio of signal to noise amplitudes while both are present together) is higher than that of the original input by up to 3 db, provided the input signal/noise ratio is not much lower than unity. The formula derived for good signal/noise ratios (i.e., $R_1 > 1$) was

output signal/noise ratio, $R_0 \approx$

$$\sqrt{2}\,R_1 - \frac{1}{4\sqrt{2}\,R_1} \quad \ldots \quad \ldots \quad (5)$$

Of course, by limiting, the amplitude discrimination between the output when the signal is present and that when it is absent is largely destroyed, but the improvement shown by equation (5) can be utilized with a periodmeter display which, as shown earlier, does not involve envelope amplitudes at all. If the simultaneous signal/noise ratio is improved, then the waveform has a smaller amount of phase-modulation during the presence of the signal pulse than it had in the original input, and thus the pulse shows more clearly on the display. It should thus be possible to obtain better detection with the periodmeter display by preceding it with a limiter and filter, than with a 'linear' detector followed by A-scan or intensity-modulated display.

Very little experimental evidence of this matter is as yet available, and it is not possible to say to what extent improved detection is obtained with very poor signal/noise ratios, nor how far further improvement can be obtained by additional stages of limiting, each followed by filtration. It seems fairly certain that a combination of phase-modulation display (periodmeter) and envelope-modulation display (A-scan, say) would give improved detection as compared with the latter alone, since more information, contained in the input signal, is used in detection.

Pulse-to-Pulse Integration and Correlation

Whatever type of detector and display is used, better detection is always possible if the signal pulse is repeated regularly. To obtain the improvement, some form of pulse-to-pulse integration or correlation is used. In all that follows it will be assumed that the pulse recurs at the same nominal point in the X-scan (or range-scan). If either the transmitter or object causing the reflection of the pulse is moving, or if the velocity of propagation varies, this will not be quite true, since the range of the received pulse will be varying; but such variation is usually slow, and a range-rate correction device will often allow integration to be performed with reasonable accuracy—and even without correction, correlation is usually possible.

True Integration

It is usually possible to arrange some device, containing a delay line or electric or magnetic storage system, which enables the incoming signals from successive transmissions to be added together after detection. Since the signal pulses recur at the same place in the scan, they add on a voltage basis, and if they are always of the same amplitude, their voltage increases by a factor n in n repetitions. The noise, on the other hand, being random, does not repeat its waveform from scan to scan, and so adds only on a power basis; that is, for n repetitions, the noise voltage is increased by a factor of $\sqrt{n}$. Thus, in terms of the signal/noise ratio for both signal and noise present simultaneously (i.e., analogous to R_1), we see that the effective signal/noise ratio is improved by a factor $\sqrt{n}$. Also, the detectability criterion R_B (whether R_{B1} or R_{B2}), being defined in a similar way, is also increased by the factor $\sqrt{n}$. For R_A, the matter is more complicated, but is easily calculated for any numerical case, the improvement being always rather less than $\sqrt{n}$.

One form of integration which is easily used when repetition periods are of the order of milliseconds rather than seconds is that provided by the persistence of the cathode-ray tube screen. It is then possible to obtain an optical integration which is quite effective. It is said by some radar engineers, too, that, even with an A-scan, true integration is obtained (quite independently of afterglow) by the effect of persistence of vision; i.e., visual integration.

Visual Correlation

Whatever may be the truth regarding visual integration when the pulse repetition period is of the order of milliseconds, there certainly is none when the period is of the order of a second, as in many asdic equipments. In such a case, the effect is more properly described as visual correlation. What seems to occur is that the observer looks at the display, then marks, either in his memory or on the screen—in the latter case, afterglow makes a suitable marking—the position of all likely pulses. On succeeding traces, he looks for any apparent pulses which repeat in the same place. After a number of repetitions he can identify, if the input signal/noise ratio is high enough, what must be the true signal pulse. It is clear that this process is correlation, not integration. Afterglow is a very big help, but the marking cannot build up appreciably unless the persistence-time is long compared with the pulse repetition period. In a case like this, the effect is very largely subjective, depending on the intelligence and experience of the observer. Consequently calculation of the improvements which can be obtained is not possible on an objective basis.

To what extent the detection of a weak pulse by continued observation of an A-scan with a rapid repetition, as in radar work, is integration or correlation is a debatable matter.

When a chemical recorder is used, each successive scan is recorded closely adjacent to, but not superposed on, the previous one. Thus, in this display, a permanent record is obtained of every scan, and if a signal pulse is present at a high enough level, it will show up as a line down the paper even though it was quite undetectable on a single trace. Improvements in detection of 10 to 20 db are readily obtained in this way. The effect is seen in Figs. 8 and 9. This is the best possible form of visual correlation, and for systems which the chemical recorder can cope with (it is far less flexible in speed, etc., than a cathode-ray display, and cannot give a two-dimensional presentation) it is much superior in performance to any cathode-ray display the authors are aware of. It has the advantage that, provided the pulse position does not change from scan to scan by

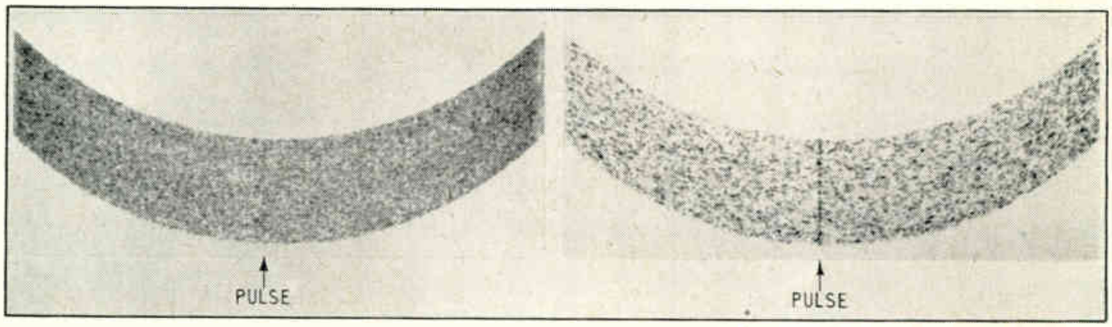


Fig. 8. *Chemical recorder traces comparing 'linear' and coherent detectors; pulse duration = 2 ms; white-noise background of 500-c/s bandwidth; input signal/noise ratio (i.e., R_1) = 0·8 (− 2 db) for both types of detector.—N.B. This is 12 db below the ratios used in Fig. 6 thus showing the benefit of pulse-to-pulse correlation.*

more than about one pulse-width, the continuity of the line is not lost and perfect correlation is still obtained; the slope of the line gives the rate of change of range of the reflecting object.

It is only in the chemical recorder that correlation over long periods of time is possible or desirable. In either integration or correlation processes of any other kind it is necessary (usually also unavoidable) to erase the record or memory after a comparatively short time.

With regard to the periodmeter display, it is evident that although integration does not apply to it, yet visual correlation is performed in exactly the same way as with an A-scan.

Instrumental Correlation

The correlation between one trace (or scan) and another can, of course, be measured precisely by an instrument which uses the mathematical definition of correlation coefficient; thus if r is the correlation coefficient, and $f_1(t)$ and $f_2(t)$ are functions representing the two traces to be correlated:—

$$r = \frac{\overline{f_1(t) \cdot f_2(t)}}{\left\{\overline{[f_1(t)]^2} \cdot \overline{[f_2(t)]^2}\right\}^{\frac{1}{2}}} \quad \ldots \quad (6)$$

where the horizontal bar indicates an average value. If $r = 0$, there is no correlation; i.e., the two traces are quite unrelated, inferring there is no signal pulse present. If $r = 1$, there is perfect

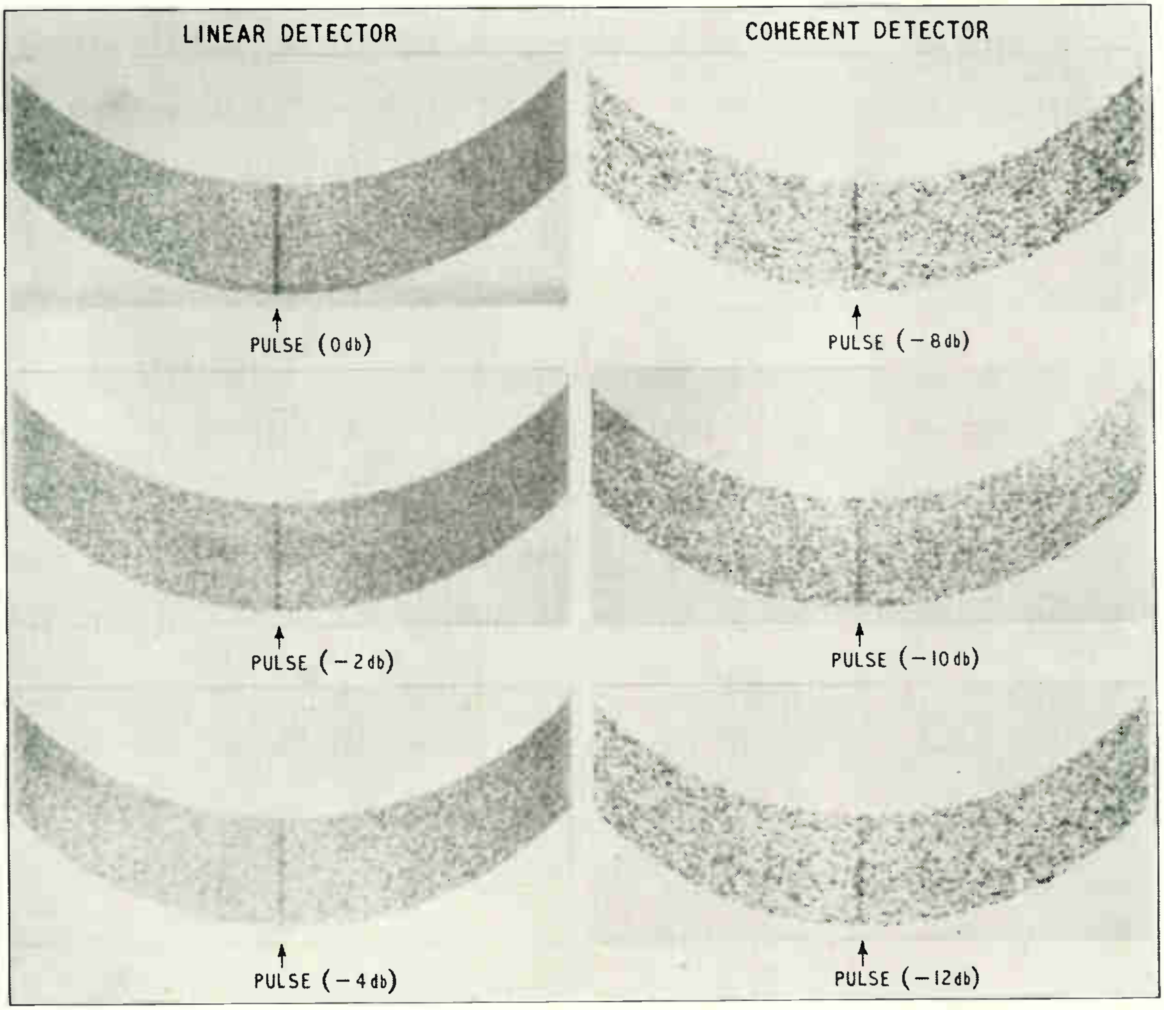


Fig. 9. *Chemical recorder traces comparing 'linear' and coherent detectors on the basis of equal output signal/noise ratios; pulse duration = 5 ms; white-noise background of 500-c/s bandwidth; input signal/noise ratios for 'linear' detector 0, −2 and −4 db, for coherent detector −8, −10 and −12 db. It can be seen that the coherent detector gives detection at an input signal/noise ratio about 8 db lower than that required in the 'linear' detector.*

correlation, inferring that the signal pulse is definitely present. For intermediate values of r, it is probable that the signal pulse is present, to an extent depending on r.

It is worth pointing out that correlation methods can be applied to the detection of a single pulse, without repetition. In this case, the transmitted waveform over the scan period is used as $f_1(t)$ and the received waveform as $f_2(t)$; but as there is now no knowledge of the range relationships between the two waveforms (i.e., the received pulse does not occur at the same time as the transmitted pulse) it is necessary to plot a correlation function instead of a single correlation coefficient. This is done by moving the time axis of one function relative to the other, recording the correlation coefficient for every position. The plot of r against time shift between the functions is the correlation function; thus, if τ is the time shift between $f_1(t)$ and $f_2(t)$:—

$$r(\tau) = \frac{\overline{f_1(t) \cdot f_2(t+\tau)}}{\{\overline{[f_1(t)]^2} \cdot \overline{[f_2(t+\tau)]^2}\}^{\frac{1}{2}}} \quad \cdots \quad (7)$$

This correlation function may show a peak at some particular value of τ, and if at this point the value of r is high enough, the existence of the received pulse at that range position may be considered probable.

This approach to detection of signals in noise is set out very clearly by Woodward and Davies,[6] in an elegant mathematical environment, and with an example.

An instrument which will actually measure the correlation coefficient and function is described by M. J. Tucker,[13] but it is in no way a rapid measurement. It is believed that instruments which will carry out the operations more or less instantaneously have been planned or built, but they are very complicated, and the authors feel that, for the time being at any rate, detection by instrumental correlation is somewhat academic.

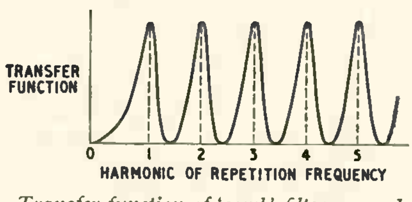


Fig. 10. *Transfer function of 'comb' filter as used after the detector.*

Optimum Filtration

So far in this paper we have considered the various problems in terms of waveform. But it is, of course, quite reasonable to think in terms of frequency spectrum. It quickly becomes clear that, if the pulse is repetitive, so that a line spectrum is obtained from it, then improved detection will result from the insertion of a filter which passes all the pulse-spectrum frequencies, but attenuates all the noise frequencies in between. Such a filter is called a 'comb' filter; a typical response is illustrated in Fig. 10.

The improvement in detection provided by such a filter is easily calculated if the response is known; it is, however, likely to be useful in practice only in special cases, because of its complications and practical difficulties.

It can perhaps be appreciated that the filtration process is very closely analogous to those processes described earlier as integration and correlation, as it utilizes the (presumed) knowledge of repetition, and is mathematically an integration process.

Conclusions

It is hoped that this review of the various ways in which the detection of pulse signals in noise can be improved has brought the whole matter into reasonable perspective. It will have been noticed that many of the processes are only different practical ways of doing the same mathematical operation; e.g., the processes described for pulse-to-pulse integration and correlation, and the comb filter, are alternatives to one another, and cannot be used additively. There are further relationships which are less restrictive. For instance, the coherent detector is really a kind of correlation-measuring instrument, for it measures the correlation between the coherent reference carrier and the received pulse, the multiplying being done in the modulator and the averaging being done by the output filter—but it operates on instantaneous values and not on envelope amplitudes. In all cases, any significant improvement in detection is obtained by utilizing some existing knowledge of the nature of the signal; the periodmeter display, for example, depends on knowledge that the signal is phase-coherent, and so is related to the coherent detector, while integration and correlation methods all depend on knowledge that the pulse is regularly repetitive.

Acknowledgment

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APPENDIX

Exact expressions for criteria of signal/noise ratio

It can be shown[2,4,6,11] that the outputs of d.c. and a.c. from the linear and coherent detectors are given by the expressions in Table 2 below:—

TABLE 2

	Linear	Coherent
D.C. output	$\frac{\alpha V_{n1}}{\sqrt{2\pi}} \cdot {}_1F_1(--)$	$kE_1 \cos\theta$
A.C. output	$\frac{\alpha V_{n1}}{\pi}\left[2(1+R_1^2) - \frac{\pi}{2}\cdot {}_1F_1^2(--)\right]^{\frac{1}{2}}$	kV_{n1}

N.B. ${}_1F_1(--)$ is short for ${}_1F_1(-\tfrac{1}{2};\ 1;\ -R_1^2)$ which is approximately

$$\left(1+\frac{R_1^2}{2}\right) \text{ when } R_1 \ll 1,$$

and

$$\frac{2R_1}{\sqrt{\pi}}\left[1+\frac{1}{4R_1^2}\right] \text{ when } R_1 \gg 1$$

and the function ${}_1F_1(--)$ is a confluent hypergeometric function, and is described in most advanced books on Functions (see, for instance, Ref. 15).

α is a constant of the linear detector; i.e., $v = \alpha e$ when $e \geqslant 0$, and $v = 0$ when $e < 0$, where e and v are the instantaneous input and output voltages respectively.

k is a constant of the coherent detector, relating the output to the input voltages.

θ is the phase angle between the input carrier and the local oscillation in the coherent detector.

E_1 = input signal peak amplitude.

V_{n1} = input noise r.m.s. amplitude.

By substituting in the expressions for the criteria given in the text, and assuming $\theta = 0$, the values in Table 3 are obtained:—

TABLE 3

	Linear	Coherent
R_{A1}	$\frac{\sqrt{(\pi/2)}.{}_1F_1(--) + 2\sqrt{[2(1+R_1^2) - (\pi/2).{}_1F_1^2(--)]}}{\sqrt{(\pi/2)} + \sqrt{(2-\pi/2)}}$	$1 + 2R_1$
R_{A2}	$\sqrt{(1+R_1^2)}$	$\sqrt{(1+2R_1^2)}$
R_{B1}	$\frac{\sqrt{(\pi/2)}.[{}_1F_1(--) - 1]}{\sqrt{[2(1+R_1^2) - (\pi/2).{}_1F_1^2(--)]}}$	$\sqrt{2}.R_1$
R_{B2}	$[\sqrt{\pi}/\sqrt{4-\pi}][{}_1F_1(--) - 1]$	$\sqrt{2}.R_1$

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The Radio Industry Council's premiums for technical writing are being continued, and submissions in respect of articles published in journals on sale to the public between January and December 1953 should be made to the Secretary, Radio Industry Council, 59 Russell Square, London, W.C.1 (from whom full particulars can be obtained), not later than 30th November, 1953.

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THE PERIODMETER

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Introduction

THIS article outlines the principles and broad design features of a little-known instrument* which measures instantaneously the period of each cycle of an input signal, the information being displayed on a cathode-ray tube as the height of the spot above an arbitrary base-line. By the use of suitably precise circuits this base-line can be effectively a long way off the screen. The instrument in use at present operates with a mean input frequency of 10 kc/s and has a maximum sensitivity of ± 30 c/s (i.e., full-scale deflection can be obtained for a change of period corresponding to a change of 60 c/s in the instantaneous frequency† of a 10-kc/s input signal). The resolution is approximately 1 c/s.

Horizontal deflection is provided by a separate time-base, externally triggered.

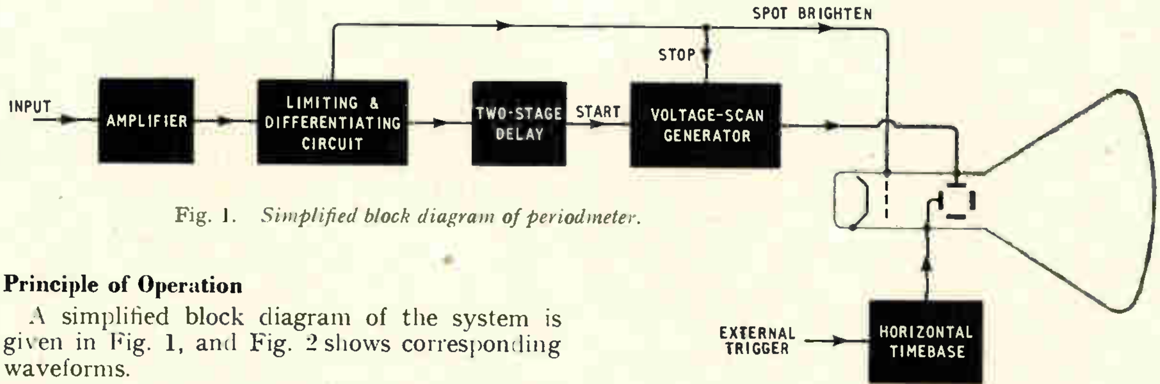

Fig. 1. *Simplified block diagram of periodmeter.*

Principle of Operation

A simplified block diagram of the system is given in Fig. 1, and Fig. 2 shows corresponding waveforms.

The input signal is amplified, limited, and differentiated, giving short pulses coinciding with the positive- and negative-going zero-crossings of the signal under examination [Fig. 2, waveforms (a), (b) and (c)]. The positive pulses (i.e., those corresponding to positive-going zero-crossings) are then delayed by an amount slightly less than the nominal period, which in the particular instrument is 100 μsec [Fig. 2(d)]. These delayed pulses trigger a voltage-scan generator, each scan of which is stopped by the following positive-going zero-crossing for the duration of the succeeding half-cycle of the input signal [Fig. 2(e)]. The remainder of the scan and the flyback take place during the negative half-cycle, the circuit then being reset in preparation for the next trigger pulse.

The output is applied to the vertical deflecting plates of a cathode-ray tube, and during the stopped period the spot is brightened not only by virtue of being stationary but also by the application of the positive half-cycles of the limited input signal [Fig. 2(b)] to the grid of the tube. During each cycle of the input, therefore, the spot is displaced from its zero position by an amount proportional to the difference between the period of the input signal and the fixed delay, and then brightened. The position of the spot is thus linearly related to the instantaneous frequency of the input signal.

The spot is also deflected laterally by an externally-triggered time-base, providing sweeps of approximately 10-msec duration. The trace obtained on the cathode-ray tube is thus as in Figs. 3 and 4.

In Fig. 3 the input signal is of constant frequency. The difference between the periodic time of successive cycles and the delay is therefore constant, and the spot-brightening occurs at a constant height.

The effect on the input waveforms of variation in the periodic time of the input signal can be understood from Fig. 2. Fig. 3 shows the display resulting from a constant-frequency input (generated by an internal 10-kc/s oscillator). Fig. 4 shows the display obtained from a 10-kc/s input signal, effectively frequency-modulated at 1 kc/s, being produced by the simultaneous application to the input circuit of 10-kc/s and 9-kc/s waves, the amplitude ratio being 25 db. This produces a frequency-modulation of 1 kc/s with a deviation of 56 c/s.

* Original development is thought to be due to the University of California. However, the present instrument works at ten times the frequency of that originally described.

† The term 'instantaneous frequency' is used throughout in the sense of the reciprocal of the period rather than the rate of change of phase.

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Accuracy and Frequency Limitations

As mentioned previously, the present model operates at 10 kc/s, with a range of ± 30 c/s and a resolution of 1 c/s. At 10 kc/s the change in period corresponding to a change in frequency of 1 c/s is 0·01 μsec, and for this to be discernible it must cause a spot deviation, on an average cathode-ray-tube screen, of at least 0·8 mm. For the full range of 60 c/s, therefore, the minimum spot deviation required is 4·8 cm, and this distance must be traversed in the time of the corresponding change in period, or 0·6 μsec.

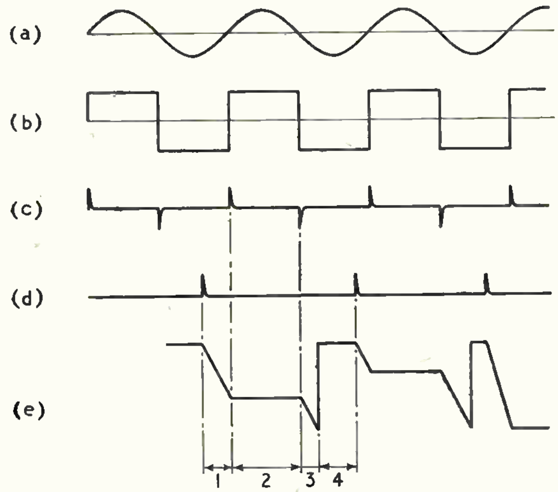

Fig. 2 (above). *Waveforms illustrating operation of periodmeter; (a) input signal, (b) input signal amplified and limited, (c) pulses derived from zero-crossings of (b), (d) positive pulses delayed, (e) output from voltage-scan generator. Period: 1, vertical scan; 2, scan stopped and spot-brightened by (b); 3, completion of scan; 4, wait.*

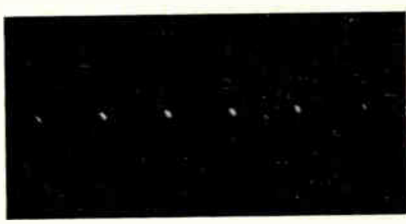

Fig. 3. *Trace obtained with a 10-kc/s constant-frequency input signal.*

Fig. 4. *Trace obtained with a 10-kc/s input signal, frequency-modulated at 1 kc/s with 56-c/s deviation.*

This high scanning velocity would of course need to be even greater for an operating frequency higher than 10 kc/s; indeed for a given resolution the former is proportional to the square of the latter so that to increase the operating frequency to 20 kc/s, for example, maintaining the same range, the scan duration would require to be reduced to 0·15 μsec.

A second limitation is set by the accuracy obtainable from the delay unit. In the instrument this consists of a phantastron circuit,[1] the required delay of very nearly a full period being effected in two stages, to accommodate the recovery time of the phantastron. The second delay is variable over a small range thus providing zero control. For 10-kc/s operation the stability of delay required to give an accuracy of 1 c/s is evidently one part in 10,000, and for similar accuracy at other frequencies the stability must clearly be proportional to frequency.

The long-term stability of the phantastron is barely sufficient for 10-kc/s operation at the required accuracy, unless specially stabilized power supplies, etc., are used, and other methods of obtaining the delay are being considered. An improvement could be effected in the first instance by the use of the sanatron circuit,[1] the accuracy of which is claimed to be higher than that of the phantastron. Another possible method would be to derive the required time delay from an accurately-controlled frequency source.

The present model, as is to be expected with an instrument of such sensitivity, is rather susceptible to hum, supply-voltage fluctuation, etc. This can be largely overcome by using batteries.

Applications

Possible applications are numerous and varied. A few are suggested below:—

(a) The detection and measurement of frequency fluctuation in a nominally pure tone from an oscillator, or from a recording device in which there may be variations in driving-motor speed, tape speed, etc.

(b) The determination of the mean frequency of a noisy signal in a narrow band.

(c) The detection of pulse signals in noise.[2]

(d) General study of phase- and frequency-modulated waves.

Acknowledgments

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REFERENCES

[1] B. H. Briggs, "The Miller Integrator", *Electronic Engineering*, Aug., Sept., and Oct., 1948.

[2] See D. G. Tucker and J. W. R. Griffiths, "Detection of Pulse Signals in Noise", *Wireless Engineer*, 1953, Vol. 30, p. 264.

HARMONIC DISTORTION DUE TO ECHO SIGNALS

Conditions for an Amplitude-Modulated Wave

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THE approximate solutions for the harmonic distortion caused by echo signals in a communication network where amplitude-modulated radio links are used have been considered in a previous paper[1] and approximate formulae and nomograms derived.

These echo signals may be produced by multipath propagation, by stations in a network using the same radio frequency, or mismatched aerial-feeder cables.

It was assumed in the earlier paper that the ratio of echo to wanted signal was small and that the modulation depth m has a value substantially less than 1. If m tends to 1 the binomial expansion used in the original paper is not justified and for this case a binomial expansion of a different form must be used. It is also shown that although the results in the original paper are correct in form when m is sufficiently small, the numerical factors are in error. The 2nd harmonic as given by equation (1) in the original paper[1] has to be multiplied by 2, and the third harmonic [equation (2)] has to be multiplied by 4. This is because cos $\omega_a t_o$ was taken as approximately equal to unity, which is not valid since the distortion is of the order of magnitude $(\omega_a t_o)^2$ and consequently if the series expansion of cos $\omega_a t_o$ is used, at least two terms should be considered. In the method employed in the present paper both sin $\omega_a t_o$ and cos $\omega_a t_o$ are expanded as power series, sufficient numbers of terms of which can be used to give accuracy adequate for all angles. The present analysis is done for angular values of $\omega_a t_o$ between 0 and $22\frac{1}{2}°$; Table 1 shows the form into which sin $\omega_a t_o$, etc., are manipulated to ensure a rapid convergence of the power series used for values of $\omega_a t_o$ greater than $22\frac{1}{2}°$.

If the arguments shown in Table 1 are employed, sufficiently accurate results are obtained by using the first two terms of the appropriate expansion.

A direct Fourier expansion of the first few terms derived from the binomial expansion has been used in the present method. This is valid for all values of m, but the maximum value of m is limited by the binomial expansion for the purpose of which it is assumed that $\frac{r(1+m)}{1-m} \leqslant 1{\cdot}0$; thus the smaller the value of r, the greater the value of m over which the expansion will be valid.

LIST OF SYMBOLS

$\omega_c = 2\pi \times$ carrier frequency
$\omega_a = 2\pi \times$ modulating frequency (f_a)
m = modulation depth (Amplitude-modulated signal)
M_d = modulation index (Frequency-modulated signal)
r = voltage ratio of $\frac{\text{Echo Signal}}{\text{Main Signal}}$
t_o = time delay of echo signal.

TABLE 1
EXPANSIONS

Range of $\omega_a t_o$	sin $\omega_a t_o$	cos $\omega_a t_o$	sin $2\omega_a t_o$	cos $2\omega_a t_o$
$22\frac{1}{2}° - 45°$	sin $\omega_a t_o$	cos $\omega_a t_o$	cos $(\pi/2 - 2\omega_a t_o)$	sin $(\pi/2 - 2\omega_a t_o)$
$45° - 67\frac{1}{2}°$	cos $(\pi/2 - \omega_a t_o)$	sin $(\pi/2 - \omega_a t_o)$	cos $(2\omega_a t_o - \pi/2)$	$-$ sin $(2\omega_a t_o - \pi/2)$
$67\frac{1}{2}° - 90°$	cos $(\pi/2 - \omega_a t_o)$	sin $(\pi/2 - \omega_a t_o)$	sin $(2\omega_a t_o - \pi)$	$-$ cos $(2\omega_a t_o - \pi)$

In order to extend the present analysis to larger delay times (such that $\omega_a t_o > \pi/8$), all expansions required should be in terms of the arguments shown above.

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Discussion of Results

Appendix 1 shows the derivation of the new equations for the harmonic distortion to a single modulating frequency. The distortion is proportional to the square of the ratio of the echo and main signals for both second and third harmonics. These values of harmonic distortion vary cyclically with changes of $\omega_a t_o$ (the angular delay of the carrier signals with no modulation applied) and hence small changes of the difference between the echo and main signal-path lengths will cause the distortion to vary cyclically and rapidly from a maximum to zero. Taking the maximum value of the harmonic distortion when $\omega_c t_o = (2n - 1)\,\pi/2$ we have

Original Signal $= E\,[1 + m \sin \omega_a t] \sin \omega_c t$

Echo Signal $= rE[1 + m \sin \omega_a\,(t + t_o)] \sin \omega_c(t + t_o)$

$$\frac{\text{2nd Harmonic}}{\text{Fundamental}} = 2r^2(\omega_a t_o)^2 \frac{m}{(1 + \sqrt{1 - m^2})^2} \{(1 - m^2) + (\omega_a t_o)^2\}^{\frac{1}{2}} \quad \ldots \quad (1)$$

$$\frac{\text{3rd Harmonic}}{\text{Fundamental}} = 2r^2(\omega_a t_o)^2 \frac{m^2}{(1 + \sqrt{1 - m^2})^2} \left\{(1 - m^2) + \frac{(\omega_a t_o)^2}{m}(1 - \sqrt{1 - m^2})^2\right\}^{\frac{1}{2}} \quad \ldots \quad (2)$$

From these equations it would appear that as $m \to 1$

$$\frac{\text{2nd Harmonic}}{\text{Fundamental}} = 2r^2(\omega_a t_o)^3$$

$$\frac{\text{3rd Harmonic}}{\text{Fundamental}} = 2r^2(\omega_a t_o)^3$$

but the assumption is not justified, as the original expansions are not valid as $m \to 1$, except for infinitely small values of r.

The ratios (expressed in db) of the 2nd- and 3rd-harmonic distortion to the fundamental are plotted in Figs. 1 and 2 for varying values of m and $\omega_a t_o$.

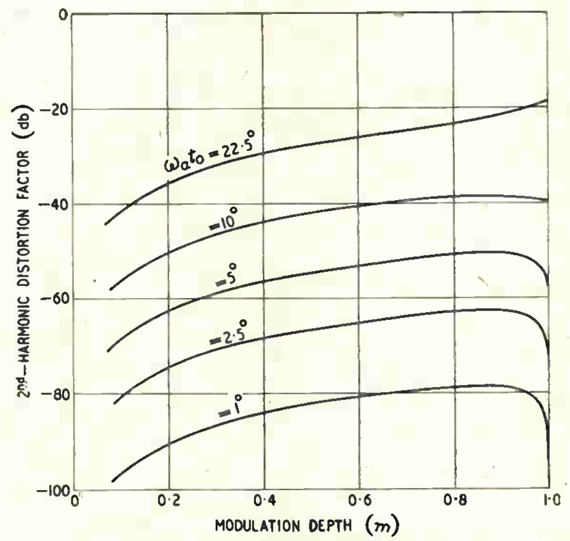

Fig. 1. *Variation of 2nd-harmonic distortion factor with modulation depth and delay time.*

The quantities so plotted are called the Distortion Factor (following the nomenclature of Ref. 1), and they should be added to the Reflection Factor (i.e., twice the value of r expressed in db) to obtain the ratio in db of the harmonic to the fundamental. As previously stated, the maximum value of m for which equations (1) and (2) are valid is governed by the value of r; and Fig. 3 shows the maximum value of m which may be used on the curves of Figs. 1 and 2 for a given value of r.

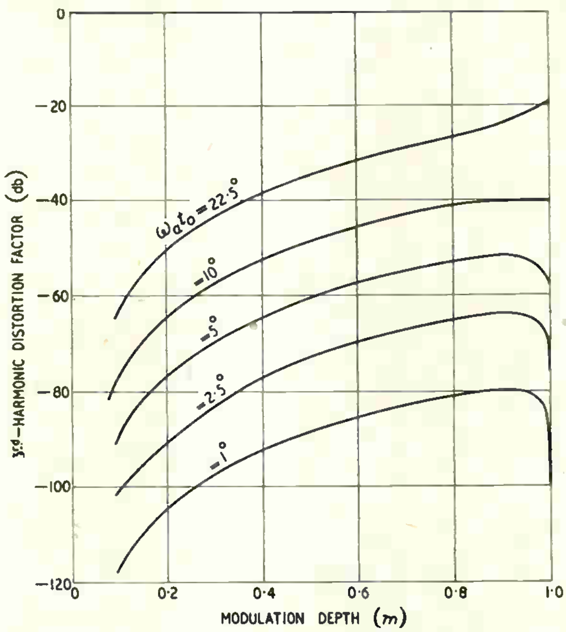

Fig. 2. *Variation of 3rd-harmonic distortion factor with modulation depth and delay time.*

From Fig. 3 it will be seen that echoes having r tending to unity can only be treated for small values of m. The case of r and m both approaching unity occurs in multipath transmission, but is not covered by the present treatment.

From Figs. 1 and 2 it will be seen that the distortion factor increases with increasing values of m for values of $\omega_a t_o$ greater than 10°.

For values of $\omega_a t_o$ less than 10° the distortion factor increases to a maximum value where m is slightly less than 1, and then decreases. In all cases as $m \to 1$ the 2nd- and 3rd-harmonic distortion factors become equal.

By differentiating the original equations for the harmonic distortion with small values of $\omega_a t_o$ it is found that maximum values of 2nd-harmonic distortion occur when $m = 0{\cdot}87$ and 3rd-harmonic distortion when $m = 0{\cdot}91$ (Appendix 2).

A point of somewhat academic interest in the behaviour of these curves very close to $m = 1$ is that Figs. 1 and 2 show finite slopes at this value of m. Such a result is hardly to be expected, since the functions plotted become complex for $m > 1$. In fact, if we differentiate expressions (1) and (2) with respect to m, and put $m = 1$, we find that for all values of $\omega_a t_o$ the gradients approach $+\infty$, so that the curves must bend upwards before they reach $m = 1$. This behaviour is illustrated in Fig. 4. It is seen that the turnover point is extremely close to $m = 1$: in Figs. 1 and 2 this 'hook' on the ends of the curves is, of course, quite unplottable. The phenomenon probably occurs because of the approximation of cos $\omega_a t_o$ and sin $\omega_a t_o$ by the first few terms of their series, and is not considered to be of practical importance.

Comparison of Echo Distortion in F.M. and A.M.

When the echo amplitude is small, the distortion problem in the frequency-modulation case can be solved exactly. It is shown in reference 1 that

$$\frac{\text{2nd Harmonic}}{\text{Fundamental}} = \frac{4r}{m_d} J_2 \left(2m_d \sin \frac{\omega_a t_o}{2} \right) \quad .. \quad (3)$$

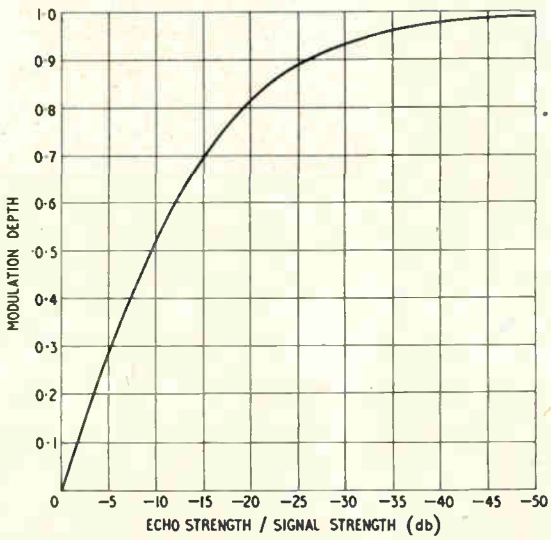

Fig. 3. *Largest modulation depth for which analysis is valid.*

and

$$\frac{\text{3rd Harmonic}}{\text{Fundamental}} = \frac{6r}{m_d} J_3 \left(2m_d \sin \frac{\omega_a t_o}{2} \right) \quad .. \quad (4)$$

It is stated in this reference that "a frequency-modulated system is much more vulnerable to the effects of echo signals than an amplitude-modulated system." Using the above relations, and Figs. 1 and 2, we can put this on a quantitative basis. Table 2 shows values of distortion factors for some typical values of the parameters. While a direct comparison is not easy, owing to the different mechanisms involved in the two systems, it is seen that in the 2nd-harmonic case the distortion factors are in general smaller for a.m. than for f.m. To these distortion-factor values the appropriate reflection factors have to be added to obtain the harmonic distortions. The reflection factor, expressed in db, has in the a.m. case twice its f.m. value [as in equations (1) and (2) an r^2 term appears compared with an r term in equations (3) and (4)] and hence the 2nd-harmonic distortion will be much less when amplitude modulation is used, especially for the smaller echoes.

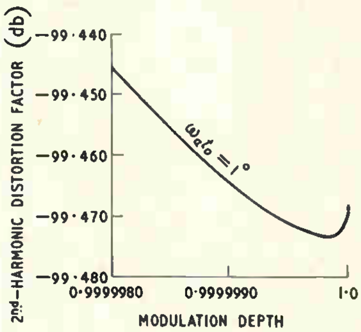

Fig. 4. *Enlarged portion of one of the curves of Fig. 1, near* $m = 1$.

From Table 2 it can also be seen that, in general, the 3rd-harmonic a.m. distortion factor is less than the f.m. for large time delays, but the reverse is true for small time delays. Taking an extreme case of a one degree time delay, an a.m. modulation depth of 0·8, an f.m. modulation index of 1, together with a large echo signal of −20 db; the f.m. system will be 20 db better for 3rd harmonic but 20 db worse for 2nd harmonic than the a.m. system. In nearly all practically

TABLE 2

	Amplitude Modulation				Frequency Modulation			
	$m = 0·1$		$m = 0·8$		$M_d = 1$		$M_d = 10$	
Values of $\omega_a t_o$	1°	22½°	1°	22½°	1°	22½°	1°	22½°
2nd-harmonic distortion (db) ..	− 96	− 42	− 79	− 23	− 76	− 23	− 56	− 16
3rd-harmonic distortion (db) ..	− 116	− 62	− 81	− 26	− 124	− 43	− 84	− 12

Comparison of echo-distortion factors for amplitude- and frequency-modulation systems in some typical cases.

equivalent systems the a.m. performance will be better than the f.m. performance in regard to harmonic distortion due to echo signals which are less than -25 db of the main signal.

APPENDIX 1

Distortion of Amplitude-Modulated Signals

Original Signal $= E(1 + m \sin \omega_a t) \sin \omega_c t$

Echo Signal $= rE[1 + m \sin \omega_a (t + t_o)] \sin \omega_o (t + t_o)$

$\therefore$ Envelope of the resultant wave is

$$= E\left\{ \begin{array}{l} [1 + m \sin \omega_a t]^2 + r^2[1 + m \sin \omega_a(t + t_o)]^2 \\ + 2r[1 + m \sin \omega_a t]\,[1 + m \sin \omega_a(t + t_o)] \cos \omega_c t_o \end{array} \right\}^{\frac{1}{2}}$$

$$= E(1 + m \sin \omega_a t) \left\{ 1 + r^2 \frac{[1 + m \sin \omega_a(t + t_o)]^2}{[1 + m \sin \omega_a t]^2} + 2r \frac{[1 + m \sin \omega_a(t + t_o)]}{[1 + m \sin \omega_a t]} \cos \omega_c t_o \right\}^{\frac{1}{2}}$$

This expression may be expanded by the binomial theorem, and the resultant series will converge rapidly if

$$\frac{r(1 + m)}{1 - m} \leqslant 1{\cdot}0$$

This can be shown as follows:—
The maximum distortion will occur when $\cos \omega_c t_o = 0$ (see later in the analysis) and the remaining term will be maximum when $\omega_a t_o = n\pi$, its value then being

$$\frac{r^2(1 + m)^2}{(1 - m)^2}$$

The analysis uses the first three terms of the binomial expansion

$$(1 + p)^{\frac{1}{2}} = 1 + \frac{p}{2} - \frac{p^2}{8}$$

and for values of p up to $1{\cdot}0$ the error is less than 3%. Taking terms of the binomial expansion to the 2nd order in r

$$\text{Envelope} \approx 1 + m \sin \omega_a t + r \cos \omega_c t_o [1 + m \sin \omega_a (t + t_o)] + \frac{r^2 \sin^2 \omega_c t_o\, [1 + m \sin \omega_a(t + t_o)]^2}{1 + m \sin \omega_a t}$$

Now $\dfrac{1}{1 + m \sin \omega_a t}$ may be expanded in a harmonic series of the form

$$\frac{1}{1 + m \cos(\omega_a t + \pi/2)} = \frac{2}{\sqrt{1 - m^2}} \left[\tfrac{1}{2} + \rho \cos\left(\omega_a t + \frac{\pi}{2}\right) + \rho^2 \cos 2\left(\omega_a t + \frac{\pi}{2}\right) + \rho^3 \cos 3\left(\omega_a t + \frac{\pi}{2}\right) \ldots \text{etc.} \right]$$

$$\text{where } \rho = \frac{m}{1 - \sqrt{1 - m^2}}$$

(reference 2, p. 78).
Using this expansion and collecting the various terms, but neglecting small terms which only contribute to the fundamental,

$$\text{Envelope} = E \left\{ \begin{array}{l} 1 + m \sin \omega_a t + \dfrac{2r^2 \sin^2 \omega_c t_o}{\sqrt{1 - m^2}} \left[\tfrac{1}{2} + m^2 - m\rho \cos \omega_a t_o + \dfrac{m^2 \rho^2}{4} \cos 2\omega_a t_o \right] \\ + \dfrac{2r^2 \sin^2 \omega_c t_o}{\sqrt{1 - m^2}} \left[\begin{array}{l} m \sin \omega_a(t + t_o) - \rho\left(1 + \dfrac{m^2}{2}\right) \sin \omega_a t \\ - \dfrac{m^2 \rho}{4} \sin(\omega_a t + 2\omega_a t_o) + m\rho^2 \sin(\omega_a t - \omega_a t_o) \\ - \dfrac{m^2 \rho^3}{4} \sin(\omega_a t - 2\omega_a t_o) \end{array} \right] \\ + \dfrac{2r^2 \sin^2 \omega_c t_o}{\sqrt{1 - m^2}} \left[\begin{array}{l} - \dfrac{m^2}{4} \cos(2\omega_a t + 2\omega_a t_o) + m\rho \cos(2\omega_a t + \omega_a t_o) \\ - \rho^2 \left(1 + \dfrac{m^2}{2}\right) \cos 2\omega_a t \\ + m\rho^3 \cos(2\omega_a t - \omega_a t_o) \\ - \dfrac{m^2 \rho^4}{4} (\cos 2\omega_a t - 2\omega_a t_o) \end{array} \right] \\ + \dfrac{2r^2 \sin^2 \omega_c t_o}{\sqrt{1 - m^2}} \left[\begin{array}{l} \dfrac{m^2 \rho}{4} \sin(3\omega_a t + 2\omega_a t_o) - m\rho^2 \sin(3\omega_a t + \omega_a t_o) \\ + \rho^3 \left(1 + \dfrac{m^2}{2}\right) \sin 3\omega_a t - m\rho^4 \sin(3\omega_a t - \omega_a t_o) \\ + \dfrac{m^2 \rho^5}{4} \sin(3\omega_a t - 2\omega_a t_o) \end{array} \right] \end{array} \right\}$$

This now requires to be expanded trigonometrically and then the trigonometric functions of $\omega_a t_o$ and $2\omega_a t_o$ must be expanded. If $\omega_a t_o$ lies between 0 and $22\frac{1}{2}°$ (i.e., $2\omega_a t_o$ lies between 0 and 45°) then they may be expanded as follows.

$$\sin \omega_a t_o \approx \omega_a t_o - \frac{(\omega_a t_o)^3}{6}$$

$$\cos \omega_a t_o \approx 1 - \frac{(\omega_a t_o)^2}{2}$$

and similarly for $\sin 2\omega_a t_o$ and $\cos 2\omega_a t_o$. If, however, the delay time or modulating frequency is sufficiently great for these expansions to be invalidated then alternative expansions must be used: for example, if $\omega_a t_o$ lies between 45 and $77\frac{1}{2}°$, expand

$$\sin \omega_a t_o \text{ as } \cos\left(\frac{\pi}{2} - \omega_a t_o\right) \text{ and}$$

$$\sin 2\omega_a t_o \text{ as } \cos\left(2\omega_a t_o - \frac{\pi}{2}\right), \text{ etc.}$$

The remainder of the analysis is evaluated for $\omega_a t_o < 22\frac{1}{2}°$ but the results can easily be obtained for other values of $\omega_a t_o$.

Expanding in this way and collecting terms we have for the 2nd- and 3rd-harmonic components,

$$\text{2nd Harmonic} = \frac{E2r^2 \sin^2 \omega_c t_o}{\sqrt{1-m^2}} \left\{ \begin{aligned} &\cos 2\omega_a t \left[-\frac{m^2}{4} + m\rho - \rho^2\left(1 + \frac{m^2}{2}\right) + m\rho^3 - \frac{m^2\rho^4}{4} \right] \\ &+ \omega_a t_o \sin 2\omega_a t \left(\frac{m^2}{2} - m\rho + m\rho^3 - \frac{m^2\rho^4}{2} \right) \\ &+ (\omega_a t_o)^2 \cos 2\omega_a t \left(\frac{m^2}{2} - \frac{m\rho}{2} - \frac{m\rho^3}{2} + \frac{m^2\rho^4}{2} \right) \\ &+ (\omega_a t_o)^3 \sin 2\omega_a t \left(-\frac{m^2}{3} + \frac{m\rho}{6} - \frac{m\rho^3}{6} + \frac{m^2\rho^4}{3} \right) \end{aligned} \right\}$$

$$\text{3rd Harmonic} = \frac{E2r^2 \sin^2 \omega_c t_o}{\sqrt{1-m^2}} \left\{ \begin{aligned} &\sin 3\omega_a t \left[\frac{m^2\rho}{4} - m\rho^2 + \rho^3\left(1 + \frac{m^2}{2}\right) - m\rho^4 + \frac{m^2\rho^5}{4} \right] \\ &+ \omega_a t_o \cos 3\omega_a t \left(\frac{m^2\rho}{2} - m\rho^2 + m\rho^4 - \frac{m^2\rho^5}{2} \right) \\ &+ (\omega_a t_o)^2 \sin 3\omega_a t \left(-\frac{m^2\rho}{2} + \frac{m\rho^2}{2} + \frac{m\rho^4}{2} - \frac{m^2\rho^5}{2} \right) \\ &+ (\omega_a t_o)^3 \cos 3\omega_{at} \left(-\frac{m^2\rho}{3} + \frac{m\rho^2}{6} - \frac{m\rho^4}{6} + \frac{m^2\rho^5}{3} \right) \end{aligned} \right\}$$

and using the identities

$m\rho^2 - 2\rho + m = 0$; $m\rho^4 = 2\rho^3 - m\rho^2$

$m^2\rho^5 = 2m\rho^4 - m^2\rho^3$; $m\rho^3 = 2\rho^2 - m\rho$

it is found that the coefficients of $\cos 2\omega_a t$; $\omega_a t_o \sin 2\omega_a t$; $\sin 3\omega_a t$; $\omega_a t_o \cos 3\omega_a t$ are zero, and the other terms reduce to

2nd Harmonic =

$$\frac{E2r^2 \sin^2 \omega_c t_o}{\sqrt{1-m^2}} \left[\frac{m^2(1-m^2)}{(1+\sqrt{1-m^2})^2} (\omega_a t_o)^2 \cos 2\omega_a t + \frac{1-\sqrt{1-m^2}-m^2}{1+\sqrt{1-m^2}} (\omega_a t_o)^3 \sin 2\omega_a t \right]$$

3rd Harmonic =

$$\frac{E2r^2 \sin^2 \omega_c t_o}{\sqrt{1-m^2}} \left[(m^2-1)\left(\frac{1-\sqrt{1-m^2}}{1+\sqrt{1-m^2}}\right) (\omega_a t_o)^2 \sin 3\omega_a t + \frac{m(1-\sqrt{1-m^2}-m^2)}{(1+\sqrt{1-m^2})^2} (\omega_a t_o)^3 \cos 3\omega_a t \right]$$

As these harmonic components are in quadrature the square root of the sum of the squares of the two coefficients, in each case, must be taken. Then the amplitudes of the harmonic distortion ratios are:—

$$\frac{\text{2nd Harmonic}}{\text{Fundamental}} = 2r^2 \sin^2 \omega_c t_o (\omega_a t_o)^2 \frac{m}{(1+\sqrt{1-m^2})^2} \{(1-m^2) + (\omega_a t_o)^2\}^{\frac{1}{2}}$$

$$\frac{\text{3rd Harmonic}}{\text{Fundamental}} = 2r^2 \sin^2 \omega_c t_o (\omega_a t_o)^2 \frac{m^2}{(1+\sqrt{1-m^2})^2} \left\{ (1-m^2) + \left(\frac{\omega_a t_o}{m}\right)^2 (1-\sqrt{1-m^2})^2 \right\}^{\frac{1}{2}}$$

When $\omega_a t_o$ is sufficiently small, these expressions become

$$\frac{\text{2nd Harmonic}}{\text{Fundamental}} = 2r^2 \sin^2 \omega_c t_o \frac{m\sqrt{1-m^2}}{(1+\sqrt{1-m^2})^2} (\omega_a t_o)^2$$

and

$$\frac{\text{3rd Harmonic}}{\text{Fundamental}} = 2r^2 \sin^2 \omega_c t_o \frac{m^2\sqrt{1-m^2}}{(1+\sqrt{1-m^2})^2} (\omega_a t_o)^2$$

If we now make m small we arrive at the case considered in Ref. 1. Then, in the worst carrier phase condition ($\sin \omega_c t_o = 1$),

$$\frac{\text{2nd Harmonic}}{\text{Fundamental}} = \frac{mr^2}{2} (\omega_a t_o)^2$$

and

$$\frac{\text{3rd Harmonic}}{\text{Fundamental}} = \frac{m^2 r^2}{2} (\omega_a t_o)^2$$

The results are respectively twice and four times as great as those given in equations (1) and (2) of Ref. 1.

APPENDIX 2

To determine maximum value of distortion for small values of r and $\omega_a t_o$ as m is varied.

$$\frac{\text{2nd Harmonic}}{\text{Fundamental}} = 2r^2(\omega_a t_o)^2 \frac{m}{(1+\sqrt{1-m^2})^2} \{(1-m^2) + (\omega_a t_o)^2\}^{\frac{1}{2}}$$

$$\approx 2r^2(\omega_a t_o)^2 \frac{m\sqrt{1-m^2}}{(1+\sqrt{1-m^2})^2}$$

This has a maximum value when $m = \sqrt{\frac{3}{4}} = 0{\cdot}866$

$$\frac{\text{3rd Harmonic}}{\text{Fundamental}} \approx 2r^2(\omega_a t_o)^2 \frac{m^2\sqrt{1-m^2}}{(1+\sqrt{1-m^2})^2}$$

This has a maximum value when $m = \sqrt{2(\sqrt{2}-1)}$ $= 0{\cdot}910$.

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[2] S. A. Schelkunoff, "Applied Mathematics for Engineers and Scientists," New York, D. Van Rostrand Co., 1948.

NOISE FACTOR OF CONVENTIONAL V.H.F. AMPLIFIERS

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1. Introduction

NOISE factor has been treated extensively in recent years, but a paper giving a simple treatment while avoiding error through over-simplification will be of some value to engineers, for there has as yet been very little production of up-to-date low-noise amplifiers.

The importance of obtaining a satisfactory signal-to-noise ratio by the reduction of receiver noise rather than by an equivalent increase in transmitter power is primarily a matter of economics. It is particularly important in any communications system using a number of transmitters comparable with the number of receivers, in much laboratory measurement work where noise sets a limit to the accuracy, and in installations where large heavy transmitters cannot be considered.

Although the improvement to be gained in the intermediate-frequency stage of a microwave receiver is not large, it is worth while, particularly as an established advance is applicable to all such systems, and hence it is economic to spend a lot on development. In addition, efforts have been made to acquire a better understanding of valve noise at these frequencies where the experimental work is not too difficult, and considerable data is available. The reader is warned against extending the arguments to circuits at much higher frequencies without checking the approximations.

2. Basic Considerations

2.1. *Noise Formulae*

An acquaintance with noise phenomenon is assumed, but for those readers who are not thoroughly familiar with the main expressions these are given below. An excellent introductory paper to the subject is that by Pierce.[1]

Thermal Noise

Equivalent series generator $v^{2\Sigma} = 4kTRdf$*

Equivalent parallel current generator $i^{2\Sigma} = 4kTGdf$

where:

$v^{2\Sigma}(i^{2\Sigma})$ = Mean square value of the effective fluctuating voltage (current).

$R(G)$ = Effective resistance (conductance) of any linear impedance at a temperature of T°K.

df = Frequency band considered.

k = Boltzmann's constant.

If $v^{2\Sigma}$ is in (volts)², R in ohms and df in c/s, the value of k is $1{\cdot}38 \times 10^{-23}$.

Shot Noise[2]

$$i = {}^{2\Sigma}F^2\, 2e\, I\, df$$

If $i^{2\Sigma}$ is in (amps)², I in amperes and df in c/s, the value of e is $1{\cdot}6 \times 10^{-19}$.

F^2 is the ratio of the value of shot noise to the value when I is temperature limited (i.e., when the current is perfectly random).

For amplifier valves, an equivalent noise 'resistance' is defined as that value of resistance giving thermal noise at the grid–cathode terminals which would give noise in the output equal to the shot noise. This resistance is assumed to be at a standard temperature of $T_0 = 290$°K.

Then $4k\,T_0 R_{equ}\, g_m^{2\Sigma} = F^2\, 2eI$ for a triode of mutual conductance g_m.

Since $g_m^{2\Sigma} = g_m^{\Sigma 2}$ for small signals,

$$R_{equ} = \frac{F^2\, 2eI}{4kT_0 g_m^2}$$

It is approximately true that $R_{equ} \propto \frac{1}{g_m}$ for a triode.

Induced Grid Noise

Grid–cathode noise generator $i^{2\Sigma} = t.4kT_0 G_T$

where G_T = input conductance due to transit time.

t = constant (usually taken as 5).

At the frequencies considered, it is assumed that the above are the only sources of noise, and that F^2, G_T and t are independent of frequency over the frequency response of the amplifier.

2.2. *Available Power*[3]

The concept of available power seems to cause considerable difficulty to the uninitiated. Once understood, it does simplify many noise calculations, particularly with simple networks, apart from its value in coaxial line and waveguide work.

The available power from a generator is equal to that which would be delivered into a matched

* $x^{2\Sigma}$ ≡ mean square of x
$x^{\Sigma 2}$ ≡ square mean of x.

MS accepted by the Editor, February 1953.

load. With other systems of rating a generator it is necessary to measure impedance, whereas available power is a unique parameter. Hence the available thermal noise power from a resistance is $kTdf$ and is independent of the resistance value.

The power gain is defined as:

$$m = \frac{\text{Available power at the output}}{\text{Available power from the input source}}$$

and is independent of the actual load impedance connected to the output terminals, but is dependent on the mismatch between the source and the input terminals. With any linear four-pole network, the actual power into the load is given by $mP\eta$ where P is the available power from the input source and η is the transfer efficiency at the output.

It is important to appreciate the significance of this definition of gain. Mismatch coefficients are eliminated and the overall gain of a multistage amplifier is $m_1 \, m_2 \, . \, . \, . \, m_r$. Moreover, the gain may be increased and yet the actual power output or voltage step-up ratio decreased because the output transfer coefficient is degraded more than the gain is increased. The available power from a network may be a function of frequency even though the available power from the source is independent of frequency. Hence a power bandwidth or noise bandwidth B is defined

$$B = \frac{\int m \, df}{M}$$

where m is the gain within a narrow frequency range df and M the mid-band value.

For noise over a frequency band the available power definition assumes a theoretical match to the source at all frequencies. Hence the ratio of actual power to available power is a function of frequency. A source with an impedance equivalent to a resistance (at temperature T) and reactance in series or parallel gives an available noise power of $kTdf$ and this is independent of frequency. Any passive network containing resistance elements at the same temperature satisfies this condition, even though the contributions to the available power from the individual elements may be frequency dependent.

2.3. *Noise Factor*

The definition used is that framed in terms of available power gain. This is not strictly equal to the measured value under all conditions, but leads to simple formulae without invalidating the arguments (see Appendix 1).

The noise factor N is a dimensionless figure of merit of the noisiness of a receiver (or part of receiver) under standard conditions, as compared with the unavoidable thermal noise arising in the source feeding the receiver. It is only applicable to linear systems in which the output signal-to-noise ratio is directly proportional to the input signal.

$$N = \frac{\text{Available noise power at the output}}{\text{That part of the available noise power arising from thermal noise in the source within the receiver passband}}$$

where

(a) the receiver (or part) is connected to the same value of source impedance as used in practice but comprising passive elements at a standard temperature of 290°K;

(b) the output is measured before demodulation. (This is implied by the linear qualification.)

It follows that

$$N = \frac{\text{Available noise power at output}}{\int kT_0 m df}$$

where m = the power gain at any frequency, and the integral is taken over the range of frequencies comprising the signal channel only.

Available noise power at output

$$= N \, kT_0 \int mdf = N \, kT_0 B \, M.$$

In effect, a receiver may be considered as noiseless and the signal source feeding it as having available noise power of N times its own thermal noise power—or rather that $(N - 1) \, kT_0 df$ is available at the input in addition to thermal noise. It should be remembered that N is a function of source impedance.

The noise factor is often expressed in decibels thus:

$$N \text{ (decibels)} = 10 \log_{10} N.$$

If we apply the definition of noise factor to a multistage receiver, the overall noise factor may be expressed in terms of the noise factors of the individual stages. To avoid confusion, different symbols have been used for the noise factor under different conditions.

The expressions are:

(a) Single-frequency noise factor (the value within a narrow frequency range df over which it is constant)

$$u = u_1 + \frac{u_2 - 1}{m_1} + \frac{u_3 - 1}{m_1 m_2} + \dots \qquad (1)$$

u_r and m_r are the noise factor and available power gain of the rth stage within the frequency range df.

(b) Mid-band noise factor (the single-frequency noise factor at the centre of the frequency response of the full amplifier)

$$n = n_1 + \frac{n_2 - 1}{M_1} + \frac{n_3 - 1}{M_1 M_2} + \dots \qquad (2)$$

M_r is the available power gain at mid-band.

(c) Full-band noise factor (the integrated effect over the frequency response of the amplifier)

$$N = N_1 + \frac{N_2 - 1}{M_1}\frac{B_2}{B_1} + \frac{N_3 - 1}{M_1 M_2}\frac{B_3}{B_1} + \ldots \quad \ldots \quad \ldots \quad (3)$$

$$\text{and } N = \frac{\int um\, df}{\int m\, df} \quad \ldots \quad \ldots \quad (4)$$

(d) The noise bandwidth

$$B_r = \frac{\int (m_r\, m_{r+1} \ldots .)\, df}{M_r\, M_{r+1} \ldots .} \quad \ldots \quad \ldots \quad (5)$$

In usual practice only the first two terms in equations (1), (2) and (3) are of importance.

2.4. *Passive Network*

If a network which is equivalent to a fixed resistance in parallel with a reactance is fed from a similar source, then the power gain (or attenuation) in available powers is independent of frequency. Hence the noise bandwidth is infinite, although of course the bandwidth of a succeeding network may be dependent on the reactances present.

The noise factor of such a network at a standard room temperature of 290°K is equal to the attenuation (reciprocal of gain in available powers). This follows immediately from the noise factor definition.

Consider the input of an amplifier with resistive damping (assumed at a temperature of 290°K).

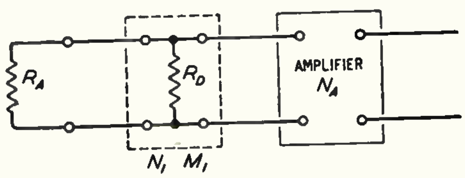


Fig. 1. *Damping on input.*

The effect on noise factor may be obtained from equation (3), by treating the damping as a stage in the amplifier of Fig. 1.

This gives:

$$N = N_1 + \frac{N_A - 1}{M_1}$$

$$= \frac{N_A}{M_1} \quad \text{since } N_1 = \frac{1}{M_1}$$

where N_A = noise factor of the amplifier when fed from the same source as the output impedance of the network under actual conditions, and hence is a function of R_D.

When R_A/R_D is small, as in all practical low-noise amplifiers, the change in N_A due to R_D is negligible, and the increase in noise factor N, due to the circuit losses R_D, is given very closely by $\frac{1}{M_1} = \left(1 + \frac{R_A}{R_D}\right)$ times.

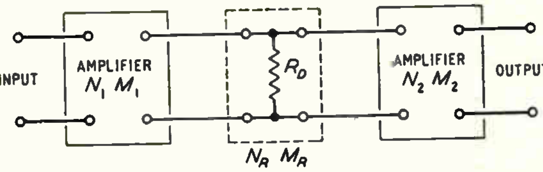


Fig. 2. *Inter-stage damping.*

The effect of losses in the coupling circuit between two stages of an amplifier may be analyzed in a similar way. Referring to Fig. 2, we have

$$N = N_1 + \frac{N_R - 1}{M_1} + \frac{N_2 - 1}{M_1 M_R}\frac{B_2}{B_1}$$

$$\approx N_1 + \frac{N_R - 1}{M_1} + \frac{N_2 - 1}{M_1 M_R}$$

neglecting bandwidth terms.

If the network is considered a part of the first stage

$$N = N'_1 + \frac{N_2 - 1}{M'_1} \quad \ldots \quad (6a)$$

where
$$N'_1 = N_1 + \frac{N_R - 1}{M_1} \quad \ldots \quad (6b)$$

$$M'_1 = M_1 M_R \quad \ldots \quad \ldots \quad (6c)$$

If the network is considered a part of the second stage

$$N = N_1 + \frac{N'_2 - 1}{M_1} \quad \ldots \quad (7a)$$

where
$$N'_2 = \frac{N_2}{M_R} \quad \ldots \quad \ldots \quad (7b)$$

If R_D is small compared with the output resistance of the first stage, then if the first stage is switched off, N'_2 is almost unchanged and the noise output is therefore $(N'_2 - 1)\, M_2\, kT\, df$.

Hence the ratio Y of the two noise outputs (with and without first stage) is given by:

$$Y = \frac{N\, M_1}{N'_2 - 1}$$

and substitution from (7a) gives

$$N_1 = N\,(1 - 1/Y) \quad \ldots \quad (8)$$

If R_D is not much smaller than the output resistance of the first stage, it is necessary to measure the ratio Y' of the normal noise output to that when the first stage is replaced by a resistance of the same value as its output resistance. This leads to:

$$N_1 = N\,(1 - 1/Y') + \frac{1}{M_1}$$

This equation is effectively the same as (8) unless $1/M_1$ is appreciable compared with N/Y'.

2.5. *Equivalent Valve Circuit*

The basic circuit for the triode without feedback is shown in Fig. 3. The single-frequency noise factor, u, may be calculated from the ratio of available noise power at the output terminals to that part of it which originates from the thermal noise in R_A.

All resistances are assumed to be at the standard reference temperature of $T_0 = 290°K$. The fictitious noise generator $4k\,T_0R_{equ}\,df$ is also defined for the same temperature.

R_A = Resistive term of the source impedance at the valve terminals.
R_D = The total grid–cathode losses due to coils, valve base and holder, etc.
jX = The total effective input reactance including the valve grid–cathode capacitance.
R_{equ} = The equivalent shot noise resistance.
R_T = The transit time input resistance.

The full-band noise factor is obtained from equation (4).

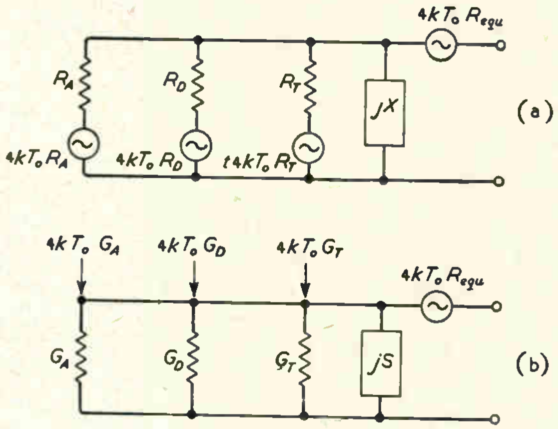


Fig. 3. *Basic equivalent noise circuit.*

The assumptions made when using the above circuit are:

(a) That noise due to transit time can be considered purely as a grid–cathode generator which is uncorrelated with shot noise, and whose magnitude is independent of frequency over the band considered.

(b) That $g_m^2 = g_m^{2\Sigma}$; i.e., that the valve is operated over a linear part of its characteristic.

If feedback is introduced from the output to input circuits by a lossless (noiseless) circuit, then the ratio of single-frequency signal/noise powers at the output is practically unchanged. This statement is not exactly true since the shot noise originates between anode and cathode and all other sources originate between grid and cathode. The correction to the shot noise term will be small and at most of the order of $\left(\frac{\mu \pm 1}{\mu}\right)^2$, where μ is the valve amplification factor.

If the feedback is frequency selective, this is allowed for in calculating the full-band noise factor by taking into account the modified frequency response. To emphasize this, the equivalent circuit can be changed to that of Fig. 4, where the feedback is represented by a finite input admittance $G_i + jS_i$, which is noiseless.

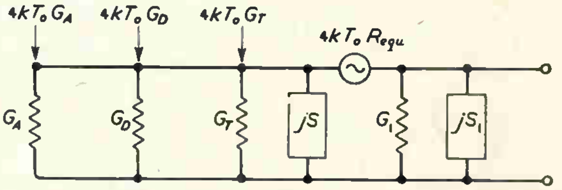


Fig. 4. *Equivalent circuit with feedback.*

The overall single-frequency noise factor of an amplifier may be affected by feedback in the first stage, even though the feedback circuit is noiseless. The output impedance and available power gain of the first stage will be changed, and hence the term due to noise from succeeding stages may be changed. The magnitude of the effect can be calculated using the method of 2.4, with sufficient accuracy for practical circuits.

When feedback is introduced by a circuit which is not in itself noiseless, the noise introduced by this circuit can be allowed for in the following way. The noise due to the feedback circuit can be represented by noise current generators in the grid–cathode and anode–cathode circuits, assuming that there is no feedback present. The equivalent circuit can then easily be modified to allow for these additional noise sources, the anode–cathode noise generator being represented as an extra source in series with $4k\,T_0R_{equ}\,.\,df$. The fact that feedback is present in practice can then be treated as before, as though the circuit is noiseless. Two precautions must be observed. The phase relation between the two new noise sources must be allowed for, and the equivalent circuit must include that part of the input admittance due to the feedback circuit, which is not due to feedback (not a function of valve gain).

Having introduced the equivalent circuit, it can now be pointed out that the effect of a damping resistance R_L in the output circuit, which was considered in 2.4, is very easily allowed for if the output is between anode and cathode. It is then equivalent to an increase in R_{equ} of $\frac{1}{R_L g_m^2}$, and, of course, subject to the same small error as is R_{equ} with certain feedback circuits.

2.6. *Variation of Single-Frequency Noise Factor over the Pass-Band*

The noise factor of a single triode stage, calculated from Figs. 3 and 4, is given by:

$$u_1 = 1 + \frac{G_D}{G_A} + t \cdot \frac{G_T}{G_A} + \frac{R_{equ}}{G_A}(G^2 + S^2) \quad \text{(9a)}$$

where $G = G_A + G_D + G_T$.

Or
$$u_1 = 1 + \frac{G_D}{G_A} + t \cdot \frac{G_T}{G_A} + G^2 \frac{R_{equ}}{G_A} (1 + Q_0^2 (f/f_0 - f_0/f)^2) \quad \text{(9b)}$$

where

Q_0 = the effective tuned circuit Q including G_D and G_T but *excluding any fed-back admittance*,

f_0 = the resonant frequency of the parallel tuned circuit *excluding any fed-back admittance*.

The expression shows that the value for u_1 increases at frequencies either side of f_0. The mid-band noise factor is a minimum when f_0 is the centre frequency of the overall frequency response.*

The full-band noise factor is also a minimum when $f = f_0$ at the centre of the overall response.* This can be clearly shown for the theoretical rectangular frequency response, when

$$N_1 = \frac{\int_{-B/2}^{+B/2} u_1 \, df}{B}$$

Fig. 5 shows the difference in N_1 when the curve for u_1 is asymmetrical about mid-band from when it is symmetrical. The shaded portion shows the increase in total noise. With the usual sloping-sided frequency response, of the same noise bandwidth, such asymmetry in u_1 will cause a greater increase in N_1. (See Appendices 2 and 3.)

When there is input susceptance due to feedback, it is difficult to obtain the best value of N_1 and retain a symmetrical frequency response. For, with u_1 a minimum at mid-band, the input circuit, including the fed-back admittance, will have an asymmetrical frequency response and it is necessary to compensate this by opposite asymmetry in later stages.

In addition to the increase in N_1 due to variation of u_1 over the pass-band, there is a similar increase in N_2, and hence the effect on the overall noise factor is slightly worse than on the first stage alone.

At the frequency f_0, formula (9) becomes:

$$u_1 = 1 + \frac{G_D}{G_A} + t \cdot \frac{G_T}{G_A} + R_{equ} \frac{G^2}{G_A} \quad \text{(10)}$$

This is of the form $u_1 = a + \frac{b}{G_A} + d G_A$, and will have a minimum value when $G_A = \sqrt{\frac{b}{d}}$. This value of G_A will usually give also the minimum value for N_1 and N.

However, it is not always possible to choose this value of G_A because the frequency response of the input circuit may then be insufficiently wide; even though it may still be possible to compensate for this circuit's reduced bandwidth, by feedback or other methods, the variation in u may be so excessive as to cause the value of N to be greater than could be achieved with a larger value of G_A.

With a wideband amplifier it is also necessary to add extra damping in the output circuit of the first stage. The extra noise due to this, and the reduction in the power gain, cause an increase in the overall noise factor.

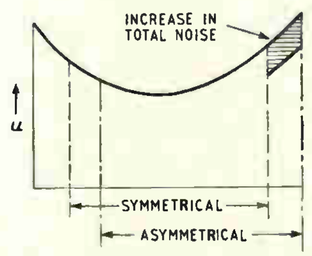


Fig. 5 (above). *Effect of variation in single-frequency noise factor.*

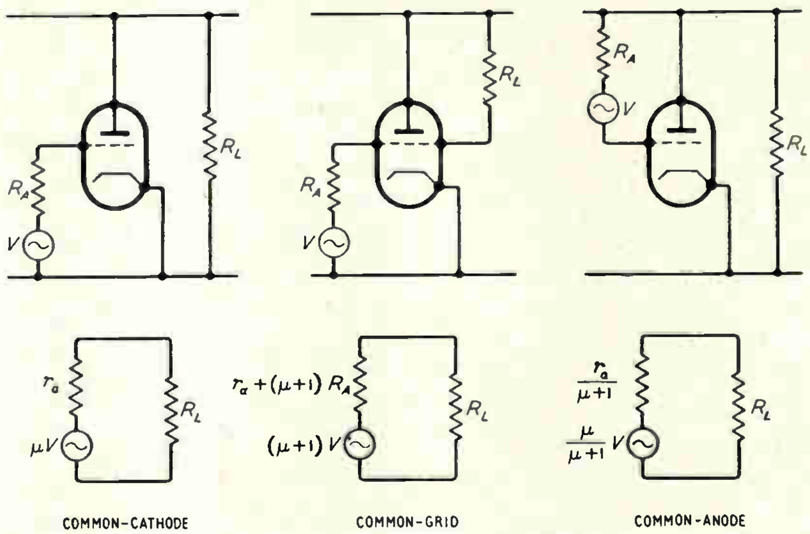


Fig. 6 (right). *Triode connections.*

* Not strictly true—see 3.5.

3. Triode Input Stages

The use of a triode for the first stage when a low-noise factor is required is now so well established that the pentode is not considered in this paper. It has been shown[4] that a pentode can be operated so that partition noise is cancelled by special feedback arrangements, but a more practical way to achieve such cancellation is to connect anode to screen.

3.1. *Comparison of Triode Circuits*

The three triode circuits are shown in Fig. 6, together with their equivalent circuits, for purely resistive circuits.

The common-cathode equivalent noise circuit is that of Fig. 4, since feedback due to grid–anode capacitance and cathode-lead inductance is effectively noiseless.

With the common-grid circuit, the feedback is direct from output to input by the load circuit, grid-lead inductance and anode–cathode capacitance. Since noise of the load circuit may be taken into account as a second stage, the same equivalent circuit may be used with appropriate modifications in S_i, G_i and S.

With the common-anode circuit, the feedback is direct from output to input via the load circuit, anode-lead inductance, and grid–cathode capacitance. Again, the noise of the feedback coupling may be considered a part of the second stage. An equivalent circuit such as shown in Fig. 7 cannot be used, however, because of the difference in feedback for shot noise and induced grid noise. It is shown in Appendix 6 that the effect of induced grid noise is reduced slightly, but this advantage is not sufficient to outweigh the disadvantages of this type of circuit.

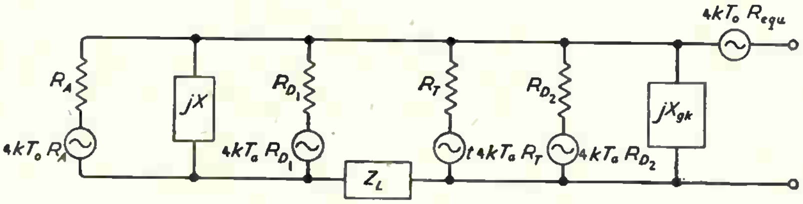


Fig. 7. *False equivalent circuit for common-anode (cathode follower).*

To compare the three circuits, we may consider a common-anode circuit in which the noise factor of the valve stage, excluding output circuit losses, is identical to the other two circuits. The effect of output circuit losses can be taken into account using equation (6). In Appendix 4 the relative merits are shown in some detail, but with simplifying approximations as follows they are more easily appreciated.

With valves used at v.h.f., R_A is of the order of 300 to 3,000 ohms, and g_m is at least 5×10^{-3}A/V. Hence g_mR_A is greater than unity. The minimum value required for R_L in wideband amplifiers is approximately 500 ohms, and therefore g_mR_L is also greater than unity. Approximating accordingly, the relative merits are shown in Table 1, on page 287, for conditions least favourable to the common-cathode circuit.

The common-cathode gives greatest gain and hence least effect due to noise from succeeding stages. The noise factor of the common-grid is more appreciably affected by the output circuit losses.

With the common-grid and common-anode circuits there is such excessive feedback that the overall (gain × bandwidth) product of the complete stage is seriously reduced, because either the input or output circuit is extravagantly wideband. For very narrow bandwidths it is possible to make all three circuits give an overall noise factor which approximates closely to n, but if the noise from succeeding stages is not negligible the common-cathode shows a marked superiority over the other two circuits.

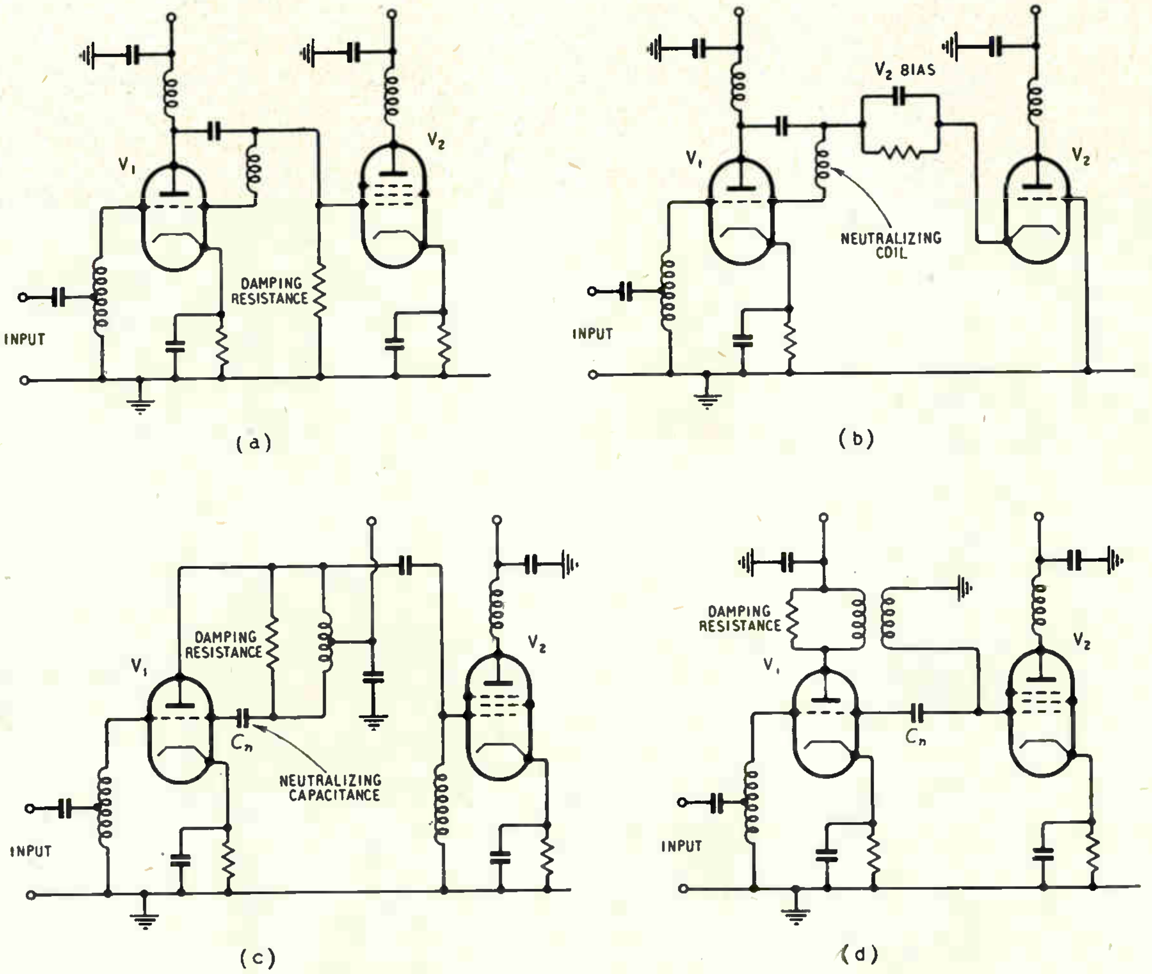
The common-grid circuit has one advantage of not requiring special neutralization, since the anode–cathode capacitance can be made very small by suitable valve design. However, there may be appreciable input susceptance due to feedback (of order $-\omega g_m R_L C_{ak}$),* and, although the effect on the frequency response is small, it will have a marked effect on noise factor if the input circuit is tuned to resonance including this.

3.2. *The Neutralized Common-Cathode Circuit*

The feedback due to the anode–grid capacitance does not directly affect the expression for u_1, but it does make the frequency response asymmetrical; and if the voltage amplification is large the response is extremely frequency sensitive. Some of the possible methods of neutralization are shown in Fig. 8.

That shown in Fig. 8(a) is not very suitable because the circuit requires excessive damping in the anode. Wallman, Macnee and Gadsden[5] have avoided this by using a common-grid second stage, Fig. 8(b), which gives the necessary damping without adding any more noise than that due to the common-grid circuit. The author prefers the circuits (c) and (d); (d) gives the more accurate phase reversal unless capacitance loading is added to match the capacitance of V_2 with the circuit (c), but the double-wound coil in (d) has a high capacitance.

* Actual expression $C_i = \dfrac{C_{ak} r_a (r_a - \mu R_L)}{(r_a + R_L)^2 + \omega^2 C_{ak} R_L^2 r_a^2}$.



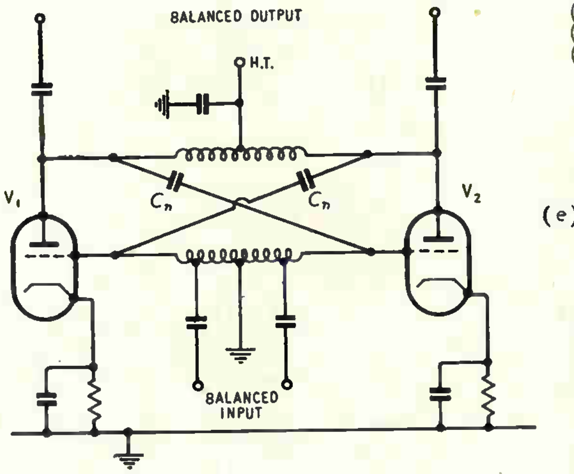


Fig. 8. *Neutralized circuits.*
(*a*) *Resonance neutralizing feeding pentode.*
(*b*) *Resonance neutralizing feeding common grid.*
(*c*) *Capacitance neutralizing with auto-transformer.*
(*d*) *Capacitance neutralizing with conventional transformer.*
(*e*) *Push-pull.*

TABLE 1

Circuit	*Available Power Gain including output damping*	*Noise Factor including output damping*
Common-cathode	$g_m^2 R_A R_L$	$n + \frac{1}{g_m^2 R_L R_A}$
Common-grid	$\frac{R_L}{R_A}$	$n + \frac{(1 + g_m R_A)^2}{g_m^2 R_L R_A}$
Common-anode	$\frac{g_m^2 R_A R_L}{1 + g_m R_L}$	$n + \frac{1}{g_m^2 R_L R_A}$

This type of neutralization is not dependent on using a low amplification circuit to avoid reduction of bandwidth.

The equivalent noise circuits for Figs. 8(b) and (d) are shown in Fig. 15. The anode circuit damping has not been included as part of the first stage.

The advantages of the capacitance-neutralized circuit are:—

The first stage gives approximately the same voltage amplification as the first two stages of Fig. 8(b).

The loss in noise factor due to neutralizing coil losses is eliminated.

The circuit is more suitable for miniature construction. (The neutralizing coil in Fig. 8(b) requires many more turns than is usual in an amplifier at these frequencies.)

The disadvantages are:—

The circuit of Fig. 8(d) increases the capacitance affecting variation of u_1 by $2C_{ga}$ as compared with C_{ga}. However, in practice, the anode-grid capacitance is appreciably increased by the neutralizing coil and leads used in the circuit in Fig. 8(b).

The circuit is more affected by slight errors in neutralizing.

The anode coil of Fig. 8(d) may have such a high capacitance that the bandwidth of the anode circuit can only be obtained by using heavy damping.

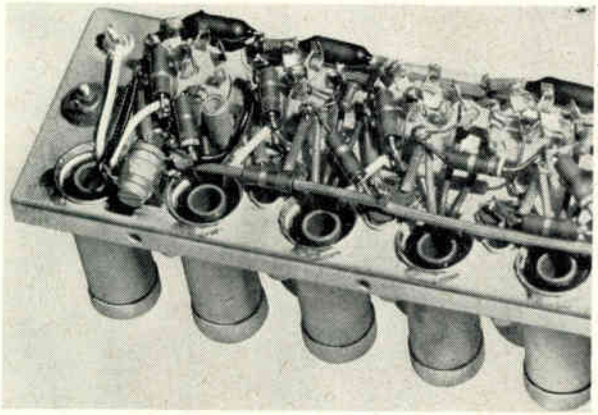


Fig. 9. *Input stage of typical unit using capacitance neutralization—circuit used in 1946.*

In practice, the coil for the anode circuit shown in Fig. 8(d) gives an increase in circuit capacitance of the order of 10 pF even when wound with spaced turns. The two windings are usually in the same sense and the secondary connections are reversed. For wideband amplifiers it is better to use the circuit of Fig. 8(c). It might be thought that the damping required in the anode will give a slight increase in overall noise factor as compared with the inductance-neutralized circuit. However, the damping conductance required in the output circuit of the common-grid stage [Fig. 8(b)] will be approximately equal to that in the anode of the common-cathode stage [Fig. 8(c)] and this will have just as serious an effect on overall noise factor as the same conductance between grid and cathode of V_2. (See Appendix 5.)

Fig. 9 is a photograph of a unit used at T.R.E. in 1946, with neutralizing circuit as Fig. 8(d), and Fig. 10 shows a recent laboratory unit. Suitable trimmer capacitors for neutralizing are the Polar C32-01 (1-5 pF) and the S5001.

The circuit of Fig. 8(e) may be useful when the input is normally push-pull, as for example with a balanced crystal mixer using crystals of like polarity, but increases the tolerances of unbalance.

With the neutralized circuit it is possible to obtain an overall noise factor which is not significantly greater than the noise factor of the first stage only. To achieve this with bandwidths of the order of 10 Mc/s it may be necessary to use two neutralized triode circuits in cascade.

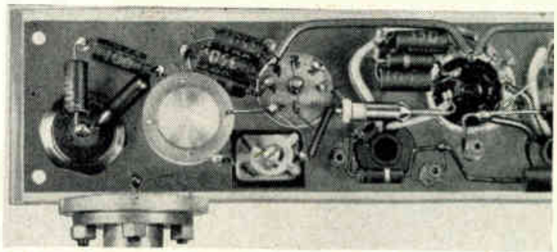


Fig. 10. *Input stage of typical unit using capacitance neutralization—recent circuit.*

3.3. *The Input Coupling*

In wideband amplifiers (greater than about 5-Mc/s bandwidth), the value of source conductance at the grid of the first stage is limited by frequency response considerations. The best value of full-band noise factor, in such circumstances, would be obtained by a compromise between improvement in mid-band value and the variation over the band. However, if this value of source conductance causes the bandwidth of the input circuit to be no longer much larger than the overall bandwidth, the design of an amplifier is uneconomical in a number of stages. In addition, the overall frequency response is made more dependent on source admittance and variations in this may have a marked effect.

This latter factor is extremely significant when the amplifier is coupled to a crystal mixer for, apart from tolerances in crystal parameters, appreciable changes can be produced by changes in the radio-frequency conditions of mixer operation.

It might be thought that an additional effect would arise due to the bandwidth terms causing an increase in noise from later stages [equation (3)], for as the 'cut' in the input circuit is increased the terms B_2/B_1, etc., are increased. However, this is counteracted by the increased power gain of the first stage when the source conductance at the grid is reduced.

A suggested compromise is that the input circuit should be designed to give a bandwidth at 1 db down at least as great as the overall bandwidth at 3 db down. This gives very little increase in full-band noise factor on the mid-band value, and does not cause a significant increase in the bandwidth and hence the number of later stages.

It is particularly important to keep the capacitance at the input as small as possible, and to use a high Q coil to minimize coil losses; it is therefore well worth the effort to design crystal-mixer decoupling and mounting arrangements to reduce capacitance. Sufficient space for a good coil is usually much cheaper than the equivalent effect in transmitter power.

It is possible to obtain a marked reduction in noise factor of a wideband amplifier by using a complex input circuit. Such improvement will be at the expense of making the response more critically dependent on the source admittance, and the previous warning about variations with crystal mixers is particularly relevant. For practical reasons the simple tuned circuit is preferred.

The equivalent noise circuit using a so-called unequal-Q transformer is shown in Fig. 11.

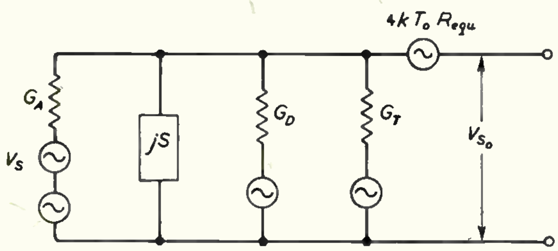


Fig. 11. *Equivalent circuit for coupled circuit input.*

Over the range of frequencies for which the response is 'flat',

$$\frac{V_S^2 G_A^2}{G^2 + S^2} = V_{S0}^2 = \text{constant}$$

and $V_S^2 G_A$ = constant (available power), since any transformer loss is included in G_D.

$$\text{Therefore} \quad \frac{(G^2 + S^2)}{G_A} = \text{constant} = D$$

$$\text{Now} \qquad u_1 = 1 + \frac{G_D + tG_T}{G_A} + \frac{(G^2 + S^2)}{G_A} R_{equ}.$$

$$\text{Therefore } u_1 = 1 + \frac{G_D + tG_T}{G_A} + D\, R_{equ}.$$

G_A is here a function of frequency, and the variation of u_1 is due to the variation of the term $(G_D + tG_T)/G_A$.

The variation in u_1 is slightly less than for the single circuit, for with a wideband amplifier the term containing R_{equ} is usually greater than the other. This advantage is hardly significant if the input circuit gives only 1 db 'cut'.

The major advantage of this circuit is obtained because of the lower effective source conductance at the grid for the same effective bandwidth as the single-tuned circuit. The price paid is critical adjustment and dubious frequency response in practice.

3.4. *Use of Feedback*

If negative feedback is used to give adequate damping of the input circuit, then the source conductance is not limited by frequency response considerations. The benefit in noise factor is not so great as deduced by Strutt and Van der Ziel[4] because of the variation in u over the frequency response. It is impossible to arrive at a general solution of the advantage obtainable with feedback damping, but it is doubtful whether any improvement can be obtained in this way, or if so, whether it would be sufficient to justify the extra complexity and unreliability. The increase in variation of u over the response and the effect of second-stage noise may in fact give a net increase in full-band noise factor.

With a neutralized-triode circuit, such feedback may be obtained by cathode-lead inductance, which does not itself add directly to the first-stage noise. The amount of feedback necessary to achieve a useful input conductance will, however, give some increase in noise factor and also produce a significant input susceptance.

It is certainly not possible to obtain any improvement by feedback using a resistance between grid and anode. The amount of feedback required will be at least such as to give an input conductance of the same order as the source at the grid. The value of resistance required is of the order of A times the input resistance, where A is the voltage amplification. The effect of this resistance on the noise factor will be to increase it about $\left(1 + \frac{1}{A}\right)$ times. Since A will not be much greater than 5 in a wideband amplifier for which feedback damping is desired, there would be a definite overall loss in noise factor. This same argument applies for feedback by a resistance from the anode of V_2 to grid of V_1 with the circuit of Fig. 8(b).

There are some circumstances in which it might be advantageous to use feedback to give damping. When the source admittance is fixed, damping may be necessary to achieve the frequency response, and feedback will be better than adding a resistor in parallel. This is rather different from the usual input circuit considerations where there is no question of adding damping but only changing the effect of the valve capacitance on the frequency response. However, if a common-grid circuit be used for this purpose, the same order of damping in its anode circuit will be required, and this will be just as serious as adding the damping in the first place. In fact, with a number of common-grid circuits in cascade, damping in the output of the last stage will have just as serious an effect on the overall noise factor as when applied to the input circuit. (See Appendix 5.)

If, with a wideband amplifier, the initial CR value of the source is greater than required to give the input circuit bandwidth, some damping by negative feedback will be valuable.

3.5. *Asymmetry Effects*

As previously indicated, the condition $f = f_0$ at mid-band does not give exactly the minimum noise factor. This inaccuracy is due to neglect of phase correlation between shot noise and the grid–cathode transit-time noise. It is an experimental fact that the single-frequency noise factor is a minimum when the source admittance is capacitive, excluding any fedback admittance.

This does not invalidate the arguments on the variation of u with frequency, but a correction must be applied.

A further advantage of the neutralized circuit may now be appreciated. By an appropriate adjustment to the neutralizing, so that the correct amount of negative capacitance is produced by feedback, it is possible to obtain minimum noise factor with a symmetrical frequency response. The amount of capacitive detuning required is dependent on the valve used.

3.6. *Valve Requirements*

Because the neutralized-triode circuit enables a noise factor to be obtained, which is effectively determined by the first valve only, further improvements are dependent on valve development.

From equation (9) it is possible to derive the criteria for the valve. These are different for n (narrow-band) and N (wideband amplifier) and are:—

For narrow-band operation: Small transit angle (grid–cathode noise).

For wideband operation: R_{equ} $(C_{ge} + 2C_{ga} + C_{strays})$ to be small.

Since, to a first approximation $R_{equ} \propto 1/g_m$, the wideband and narrow-band requirements are both dependent on the same valve techniques; namely, reduced grid–cathode clearance, fine wire grids, high current density at the cathode, smooth cathode surface and miniature construction. For both requirements, any improvement will also lead to higher power gain and reduced noise effect from succeeding stages. It can easily be shown that operation of two valves in parallel or push-pull does not change the optimum noise factor, or power gain with source impedance giving optimum noise factor.

Although of little importance, the requirement for output parameters is a low value of $r_a\, C_{ak}$. There is a slight advantage in a low value of r_a (other parameters being unchanged), except when a value less than necessary for the anode-circuit bandwidth is obtained. This is easily shown, the condition being that the total anode damping be constant.

We get:—

$$N = N_1 + \frac{r_a}{R_L M_1} + \frac{(N_2 - 1)(1 + r_a/R_L)}{M_1}$$

Now $M_1 = \alpha\, r_a$

$G_L + g_a = \text{Constant } \beta$

$$\text{Hence,}\quad N = N_1 + \frac{1}{\alpha R_L} + \frac{(N_2 - 1)\dfrac{(R_L + r_a)}{R_L}}{\alpha r_a}$$

$$= N_1 + \frac{1}{\alpha R_L} + \frac{\beta(N_2 - 1)}{\alpha}$$

That is, the maximum value of R_L (minimum r_a) gives the lowest value for N.

In the foregoing, an effect which may have an opposing bearing on valve design has been neglected. This is the possible existence of an additional noise source in the grid–cathode circuit, which may not be minimized by the same techniques.

3.7. *Useful Approximate Formulae*

If input circuit coil losses are small, then, assuming there is no correlation between induced grid noise and shot noise, the optimum single-frequency noise factor is of the form

$$n = 1 + Af$$

where f is the frequency and A is a constant for the particular valve. The source resistance at the grid to give this optimum noise factor is inversely proportional to frequency even though input circuit losses are not negligible, for with a constant Q for the input coil the losses will give an increased ratio for n which is independent of frequency. (On the other hand it is more logical to allow a fixed volume for the input coil.)

The equivalent shot noise resistance for modern high-slope triodes is in the region $3{\cdot}0/g_m$ to $4{\cdot}0/g_m$.

(*To be concluded in December issue*)

NEW BOOKS

Small Transformers and Inductors

By K. A. MACFADYEN, M.Sc.(Lond.), F.Inst.P. Pp. 237 + xii. Chapman & Hall, Ltd., 37 Essex Street, London, W.C.2. Price 37s. 6d.

This book covers most forms of small iron-cored transformers, including power-supply types, and it does so in an unusual way. The treatment, even of power transformers, is throughout mainly on the lines which are normally adopted only for audio-frequency types. Instead of working in terms of magnetizing currents and reactive voltage drops, the treatment is on an impedance basis and the shunt and leakage inductances form the equivalent entities.

This treatment is one which will undoubtedly appeal to the communications engineer, for he is on familiar ground and is not confused by the complicated vector diagrams which appear in so much of the literature written by the power engineer. One matter, however, he will find rather strange. The author makes great use of the concept of complex permeability ($\mu = \mu' - j\mu''$) and takes account of core losses in this way.

The method is unfamiliar and it takes a little time and effort to get used to it, but it does appear to have advantages. It is, of course, necessary to have graphs of μ' and μ'' for various core materials before any practical use can be made of this method. The author provides these graphs for a large number of core materials in common use and he also gives tables of standard laminations in a form well suited to help the designer. He has obviously been to a great deal of trouble, not only to collect the necessary design data but to present it in the most useful way.

After an introductory chapter on the principles of design, there is a chapter on circuit theory, followed by one on the magnetic circuit. The following three chapters cover:—Equivalent circuits for imperfect transformers, effects of transformer imperfections and the measurement, calculation and control of inductances.

The telecommunications engineer is likely to be on familiar ground through most of these six chapters, except perhaps in Chapter 3 where complex permeability is treated. After this general treatment, the rest of the book deals with more specific cases of transformers. There are separate chapters for power, wideband, instrument and pulse transformers, and also for transformers and inductors with magnetic polarization on the one hand and of high Q on the other.

The final chapter proper covers materials, mechanical design and construction. Chapter 14 is an appendix in which a lot of miscellaneous items are covered.

Television scanning transformers receive only a brief mention under Instrument Transformers and what the author does say about them is not very helpful. He says that "The phase errors of such transformers must be kept low if distortion of the picture is to be avoided." This is true only if the transformer must introduce negligible distortion. At frame frequency it is usually impcssible to build such a transformer and the usual practice is to permit the transformer to introduce a large amount of distortion and to obtain the desired output waveform by introducing an inverse pre-distortion of the input.

It would have been helpful if a chapter dealing specifically with such transformers had been included and, in particular, if it had treated the effect of the core on the distortion. This is, however, an omission which will concern only a few of the users of the book and in all other respects it is an extremely useful one.

W. T. C.

Wireless World Diary 1954

Pp. 80 of reference material and pp. 2 per week diary. Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1. Price 4s. 1d. (rexine), 5s. 10d. (morocco leather).

Television and Radio Repairing

By JOHN MARKUS. Pp. 556 + ix. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price 48s.

Dipole Moments

By R. J. W. LE FÈVRE. Pp. 140 + vii. Methuen & Co. Ltd., 36 Essex Street, London, W.C.2. Price 8s. 6d.

Radio Engineering (2nd Edition)

By E. K. SANDEMAN. Vol. 1. Pp. 779 + xxiv. Price 60s. Vol. 2. Pp. 613 + xxi. Price 55s. Chapman & Hall, Ltd., 37 Essex Street, London W.C.2.

Electronique Générale

By A. BLANC-LAPIERRE, R. GOUDET and P. LAPOSTOLLE. Pp. 394. Editions Eyrolles, 61 Boulevard Saint-Germain, Paris Ve. Price 3300 fr.

Synchronization of Reflex-Oscillators

By ALY H. ABDEL DAYEM, Mitteilungen aus dem Institut für Hochfrequenztechnik, Nr. 18. Pp. 110 (In English). Verlag Leeman, Zürich. Price 10.40 fr.

Principles of Electronics

By H. BUCKINGHAM, Ph.D., M.Sc., A.M.I.E.E., and E. M. PRICE, M.Sc.(Tech.). Pp. 335. Cleaver-Hume Press Ltd., 42a South Audley Street, London, W.1. Price 15s.

Schweizerisches Bezugsquellen-Lexikon für Elektro-Industrie und- Handel 1952/56 (2nd Edition)

Pp. 634. In French and German. Fritz Lindner Verlag, Urdorf-Zürich. Price 23.50 fr.

CORRESPONDENCE

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

Harmonic Distortion and Negative Feedback

SIR,—The valuable article on "Distortion and Negative Feedback" in the June 1953 issue of *Wireless Engineer* concludes with an approximation which comes as rather a disappointment. Fortunately, it turns out that this approximation is not necessary.

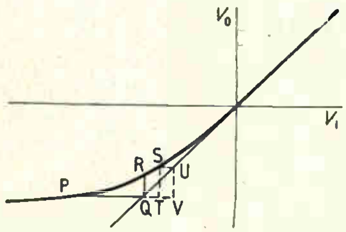
In Fig. 3 (reproduced here) of the article, ST (d_0) is the instantaneous distortion for a given output without feedback. RQ (d_0') is the distortion for the same output with feedback. It is shown that QT = βd_0. The author then says, "the points P, R and S lie approximately on a straight line" and he makes use of the similar triangles PTS and PQR. As we are dealing with distortion it is of the essence of the problem that P, R and S do not lie on a straight line.

Let the points on the graph for no feedback corresponding to R and S be R′ and S′, and call the mean slope between them A_d. Then the mean slope between R and S is

$$\frac{A_d}{1 + \beta A_d}$$

and $$\frac{ST - RQ}{QT} = \frac{d_0 - d_0'}{\beta d_0} = \frac{A_d}{1 + \beta A_d}$$

from which the author's result, $d_0' = \dfrac{d_0'}{1 + \beta A_d}$ follows. In the case of an amplifier with cut-off, for points below cut-off, where distortion occurs, RQ = ST.



As the slope below the point of change of slope tends to zero a finite $(d_0 - d_0')$ tends to demand a greater and greater feedback. An infinitely amplified feedback supplies the theoretical solution in the extreme case, as appears also from the author's approximate treatment.

G. F. Nicholson.

Dept. of Physics and Electrical Engineering,
Royal Naval College,
Greenwich, London.
19th October 1953.

Sir,—As with girls in W. S. Gilbert's Japan, so with some engineers in England; "from 17 to 49 are considered years of indiscretion." Mr. Rowlands is, of course, quite right about the error in my letter on Harmonic Distortion and Negative Feedback, published in the September issue of *Wireless Engineer*. Perhaps in 25 years' time, by W.S.G.'s reckoning, I shall be able to get things right before bursting into print with them. So now, like Ko-Ko, I "beg to offer an unqualified apology" to you, to Mr. Rowlands, and to anyone who may have used my results without having spotted my mistake.

Frank G. Kerr.

Richmond, Surrey.
8th October 1953.

MEETINGS

I.E.E.

9th November. "What are the Requirements of Electrical Engineering Textbooks?" Discussion to be opened by Instr.-Cdr. D. K. McCleery, M.Sc. Commencing at 6 p.m.

11th November. "Some Aspects of the Design of V.H.F. Mobile Radio Systems," by E. P. Fairbairn, M.C., B.Sc.

12th November. "Safety Precautions in Electronic Apparatus with Particular Reference to Medical Applications." Discussion to be opened by H. W. Swann, O.B.E., and H. W. Shipton.

23rd November. "Loudspeaker Systems—Recent Trends in Design," by Major A. E. Falkus, B.Sc.(Eng.).

30th November. Radio Section Soirée, commencing at 6.30 p.m.

2nd December. "Telegraph Codes and Code Convertors," by T. Hayton, B.Sc., C. J. Hughes, B.Sc., and R. L. Saunders, B.Sc.(Eng.). "Code Convertors for the Interconnection of Morse and Teleprinter Systems," by R. O. Carter, M.Sc.(Eng.), and L. K. Wheeler, B.Sc. (Eng.).

3rd December. "Technical Arrangements for the Sound and Television Broadcasts of the Coronation Ceremonies on 2nd June 1953," by W. S. Proctor, M. J. L. Pulling, M.A., and F. Williams.

These meetings will be held at the Institution of Electrical Engineers and will commence at 5.30, except where otherwise stated.

Brit.I.R.E.

11th November. "A High-Definition General-Purpose Radar," by J. W. Jenkins, J. H. Evans, G. A. G. Wallace and D. Chambers, B.Sc. To be held at the London School of Hygiene and Tropical Medicine, Keppel St., Gower St., London, W.C.1, at 6.30 p.m.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for September 1953

Date 1953 Sept.	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	− 0·7	+ 2	+ 4·0
2	− 0·7	+ 3	+ 2·6
3	− 0·6	+ 2	+ 1·8
4	− 0·6	+ 2	+ 1·1
5	N.M.	+ 2	N.M.
6	N.M.	N.M.	N.M.
7	− 0·7	0	− 1·7
8	− 0·5	+ 2	− 4·5
9	− 0·6	+ 2	− 4·7
10	− 0·5	+ 2	N.M.
11	− 0·5	+ 2	− 6·5
12	− 0·5	+ 1	N.M.
13	N.M.	+ 1	N.M.
14	− 0·4	+ 3	− 10·2
15	− 0·4	+ 3	− 12·7
16	N.M.	+ 3	N.M.
17	− 0·5	+ 4	−15·9
18	− 0·4	+ 3	− 17·9
19	− 0·5	+ 4	N.M.
20	− 0·4	+ 4	N.M.
21	− 0·6	+ 5	− 24·1
22	− 0·4	+ 5	− 27·5
23	− 0·3	+ 5	− 28·7
24	N.M.	+ 5	N.M.
25	− 0·3	+ 4	− 32·1
26	− 0·4	+ 4	N.M.
27	− 0·4	+ 5	N.M.
28	− 0·3	− 5	− 36·9
29	N.M.	− 4	− 39·2
30	− 0·4	− 4	− 40·5

The values are based on astronomical data available on 1st October 1953.

N.M. = 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 list of journals abstracted, the abbreviations of their titles and their publishers' addresses.

ACOUSTICS AND AUDIO FREQUENCIES

534.133 : 534.232 **3164**

Graphical Aids in Interpreting the Performance of Crystal Transducers.—W. G. Cady. (*J. acoust. Soc. Amer.*, July 1953, Vol. 25, No. 4, pp. 687–696.) When the mechanical damping of a transducer is large, as by acoustic radiation from one or both faces into a liquid or solid, the circular diagram that represents its characteristics requires special treatment. As a background for this treatment, the uses and limitations of the conventional circle for a resonator with small losses is first reviewed. The problem of the transducer with large losses is then considered with special reference to the equations and graphs for a thickness-type transducer with unsymmetrical loading. For plane-wave transducers the expressions are exact for all loads and at all frequencies, including harmonics. Either the voltage or the current may be constant. From the admittance or impedance diagrams the magnitude and phase of current, voltage, particle velocity, and vibrational amplitude at any frequency can be obtained immediately. Similar results would be found with plates in lengthwise vibration. A new type of diagram is developed for representing vibrational amplitudes. As an illustration, the case of a quartz plate radiating into three liquids of widely different acoustic properties is treated. When the load is unsymmetrical, there is no true node anywhere in the crystal except when the load is zero or infinity. There is, however, a plane of minimal vibration, the amplitude and location of which are derived. The equations indicate certain peculiar effects when the specific acoustic resistance of the medium is just twice that of the crystal.

534.231 **3165**

The Problem of the Momentum of a Sound Wave.—A. Schoch. (*Z. Naturf.*, March/April 1952, Vol. 7a, Nos. 3/4, pp. 273–279.) The relation between the radiation pressure and momentum of a travelling sound wave is investigated; it is found that a wave packet does possess momentum, but that in a stationary wave the time average of the momentum is zero.

534.231 : 532.527 **3166**

The Theory of Steady Rotational Flow Generated by a Sound Field.—P. J. Westervelt. (*J. acoust. Soc. Amer.*, July 1953, Vol. 25, No. 4, p. 799.) Corrections to paper noted in 1552 of June.

534.232 **3167**

Design Techniques for a High-Frequency Transducer with a Wide-Beam Searchlight Pattern.—A. L. Lane. (*J. acoust. Soc. Amer.*, July 1953, Vol. 25, No. 4, pp. 697–702.) Experiments showed that a properly designed $BaTiO_3$ spherical-shell sector will give the required wide-angle radiation with negligible side lobes. Design details are discussed and the effects of various types of baffle are shown graphically.

534.232 : 546.431.824-31 **3168**

Electromechanical Response and Dielectric Loss of Prepolarized Barium Titanate under Maintained Electric Bias: Part 1.—H. G. Baerwald & D. A. Berlincourt. (*J. acoust. Soc. Amer.*, July 1953, Vol. 25, No. 4, pp. 703–710.) For moderate driving fields, operation on the retained polarization, without additional bias voltage, is satisfactory, but at higher driving fields the dielectric losses increase inordinately and lead eventually to depolarization and loss of response. This can be remedied by application of a comparatively modest direct-voltage bias. The losses of various $BaTiO_3$ ceramics for considerable ranges of temperature, applied field and bias voltage are shown graphically. Other effects obtained with bias-voltage operation, such as increase of electromechanical coupling, are also considered.

534.26 **3169**

On the Diffraction of a Plane Sound Wave by a Paraboloid of Revolution: Part 2.—C. W. Horton. (*J. acoust. Soc. Amer.*, July 1953, Vol. 25, No. 4, pp. 632–637.) Numerical values are tabulated of functions which occur in connection with the scalar wave equation in rotational paraboloidal coordinates. Application is made to analysis of the scattering of a plane wave by a rigid convex paraboloid of revolution. The asymptotic expansions of the scattered waves are discussed and their amplitudes are tabulated. The magnitude and phase angle of the total pressure are evaluated for points on the surface of the paraboloid. Part 1: 1047 of 1951 (Horton & Karal)

534.26 **3170**

Diffraction of Acoustic Waves at a Small Circular Aperture.—T. Anders. (*Z. Phys.*, 2nd June 1953, Vol. 135, No. 2, pp. 219–224.) The integral equations derived by Hönl (2182 of 1952) are applied to the case of a plane acoustic wave incident normally on a circular aperture in a plane screen impervious to sound. They are solved as a first approximation with the help of a suitable series representing the wave amplitude.

534.321.9 : 534.61 **3171**

A Thermoelectric Method of Comparing Intensities of Ultrasonic Fields in Liquids.—R. B. J. Palmer. (*J. sci. Instrum.*, June 1953, Vol. 30, No. 6, pp. 177–179.) A sensitive detector causing very little disturbance of the field, and useful at high ultrasonic frequencies, is based on the rise of temperature of certain materials due to the absorption of incident radiation. A comparison is made between this device and the radiation-pressure detector. Experimental procedure and some results are given.

534.414 **3172**

The Effects of Viscous Dissipation in the Spherical Acoustic Resonator.—H. G. Ferris. (*J. acoust. Soc. Amer.*, July 1953, Vol. 25, No. 4, p. 799.) Corrections to paper noted in 1564 of June.

534.612.4 : 621.395.61 **3173**

Absolute Calibration of Microphones at Audible and Ultrasonic Frequencies.—V. Gavreau & A. Calaora. (*Ann. Télécommun.*, May 1953, Vol. 8, No. 5, pp. 150–157.) Two methods are discussed: (*a*) the vibrating-piston method noted in 2690 of 1952 (Gavreau & Calaora), (*b*) a reciprocity method described by Beranek (1857 of 1950), with the formula corrected to take account of the variation of acoustic impedance due to the 'baffle effect' at high frequencies.

534.833.4 : 621.397.7 **3174**

Investigations on Sound Absorbers for Television Studios.—G. Venzke. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, March/April 1953, Vol. 5, Nos. 3/4, pp. 41–46.) Results of the experimental determination of sound absorption characteristics of perforated bricks backed with rock-wool, and of rock-wool boards, are shown graphically. The design and construction of sound absorbers for the Hamburg television studios are described.

534.84 **3175**

Acoustic Design of Auditoria.—P. H. Parkin & W. A. Allen. (*Nature, Lond.*, 18th July 1953, Vol. 172, No. 4368, pp. 98–99.) A survey of design problems and their solutions, with reference to the requirements for both speech and music.

534.844.1/.2 **3176**

Reverberation Times of Some Australian Concert Halls.—A. F. B. Nickson & R. W. Muncey. (*Aust. J. appl. Sci.*, June 1953, Vol. 4, No. 2, pp. 186–188.) Measurements were made of sound-level decay rate of fortissimo chords recorded on magnetic tape from concerts broadcast from several Australian halls and from the Usher Hall, Edinburgh. The reverberation times for the Usher Hall were within 0·1 sec of those obtained by Parkin et al. (3320 of 1952). All the Australian halls have reverberation times long for their respective volumes, the Sydney hall being the worst in this respect, with an average reverberation time of 2·9 sec. This may account for complaints of the poor musical quality of the halls.

534.861.1 **3177**

Electroacoustic Means for the Reproduction of Sound Effects.—F. Enkel. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, March/April 1953, Vol. 5, Nos. 3/4, pp. 47–50.)

621.395.623.7 **3178**

Loudspeaker Developments.—P. W. Klipsch. (*Trans. Inst. Radio Engrs*, May/June 1953, Vol. AU-1, No. 3, pp. 16–22.) The historical development of the corner type of loudspeaker unit for high-fidelity reproduction of sound is described.

621.395.623.7 : 534.373 **3179**

Acoustic Damping for Loudspeakers.—B. B. Bauer. (*Trans. Inst. Radio Engrs*, May/June 1953, Vol. AU-1, No. 3, pp. 23–34.) The transient response of loudspeakers and the cabinets in which they are fitted can be controlled by acoustic damping, which can be determined from the acoustical constants of the system by application of the results of an equivalent-circuit analysis. The performance characteristics of acoustically damped loudspeakers are largely independent of amplifier impedance. Details are given of the method of obtaining the required acoustic damping for a loudspeaker (*a*) in a flat baffle, (*b*) mounted in a cabinet.

621.395.625.3 **3180**

Apparatus for the Continuous Very-Long-Period Recording of Sound.—A. M. Springer. (*Fernmeldetech. Z.*, May 1953, Vol. 6, No. 5, pp. 218–219.) An endless wire on a storage cylinder is continuously moved axially along the cylinder with the aid of a skewed auxiliary cylinder, and traverses a recording head. A wire of length 1·4 km gives a recording time of 3 hours when run at 13 cm/s.

681.85 : 534.851 **3181**

The Lateral Mechanical Impedance of Phonograph Pickups.—J. G. Woodward & J. B. Halter. (*Audio Engng*, June & July 1953, Vol. 37, Nos. 6 & 7, pp. 19–20 . . 54 & 23–24 . . 43.) An outline of the experimental method of determining the dependence of pickup mechanical impedance and response on frequency is given. Curves for several types of gramophone pickup manufactured between 1901 and 1951 illustrate the improvement in performance during this period.

AERIALS AND TRANSMISSION LINES

621.315.2 : 621.396.822 **3182**

Noise Measurements on Telecommunication Cables.—E. Widl. (*Fernmeldetech. Z.*, June 1953, Vol. 6, No. 6, pp. 261–268.) Circuits used in the determination of the noise sensitivity factor of long-distance telecommunication cables are given. The theory of noise investigations in artificially influenced cables has previously been given (*Frequenz*, Jan. 1952, Vol. 6, No. 1, p. 1). Measurements on such cables under various load conditions are described and the results are discussed.

621.315.212 **3183**

Characteristics of Coaxial Cables with Disk Insulators in the Frequency Range above 1 kMc/s.—G. Günther. (*Arch. Elektrotech.*, 1953, Vol. 41, No. 1, pp. 40–45.) The h.f. properties of a coaxial cable, with 13-mm inner conductor and an outer conductor of 40-mm internal diameter, were investigated theoretically and experimentally at wavelengths of 10, 20 and 30 cm. For a disk separation of 40 mm, the resonance wavelength, which is equal to the critical wavelength below which losses due to the production of other wave modes increase very rapidly, is 8·3 cm. At 10-cm wavelength the attenuation due to thermal losses is 0·6% higher, the leakage attenuation 10% higher and the characteristic impedance 7% lower than in an equivalent homogeneous cable.

621.315.212 **3184**

Characteristics of Coaxial Cables with Helical-Strip Insulation in the Frequency Range above 1 kMc/s.—

H. Kaden. (*Arch. Elektrotech.*, 1953, Vol. 41, No. 1, pp. 45–64.) The characteristics of this type of cable, determined theoretically and experimentally, are compared with the characteristics of the disk-insulator type (see 3183 above). The attenuation due to thermal losses at 10-cm wavelength is 11% higher than in a homogeneous cable, the leakage attenuation 78% higher and the characteristic impedance 18% higher, for cables with conductors of the same dimensions.

621.315.212 **3185**

Composite-Dielectric Coaxial Line.—J. A. Kostriza. (*Elect. Commun.*, June 1953, Vol. 30, No. 2, pp. 155–163.) Analysis is given for wave propagation in a coaxial line in which the conductors are separated by two coaxial dielectrics of different permittivity; the possible modes are indicated and equations are derived for the cut-off frequencies. A comparison is made with the single-dielectric line as regards the ratios of the cut-off frequencies of higher-order modes. The effective permittivity of the equivalent single-dielectric line is computed from electrostatic considerations and the calculated value is verified by an experimental determination at 2·2 kMc/s for a line using air and pyralin as the two dielectrics.

621.392.09 **3186**

Experiments with Single-Wire Transmission Lines at 3-cm Wavelength.—D. G. Kiely. (*J. Brit. Instn Radio Engrs*, April 1953, Vol. 13, No. 4, pp. 194–199.) An account of experiments carried out in 1950 on various dielectric-covered wires. Attenuation values obtained by s.w.r. measurements are in agreement with calculated values. A tapered cylinder of soft wood or a carbon-coated card was found suitable for loading the wire for measurement purposes. Polystyrene disks supported the stretched wire. The effect of moisture and rain on the transmission loss is considerable and limits the applications of these lines at 3-cm λ.

621.392.2 **3187**

The 'Exponential' Transmission Line and its Radiation Loss.—G. Piefke. (*Arch. elekt. Übertragung*, May & June 1953, Vol. 7, Nos. 5 & 6, pp. 229–235 & 274–280.) The 'exponential' transmission line, defined by $L \propto e^{-2\beta z}$, $C \propto e^{2\beta z}$ and propagation constant $\gamma_0 = \pm i\alpha + \beta$, cannot be realized in practice. A transmission line having a similar field distribution is considered and its characteristics and field distribution are derived by a new method. Maxwell's e.m. equations are solved for the radiation loss from this line.

621.392.21.028.4 **3188**

A Method of Calculating the High-Frequency Resistance of Cylindrical Conductors of Arbitrary Cross-Section.—F. Lettowsky. (*Arch. Elektrotech.*, 1953, Vol. 41, No. 1, pp. 64–72.) An expression for the h.f. resistance of an infinitely long conductor is derived and applied to the Lecher-wire system and to a conductor of rectangular cross-section.

621.392.26 **3189**

Travelling Waves between Two Parallel Diaphragm-Loaded Reflector Planes.—W. Dällenbach. (*Arch. elekt. Übertragung*, June 1953, Vol. 7, No. 6, pp. 297–304.) The phase velocity of waves travelling between two parallel planes can theoretically be made greater than, equal to, or less than the velocity of light in free space, by using diaphragms normal to the planes and to the direction of propagation of the waves. The theory of wave propagation in such guides with and without diaphragms is given.

621.392.26 **3190**

Thick Obstacles in Waveguides, and Applications.—P. Chavance & P. Salort. (*Ann. Télécommun.*, May 1953, Vol. 8, No. 5, pp. 171–183.) The system is considered in which a rectangular conductive block is fixed by means of screws to one of the broad sides of a rectangular waveguide; such an arrangement can be used for impedance matching in several frequency channels simultaneously. From the point of view of analysis the obstacle resembles a low-Q resonant cavity, but simplifying approximations admissible in dealing with the latter are not valid in the present case. Charts useful for dealing with design problems for two or three channels are presented; the extension to deal with n channels or a wide frequency band is discussed. The measurements were made with standard French waveguide of cross-section 66·37 × 29·50 mm.

621.392.26 **3191**

A Simple Graphical Analysis of a Two-Port Waveguide Junction.—J. E. Storer, L. S. Sheingold & S. Stein. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 1004–1013.) Graphical analysis based on the original work of Deschamps is presented for obtaining the scattering matrix of a two-port waveguide junction from standing-wave measurements. The section may have losses, and can be asymmetrical. In addition, a method is outlined whereby the reflection coefficient of a load terminating the junction can be obtained graphically from the measurement of the reflection coefficient as seen through the junction.

621.392.26 : 621.392.5 **3192**

Wide-Band Phase-Delay Circuit.—H. Sohon. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 1050–1052.) If signals of the same frequency are applied at the inputs of two waveguides, the relative phase at the two outputs depends on the lengths of the waveguides and on the cut-off frequencies. The phase-delay circuit described is based on these considerations. Four relations are developed which uniquely determine the four parameters constituted by the lengths of the waveguides and the cut-off frequencies. A numerical example is worked out.

621.392.26 : 621.396.611.3 **3193**

Slot Coupling of Rectangular and Spherical Wave Guides.—L. B. Felsen & N. Marcuvitz. (*J. appl. Phys.*, June 1953, Vol. 24, No. 6, pp. 755–770.) A dominant-mode waveguide radiating through a slot into a half-space may be regarded from a network viewpoint, the half-space being represented by a number of spherical transmission lines, the waveguide feed by a uniform transmission line and the slot by a coupling network. An approximate evaluation is made of the equivalent-circuit parameters of a slot-coupled junction of a rectangular and a spherical waveguide, when the far field can be satisfactorily represented by the dominant spherical mode. The results are useful in connection with the measurement of e.m. scattering by obstacles located in a half-space illuminated by a slot aerial.

621.396 **3194**

Portable Aluminium Mast.—(*Wireless World*, Sept. 1953, Vol. 59, No. 9, p. 424.) The 200-ft mast, of Al-Si-Mg alloy, is in 8-ft 4-in. sections each weighing 110 lb, and can be erected in 8 hours by a team of six men.

621.396.67 **3195**

A New Solution for the Current and Voltage Distributions on Cylindrical, Elliptical, Conical or Other Axisymmetrical Antennas.—O. Zinke. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 1048–1049.) Abstract only. See 2437 of 1952.

621.396.67 **3196**

Designing Discone Antennas.—J. J. Nail. (*Electronics*, Aug. 1953, Vol. 26, No. 8, pp. 167–169.) Discussion enabling a designer to select the least flare-angle compatible with bandwidth requirements, to determine disk size and disk-to-cone spacing for optimum matching to a 50-Ω line, and to predict radiation-pattern characteristics.

621.396.67 **3197**

Mutual Impedance of Rhombic Antennas Spaced in Tandem.—J. G. Chaney. (*J. appl. Phys.*, June 1953, Vol. 24, No. 6, pp. 751–755.) The mutual impedance formula for separately driven collinear standing-wave aerials may be used directly in the determination of the radiation impedance of these aerials when connected in cascade, but modifications are necessary for the case of travelling-wave aerials under similar circumstances. Formulae are derived for two identical, coaxial and coplanar rhombic aerials in tandem. These formulae are considerably simplified for the case where the aerials are closely spaced and connected in series.

621.396.67 : 621.317.328 **3198**

An Aerial Analogue Computer.—W. Saraga, D. T. Hadley & F. Moss. (*J. Brit. Instn Radio Engrs*, April 1953, Vol. 13, No. 4, pp. 201–224.) Problems of aerial-array design are discussed and a general expression for the field is derived on which the computer design can be based. A description is given of experimental apparatus, demonstrated in 1950 at the Physical Society Exhibition, by which the radiation pattern of a 2- or 3-element array is traced instantaneously on a c.r. tube screen. In array design, a satisfactory approximation to the required pattern is made by a direct method of curve fitting, the position of the elements, the current amplitude and phase being determined directly from the settings of the computer controls. Typical oscillograms are shown and explained. See also 1337 of 1947 (Brown & Morrison) and 282 of 1951 (Todd).

621.396.67.011.21 **3199**

Simplification for Mutual Impedance of Certain Antennas.—J. G. Chaney. (*J. appl. Phys.*, June 1953, Vol. 24, No. 6, pp. 747–750.) The formula for mutual impedance obtained by the generalized circuit method (1854 of 1952) is reduced to a form requiring fewer integrations when calculating the mutual impedance of various combinations of open and terminated wire aerials. The new formula is used to determine the mutual impedance of the legs of an X-type crossed-wire aerial having sinusoidal current distribution. From this result the driving point impedance of a biconical aerial can be deduced.

621.396.671 **3200**

A Note on the Impedance Transformation Properties of the Folded Dipole.—M. Zakhaim. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 1061–1062.) Comment on 34 of 1951 (Guertler).

621.396.671 **3201**

The Theory of a Linear Antenna: Part 1.—Y. Nomura & T. Hatta. (*Technol. Rep. Tohoku Univ.*, 1952, Vol. 17, No. 1, pp. 1–18.) The expressions for the current distribution and the field intensity along an aerial are expanded in Fourier series, and an impedance matrix is introduced to connect the coefficients of the two expansions. The current distribution and the feeding-point impedance are calculated and the results tabulated and shown graphically. A good agreement with Hallén's results (2763 of 1939) was found.

621.396.676 **3202**

The Fields of an Oscillating Magnetic Dipole Immersed in a Semi-Infinite Conducting Medium.—J. R. Wait & L. L. Campbell. (*J. geophys. Res.*, June 1953, Vol. 58, No. 2, pp. 167–178.) Expressions for the fields are derived for the case when the axis of the dipole is parallel to the interface between the conducting medium and the semi-infinite insulating space above it. Various special cases are discussed in detail. An estimate of the field for a frequency of 160 kc/s shows that the attenuation in sea-water is very great if the transmitting dipole is more than a few metres below the surface. See also 39 of January (Wait) and 2109 of July.

621.396.677 **3203**

Lens Aerials at Centimetric Wavelengths.—J. P. A. Martindale. (*J. Brit. Instn Radio Engrs*, May 1953, Vol. 13, No. 5, pp. 243–259.) A survey paper. Compared with systems using reflectors, lens aerials have the advantage of rear feed; scanning can be achieved by movement of the feed only, without any great change of the beam shape or loss of efficiency. Criteria are stated for assessing lens aerials, and a brief description is given of various types in use or under development.

621.396.677 **3204**

The Theory [of the] Convex-Waveguide Lens.—T. Sakurai. (*J. phys. Soc. Japan*, May/June 1953, Vol. 8, No. 3, pp. 372–377.) Theory and design data are given for a device forming part of a reflex e.m. horn radiator. The transformation by the lens of a cylindrical wave into a plane wave is independent of frequency. An outline of the use of the lens in the construction of a very-wide-band microwave radiator is given.

621.396.677 **3205**

Nonresonant Sloping-V Aerial.—J. S. Hall. (*Wireless Engr*, Sept. 1953, Vol. 30, No. 9, pp. 223–226.) Explicit formulae are derived for the components of the distant electric field of the apex-driven sloping-V aerial, assuming uniform current distribution and infinite ground conductivity. Calculated patterns are in agreement with experiment for a typical aerial, but there is considerable vertically polarized radiation off the line of the main beam.

621.396.677 **3206**

The Characteristics of Parabolic Reflectors in Absorbing Media.—A. Esau. (*Fernmeldetech. Z.*, May 1953, Vol. 6, No. 5, pp. 197–201.) Formulae are derived for the gain and the radiation characteristics of an omnidirectional dipole used with a parabolic reflector in an absorbing medium. Analogous formulae are derived for the case of the normal type of dipole. In an absorbing medium the gain decreases, the magnitudes of the subsidiary minima increase and the maxima decrease with increase of $\alpha\rho_0$, where α is the absorption coefficient of the medium and ρ_0 is the radius of the aperture of the mirror. Radiation characteristic curves are given. For experimental work at 14 cm, see 38 of 1938 and 1795 of 1939 (Brüne).

621.396.677 : 621.396.933 **3207**

Stagger-Tuned Loop Antennas for Wide-Band Low-Frequency Reception.—D. K. Cheng & R. A. Galbraith. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 1024–1031.) Design calculations are made for an experimental 100-kc/s system which consists of 12 identical small loop aerials, arranged in two groups at right angles to each other and stagger tuned to different frequencies within the required frequency band. The outputs are applied to a squaring circuit and then added in a parallel-plate summing amplifier before being passed via a grounded-grid amplifier and a cathode-follower stage to the receiver. The system has a 3-db bandwidth of 16·5 kc/s. An electrolyte-tank method of analysing the response of such systems is discussed in an appendix.

621.396.677.012.12 **3208**

A Simple Model for the Representation of the Directional Action of Two Vertical Radiators.—R. Walter. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, March/April 1953, Vol. 5, Nos. 3/4, pp. 37–40.) The intersection of a right circular cylinder with a corrugated surface representing a plane wave is shown to represent the directional pattern of radiation from two vertical radiators for any given phase and amplitude conditions. A graphical representation of a 3-dimensional model is used to obtain numerical results for particular cases.

621.396.677.1 : 523.72 : 621.396.822 **3209**

The Distribution of Radio Brightness over the Solar Disk at a Wavelength of 21 cm: Part 1 — A New Highly Directional Aerial System.—W. N. Christiansen & J. A. Warburton. (*Aust. J. Phys.*, June 1953, Vol. 6, No. 2, pp. 190–202.) Detailed description of the aerial system and method of use. A shorter account was noted in 2573 of September (Christiansen).

CIRCUITS AND CIRCUIT ELEMENTS

621.314.3† **3210**

Note on the Optimum Input-Winding Resistance of a Magnetic Amplifier employing Voltage Feedback.—P. D. Atkinson. (*Elliott J.*, May 1953, Vol. 1, No. 4, pp. 102–103.)

621.314.3† : 621.314.7 **3211**

Transistor-Controlled Magnetic Amplifier.—R. H. Spencer. (*Electronics*, Aug. 1953, Vol. 26, No. 8, pp. 136–140.) A circuit is described in which the collector electrode of a transistor is connected to a winding on a toroidal core, this part of the circuit being completed via the secondary of a transformer (primary voltage 12·5 V at 60 c/s) and the load resistor back to the transistor base electrode. With this arrangement, output currents up to 100 mA peak can be obtained in the load for emitter input-signal currents $< 0{\cdot}5$ mA peak. Complete response to a change of input signal is obtained in one cycle of the applied alternating voltage.

621.314.7 **3212**

Collector-Base Impedance of a Junction Transistor.—R. L. Pritchard. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, p. 1060.) Comment on 874 of March (Early).

621.314.7 : [621.396.645 + 621.318.57 **3213**

Transistor Circuits and Applications.—G. C. Sziklai. (*Electronic Engng*, Sept. 1953, Vol. 25, No. 307, pp. 358–364.) See 2583 of September.

621.316.726.078.3 **3214**

Theory of A.F.C. Synchronization.—W. J. Gruen. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 1043–1048.) The performance of an a.f.c. system can be described in terms of three parameters: (*a*) the gain constant, (*b*) the damping ratio, and (*c*) the resonance or cut-off frequency. Using these parameters, expressions for the performance under conditions of small disturbance to the input phase and for the pull-in performance are derived, two different types of control-network transfer function being considered.

621.316.86 : 537.312.6 **3215**

The Characteristics and Applications of Thermally Sensitive Resistors or Thermistors.—J. W. Howes. (*J. Brit. Instn Radio Engrs*, April 1953, Vol. 13, No. 4, pp. 228–239.) Basic properties of thermistors are reviewed and terms used to specify their characteristics are defined. Outline descriptions are given of their applications in measurement, control and protection circuits, etc.

621.318.435.3 **3216**

A Range of 400-c/s and 1 600-c/s Transductors for Service Use.—A. G. Milnes & C. S. Hudson. (*Electronic Engng*, Aug. 1953, Vol. 25, No. 306, pp. 322–326.) A survey of common transductor types is made and details are given, together with particular applications, of input-type transductors with mumetal or permalloy cores for supply voltages of 13, 25 and 50 V r.m.s., and power-type units with H.C.R. or permalloy-F cores for supply voltages of 115 and 200 V r.m.s. All are of the automatic self-excitation type.

621.319.4 **3217**

Stray Capacitance with High-Permittivity Dielectrics.—W. Heywang. (*Z. angew. Phys.*, May 1953, Vol. 5, No. 5, pp. 161–163.) Expressions are derived for the stray e.s. field and stray capacitance of a parallel-plate capacitor. The corrections for stray capacitance of circular parallel-plate and cylindrical capacitors are determined.

621.387 : 621.316.721 **3218**

Control of Thyratrons by Small Signals.—R. Bailey. (*Electronic Engng*, Sept. 1953, Vol. 25, No. 307, pp. 374–377.) Variation of the phase of the control-grid voltage enables the power supplied by a thyratron to an external circuit to be varied continuously over a wide range. The results obtained with small control voltages indicate that although signals as small as 1–2 V may be permissible when the thyratron forms part of a feedback system, large signals should be used whenever possible.

621.392.26 **3219**

A Circular-Waveguide Magic Tee and its Application to High-Power Microwave Transmission.—B. E. Kingdon. (*J. Brit. Instn Radio Engrs*, May 1953, Vol. 13, No. 5, pp. 275–287.) The magic-T discussed comprises a circular-section waveguide with two mutually perpendicular side arms of rectangular section, spaced longitudinally at a distance λ_g; this system constitutes a pair of H_{01}–H_{11} mode transformers. The device can be used as a variable power-dividing or power-combining bridge, and one of its main uses is for combining feedback power with power from the source at the input to a linear electron accelerator. When two similar sections are joined via a rotatable coupling, the resulting system is suitable for external connection via the rectangular side arms; the power-dividing ratio depends on the angle of rotation between the two sections. Other applications include use with a circular polarizer to act as a phase-shifter or variable impedance.

621.392.4 **3220**

A Contribution to the Theory of Nonlinear Systems.—L. A. Zadeh. (*J. Franklin Inst.*, May 1953, Vol. 255, No. 5, pp. 387–408.) A system of classification of nonlinear 2-terminal networks is introduced and basic properties of various classes are established. The system is such that each class in the sequence N_1, N_2, N_3 . . . contains as a member the class before it. A general nonlinear network of class N_1 is completely defined by its responses to a family of step functions with amplitudes ranging over all real values. An explicit expression is developed for the response of a class-N_1 network to a specified input. Modes of realization and characterization of networks of class N_n are outlined and a procedure for determining the optimum filter of any class is indicated.

621.392.5 + 621.396.615 **3221**

The Equivalent *Q* of *RC* Networks.—P. Tenger; A. P. Bolle. (*Electronic Engng*, Sept. 1953, Vol. 25, No. 307, pp. 394–395.) Comments on 2919 of October (Brown) and author's reply.

621.392.5 3222

Response Characteristics.—L. Storch. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, p. 1061.) Comment on 2374 of 1951 (Kenyon).

621.392.5 3223

A Note on the Analysis of Vacuum-Tube and Transistor Circuits.—L. A. Zadeh. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 989–992.) The setting up of node equations for a network containing one or more active elements is reduced essentially to the determination of the admittance coefficients for the passive network resulting from removal of the active elements, and adding to these the corresponding admittance coefficients for the active elements, the latter being obtained from tables given in the text. Mesh equations are obtained in an analogous way, using impedance-coefficient tables.

621.392.5 3224

Tolerance Coefficients for *RC* Networks.—C. Belove. (*J. appl. Phys.*, June 1953, Vol. 24, No. 6, pp. 745–747.) A method is presented for determining the effect on network design characteristics of the use of nonideal components. A set of tolerance coefficients is derived relating percentage changes of the positions of the poles and zeros of the network function to percentage changes of the network components. Changes of gain or phase are then easily calculated. An exact solution is obtainable only when the network contains at most three independent capacitors. Two theorems are proved which serve to check approximations made for more complex networks.

621.392.5 3225

Realizability Conditions for the Series-Parallel Matrix and Canonical Series-Parallel Circuits for Reactance Quadripoles.—F. M. Pelz. (*Frequenz*, June 1953, Vol. 7, No. 6, pp. 160–166.)

621.392.5 3226

Spinor Theory of Four-Terminal Networks.—W. T. Payne. (*J. Math. Phys.*, April 1953, Vol. 32, No. 1, pp. 19–33.) A spinor can be described as a geometrical object in 3-dimensional space having a magnitude (*S*) and three Eulerian angles. In the application of spinor theory to 4-terminal networks the four spinor components represent the complex current and complex voltage, the associated vector represents power and the direction ratio represents impedance. Applications of the spinor theory are shown, but the study is restricted to steady-state conditions. Negative resistance is not excluded from the considerations.

621.392.5 3227

The Four-Pole Transmission Matrix.—S. R. Deards. (*Electronic Engng*, Aug. 1953, Vol. 25, No. 306, p. 351.) Comment on 1605 of June (Hinton), and author's reply.

621.392.5 3228

The Gyrator as a 3-Terminal Element.—J. Shekel. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 1014–1016.) A 3-terminal gyrator, forming the nucleus of any 3-terminal network that violates the reciprocity relation, is considered. A method is developed which realizes such an element with any unilateral transducer such as a valve or transistor. The effect of loading by parallel or series admittance is investigated.

621.392.5(.6] : 512.831 3229

The Algebraic Theory of Linear Transmission Networks.—M. G. Arsove. (*J. Franklin Inst.*, April & May 1953, Vol. 255, Nos. 4 & 5, pp. 301–318 & 427–444.) The theory is based on the series combination of two networks, with equal numbers of input and output terminals, to form a 'semi-group'. The theory is developed in a series of definitions, theorems and proofs. Fundamental properties of transmission networks are derived and principal types of network are classified. By means of a factorization theorem a simple criterion for symmetry can be derived. The necessary and sufficient conditions for the existence of a characteristic impedance are determined. The theory provides a concise definition and a method of rigorous treatment of the general transmission line.

621.392.5 : 621.314.25 3230

Simplified Solution of Phase-Shift Networks.—R. D. Trigg. (*Electronic Engng*, Aug. 1953, Vol. 25, No. 306, pp. 331–332.) The method is particularly applicable to phase-shift circuits in *RC* oscillators and selective amplifiers. The arbitrary initial assumption is made that all reactive network elements can be treated algebraically as resistances, i.e. as scalar quantities. This enables mesh equations to be written down and solved simply, the results being then interpreted in terms of the complex quantities involved. Three illustrative examples are worked out.

621.392.5.015.3 3231

A Simple Connection between Closed-Loop Transient Response and Open-Loop Frequency Response.—J. C. West & J. Potts. (*Proc. Instn elect. Engrs*, Part II, June 1953, Vol. 100, No. 75, pp. 201–208. Discussion, pp. 209–212.) The phase-margin concept of the characteristics of the Nyquist diagram in the vicinity of the critical point is extended to give a more generalized formula. This relates the damping of the principal oscillatory mode of a closed-loop feedback system to the shape of the Nyquist diagram. All the quantities involved can be obtained from this diagram without further mathematical analysis or graphical construction on the diagram.

621.392.5.029.64 : 538.614 3232

New Linear Passive Nonreciprocal Microwave Circuit Component.—L. Goldstein & M. A. Lampert. (*Elect. Commun.*, June 1953, Vol. 30, No. 2, pp. 164–165.) Reprint. See 1265 of May.

621.392.52 3233

Termination Variation in the Constant-*K* Filter.—S. C. Dunn. (*Wireless Engr*, Sept. 1953, Vol. 30, No. 9, pp. 227–231.) The problem treated is that of finding the modification required in a conventional filter when the terminations are resistive, but otherwise quite general. The filter elements are given normalized values and, in addition, modified by factors x_1 and x_2 to correspond with the change in termination from equal values to those modified by factors α and β. From the expressions for the insertion transfer ratio of the original and of the modified circuit, by equating appropriate terms and solving, a diagram is constructed relating the four factors x_1, x_2, α and β. Two numerical examples illustrate the practical application of the diagram in the design of filter half-sections. Full-section and multi-section filters are treated in similar fashion, the calculations being correspondingly more complex.

621.392.52 3234

Synthesis of Narrow-Band Direct-Coupled [waveguide] **Filters.**—H. J. Riblet. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 1058–1059.) Discussion on 58 of January.

621.392.52 3235

Design of Symmetrical Bridge-Type Electrical Filters by the Operating-Parameter Theory.—F. M. Pelz. (*Arch. elekt. Übertragung*, June 1953, Vol. 7, No. 6, pp. 290–296.) Design formulae are developed and tabulated for symmetrical low-pass filters of degrees 1, 3, 5 and 7, for attenuation characteristics with given zeros and poles. A review of the theoretical foundations is based on work by Cauer (392 of 1942) and by Darlington (1361 of 1940).

621.392.52 **3236**

A Unitary Design System for Band-Pass Filters of the Zobel and Laurent Types.—R. v. Brandt. (*Frequenz*, June 1953, Vol. 7, No. 6, pp. 167–180.) A systematic representation of the properties of band-pass filters, based on the wave-parameter theory, is followed by the development of general design formulae for band-pass half-sections, including the zigzag filters of Laurent. The system is based on Cauer's classification of Q functions, supplemented by some intermediate functions.

621.392.52.029.42 **3237**

A Band-Pass Filter for Low Frequencies.—G. W. Morris & P. G. M. Dawe. (*Electronic Engng*, Sept. 1953, Vol. 25, No. 307, pp. 365–369.) Description, with circuit diagrams, of a filter with a pass band of 8–13 c/s, consisting of four stagger-tuned *RC*-amplifier circuits with inputs and outputs connected in parallel by resistor networks. The filter was developed for α-band encephalography.

621.392.6 **3238**

Synthesis of 2*n*-Poles by Networks Containing the Minimum Number of Elements.—B. D. H. Tellegen. (*J. Math. Phys.*, April 1953, Vol. 32, No. 1, pp. 1–18.) The method of Brune (1932 Abstracts, p. 280) is extended to the synthesis of passive $2n$-poles; the procedure is illustrated for $n = 3$. After splitting off a series resistance, the number of elements in the $2n$-pole can be reduced, and by repeating this procedure the $2n$-pole of zero order can be realized as shown.

621.395.645 : 621.395.44 **3239**

The L3 Coaxial System: Amplifiers.—Morris, Lovell & Dickinson. (See 3411.)

621.396.61.029.62 : 621.396.933 **3240**

High-Frequency Oscillators designed for Regulation and Control of Aircraft V.H.F. Equipment.—R. Olivier. (*Onde élect.*, May 1953, Vol. 33, No. 314, pp. 343–346.) Description of a quartz-controlled 75-Mc/s fixed-frequency unit, and a 108–132-Mc/s v.f.o. with a frequency converter for the range 329–335 Mc/s. A 1-Mc/s quartz crystal provides check points for all harmonics of 1 Mc/s in the range of the v.f.o.

621.396.611.1 **3241**

Action of an Unlimited Train of Telegraphic Signals on a Resonant *RLC* Circuit.—J. Marique. (*HF*, *Brussels*, 1953, Vol. 2, No. 6, pp. 145–156.) The response of a series *RLC* circuit to a train of pulse signals is considered for three types of pulse: rectangular, symmetrical trapezoidal, and of sine-squared form. Analysis shows that in the steady state, whatever the degree of mistuning of the circuit with respect to the signal h.f., the amplitude of the current varies continually with time. The variations are due partly to energy dissipation during the intervals between the signals, and partly to beating of the forced oscillations due to the signals with the natural oscillations of the circuit. The latter effect is particularly noticeable in a very selective circuit, but its magnitude is largely dependent on the degree of mistuning of the circuit. See also 3041 of 1952 and 1941 of July.

621.396.611.1 : 621.3.016.35 **3242**

Amplitude Stability in Oscillating Systems.—N. R. Scott. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 1031–1034.) As a supplement to the Kryloff and Bogoliuboff method of determining amplitude of oscillation in quasilinear systems, a method based upon energy balance is presented. From the energy-balance condition, a criterion for stability of oscillation is deduced for the case of one degree of freedom. The treatment is then generalized to systems of n degrees of freedom.

621.396.611.1 : 621.396.822 : 530.145 **3243**

Quantum Theory of a Damped Electrical Oscillator and Noise.—J. Weber. (*Phys. Rev.*, 1st June 1953, Vol. 90, No. 5, pp. 977–982.) "Field quantization is applied to an electrical oscillating circuit. Damping effects are treated by perturbation theory. Quantum effects occur both in the damping and in the noise, and are discussed in detail . . . The vacuum fluctuations are shown to be observable in certain [low-temperature] noise experiments."

621.396.615 : 621.314.7 **3244**

Junction-Transistor Circuit Applications.—P. G. Sulzer. (*Electronics*, Aug. 1953, Vol. 26, No. 8, pp. 170–173.) The basic circuits described include amplifiers, impedance-changing circuits, phase inverters, oscillators, multivibrators and sawtooth frequency-sweep oscillators.

621.396.615.14 **3245**

Self-Excitation with Disk-Seal Valves in a Grounded-Grid Circuit.—E. Willwacher. (*Fernmeldetech. Z.*, June 1953, Vol. 6, No. 6, pp. 243–249.) A circle diagram is derived for the internal capacitive coupling between anode and cathode in a grounded-grid oscillator. The phase of the transconductance and the input admittance, resulting from the long transit times of electrons, are taken into account. Calculations for several circuits with external feedback are made with the aid of the diagram.

621.396.619.13 : 621.392 **3246**

A.M.-F.M. Analogy.—H. C. Harris. (*Sylvania Technologist*, July 1952, Vol. 5, No. 3, pp. 64–69.) A method is described for analysing the response of a circuit to a f.m. signal, and for determining spectral distribution, based on considering the equivalent signal produced by a series of sequentially pulsed a.m. carriers whose frequencies range between the extremes of the f.m. deviation. The required f.m. response is obtained as the synthesis of the responses to these separate a.m. signals. Both aperiodic and periodic signals are considered. The method is particularly appropriate and gives exact results when the modulating waveform is rectangular; for other cases the results are approximate.

621.396.645 **3247**

The Theory and Design of Cathode-Follower Output Stages.—E. T. Emms. (*Electronic Engng*, Sept. 1953, Vol. 25, No. 307, pp. 386–387.) Limitations imposed on the design of cathode-follower stages feeding loads with an earthy end are noted. Procedures are outlined for choice of valve for a particular service and then completing the circuit design, under the conditions that the specified maximum anode current and anode dissipation are not exceeded and that linear operation is achieved.

621.396.645 : 621.314.7 **3248**

Transistor Operation: Stabilization of Operating Points.—R. F. Shea. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, p. 992.) Correction to paper abstracted in 688 of March.

621.396.645 : 621.396.619.23 **3249**

Why Fight Grid Current in Class B Modulators?—J. L. Hollis. (*Trans. Inst. Radio Engrs*, March/April 1953, Vol. AU-1, No. 2, pp. 26–32.) By using triode valves with low anode resistance and low amplification factor, efficient operation of audio amplifiers can be obtained without allowing the grid voltage to swing positive; grid current is thus avoided. Direct coupling may be used between the driving amplifier and a modulator operated in this way, permitting the use of a large amount of negative feedback. Measured response and distortion are presented for a modulator designed on these lines.

621.396.645.029.3 **3250**

A Single-Ended Push-Pull Audio Amplifier.—A. Peterson & D. B. Sinclair. (*Proc. Instn Radio Engrs, Aust.*, May 1953, Vol. **14**, No. 5, pp. 118–122.) Reprint. See 1250 of 1952.

621.396.645.029.3 **3251**

Analysis of a Single-Ended Push-Pull Audio Amplifier. —Chai Yeh. (*Trans. Inst. Radio Engrs*, March/April 1953, Vol. AU-1, No. 2, pp. 9–19.) See 2613 of September.

621.396.645.35 **3252**

Coupling of Cathode Followers in D.C. Amplifiers.—L. A. Vallet-Cerisier. (*Électronique, Paris*, May 1953, No. 78, pp. 22–27.) A symmetrical phase-reversal circuit is described which, together with screen-coupled cathode-follower stages, has resulted in the development of an amplifier operated from a single 250-V source, giving a gain of 60 db and having a flat response up to 20 kc/s. The level of the fluctuation-noise voltage is almost too low to be measurable.

621.396.645.371.081.75 **3253**

Harmonic Distortion and Negative Feedback.—F. G. Kerr; S. Uminski. (*Wireless Engr*, Sept. 1953, Vol. 30, No. 9, pp. 232–233.) Comments on 2278 of August (Rowlands) and author's reply.

GENERAL PHYSICS

530.145.7 **3254**

Accuracy of Perturbation Calculated from Inaccurate Unperturbed Wave Function.—A. Rahman. (*Physica*, May 1953, Vol. 19, No. 5, pp. 377–384.)

535.37 : 527.52 **3255**

Excitation of Luminescence by Variable Electric Fields. Primary Effect.—G. Destriau. (*J. Phys. Radium*, May 1953, Vol. **14**, No. 5, pp. 307–310.) Reply to Herwelly's criticisms (1644 of June), and description of further experiments showing clearly that luminescence due to glow discharge can be distinguished from luminescence induced by an electric field.

535.37 : 621.32 **3256**

Electroluminescence: a New Source of Light.—G. Destriau. (*Bull. Soc. franç. Élect.*, June 1953, Vol. 3, No. 30, pp. 381–387.) A review of developments since the discovery of the phenomenon by the author in 1936, and an account of recent experimental work. See also 110 of 1949 and **1341 of 1951** (Payne et al.).

537.224 **3257**

Fundamentals in the Behavior of Electrets.—W. F. G. Swann. (*J. Franklin Inst.*, June 1953, Vol. 255, No. 6, pp. 513–530.) Further mathematical development. For previous work see 608 of 1951 and back reference.

537.311.4 **3258**

The Resistance of an Imperfect Contact between Two Metals. Comparison with Experimental Results for Thin Granular Films.—N. Nifontoff & M. Perrot. (*C. R. Acad. Sci., Paris*, 20th July 1953, Vol. 237, No. 3, pp. 228–231.) See also 2628 of September (Nifontoff).

537.311.4 : 537.315 **3259**

A Simple Varying-Capacitor Method for the Measurement of Contact Potential Difference in High Vacuum.—H. P. Myers. (*Proc. phys. Soc.*, 1st June 1953, Vol. 66, No. 402B, pp. 493–499.) Apparatus based on Kelvin's original method and suitable for use at a pressure of 10^{-8} mm Hg is described. The value found for the contact-potential difference between Cu and Ag films evaporated on to W sheets was $0{\cdot}28 \pm 0{\cdot}03$ V, the Ag being positive with respect to the Cu.

537.311.4 : 537.315 : 546.289 **3260**

On the Changes in Contact Potential Difference of a Germanium Rectifier during the Electrical Forming.—T. Niimi. (*J. phys. Soc. Japan*, May/June 1953, Vol. 8, No. 3, pp. 324–330.)

537.52 **3261**

Breakdown of a Gas subject to Crossed Electric Fields.—W. A. Prowse & P. E. Lane. (*Nature, Lond.*, 18th July 1953, Vol. 172, No. 4368, pp. 116–117.) Experiments with various gases show that breakdown in a resonator subjected to 1-μs voltage pulses with a repetition frequency of 10^{10}/sec is not helped by application of alternating fields, of frequency 1, 3 or 10 Mc/s, perpendicular to the main field, even when such fields are only a little below the value at which they would cause breakdown if acting alone.

537.523 **3262**

Formative Time-Lags in the Electrical Breakdown of Gases.—J. Dutton, S. C. Haydon & F. L. Jones. (*Brit. J. appl. Phys.*, June 1953, Vol. 4, No. 6, pp. 170–175.) The time rate of growth of ionization currents in a uniform field greater than that corresponding to the static sparking potential is investigated theoretically. Curves showing the dependence of the formative time lag on overvoltage are given, with an example of their use in the elucidation of secondary ionization processes operative in the breakdown mechanism.

537.525 **3263**

Microwave Technique for Studying Discharges in Gases.—M. A. Lampert & A. D. White. (*Elect. Commun.*, June 1953, Vol. 30, No. 2, pp. 124–128.) Experiments were made with a neon-filled tube inserted through a 'pancake' waveguide (internal height 1 mm), the variation of the discharge d.c. due to the r.f. field being observed on a c.r.o. for various relative positions of tube and guide. The frequency used was about 5 kMc/s and the r.f. power < 100 mW.

537.525 : 621.316.721 : 535.215 **3264**

The Control of Self-Maintained Discharge Currents by Illumination of the Cathode.—W. Kluge. (*Z. angew. Phys.*, May 1953, Vol. 5, No. 5, pp. 173–177.) The discharge characteristics of neon-filled Ni/Cs and Al/K photocells, illuminated with a Hg-discharge lamp, were obtained experimentally. The magnitude of the controllable current in the Townsend discharge space depends on the illumination of the cathode, the driving potential applied and the value of the stabilizing resistance used.

537.527.2 **3265**

Discharge between Positive Point and Plane in Compressed Air.—M. Toitot & A. Boulloud. (*C. R. Acad. Sci., Paris*, 27th July 1953, Vol. 237, No. 4, pp. 322–323.) At gas pressures of several tens of atmospheres, the spark discharge from a point may be preceded by relatively large currents, which are due either to the corona effect or to a dark discharge resulting from the emission from the cathode.

537.531.8 + 535.215 3266

Electron Emission from Metals as an After-Effect of Irradiation.—K. Seeger. (*Z. Phys.*, 2nd June 1953, Vol. 135, No. 2, pp. 152–162.) The emission of electrons from metals was measured after irradiation with X rays, ultraviolet rays and visible rays and after glow-discharge had taken effect. The dependence of the secondary emission on the wavelength of the incident radiation, and on the structure of the metal surface, was investigated. The time and temperature relation is independent of the manner of excitation, which may even be produced by mechanical means. The magnitude of the secondary emission depends on the state of oxidation of the surface. It is further discussed, how far the secondary emission depends on the nature of the metal or the nature of the surrounding gas. The work function of electrons after irradiation is also considered.

537.533 : 539.232 3267

Surface Films and Field Emission of Electrons.—F. L. Jones & C. G. Morgan. (*Proc. roy. Soc. A*, 9th June 1953, Vol. 218, No. 1132, pp. 88–103.) The mechanism of cold emission of electrons from surfaces covered with a tarnish film, under electric fields of the order of 10 kV/cm, was investigated. The results obtained are consistent with the view that the electrons are extracted from the metal substrate by the high electric field set up across the thin surface film when covered with a layer of positive ions.

537.567 3268

Transmission of Radio Waves through Highly Ionized Gases.—P. Verzaux. (*J. Phys. Radium*, May 1953, Vol. 14, No. 5, pp. 310–316.) Preliminary theoretical discussion of the effect of highly ionized media on the transmission of e.m. waves, with special reference to the possible use of centimetre waves in the study of the intense thermal ionization produced in a gas subject to high-pressure shock waves.

538.56 : 535.42 3269

Intensity and Polarization of Electromagnetic Waves diffracted at a Slit: Part 2.—H. Hönl & E. Zimmer. (*Z. Phys.*, 2nd June 1953, Vol. 135, No. 2, pp. 196–218.) Diffraction of electromagnetic waves at a slit is investigated under the supposition that the electric field vector is perpendicular to the slit edges (magnetic case). This supplements the investigation of the same problem for the case where the electric field vector is parallel to the slit edges (electric case) [2183 of 1952 (Groschwitz & Hönl)]. The combination of both cases makes it possible to calculate the intensity and polarization of the wave field for incident polarized and unpolarized radiation. In the case of very narrow slits, to which the Fresnel-Kirchhoff theory is no longer applicable, the electric component normal to the slit edges is determinative for the diffracted field.

538.566 : 535.42 3270

The Diffraction of Electromagnetic Waves at a Conducting Circular Disk and at a Circular Aperture in a Conducting Plane Screen.—W. Andrejewski. (*Z. angew. Phys.*, May 1953, Vol. 5, No. 5, pp. 178–186.) Numerical results of calculations based on the rigorous theory of diffraction of Meixner & Andrejewski (2767 of 1950) are given. These describe the characteristics of diffracted waves in the near and far fields and are valid when the circumference of the disk and the wavelength are of comparable magnitude. The methods of approximation commonly used are critically examined.

541.183.26 : 539.234 : 546.74 3271

Sorption Properties of Thin Nickel Films.—W. Scheuble. (*Z. Phys.*, 2nd June 1953, Vol. 135, No. 2, pp. 125–140.) The sorption of O and H by Ni films is studied. Oxygen sorption, which is much greater than that of hydrogen, takes place in two stages: (*a*) instantaneous covering of the film by a monatomic layer, (*b*) gradual penetration of the oxygen atoms into the film. The quantity of sorbed oxygen is independent of the gas pressure but depends on temperature. Oxygen-covered Ni films have catalytic reactions with hydrogen; on the other hand, hydrogen films have no effect on the sorption of oxygen.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.72 3272

Distribution of Radio Brightness on the Solar Disk at 9·35 kMc/s.—I. Alon, J. Arsac & J. L. Steinberg. (*C. R. Acad. Sci., Paris*, 27th July 1953, Vol. 237, No. 4, pp. 300–302.) Observations during the first six months of 1953, made with interferometer equipment, confirm the results obtained at the annular eclipse of 1st September 1951 [1282 of 1952 (Bosson et al.)] concerning the increased brightness at the edge of the disk. Further observations will be required to determine whether the radio brightness increases at first uniformly from the centre and then more rapidly to a maximum near the edge, or whether there is a secondary maximum at the centre.

523.72 : 621.396.822 3273

Asymmetry in the Decimetre-Wave Radiation from the Sun.—E. J. Blum. (*C. R. Acad. Sci., Paris*, 15th July 1953, Vol. 237, No. 2, pp. 135–137.) Data obtained in Australia at the eclipse of 1st November 1948, and in Khartoum at the eclipse of 25th February 1952, are adduced in support of the conclusion that radiation from the quiet sun at wavelengths around 50 cm is best accounted for by a model sun having an equatorial-axis/polar-axis ratio of 3:2. See also 2786 of 1952 (Blum et al.) and back reference.

523.72 : 621.396.822 : 621.396.621 3274

Radio-Noise Receivers.—Steinberg. (See 3402.)

523.746"1953.01/.03" 3275

Provisional Sunspot-Numbers for January to March, 1953.—M. Waldmeier. (*J. geophys. Res.*, June 1953, Vol. 58, No. 2, p. 266.)

523.75 : 550.385 3276

Solar-Flare Effects and Magnetic Storms.—D. van Sabben. (*J. atmos. terr. Phys.*, June 1953, Vol. 3, No. 5, pp. 270–273.) In the years 1949–1951 there appeared to be no increase in storm probability after the occurrence of a solar-flare effect, except perhaps in the probability of severe magnetic storms.

523.755 : 621.396.822 3277

Thermal Emission from the Solar Corona in the Wavelength Range 10 cm–10 m.—A. Reule. (*Z. Naturf.*, March/April 1952, Vol. 7a, Nos. 3/4, pp. 234–247.) Calculations are made of the sun's radiation for various models of the corona. A method is developed for determining the temperature of the corona from the measured intensity distribution over the solar disk. Radiation from the corona may introduce appreciable irregularities in the intensity distribution. Failure to take account of deviations from radial symmetry may lead to incorrect interpretation of results; Stanier's conclusions (1401 of 1950) may be affected in this way.

523.854 : 621.396.822 3278

An Investigation of the H II Regions by a Radio Method.—P. A. G. Scheuer & M. Ryle. (*Mon. Not. R. astr. Soc.*, 1953, Vol. 113, No. 1, pp. 3–17.) The distribution of brightness near the galactic plane was determined from measurements made with a radio interferometer at 81·5 Mc/s and 210 Mc/s. The results obtained are discussed.

550.38 : 621.317.353.2 3279

The Harmonic Analysis of the Earth's Magnetic Field, for Epoch 1942.—H. Spencer Jones & P. J. Melotte. (*Mon. Not. R. astr. Soc., geophys. Supplement*, June 1953, Vol. 6, No. 7, pp. 409–430.) There is no evidence of a dipole field of external origin greater than 0·1% of the field of internal origin. The intensity of the latter is, at present, decreasing by about 5% per century. The geomagnetic poles have a westerly drift of 4·5° per century; the mean position of the north magnetic pole at present is 73·5°N, 100°W.

550.38"1953.01/.03" 3280

Cheltenham Three-Hour-Range Indices *K* for January to March 1953.—R. R. Bodle. (*J. geophys. Res.*, June 1953, Vol. 58, No. 2, p. 266.)

550.384 : 525.35 3281

On Variations of the Geomagnetic Field, Fluid Motions, and the Rate of the Earth's Rotation.—E. H. Vestine. (*J. geophys. Res.*, June 1953, Vol. 58, No. 2, pp. 127–145.) Full paper. See 1338 of May.

550.384.3 (931) 3282

The Magnetic Secular Variation in New Zealand.—A. L. Cullington. (*N.Z. J. Sci. Tech.*, March 1952, Vol. 33, Sec. B, No. 5, pp. 355–372.) An account is given of the work on secular variation from 1941 to 1950. The observations are tabulated and the results are presented in the form of isoporic charts for each magnetic element.

550.385 : 523.72 : 621.396.822] : 621.396.11 3283

Relationship between Radio-Propagation Disturbance, Geomagnetic Activity and Solar Noise.—D. van Sabben. (*J. atmos. terr. Phys.*, May 1953, Vol. 3, No. 4, pp. 194–199.) Data on ionospheric disturbance of radio communication between New York and Amsterdam during the years 1948–1950 are compared with geomagnetic character figures. A general correlation is established, the maximum radio disturbance showing a mean time lag of 7 hr behind the maximum geomagnetic disturbance. Investigation of the relation between solar r.f. radiation and geomagnetic storms showed that a marked increase of the radiation on at least one of the frequencies 80, 175 and 200 Mc/s occurred at some time during the five days preceding a storm, except in the case of recurrent storms.

550.385"1952.10/1953.03" 3284

Principal Magnetic Storms [Oct. 1952–March 1953].—(*J. geophys. Res.*, June 1953, Vol. 58, No. 2, pp. 267–269.)

550.386"1952.10/.12" 3285

International Data on Magnetic Disturbances, Fourth Quarter, 1952.—J. Bartels & J. Veldkamp. (*J. geophys. Res.*, June 1953, Vol. 58, No. 2, pp. 261–265.)

551.51 3286

Physical Properties of the Atmosphere between ~ 80 km and ~ 250 km.—H. K. Kallmann. (*J. geophys. Res.*, June 1953, Vol. 58, No. 2, pp. 209–217.) The calculated values of temperature, pressure, molecular weight, density and particle concentration are in fair agreement with available experimental results.

551.510.3 : 535.325 3287

The Constants in the Equation for Atmospheric Refractive Index at Radio Frequencies.—E. K. Smith, Jr, & S. Weintraub. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 1035–1037.) See 1990 of July.

551.510.535 3288

Ionospheric Disturbance Forecasting.—L. H. Martin. (*J. Brit. Instn Radio Engrs*, June 1953, Vol. 13, No. 6, pp. 291–301.) Causes and effects of solar-flare disturbances, prolonged periods of low-layer absorption, and ionospheric storms, are reviewed. In an examination of the bases for predicting ionospheric storms, sunspot classification is explained and the correlation of ionospheric storms with solar-flare effects, geomagnetic data and 'precursor' disturbances in high latitudes is discussed. In general the accuracy of predictions 10 days in advance is 65–80%. Disturbance ratings used in the storm warning service are given and methods of prediction adopted during sunspot-maximum and sunspot-minimum periods are outlined.

551.510.535 3289

A Note on the 'Sluggishness' of the Ionosphere.—E. V. Appleton. (*J. atmos. terr. Phys.*, June 1953, Vol. 3, No. 5, pp. 282–284.) By analogy with the *LR* circuit, a 'time of relaxation' $1/2\alpha N$, analogous to a time constant, is obtained from the equation for the variation of ionization, where N is the electron density and α the recombination coefficient. Two methods of determining αN are described and its value is calculated for the D, E, F_1 and F_2 (winter) layers. Although the value of N varies by a factor of over 100, αN is constant within a factor of about 5. This result may be interpreted as indicating either that α must fall steadily from the D to the F_2 layer, or that the physical process of electron disappearance is not merely one of attachment.

551.510.535 3290

Method of Determining the True Height of the Ionospheric Layers, taking account of the Effect of the Geomagnetic Field: Part 1—Application of an Approximate Expression for the Refractive Index (Ordinary-Ray Case).—É. Argence & M. Mayot. (*J. geophys. Res.*, June 1953, Vol. 58, No. 2, pp. 147–165. In French.) Expressions are derived for the characteristic parameters of an ionospheric region, assuming a parabolic law for the ionization. The values of the virtual heights of the F_2 layer, for frequencies in the range 1·45–3·80 Mc/s are calculated and compared with the observed heights and those calculated by Appleton's method. The correction term in the expression for the thickness of the layer shows that the effect of the geomagnetic field is such as to reduce the value of the thickness. Numerical results are in good agreement with the corrections indicated by Shinn & Whale (1405 of 1952).

551.510.535 3291

Dynamic Probe Measurements in the Ionosphere.—G. Hok, N. W. Spencer & W. G. Dow. (*J. geophys. Res.*, June 1953, Vol. 58, No. 2, pp. 235–242.) Measurements made during a V2 rocket flight showed a rapid rise of probe current between 90- and 105-km height, which indicated a positive-ion/electron density ratio of approximately 10:1.

551.510.535 3292

Nature and Origin of Sporadic E Regions as observed at Different Hours over Calcutta.—B. Chatterjee. (*J. atmos. terr. Phys.*, June 1953, Vol. 3, No. 5, pp. 229–238.) Typical results of the experimental determination of the variation of the reflection and transmission coefficients of the E_s region are shown graphically, and a statistical study is made of the variation of echo amplitude. The amplitude distribution of E_s echoes from a thin layer is of the Gauss type, that from an ion cloud of the Rayleigh type, whilst, in general, both the steady and the scattered reflection components are present to give a Rice type of distribution (2168 of 1945). The results of observations made in the early morning, at sunrise, in the afternoon and during thunderstorms are discussed and several ionizing agents are suggested.

551.510.535 **3293**

Ionospheric Storms in the Auroral Zone.—T. Nagata & T. Oguti. (*Rep. Ionosphere Res. Japan*, March 1953, Vol. 7, No. 1, pp. 21–28.) The values of f_0F_2 observed at College, Alaska, are statistically analysed. When the F_2 layer in the auroral zone is sunlit, the value of f_0F_2 usually begins to decrease just after the commencement of a geomagnetic storm. This is attributed to an expansion of the layer due to heating caused by impinging corpuscles. When the F_2 layer is in darkness, an increase of f_0F_2 is observed.

551.510.535 **3294**

Travelling Disturbances in the Ionosphere.—R. E. Price. (*Nature, Lond.*, 18th July 1953, Vol. 172, No. 4368, pp. 115–116.) Pulse measurements at 5·8 Mc/s have been made during the past three years at Perth, Western Australia, using Munro's method (2504 of 1950). Results agree in general with those of Munro. The velocities observed ranged between 5 and 20 km/min. The direction of travel in summer lies between 0° and 90° S of E and in winter between 0° and 60° E of N, with a rapid change-over in the equinoctial months.

551.510.535 **3295**

F-Region Triple Splitting.—G. R. Ellis. (*J. atmos. terr. Phys.*, June 1953, Vol. 3, No. 5, pp. 263–269.) "Measurements of the direction of arrival of *Z* echoes have been made at Hobart, Tasmania. The results indicate that F-region triple splitting is caused by back scattering from a rough layer. The directions observed are consistent with the assumption that reflection at the *Z* level occurs when the angle of incidence is such that the wave normal becomes parallel to the geomagnetic field at the ordinary level of reflection."

551.510.535 **3296**

Continental Maps of Four Ionospheric Disturbances.—R. S. Lawrence. (*J. geophys. Res.*, June 1953, Vol. 58, No. 2, pp. 219–222.) The changes with time of the geographical distribution of f_0F_2 deviations from normal are shown.

551.510.535 **3297**

Sweep-Frequency *h't* Measurement of the Ionosphere.—Y. Nakata, M. Kan & H. Uyeda. (*Rep. Ionosphere Res. Japan*, March 1953, Vol. 7, No. 1, pp. 1–6.) Continuous records of F_2-layer heights were obtained with equipment covering the frequency band 1·8–3·5 Mc/s in 10 sec. The records show marked discontinuities at approximately 1700, 0000 and 0400 local time. The corresponding graphs of F_2-layer heights for several stations in Japan, and of the geomagnetic variations at one station, are shown.

551.510.535 **3298**

A Consideration of the Mechanism of Electron Removal in the F_2 Layer of the Ionosphere.—T. Yonezawa. (*Rep. Ionosphere Res. Japan*, March 1953, Vol. 7, No. 1, pp. 15–20.) The transformation of atomic ions into molecular ions by a transfer of charge and the subsequent recombination of molecular ions with electrons are considered theoretically. Such a mechanism may possibly explain the attachment type of electron removal, but the theory fails to give the value of the attachment coefficient and of its variation with height.

551.510.535 : 537.568 **3299**

The Collision Frequency of Electrons in the Ionosphere.—M. Nicolet. (*J. atmos. terr. Phys.*, May 1953, Vol. 3, No. 4, pp. 200–211.) Theoretical analysis indicates that the electron collision frequency in the ionosphere depends on the concentration of neutral particles in the D and E regions and on the electron concentration in the F region. Any atmospheric model for the region above the E layer should not, therefore, be based on the frequency of electron collisions with neutral particles.

551.510.535 : 550.384/.385 **3300**

Diurnal and Storm-Time Variations of Geomagnetic and Ionospheric Disturbance.—R. P. W. Lewis & D. H. McIntosh. (*J. atmos. terr. Phys.*, May 1953, Vol. 3, No. 4, pp. 186–193.) Analysis of records obtained at Abinger and Slough during a period of 46 months shows the diurnal variations to be very complex, with significant differences dependent on season and on activity level. The 24-hr component is important in both geomagnetic and ionospheric phenomena, but a 12-hr component is found only in ionospheric f_0F_2 disturbance. The storm-time variations of f_0F_2 and of horizontal magnetic force H are statistically closely linked.

551.510.535 : 550.384/.385 **3301**

The Morphology of the Ionospheric Variations associated with Magnetic Disturbance: Part 1 — Variations at Moderately Low Latitudes.—D. F. Martyn. (*Proc. roy. Soc. A*, 9th June 1953, Vol. 218, No. 1132, pp. 1–18.) Graphs of the ionospheric variations with local time and with storm time at Watheroo, Canberra and Washington are given. The local-time variations are mainly diurnal; the storm-time variations are appreciable for about three days after the commencement of the magnetic storm, and the initial shape of the curve depends on the local time of the commencement. A theory of these variations is developed. All ionospheric variations due to magnetic disturbance are attributed to the e.s. field produced by the intense impressed current system in the auroral regions. See also 1673 of June.

551.510.535 : 550.384 **3302**

An Ionospheric-Disturbance Index.—J. M. Bullen. (*N.Z. J. Sci. Tech.*, March 1952, Vol. 33, Sec. B, No. 5, pp. 348–354.) A three-hourly index *I*, with a range 0–9, based on the variation of the critical frequency and the height of the F_2 layers, has been developed. A positive correlation of $\sim 0{\cdot}6$ was found between the Lincoln (N.Z.) *I*-indices and the Amberley (N.Z.) geomagnetic *K*-indices for the period between October 1949 and September 1950. The seasonal variation of the correlation between the *I* and *K* indices was investigated and a comparison with radio disturbance conditions made.

551.510.535 : 550.385 **3303**

On the Variation of the F_2 Layer accompanying Geomagnetic Storms.—K. Sinno. (*Rep. Ionosphere Res. Japan*, March 1953, Vol. 7, No. 1, pp. 7–14.) The universal-time dependent and local-time dependent parts of the variations of f_0F_2 and $h'F_2$ during geomagnetic storms have been calculated from data for several widely separated stations. The local-time variations appear to be due to the S_D current associated with the storm. No indication of a moving disturbance from the auroral zone was found.

551.510.535 : 550.385 **3304**

Storm Phenomena in the Ionosphere.—E. V. Appleton. (*Arch. elekt. Übertragung*, June 1953, Vol. 7, No. 6, pp. 271—273. In English.) Recent advances in our knowledge of F_2-layer ionospheric perturbations accompanying a magnetic storm are summarized. The diurnal control is shown and differences between phenomena at high, medium and low latitudes are noted. For medium latitudes the effect of a storm is to exaggerate the geomagnetic distortion of the F_2 layer already present. See also 2308 of August (Appleton & Piggott).

551.510.535 : 621.396.812 **3305**

Investigation of Ionospheric Absorption.—S. J. Estrabaud. (*Electronic Engng*, Sept. 1953, Vol. 25, No. 307, p. 395.) Comment on 1680 of June (Jenkins & Ratcliff) and authors' reply.

551.594.21 : 621.396.9 **3306**

Radar Echoes associated with Lightning.—V. G. Miles. (*J. atmos. terr. Phys.*, June 1953, Vol. 3, No. 5, pp. 258–262.) Echoes from thunderstorms overhead and from distant storms were observed using 500-kW, 1·9-μs radar pulses operating on a wavelength of 10 cm. The echoes from a storm overhead frequently occur in pairs, corresponding to reflections from cloud layers at different heights.

551.594.5 **3307**

Orientations of Auroral Displays in West-Central Canada.—R. E. Jensen & B. W. Currie. (*J. geophys. Res.*, June 1953, Vol. 58, No. 2, pp. 201–208.) A seasonal variation of orientation was found by statistical analysis of observations made at Saskatoon in 1949–1951.

551.594.5 **3308**

Radio Reflections from Aurora.—B. W. Currie, P. A. Forsyth & F. E. Vawter. (*J. geophys. Res.*, June 1953, Vol. 58, No. 2, pp. 179–200.) Echoes, at 56 and 106 Mc/s, were observed when the auroral form exhibited some ray structure and then only from parts of the aurora at elevations $< 15°$ above the horizon. No echoes were recorded at 3 kMc/s. The 106-Mc/s echoes occur most frequently within the auroral zone, the 56-Mc/s echoes some distance south of it. It is suggested that echoes arise by critical reflection from centres of high electron density, $1{\cdot}4 \times 10^8/\text{cm}^3$ for 106-Mc/s echoes and $4 \times 10^7/\text{cm}^3$ for the 56-Mc/s echoes.

551.594.5 **3309**

Scale-Height Determinations and Auroras.—D. R. Bates & G. Griffing. (*J. atmos. terr. Phys.*, May 1953, Vol. 3, No. 4, pp. 212–216.) The method hitherto used for determining scale heights from auroral luminosity curves is shown to be invalid. The possibility of protons being responsible for the production of aurorae is considered.

551.594.52(99) **3310**

The Southern Auroral Zone as Defined by the Position of Homogeneous Arcs.—F. Jacka. (*Aust. J. Phys.*, June 1953, Vol. 6, No. 2, pp. 219–228.)

LOCATION AND AIDS TO NAVIGATION

526.94 : 621.396.9 **3311**

Nonquantized Frequency-Modulated Altimeter.—H. P. Kalmus, J. C. Cacheris & H. A. Dropkin. (*J. Res. nat. Bur. Stand.*, April 1953, Vol. 50, No. 4, pp. 215–221.) A frequency shifter, connected between the microwave f.m. oscillator feeding the transmitting aerial and the mixer of the receiver, makes continuous altitude measurement possible. The theory of operation is given and three types of the instrument are described. Experimental results, shown graphically, indicate that linear operation down to 10 ft is possible. A suitable frequency shifter for microwave equipment, utilizing a gyrator of the ferrite type, is described in an appendix.

621.396.9 : 551.578.1 **3312**

Radar Observations of Rain at Sydney, New South Wales.—G. A. Day. (*Aust. J. Phys.*, June 1953, Vol. 6, No. 2, pp. 229–239.) Observations made at 9·1-cm wavelength are described and a correlation between the type of rain echo and weather conditions is established.

621.396.932/.933 **3313**

Latest Developments of the Decca Navigation System.—P. Giroud & A. Gayffier. (*Onde élect.*, May 1953, Vol. 33, No. 314, pp. 300–308.) The service area and error curves are shown for the French Decca chain to be brought into service in 1953. The operation of the Mark VII receiver and route tracer for aircraft use are described.

621.396.932/.933 **3314**

Rana Radio-Navigation Equipment.—É. Honoré & É. Torcheux. (*Onde élect.*, May 1953, Vol. 33, No. 314, pp. 319–327.) Considerable advantages are claimed for the system, which was first demonstrated in 1952. In its simplest form an installation comprises a 'free' transmitter radiating on two frequencies F_1 and F_2 near 1 600 kc/s, a 'slave' transmitter on frequencies 40 c/s above F_1 and 40 c/s below F_2, and a control receiver connected by land line to the slave transmitter for frequency control. In a mobile receiver, measurement is made of the phase difference between the two 40-c/s beat notes. With each transmitter radiating on four different frequencies, three signals for phase comparison are derived, providing three degrees of sensitivity in position determination. A description is given of various units of the equipment.

621.396.933 **3315**

Distance Measuring Equipment D.M.E.—F. Penin & G. Phélizon. (*Onde élect.*, May 1953, Vol. 33, No. 314, pp. 309–318.) Illustrated description of direct-reading equipment produced in France. Results of tests show the accuracy of the system to be within about ± 250 m at distances up to 260 km.

621.396.933 **3316**

Regional-Control Radar Equipment at Orly.—P. Bouvier. (*Onde élect.*, May 1953, Vol. 33, No. 314, pp. 328–336.) Description of 10-cm equipment for moving-target indication now being installed. Altitude range is 10 km and extreme horizontal range 150 km.

621.396.933 **3317**

Telecommunications and Radio Aids in Civil Aviation.—(See 3419.)

621.396.933 : 621.396.677 **3318**

Stagger-Tuned Loop Antennas for Wide-Band Low-Frequency Reception.—Cheng & Galbraith. (See 3207.)

621.396.933.4 : 373.62 **3319**

Training Apparatus for Air-Traffic Control by Radar.—(*Engineering, Lond.*, 1st May 1953, Vol. 175, No. 4553, pp. 572–574.) Apparatus is described which produces artificially, by means of an electromechanical system, aircraft traces on a p.p.i. display screen. Block diagrams are given. The heading, rate of turn, and airspeed of four aircraft may be regulated from a remote-control box and the effect of wind simulated. The pulse width, aerial beam width and speed of rotation can also be adjusted.

MATERIALS AND SUBSIDIARY TECHNIQUES

533.5 **3320**

Methods of Obtaining High Vacuum by Ionization. Construction of an 'Electronic Pump'.—H. Schwarz. (*Rev. sci. Instrum.*, May 1953, Vol. 24, No. 5, pp. 371–374.) English version of 2011 of July.

535.37 3321

Field Emission of Crystal Phosphors.—M. Ueta. (*J. phys. Soc. Japan*, May/June 1953, Vol. 8, No. 3, pp. 429–431.) The relation between emission intensity and applied potential for Zn_2SiO_3-Mn and ZnS-Cu phosphors is shown graphically.

535.37 : 546.471.61 : 537.533.9 3322

The Decrease of Luminescence resulting from Irradiation by Electrons.—K. H. J. Rottgardt. (*Naturwissenschaften*, June 1953, Vol. 40, No. 11, pp. 315–316.) The observed change in luminescence of ZnF_2 screens with the number of the impinging electrons is in close agreement with a relation given by Broser & Warminsky (1371 of May). The electron traps produced cause the transition of electrons, without radiation, from the conduction band into the filled band.

537.224 3323

Plastic Electrets and their Applications.—H. H. Wieder & S. Kaufman. (*Elect. Engng, N.Y.*, June 1953, Vol. 72, No. 6, pp. 511–514.) See 2025 of July.

537.311.33 3324

Space-Charge-Limited Emission in Semiconductors.—W. Shockley & R. C. Prim. (*Phys. Rev.*, 1st June 1953, Vol. 90, No. 5, pp. 753–758.) For an *n-i-n* structure comprising a plane parallel layer of intrinsic semiconductor between two layers of *n*-type material, an expression for current density is derived which is analogous to Child's law for thermionic emission. In obtaining this expression, both diffusion and dependence of mobility on field have been neglected. An exact solution can be given for both *n-i-n* and *p-i-p* structures, when drift velocity is proportional to field.

537.311.33 3325

Preparation and Properties of Arsenide Semiconductors.—F. Gans, J. Lagrenaudie & P. Seguin. (*C. R. Acad. Sci., Paris*, 27th July 1953, Vol. 237, No. 4, pp. 310–313.) Arsenides of Ga and In were prepared by heating the constituent materials to controlled temperatures in an evacuated fused-quartz tube. The arsenide of Ga has good rectifying properties. One *n*-type sample would rectify up to 20 or 25 V. Another sample showed *p*-type rectification. The material is also photoconductive, with a sharp cut-off near 1·1 μ.

537.311.33 : [546.28 + 546.289 3326

Mobility of Holes and Electrons in High Electric Fields.—E. J. Ryder. (*Phys. Rev.*, 1st June 1953, Vol. 90, No. 5, pp. 766–769.) Pulse measurements of conductivity were made for *n*- and *p*-type Ge filaments at temperatures of 77°K, 193°K, and 298°K, and for *n*- and *p*-type Si filaments at 298°K. The specimens consisted of short slender filaments with two relatively massive ends. At 298°K, the observed critical field, above which mobility varies as $E^{-\frac{1}{2}}$, E being the electric field, is 0·9 kV/cm for *n*-type Ge, 1·4 kV/cm for *p*-type Ge, 2·5 kV/cm for *n*-type Si, and 7·5 kV/cm for *p*-type Si. Results of conductivity measurements on an *n*-type Ge specimen at 20°K are also given.

537.311.33 : 546.28 3327

Energy-Band Structure in Silicon Crystal.—E. Yamaka & T. Sugita. (*Phys. Rev.*, 1st June 1953, Vol. 90, No. 5, p. 992.) Calculations for the lowest three energy levels give results in agreement with experimental values. See also 421 of February (Holmes).

537.311.33 : 546.289 3328

High-Field Mobility in Germanium with Impurity Scattering Dominant.—E. M. Conwell. (*Phys. Rev.*, 1st June 1953, Vol. 90, No. 5, pp. 769–772.) Discussion of Ryder's results (3326 above) for an *n*-type Ge specimen at 20°K. A combination of impurity scattering and lattice scattering accounts semiquantitatively for the observed variations in mobility, provided the rate of energy loss in collisions is greater by a factor of about 9 than that given by theory based on the assumption of spherical surfaces of constant energy.

537.311.33 : 546.289 3329

Space-Charge Limited Hole Current in Germanium.—G. C. Dacey. (*Phys. Rev.*, 1st June 1953, Vol. 90, No. 5, pp. 759–763.) The theory of Shockley & Prim (3324 above) is extended to the case of high electric fields, and the corresponding expression for current density is deduced. Experiments on Ge at 77°K give results in good agreement with theory, when recently determined values of mobility, critical field and 'punch-through' voltage are used.

537.311.33 : 546.289 3330

Forming of Germanium Surfaces.—R. Thedieck. (*Z. angew. Phys.*, May 1953, Vol. 5, No. 5, pp. 163–165.) The barrier potential distribution in the vicinity of a forming crater was determined experimentally. A continuous *p*-type surface was obtained from *n*-type Ge by repeated point forming so that the formed areas overlapped. The barrier potential of the formed surface was independent of the shape and material of the metal point contact used in the forming process.

537.311.33 : 546.289 : 621.314.7 3331

Mechanism of Point-Contact Transistors.—R. Thedieck. (*Z. angew. Phys.*, May 1953, Vol. 5, No. 5, pp. 165–166.) The thickness of the *p*-type layer and that of the barrier layer obtained by the point-contact forming of *n*-type Ge (3330 above) were estimated as about 14 μ and < 2 μ respectively. A 2-point-contact transistor with characteristics similar to those of a *p-n-p*-junction type was produced.

537.311.33 : 621.314.7 3332

***p-n* Junction revealed by Electrolytic Etching.**—E. Billig & J. J. Dowd. (*Nature, Lond.*, 18th July 1953, Vol. 172, No. 4368, p. 115.) The specimen is immersed in a suitable electrolyte, which has an inert electrode as cathode and the *n*-type region as anode. A current of about 1 mA/mm² maintained for 1–2 minutes etches the *n*-type region sufficiently to reveal the potential barrier. The method is also applicable to *n-p-n* specimens. This type of etching is much less effective than a strong chemical etch in improving the rectifying properties of the junction.

537.581 : 537.311.32 : 546.28 3333

Electrical Resistivity and Thermionic Emission of Silicon.—L. Esaki. (*J. phys. Soc. Japan*, May/June 1953, Vol. 8, No. 3, pp. 347–349.) Experimental investigation in the range 1 100–1 350°C.

538.221 3334

Developments in Sintered Magnetic Materials.—J. L. Salpeter. (*Proc. Instn Radio Engrs, Aust.*, May 1953, Vol. 14, No. 5, pp. 105–118.) A review of ferromagnetic theory underlying the development of materials such as ferroxcube and ferroxdure.

538.221 3335

A Neutron Diffraction Study of Magnesium Ferrite.—L. M. Corliss, J. M. Hastings & F. G. Brockman. (*Phys. Rev.*, 15th June 1953, Vol. 90, No. 6, pp. 1013–1018.) Application of a strong magnetic field enables the coherent scattering to be separated into its nuclear and magnetic parts. The magnetic scattering is in good agreement with the Néel model of ferrimagnetism. Analysis of the nuclear scattering gave a value of 0·88 ± 0·01 for the degree of inversion and 0·381 ± 0·001 for the *u* space group parameter.

538.24 : 538.662 3336
The Temperature Dependence of the Spontaneous Magnetization in an Antiferromagnetic Single Crystal.—N. J. Poulis & G. E. G. Hardeman. (*Physica*, May 1953, Vol. 19, No. 5, pp. 391–396.)

538.662 3337
Condition for Vanishing Spontaneous Magnetization below the Curie Temperature.—K. F. Niessen. (*Physica*, May 1953, Vol. 19, No. 5, pp. 445–450.) Theory is given of a method of determining the temperature T_0 at which the spontaneous magnetization of a particular spinel structure vanishes. A condition is derived for the extreme case where T_0 coincides with the Curie temperature; this condition can be applied to experiments for finding the ratio of mutual interactions of magnetic ions.

538.662 3338
On the Temperature Sensitivity of Special Magnetic Materials.—T. A. Heddle. (*Brit. J. appl. Phys.*, June 1953, Vol. 4, No. 6, pp. 161–166.) A survey of thermomagnetic materials sensitive in the range − 60°C to + 120°C. Applications of these materials in temperature-sensitive and compensating devices are considered.

539.23 : 537.311.31 3339
Complement to the Paper on the Variation of the Electrical Resistance of Very Thin Metal Films as a Function of Applied Potential.—J. Romand, R. Aumont & B. Vodar. (*C. R. Acad. Sci., Paris*, 6th July 1953, Vol. 237, No. 1, pp. 33–35.) Complementary to 2535 of 1950 (Vodar & Mostovetch). Further experiments on evaporated Pt films, using Aumont & Romand's sensitive null detector (2726 of September), confirm the previous results, but suggest that theory will probably have to be developed for films of thickness such that the space occupied by aggregates is not negligible.

539.231 : 537.311.31/.32 3340
Conducting Films on Glass.—(*Elect. Rev., Lond.*, 8th May 1953, Vol. 152, No. 19, pp. 1069–1072.) The development of the cathodic sputtering process at the National Physical Laboratory for the preparation of oxide and low-resistance Au films on glass is reviewed. The thickness of the Au films is about 25×10^{-8} in. and test samples have been operated with a dissipation of 3 kW/ft² without failure. Possible applications of such films to prevent electrification of the glass in sensitive electrical instruments, as fixed resistors in radar circuits, and in fluorescent lamps and Hg-vapour rectifiers are noted.

546.28 + 546.289 + 546.832 3341
The Isotopic Constitution of Silicon, Germanium, and Hafnium.—J. H. Reynolds. (*Phys. Rev.*, 15th June 1953, Vol. 90, No. 6, pp. 1047–1049.)

546.28 3342
Electron-Spin Resonance in a Silicon Semiconductor.—A. M. Portis, A. F. Kip, C. Kittel & W. H. Brattain. (*Phys. Rev.*, 1st June 1953, Vol. 90, No. 5, pp. 988–989.)

546.289 3343
Solute Distribution in Germanium Crystals.—W. P. Slichter & E. D. Kolb. (*Phys. Rev.*, 1st June 1953, Vol. 90, No. 5, pp. 987–988.)

546.289 : [535.32 + 535.34 3344
The Optical Constants of a Single Crystal of Germanium.—D. G. Avery & P. L. Clegg. (*Proc. phys. Soc.*, 1st June 1953, Vol. 66, No. 402B, pp. 512–513.)

621.314.632.1 : 537.312.6 3345
The Temperature Dependence of the Zero-Bias Resistance of Cuprous-Oxide Rectifiers.—A. Okazaki, H. Tubota & H. Suzuki. (*J. phys. Soc. Japan*, May/June 1953, Vol. 8, No. 3, pp. 431–432.) The mobility of current carriers in Cu_2O, in the range 250°–400°K, was determined from measurements of the zero-bias resistance. Landsberg's formula (3472 of 1952) was used. Comparison is made with results obtained by other workers from Hall-effect measurements.

MATHEMATICS

511 : 53 : 621.39 3346
Radio Technology and the Theory of Numbers.—B. van der Pol. (*J. Franklin Inst.*, June 1953, Vol. 255, No. 6, pp. 475–495.) The application of results of the theory of numbers to problems in physics and radio technology is illustrated by particular examples.

516.6 : 517.7 3347
Some Coordinate Systems Associated with Elliptic Functions.—P. Moon & D. E. Spencer. (*J. Franklin Inst.*, June 1953, Vol. 255, No. 6, pp. 531–543.) Results are given of a study of cylindrical and rotational coordinate systems based on elliptic-function transformations. The six transformations listed are the ones most likely to be of practical utility.

519.272.119 3348
On a Class of Stochastic Operators.—L. A. Zadeh. (*J. Math. Phys.*, April 1953, Vol. 32, No. 1, pp. 48–53.) A relation between autocorrelation functions, and a product relation for autocorrelation functions, are derived for given linear stochastic operators.

681.142 3349
An Automatic Analogue Computor for the Solution of Mine Ventilation Networks.—D. R. Scott & R. F. Hudson. (*J. sci. Instrum.*, June 1953, Vol. 30, No. 6, pp. 185–188.)

681.142 : 519.272.119 3350
A Thermistor-Bridge Correlator.—V. C. Anderson & P. Rudnick. (*Rev. sci. Instrum.*, May 1953, Vol. 24, No. 5, pp. 360–361.) The correlator described is designed to give a direct indication of the correlation coefficient between two c.w. signals. Two thermistors are used as mean-square elements in a circuit giving correlation coefficients to within about 1%.

681.142 : 538.221 3351
Digital Storage using Ferromagnetic Materials.—A. E. De Barr. (*Elliott J.*, May 1953, Vol. 1, No. 4, pp. 116–120.) Four types of magnetic digit-storage system are described.

681.142 : 538.221 3352
An Analysis of Magnetic-Shift Register Operation.—E. A. Sands. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 993–999.)

681.142 : 621.385.832 3353
A Method for Improving the Read-Around Ratio in Cathode-Ray Storage Tubes.—J. Kates. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 1017–1023.)

MEASUREMENTS AND TEST GEAR

529.786 3354
Quartz Clocks of the Greenwich Time Service.—H. M. Smith. (*Mon. Not. R. astr. Soc.*, 1953, Vol. 113, No. 1, pp. 67–80.) A general account of the development and of

the assessment of performance of the quartz clocks at Greenwich and Abinger is given. The criterion of performance, defined as the change in rate expressed in milliseconds per day per month per month, is of the order of 0·1 in modern ring-crystal-oscillator clocks, compared with a criterion value of 3·4 for the best pendulum clock of the Paris observatory. See also 2223 of 1951.

538.566 : 535.222 **3355**

Proposed Use of a Cylindrical Surface-Wave Resonator for the Determination of the Velocity of Short Electromagnetic Waves.—H. M. Barlow & A. E. Karbowiak. (*Brit. J. appl. Phys.*, June 1953, Vol. 4, No. 6, pp. 186–187.) The use of the cylindrical surface-wave resonator at a frequency between 1 and 40 kMc/s for the determination of *c* is suggested. The method of calculating the wave velocity is outlined, but no estimate of the ultimate accuracy of the method is made.

621.3.087.4 : 551.510.535 **3356**

Automatic Ionospheric-Height Recorder.—C. Clarke & E. D. R. Shearman. (*Wireless Engr*, Sept. 1953, Vol. 30, No. 9, pp. 211–222.) Description of "commercial equipment based on a design by Naismith & Bailey [1191 of 1951], for measuring the virtual heights of reflection of ionospheric echoes as a function of transmitted frequency. A pulse transmitter consisting of a master oscillator and power amplifier is tuned through the frequency range in five bands. The receiver is separately tuned and is kept in step with the transmitter by a frequency discriminator and servo-mechanism. The receiver output is presented on two cathode-ray tubes, one for monitoring and one for photographic recording, each displaying a linear time-base sweep. Height and frequency calibrations, which are derived from a crystal source, are displayed, and a crystal-controlled time switch is incorporated for the automatic operation of the equipment."

621.314.25 **3357**

A Phase Shifter for Use from 10–100 Mc/s.—W. P. Melling. (*Elliott J.*, May 1953, Vol. 1, No. 4, p. 115.) Description of a goniometer type of phase shifter using crossed conductors in a metal cylinder closed by a rotatable cap carrying a pickup loop. See also 1735 of June (Thirup).

621.314.7 : 621.317.3 **3358**

Measurement of the Small-Signal Parameters of Transistors.—G. Knight, Jr, R. A. Johnson & R. B. Holt. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 983–989.) With input current and output voltage as independent variables, the set of parameters most appropriate for the description of circuit operation of junction transistors comprises (*a*) the short-circuit input conductance, (*b*) a specified voltage feedback ratio, (*c*) the short-circuit current gain, and (*d*) the open-circuit output resistance. Grounded-base connection only is considered. The measurement method applied in obtaining the dynamic impedance parameters [2863 of 1949 (Lehovec)] is also applied here. Circuit diagrams and a table giving measurement details for all parameters are presented. Sources of error and their elimination are discussed.

621.314.7 : 621.317.733 **3359**

Equipments for Measuring Junction-Transistor Admittance Parameters for a Wide Range of Frequencies.—L. J. Giacoletto. (*RCA Rev.*, June 1953, Vol. 14, No. 2, pp. 269–296.) Description of the use of a commercially available bridge for the determination of conductance parameters associated with a junction transistor, at 1 kc/s, and also of the construction of four admittance bridges for the determination of transistor conductance and suceptance parameters as a function of frequency from 1 kc/s to 1 Mc/s.

621.316.842/.843].025(083.74) **3360**

Alternating-Current Resistance Standards.—A. H. M. Arnold. (*Proc. Instn elect. Engrs*, Part II, June 1953, Vol. 100, No. 75, pp. 319–328.) Basic design principles for resistance standards having a resistance within 0·01% of the nominal value and a phase angle $< 10^{-4}$ radian at frequencies up to about 20 kc/s are described. Ni-Cr-Al alloys appear suitable for standards in which self-heating is considerable. Details are given of a 1-Ω standard comprising 21 bifilar units of Cu-Ni wire in parallel; the calculated phase angle is $< 10^{-4}$ radian up to 30 kc/s. Formulae for eddy-current losses, inductance and capacitance of standards of various types are given.

621.317.333.4.015.7 : 621.315.2 **3361**

A Portable Pulse Test-Set for the Measurement of Impedance Irregularities in Coaxial Cables used for the Transmission of Television Signals.—F. A. Vitha. (*Commun. News*, June 1953, Vol. 13, No. 4, pp. 117–127.) The theory of reflection at cable irregularities is reviewed. The equipment described comprises a transmitter unit and c.r.o. Pulse duration can be either 0·05 or 0·25 μs. Pulse rate is controlled by an oscillator at one of nine frequencies between 10 and 300 kc/s or by triggering from an outside timing source. Cables of length from 50 m to 15 km can be tested. Accuracy is within ± 3 m or ± 5 m according to total cable length.

621.317.7.029.6 **3362**

Instruments for use in the Microwave Band.—A. F. Harvey. (*Proc. Instn elect. Engrs*, Part II, June 1953, Vol. 100, No. 75, p. 244.) Discussion on paper abstracted in 1969 of 1952 (Harvey).

621.317.715 **3363**

A Logarithmic-Scale Valve Galvanometer.—G. Heiland & G. Rupprecht. (*Z. angew. Phys.*, May 1953, Vol. 5, No. 5, pp. 167–171.) Theory, design and applications of instruments with a working range of 10^{-12}–10^{-6} A.

621.317.723 **3364**

A Logarithmic-Scale Electrometer.—W. Waidelich. (*Z. angew. Phys.*, May 1953, Vol. 5, No. 5, pp. 171–173.) The design of an e.s. quadrant-type voltmeter for the range 10 V–1 kV is described.

621.317.733 **3365**

High-Resistance Bridge for Conductivity Measurements.—M. Unz. (*J. sci. Instrum.*, June 1953, Vol. 30, No. 6, pp. 179–184.) The bridge described is suitable for measuring the conductivity of the ground or other specimens with unreliable contact surfaces.

621.317.733 : [621.316.86 : 537.312.6 **3366**

Direct-Reading Thermistor Bridge.—K. F. Treen. (*Electronic Engng*, Aug. 1953, Vol. 25, No. 306, pp. 350–351.) Comment on 1749 of June (Pearson & Benson), and authors' reply.

621.317.733.011.21 **3367**

Wheatstone Bridge for Admittance Determinations.—H. P. Schwan & K. Sittel. (*Elect. Engng, N.Y.*, June 1953, Vol. 72, No. 6, p. 483.) Digest only. Description of a bridge for measurement (to within 0·1%) of resistances in the range 10 Ω–100 kΩ and capacitances from zero to 1 000 pF. The reactance calibration is independent of frequency for all resistors > 100 Ω.

621.317.74 **3368**

Design and Construction of an Accurate Standing-Wave-Ratio Meter for the 9·5-kMc/s Band.—J. Le Bot & S. Le Montagner. (*J. Phys. Radium*, May 1953, Vol. 14, No. 5, pp. 299–303.) A general discussion is given of the effects on wave propagation of the usual type of slotted guide and probe arrangement. Accuracy of machining and finishing the equipment described was such that the positions of the voltage minima were defined to within ± 5 μ.

621.317.755 **3369**

Automatic C.R.T. Trace Brightening for Varying-Amplitude R.F. Signals.—J. de Klerk. (*Electronic Engng*, Sept. 1953, Vol. 25, No. 307, pp. 388–389.) In photography of short trains of pulses of varying amplitudes, if the exposure is correct for the large-amplitude pulses, the small-amplitude pulses will be over-exposed. This difficulty is overcome by means of a circuit which automatically increases the brightness of the trace by an amount proportional to the signal amplitude. Typical oscillograms illustrate the use of the circuit.

621.317.755 **3370**

The Oscilloscope, Type GM 5660.—T. M. W. van Velthoven. (*Commun. News*, June 1953, Vol. 13, No. 4, pp. 139–146.) Description, with circuit diagrams of the various units, of an instrument with a 10-cm screen and a special synchronization system for examination of transients of duration down to 0·1 μs.

621.317.755 **3371**

Some Special Oscillograph Techniques.—F. M. Bruce. (*J. Brit. Instn Radio Engrs*, June 1953, Vol. 13, No. 6, pp. 303–314.) Techniques and apparatus in use for investigation of transients in heavy-current equipment are described, including (*a*) a continuously evacuated oscillograph with the beam impinging directly on photographic film, (*b*) high-speed oscillographs for microsecond transients, (*c*) equipment for multiple recording with a high-speed drum camera, (*d*) a recurrent-surge c.r.o. for pulse tests on transformers, (*e*) a recovery-voltage indicator for testing circuit-breakers.

621.317.755 : 512.99 **3372**

Study of a Vectorial Analyser.—J. van Geen. (*HF, Brussels*, 1953, Vol. 2, No. 6, pp. 157–161.) Description of equipment for vectorial c.r.o. display of sinusoidal voltages. A circular-sweep timebase is used, the frequency being 1 kc/s. Applications to e.m. wave propagation along an artificial transmission line and to sound-wave propagation in an echo-free room are illustrated.

621.317.755 : 621.397.6 **3373**

Testing of Television Studio Equipment by means of Synchronizing Pulses of Variable Phase.—E. Demus. (*Fernmeldetech. Z.*, May 1953, Vol. 6, No. 5, pp. 208–213.) A 'line selector' oscilloscope and its applications to 625-line equipment are described. Block diagrams and oscillograms illustrating various faults are given. Similar equipment was described by Fisher (1762 of 1952).

621.317.756 + 621.317.77 **3374**

A Harmonic-Response-Testing Equipment for Linear Systems.—D. O. Burns & C. W. Cooper. (*Proc. Instn elect. Engrs*, Part II, June 1953, Vol. 100, No. 75, pp. 213–221. Discussion, pp. 221–222.) Detailed description of equipment for measurements in the frequency range 0·1–100 c/s. Phase shift and attenuation are indicated on dials. The phase meter is an air-cored dynamometer with oil damping. With purely sinusoidal excitation of the fixed coil, the moving coil responds only to the fundamental component of the applied signal. Sinusoidal signals at the test frequency are obtained by demodulating a carrier-frequency signal which has been modulated electromechanically. The equipment includes a special amplifier for low-impedance coupling and a calibrated amplifier for attenuation measurements.

621.317.761.029.422 **3375**

Automatic Frequency Meter for Very Low Frequencies.—F. Hubl. (*NachrTech.*, May 1953, Vol. 3, No. 5, pp. 222–225.) A description is given of the principles and design of a direct-reading meter suitable for measurement of the frequency difference between two standard-frequency sources. A system of sensitive relays is used to control the operating time of a motor coupled to a d.c. potentiometer, the current through which is a measure of the difference frequency.

621.317.794 : 621.362 : 537.311.33 **3376**

The Construction of Radiation Thermocouples using Semiconducting Thermoelectric Materials.—D. A. H. Brown, R. P. Chasmar & P. B. Fellgett. (*J. sci. Instrum.*, June 1953, Vol. 30, No. 6, pp. 195–199.) An investigation is reported of methods of construction of thermocouples of the type described by E. Schwarz (British Patent Specifications Nos. 578187 and 578188). The composition of the thermoelectric materials is discussed. Performance figures for experimental thermocouples are given together with those obtained on commercial couples made by Schwarz.

621.396.615.17.018.75 : 621.397.62 **3377**

An Introduction to the Sine-Squared Pulse.—G. J. Hunt & E. W. Elliott. (*J. Televis. Soc.*, April/June 1953, Vol. 7, No. 2, pp. 49–59. Discussion, p. 59.) The sine-squared, or 'raised cosine', pulse is one in which each ordinate is the square of the corresponding ordinate of a half sine-wave. A description is given of a complete generator using in the squaring stage a r.f. pentode with signal applied simultaneously to control and suppressor grids; pulses of 0·34, 0·17, 0·1 and 0·05 μs are available. For testing the transient response of television or other low-pass systems, this type of pulse offers many advantages.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

620.179.1 : 677.72 **3378**

Electromagnetic Testing of Winding Ropes.—A. Semmelink. (*Trans. S. Afr. Inst. elect. Engrs*, May 1953, Vol. 44, Part 5, pp. 113–129. Discussion, pp. 130–145.) Details are given of an a.c. test method in which a wire rope is magnetized longitudinally and flux variations are detected by means of a search coil.

621.316.71 : 666.16 **3379**

Electronic Control of Glass-Grinding Machines.—(*Engineering, Lond.*, 1st May 1953, Vol. 175, No. 4553, pp. 574–575.) Glass windows whose edges are to be ground are mounted on a rotating chuck whose rate of rotation is controlled by a cam-operated servomechanism to prevent excessive grinding speeds at critical points of the profile.

621.384.612 **3380**

Radiation by Electrons in Large Orbits.—D. R. Corson. (*Phys. Rev.*, 1st June 1953, Vol. 90, No. 5, pp. 748–752.) Measured values of electron energy loss per orbit revolution in the Cornell University synchrotron are in excellent agreement with values calculated from classical e.m. theory.

621.384.613 **3381**

The Betatron.—R. Wideröe. (*Z. angew. Phys.*, May 1953, Vol. 5, No. 5, pp. 187–200.) A clear account of the theory, development and applications of the betatron to date. 70 references.

621.385.833 **3382**

An Improved Scanning Electron Microscope for Opaque Specimens.—D. McMullan. (*Proc. Instn elect. Engrs*, Part II, June 1953, Vol. 100, No. 75, pp. 245–256. Discussion, pp. 257–259.)

621.385.833 **3383**

Potential of an Electrostatic Electron Lens. Comparison of the Results of Calculations and of Measurements made in an Electrolyte Tank.—M. Laudet & P. Pilod. (*J. Phys. Radium*, May 1953, Vol. 14, No. 5, pp. 323–328.)

621.385.833 **3384**

Space-Charge Requirements in Some Ideally Focused Electronoptical Systems.—N. Wax. (*J. appl. Phys.*, June 1953, Vol. 24, No. 6, pp. 727–730.)

621.385.833 **3385**

The Magnetic Circuit in Electron-Microscope Lenses.—T. Mulvey. (*Proc. phys. Soc.*, 1st June 1953, Vol. 66, No. 402B, pp. 441–447.)

621.385.833 **3386**

The Effect of Pole-Piece Saturation in Magnetic Electron Lenses.—G. Liebmann. (*Proc. phys. Soc.*, 1st June 1953, Vol. 66, No. 402B, pp. 448–458.)

621.385.833 **3387**

Focusing of High-Energy Particles by Grid Lenses: Part 1—The Convergence of Grid Lenses.—M. Y. Bernard. (*J. Phys. Radium*, June 1953, Vol. 14, No. 6, pp. 381–394.)

621.385.833 : 537.12/.13 **3388**

Obtaining Electron Images of Surfaces Bombarded by Ions.—A. Septier. (*C. R. Acad. Sci., Paris*, 20th July 1953, Vol. 237, No. 3, pp. 231–233.)

621.387.424 **3389**

Application of Wilkinson's Theory to G-M Counters with External Cathode.—D. Blanc & H. Zyngier. (*C. R. Acad. Sci., Paris*, 6th July 1953, Vol. 237, No. 1, pp. 38–39.)

621.387.424 **3390**

Limitation of the Propagation of the Discharge in G-M Counters.—E. Picard & A. Rogozinski. (*J. Phys. Radium*, May 1953, Vol. 14, No. 5, pp. 304–306.)

PROPAGATION OF WAVES

538.566 **3391**

A Note on Sommerfeld's 1909 Paper.—B. M. Fannin. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 1059–1060.) Comment on 2871 of 1950 (Kahan & Eckart).

538.566 **3392**

Concerning Green's Reinterpretation of the Magneto-ionic Theory.—C. G. McCue. (*J. atmos. terr. Phys.*, June 1953, Vol. 3, No. 5, pp. 239–244.) In Green's equations of motion of an electron under the influence of the electric component of the radio wave, which were given in a handbook of the Ionospheric Prediction Service, N.S.W., Australia, October 1950, the x component of the electric wave vector is neglected. This invalidates the remainder of Green's analysis. Appleton's interpretation of the magneto-ionic theory appears to be sufficient at present. The calculation of the m.u.f., when the magnetic field of the earth is taken into consideration, is discussed.

538.566 **3393**

Wave Propagation in an Anisotropic Inhomogeneous Medium.—J. Feinstein. (*J. geophys. Res.*, June 1953, Vol. 58, No. 2, pp. 223–230.) The results are given of wave-theory calculations for the characteristic-mode polarizations and reflection coefficients for the case of finite gradients of electron density and an arbitrarily oriented geomagnetic field. Consideration of the variation of polarization with distance within the medium leads to a simple interpretation of the departures from geometrical optics. The treatment is extended to collision frequencies near the critical value. Modifications introduced by the wave principles used are discussed and the results of l.f. polarization measurements are explained.

538.566 : 551.551 **3394**

Electromagnetic-Field Fluctuations due to Turbulence, at the End of a Line-of-Sight Propagation Path.—J. Voge. (*C. R. Acad. Sci., Paris*, 27th July 1953, Vol. 237, No. 4, pp. 351–353.) Megaw (1105 of April) treated this problem on the basis of a turbulence spectrum physically the most probable. A Taylor type of spectrum is here assumed; this may be less correct, but is often simpler to use. Analysis for two cases considered, (*a*) ultra-short waves transmitted over a moderately great distance, (*b*) light waves, leads to formulae analogous to those of Megaw.

621.396.11 : [550.385 : 523.72 : 621.396.822 **3395**

Relationship between Radio-Propagation Disturbance, Geomagnetic Activity and Solar Noise.—van Sabben. (See 3283.)

621.396.11.029.6 **3396**

Large Reductions of V.H.F. Transmission Loss and Fading by the presence of a Mountain Obstacle in Beyond-Line-of-Sight Paths.—F. H. Dickson, J. J. Egli, J. W. Herbstreit & G. S. Wickizer. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 967–969.) A graph of transmission-loss/obstacle-height for paths of 50 and 150 miles and frequency 100 Mc/s is shown which is based on knife-edge diffraction theory. As compared with paths having no mountain obstacles, paths with such obstacles will show considerable transmission gains provided particular combinations of aerial height, obstacle height and frequency are chosen. Experimental results for a 38-Mc/s, 160-mile communication link in Alaska, with effective aerial height of 50 ft and obstacle height $>$ 8 000 ft, were in reasonable agreement with calculations. Field-strength records also showed absence of severe tropospheric fading.

621.396.11.029.62 **3397**

Study of Ultra-Short-Wave Propagation over the Barrier presented by the Alps.—J. Dufour. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, 1st May 1953, Vol. 31, No. 5, pp. 124–130. In French.) The signal strengths of North Italian and South German f.m. transmitters operating in the 90–95-Mc/s band were measured during the summer of 1952 at five stations in Switzerland. Recordings were made at three stations. The expected field-strength calculated from the free-space field, but allowing for diffraction at intervening ridges, agreed within 10 db with the measurement results which, apart from line-of-sight paths, show fair agreement with the C.C.I.R. curves. Rapid signal-strength variations were related to interference between the direct and the ground-reflected wave. Slow variations seemed to be due to refraction conditions. The absence of variations always corresponded to rainy windy weather. On two occasions the passage of a cold front coincided with marked signal increase. Apart from improved reception at high-altitude stations, the Alps are not found to introduce any new propagation effect.

RECEPTION

621.396.62 + 621.397.62] : 061.4 **3398**

Radio and Television at the Paris Fair.—P. A. François. (*TSF et TV*, June–Aug. 1953, Vol. 29, Nos. 296–298, pp. 202–206 & 241–242.) 133 radio receivers exhibited, including one f.m. receiver, are classified, with indications of ranges, number of valves, etc. Trends in design are noted.

621.396.621 **3399**

The Reception of Frequency-Shift Signals from Short-Wave Transmitters.—H. Bohnenstengel. (*Fernmeldetech. Z.*, June 1953, Vol. 6, No. 6, pp. 249–253.) The effect of interference on the reception of frequency-shift signals is analysed. A table is given relating bandwidth, mean effective noise voltage, the minimum voltage required for reception for (*a*) recorder operation with amplitude keying, (*b*) frequency-shift keying, (*c*) printer operation, and the permissible frequency shift using a minimum keying-pulse time of 20 ms. The effect of fading on reception is also investigated. The efficacy of normal frequency-shift systems is attributed to amplitude limitation. A frequency-diversity system using two values of frequency shift, or additional f.m. of the frequency radiated, is recommended.

621.396.621 **3400**

Magnetic Demodulation.—L. Pungs & G. Meinshausen. (*Frequenz*, June 1953, Vol. 7, No. 6, pp. 153–160.) The magnetic-flux/field-strength characteristic is made use of for demodulation which results from the rectification of the induction flux. Demodulation of A and B types, depending on the point of operation on the Φ/H curve, is discussed by analogy with demodulation by means of nonlinear resistors. The circuit and method used in the determination of the rectification characteristic curves are described and examples of load-line determination of the dynamic characteristics are given.

621.396.621 : 621.396.822 **3401**

Signal-to-Noise Ratios in Band-Pass Limiters.—W. B. Davenport, Jr. (*J. appl. Phys.*, June 1953, Vol. 24, No. 6, pp. 720–727.) A general analysis is made of the relation between the output signal and noise powers and the input signal and noise powers for band-pass limiters whose transfer characteristic is a nondecreasing odd function of its argument. Specific results are given for the case where the limiter output is proportional to the *n*th root of its input; they include the ideal symmetrical limiter as a limiting case. The output signal/noise power ratio is essentially directly proportional to the input signal/noise power ratio for all values of the latter. This is due to the band-pass characteristics, rather than to the symmetrical limiting action.

621.396.621 : 621.396.822 : 523.72 **3402**

Radio-Noise Receivers.—J. L. Steinberg. (*Onde élect.*, Nov. & Dec. 1952, Vol. 32, Nos. 308 & 309, pp. 445–454 & 519–526; June 1953, Vol. 33, No. 315, pp. 274–284.) A detailed account of work carried out from 1947 to 1950. The conditions were determined which receiving equipment must satisfy if it is to be used for measurement of u.h.f. radiation. The known characteristics of solar r.f. radiation are reviewed and the operation of a receiver is analysed for the case when the level of the received r.f. noise is small compared with the valve and circuit noise of the receiver itself. With regard to the measurement of temperatures, noise factor is defined, and statistical analysis is presented of the fluctuations in the measurement apparatus at the output of the receiver after the detector. A method of eliminating fluctuations of gain by use of a permutation system at the input of the receiver is described. This system of modulation results in an increase of the ratio of useful signal to noise and hence of the stability in all cases, even when gain fluctuations are absent. Various modulation systems are critically discussed and a new system is described. The relation between noise factor N and bandwidth Δf is considered. A minimum value of the quantity $N/\sqrt{\Delta f}$ is required; this condition introduces circuit problems different from those met with in normal radiocommunication practice. Measurements of the spectrum of gain fluctuations, made with a selective amplifier and a noise generator modulated with square waves, show that the spectrum extends much farther towards high frequencies than is indicated by American investigators. A complete description is given of r.f. noise-measurement equipment operating on 1·2 kMc/s, with a noise factor of 11 db and an input bandwidth of 10 Mc/s. With this equipment a variation of apparent aerial temperature of ± 1°K can be detected. With the aerial actually used, a received power of 4×10^{-23} W/cm² can be detected. Apparatus operating with a larger aerial on 158 Mc/s, but with a smaller bandwidth, can detect a power of $2{\cdot}5 \times 10^{-23}$ W/cm². Results of observations made with this equipment during the eclipse of the sun on 28th April 1949 are described and discussed in relation to optical measurements.

621.396.621.54 : 621.314.7 **3403**

Transistorized Superhet Receiver.—(*Electronics*, Aug. 1953, Vol. 26, No. 8, pp. 202 . . 205.) Description abstracted from a paper entitled 'Application of Transistors to Radio-Receiver Circuitry', by E. Toth. The special problems arising in receiver design owing to transistor characteristics such as low input impedance, high output impedance, low power-handling capacity, etc., are discussed. The receiver described has one r.f. amplifier stage (550 kc/s–1·55 Mc/s), mixer, heterodyne oscillator operating 455 kc/s above the signal frequency, three 455-kc/s i.f. amplifier stages, crystal-diode second detector, a.f. pre-amplifier and a.f. output stage, eight transistors being used. Gain control is effected by ganged potentiometers at the inputs of the r.f. amplifier and the first i.f. amplifier. An input of about 200 μV is required for 6 mW output at 1 kc/s, with 10 db output signal/noise ratio. Maximum power output is about 15–20 mW for 5% harmonic distortion at 1 kc/s. The total d.c. power required is about 1 W: 3 V, 8 mA for the emitter bias circuits, and 30 V, 30 mA for the collector circuits.

621.396.622 : 621.396.822 **3404**

Decrease of the Low-Frequency Signal-to-Noise Ratio when Increasing the Intermediate-Frequency Bandwidth, using a Square-Law or a Linear A.M. Detector.—H. de Lange Dzn. (*Commun. News*, June 1953, Vol. 13, No. 4, pp. 128–138.) Mechanical analogy suggests that for calculations relating to linear detection, a l.f. noise voltage can be treated as an a.m. and ph.m. carrier having the same frequency as that of the system. This provides a simpler method of calculation than that of Fränz (3026 of 1941 and 443 of 1944). Taking particular account of the relation between l.f. and i.f. noise spectra, Burgess's analysis for a linear detector (3098 of 1951) is modified to give results in closer agreement with experiment. Calculations of the relation between bandwidth and signal/noise ratio for different levels of modulation show that i.f. bandwidth can be increased considerably with only a moderate reduction of signal/noise ratio, particularly with linear detection. The use of pre-amplification to ensure linear operation of a detector is justified even for threshold signals.

621.396.622 : 621.396.822 **3405**

The Output Signal-to-Noise Ratio of a Power-Law Device.—N. M. Blachman. (*J. appl. Phys.*, June 1953, Vol. 24, No. 6, pp. 783–785.) First-order statistics are applied to the problem of a power-law device fed by a sinusoidal signal and narrow-band random noise, to obtain an expression for the signal/noise power ratio for the output components in the vicinity of any harmonic of the input signal, in terms of the input signal/noise power ratio. Formulae are given for the cases of large and small values of the input ratio. See also 2168 and 2169 of 1945 (Rice) and 1175 of 1949 (Middleton).

621.396.822 : 621.317.34 **3406**

The Measurement and Assessment of Background Noise.—E. Belger. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, March/April 1953, Vol. 5, Nos. 3/4, pp. 51–59.) The characteristics of noise meters are surveyed and experimental results of subjective tests made to determine the permissible signal/noise ratio are given. The average ratio to be aimed at is ~ 55 db, but 45 db is tolerable for most types of modulation and even at 35 db the quality of many programmes is satisfactory.

STATIONS AND COMMUNICATION SYSTEMS

016 : 621.396.931 **3407**

Metre Waves in Mobile Services.—R. Hermann. (*Onde élect.*, May 1953, Vol. 33, No. 314, pp. 347–352.) Classified bibliography of papers and books published in U.S.A., Europe and Australia before February 1952, dealing with mobile communication systems and equipment.

621.394.333 : 621.018.78 **3408**

Oscillograph Representation of the Degree of Distortion in Teletype Signals.—K. W. Seiffert. (*Fernmeldetech. Z.*, May 1953, Vol. 6, No. 5, pp. 214–217.) The degree of general distortion and the degree of relative distortion are defined and two c.r. instruments used to measure them are described.

621.395.44 **3409**

The L3 Coaxial System: System Design.—C. H. Elmendorf, R. D. Ehrbar, R. H. Klie & A. J. Grossman. (*Bell Syst. tech. J.*, July 1953, Vol. 32, No. 4, pp. 781–832.) Design problems and requirements for the system, which provides 1 860 telephony channels or 600 telephony channels and a television channel in each direction on a pair of coaxial cables, are discussed and methods adopted to meet the requirements are described. An account is given of the main features of the terminal and repeater equipment.

621.395.44 : [621.395.521.3 + 621.395.664 **3410**

The L3 Coaxial System: Equalization and Regulation.—R. W. Ketchledge & T. R. Finch. (*Bell Syst. tech. J.*, July 1953, Vol. 32, No. 4, pp. 833–878.) A theory of the equalization of complex systems is outlined and the location and function of the various equalizers are explained. The analogue computer used in the regulation system is described and also the cosine-equalizer adjustment technique used with manual equalizers. Details of the circuits and operation of the regulation system are given.

621.395.44 : 621.395.645 **3411**

The L3 Coaxial System: Amplifiers.—L. H. Morris, G. H. Lovell & F. R. Dickinson. (*Bell Syst. tech. J.*, July 1953, Vol. 32, No. 4, pp. 879–914.) The circuits and mechanical design of the line amplifiers and the flat-gain amplifiers are described. The two types are basically similar, consisting of two feedback amplifiers in tandem coupled by a network which, in the case of the line amplifier, is variable and is automatically adjusted to compensate for variations in cable temperature and for small deviations from the nominal 4-mile spacing of repeaters. All important components are subject to strict quality control to ensure uniformity of amplifier performance.

621.396.4 : 621.396.65 **3412**

Experimental Radio Bearer Equipment for Carrier Telephone Systems.—W. S. McGuire & A. G. Bird. (*Proc. Instn Radio Engrs, Aust.*, June 1953, Vol. 14, No. 6, pp. 135–147.) Two f.m. bearer systems are described, for the frequency ranges 420–470 Mc/s and 860–960 Mc/s respectively. Both are crystal controlled, with a frequency deviation of ±180 kc/s and a carrying capacity of 12–17 telephone channels. The performance of both systems meets C.C.I.F. requirements when used with the appropriate carrier equipment. Intermediate relay stations re-radiate on a frequency slightly different from that of the received signal. Details are given of a typical installation comprising transmitter, repeater and receiver for the 420–470-Mc/s band, and its performance over a 100-mile circuit is reported.

621.396.619.13 : 621.392 **3413**

A.M.-F.M. Analogy.—Harris. (See 3246.)

621.396.619.16 **3414**

Coding by Feedback Methods.—B. D. Smith. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 1053–1058.) The feedback coder converts an analogue quantity, such as a voltage, into a digital quantity. It comprises an error amplifier, control circuits and a decoding network which is used as the feedback element in the amplifier. The binary coder is considered in detail, and the principles of binary-coded decimal systems are briefly mentioned. The feedback coding method is compared with the counting and coding-tube methods, and a system of nonlinear coding is outlined.

621.396.619.16 **3415**

New Method of Modulation with Reduced Bandwidth.—F. Benz. (*Öst. Z. Telegr. Teleph. Funk Fernsehtech.*, May/June 1953, Vol. 7, Nos. 5/6, pp. 66–75.) Two equal carriers 90° out of phase are modulated by the outputs of two valves in push-pull. These valves are driven by an alternating pulsed input, the modulation voltage being applied to the screen grids in parallel. The two modulated carriers are combined and amplified. For the same h.f. bandwidth, the received l.f. bandwidth is twice that obtaining in other modulation systems. For reception, two i.f. signals are derived and demodulated separately before being applied to a phase-discriminator combining circuit. Reception is possible using a normal a.m. receiver. Distortion due to inaccurate phase and frequency transformations in the receiver is calculated and the application of the method in p.w.m., p.ph.m., and p.c.m. systems is discussed.

621.396.722 **3416**

International Monitoring.—J. T. Dickinson. (*Wireless World*, Sept. 1953, Vol. 59, No. 9, pp. 422–423.) Functions and equipment of the receiving station of the European Broadcasting Union, opened in July 1953 at Jurbise-Masnuy, near Mons, are briefly described. Continuous watch is kept, covering all broadcasting bands. Two frequency standards are housed in an underground compartment, one of modified Telefunken design with outputs of 1, 10, 100 and 1 000 kc/s derived from a 500-kc/s quartz crystal, the other of American manufacture. Frequency monitoring, accurate to within 4 or 5 parts in 10^7, is based on heterodyning the received signal with an appropriate r.f. signal of known frequency injected into the receiver. For l.f. and m.f. reception, inverted-L aerials are used; for h.f., elevated horizontal dipoles; for v.h.f., a rigid horizontal dipole. Large rotating frame aerials facilitate reception in crowded channels.

621.396.931 **3417**

Problems concerning Radio Transmission for Telephone Links with Mobile Stations.—W. Klein. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, 1st June 1953, Vol. 31, No. 6, pp. 145–168.) French version of paper abstracted in 1486 of May.

621.396.932 **3418**

Radiotelegraphy, Radiotelephony and Navigational Aids aboard Merchant Ships and Fishing Vessels.—L. Lahure & J. Fontaine. (*Onde élect.*, May 1953, Vol. 33, No. 314, pp. 289–299.) A review of the development of ships' radio apparatus from the first transmitters to modern equipment based on recommendations of recent conventions.

621.396.933 **3419**

Telecommunications and Radio Aids in Civil Aviation.—(*Onde élect.*, May 1953, Vol. 33, No. 314, pp. 249–288.) Seven papers reviewing post-war developments:—

Organization of Civil Aviation.—Portier. An outline of national and international arrangements.

Telecommunications in Civil Aviation.—G. Hoerter. An account of traffic and navigation systems required for different air services.

Control Towers.—Macelloni & Vannel. Description of equipment lay-out and v.h.f. R/T, recording and d.f. apparatus.

Central Telecommunications Office.—Quiquandon & Lalmi. Operation of telegraph and telephony services and receiving equipment of a typical centre are described.

H.F. Transmitters for Aerodromes.—Nill. Illustrated general description of a standard series of fixed-frequency transmitters for 50 W, 300 W, 1 kW and 10 kW.

Radio Aids to Air Navigation.—Villiers. A review of past and present systems, in particular loran and consol, V.O.R./D.M.E. and Decca, and instrument landing systems I.L.S. and G.C.A.

Radio Installations at French Overseas Aerodromes.—Bargain. A note on the phases of development, with a map showing the services planned for different airfields, particularly in N.W. Africa.

SUBSIDIARY APPARATUS

621-526 **3420**

A Study of a Second-Order Sampling Servo.—S. R. Cooper. (*Electronic Engng*, Aug. 1953, Vol. 25, No. 306, pp. 342–349.)

621-526 **3421**

Operating Modes of a Servomechanism with Nonlinear Friction.—H. Lauer. (*J. Franklin Inst.*, June 1953, Vol. 255, No. 6, pp. 497–511.)

621-526 **3422**

Considerations on Discriminators in Airborne Servo Systems.—J. C. Gille. (*Onde élect.*, May 1953, Vol. 33, No. 314, pp. 337–342.) Discussion of performance requirements for automatic-pilot mechanisms.

621-526 **3423**

Reduction of Forced Error in Closed-Loop Systems.—L. H. King. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 1037–1042.)

TELEVISION AND PHOTOTELEGRAPHY

621.397.242 : 621.395.44 **3424**

The L3 Coaxial System: Television Terminals.—J. W. Rieke & R. S. Graham. (*Bell Syst. tech. J.*, July 1953, Vol. 32, No. 4, pp. 915–942.) The special requirements of terminal equipment for the transmission and reception of television signals are discussed. The transmitting and receiving equipments are described, with details of the modulation process, vestigial-sideband operation, filter characteristics, pilot-frequency generator, etc.

621.397.26 : 621.396.65 **3425**

The Hinsbeck Relay Station in the International Television Link on the occasion of the Coronation.—H. Ehlers & G. Dröscher. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, March/April 1953, Vol. 5, Nos. 3/4, pp. 33–36.) The planning of the station is discussed and the equipment used in the relay described, with a block diagram showing the arrangements for duplicate reception and retransmission to Wuppertal for distribution to the N.W.D.R. network.

621.397.26 : 621.396.65 **3426**

New Television Directional Radio Links of the Federal German Post Office.—(*Fernmeldetech. Z.*, May & June 1953, Vol. 6, Nos. 5 & 6, pp. 220–233 & 269–279.) A series of seven articles by various authors giving a description of the complete system linking Hamburg and Cologne, with the recent extension to Frankfurt am Main. See also 1431 and 2033 of 1952 (Schmidt) and 2160 of July (Behling et al.)

621.397.335 : 535.623 **3427**

A Subjective Study of Color Synchronization Performance.—M. I. Burgett, Jr. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 979–983.) The circuits associated with five main colour-receiver functions are so grouped that noise can be introduced into each separately in amounts controllable by the observer, who varies the noise to match each of seven standard comments ranging from 'not perceptible' to 'not usable'. N.T.S.C. signals are applied, and a monochrome receiver is also used for comparison. Results obtained by ten observers indicate that noise in the colour-synchronization circuit has relatively little adverse effect on picture quality, compared with noise in the luminosity-information or deflection-synchronization circuits.

621.397.5 : 535.623 **3428**

Principles and Development of Color-Television Systems.—G. H. Brown & D. G. C. Luck. (*RCA Rev.*, June 1953, Vol. 14, No. 2, pp. 144–204.) A review of the development of compatible colour-television systems by the R.C.A. from 1940 to 1953 is given and the fundamentals of colorimetry and the physiology of vision are discussed. The 1953 N.T.S.C. field-test specifications are given in an appendix.

621.397.5 : 535.623 **3429**

Colorimetric Analysis of R.C.A. Color-Television System.—D. W. Epstein. (*RCA Rev.*, June 1953, Vol. 14, No. 2, pp. 227–258.) An outline of the principles of colorimetry is given, and the effects of camera spectral characteristics and studio lighting on the fidelity of reproduction in the R.C.A. compatible colour-television system are analysed.

621.397.6 : 621.317.755 **3430**

Testing of Television Studio Equipment by means of Synchronizing Pulses of Variable Phase.—Demus. (See 3373.)

621.397.61 : 535.623 **3431**

Optimum Utilization of the Radio-Frequency Channel for Color Television.—R. D. Kell & A. C. Schroeder. (*RCA Rev.*, June 1953, Vol. 14, No. 2, pp. 133–143.) Discussion of the technical and physiological considerations on which the N.T.S.C. specifications of the field-test signal are based.

621.397.611.2 **3432**

Standards Converter for International TV.—A. V. Lord. (*Electronics*, Aug. 1953, Vol. 26, No. 8, pp. 144–147.) Description of the principles and construction of equipment of the type used for conversion of programmes from the French 819-line standard to the British 405-line standard. See also 2469 of August.

621.397.611.2 **3433**

Signal Generation in Television Camera Tubes: Part 2—Construction, Operation and Performance of the Different Types of Tube.—R. Theile. (*Arch. elekt. Übertragung*, June & July 1953, Vol. 7, Nos. 6 & 7, pp. 281–290 & 328–337.) See 1828 of June. Part 1: 1829 of June.

621.397.62 **3434**

A 427/45-Mc/s Converter for the Society's Television Transmissions.—D. N. Corfield. (*J. Televis. Soc.*, April/June 1953, Vol. 7, No. 2, p. 86.) Corrections to paper abstracted in 2809 of September.

621.397.62 + 621.396.62] : 061.4 **3435**

Radio and Television at the Paris Fair.—François. (See 3398.)

621.397.62 : 535.623 **3436**

Color-Television-Signal Receiver Demodulators.—D. H. Pritchard & R. N. Rhodes. (*RCA Rev.*, June 1953, Vol. 14, No. 2, pp. 205–226.) The basic concepts of a simultaneous subcarrier colour system are described, with particular reference to the receiver demodulator problem. The design of demodulators for the N.T.S.C. type of signal is discussed and examples are given of practical circuits.

621.397.62 : 621.396.615.17.018.75 **3437**

An Introduction to the Sine-Squared Pulse.—Hunt & Elliott. (See 3377.)

621.397.62 : 621.396.662 **3438**

Factors Affecting the Design of V.H.F.-U.H.F. Tuners.—E. H. Boden. (*Sylvania Technologist*, July 1953, Vol. 6, No. 3, pp. 64–67.) Design requirements for the stages of a single tuner for U.S. channels 2–83, utilizing Type-6AN4 triodes in the r.f. amplifier and mixer stages and a Type-6T4 triode as oscillator valve, are considered. Three possible tuner circuits are examined. Noise-figure and gain measurements made on the compromise tuner gave values of 7–14 db and 25–20 db respectively, in the frequency range from channel 2 to channel 83.

621.397.621 **3439**

In Search of the Perfect Raster.—P. J. Edwards. (*J. Televis. Soc.*, April/June 1953, Vol. 7, No. 2, pp. 60–76.) Defects considered are: (*a*) inaccurate interlace; (*b*) deformities of the complete raster and of the individual scanning lines; (*c*) nonlinearity of scan; (*d*) nonuniformity of focus. The adverse influence of incorrect synchronization is considered and a description is given of a synchronization separator circuit using a cathode-coupled limiter. Good raster shape and linear scanning fields can be obtained by using deflection coils with suitably graded windings.

621.397.621.2 : 535.623 **3440**

The Preparation of Phosphor Screens for Color-Television Tubes.—S. Levy & A. K. Levine. (*Sylvania Technologist*, July 1953, Vol. 6, No. 3, pp. 60–63.) A photographic method for the preparation of three-dot screens for colour-television tubes is described. A paste consisting of a green, red or blue phosphor mixed with a photosensitive binder is applied to the glass screen and is illuminated by a point source of light through a mask of the desired pattern. The screen is then developed, fixed and the unexposed areas are washed away by a solvent. The process is repeated for the other colours. A permanent silicate binder is sprayed on, after removing the photosensitive binder by baking at 400°C.

621.397.826 **3441**

Influence of Echoes on Television Transmission.—P. Mertz. (*J. Soc. Mot. Pict. Telev. Engrs*, May 1953, Vol. 60, No. 5, pp. 572–596.) Image distortion due to echoes is classified in terms of the characteristics of the echo signals, and tolerances for small- and large-screen television pictures in respect of overall phase drift, envelope delay and phase delay are estimated.

TRANSMISSION

621.396.619.14 **3442**

Susceptance Valves and Reactance Valves as Phase Modulators.—A. van Weel. (*J. Brit. Instn Radio Engrs*, June 1953, Vol. 13, No. 6, pp. 315–320.) In two of the three basic ways of connecting a triode to operate as a variable impedance, the grid-anode capacitance is effectively in parallel with the circuit impedance. The application of these 'susceptance' valves for ph.m. is described, only one valve being necessary to introduce phase variations up to 45° with mutual-conductance variations $< 0{\cdot}5$ mA/V. A pentode grounded-anode susceptance-valve circuit and a pentode reactance-valve circuit, both providing a ph.m. output current, are described.

621.396.932 **3443**

Some Problems in the Design of Marine Transmitters.—D. J. Spooner. (*J. Brit. Instn Radio Engrs*, June 1953, Vol. 13, No. 6, pp. 325–330.) Technical requirements in transmitter design in respect of frequency stability, bandwidth, power output and keying, based on official performance specifications, are discussed. Recommended 'type-approval' tests, including a sequence of climatic and durability tests, are described.

VALVES AND THERMIONICS

537.311.33 : 621.314.7 **3444**

Transistors: Theory and Application: Part 6—Operation of Junction Transistors.—A. Coblenz & H. L. Owens. (*Electronics*, Aug. 1953, Vol. 26, No. 8, pp. 156–161.) Discussion of the physical and electrical properties of transistor triodes and tetrodes, *p-n-p-n* junctions, and the phototransistor. Part 5: 3021 of October.

621.314.7 **3445**

Unipolar 'Field-Effect' Transistor.—G. C. Dacey & I. M. Ross. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 970–979.) The field-effect transistor is essentially a structure containing a semiconducting current path, the conductivity of which is modulated by the application of a transverse electric field. Modifications are made to Shockley's ideal theory to take account of the following factors:—(*a*) series resistance at the source and/or drain contacts, (*b*) carrier depletion, (*c*) negative gate resistance, (*d*) temperature effects. Design charts are then developed and their use explained. Details of results obtained with experimental units differing slightly in their dimensions are given. The results are in substantial agreement with the modified theory. Units having stable characteristics, transconductances up to 0·3 mA/V and flat frequency response to 3 Mc/s have been produced.

621.314.7 : 537.311.33 : 546.289 **3446**

Mechanism of Point-Contact Transistors.—Thedieck. (See 3331.)

621.383.032.2 **3447**

Development of the Cs-AgO Photocathode during Thermal Treatment.—V. Schwetzoff. (*C. R. Acad. Sci., Paris*, 27th July 1953, Vol. 237, No. 4, pp. 320–322.) The observed changes of photoelectric and secondary emission are shown graphically and discussed.

621.383.2.014.33 **3448**

Pulse Irradiation of Composite Photocathodes with Intermediate Semiconducting Layers.—W. Kluge & S. Weber. (*Naturwissenschaften*, June 1953, Vol. 40, No. 11, p. 315.) Pulse illumination of Ag-Cs_2O-Cs cathodes in vacuum photocells resulted in no fatigue. The illumination with white light was varied up to 10^7 lux, with a pulse width at half maximum intensity of 15 μs. The saturation pulse current was proportional to the intensity of illumination.

621.383.42 **3449**

The Modern Single-Layer Selenium Photoelectric Cell.—G. A. Veszi. (*J. Brit. Instn Radio Engrs*, April 1953, Vol. 13, No. 4, pp. 183–189.) Review of photocell development and applications.

621.385 : 537.525.92 **3450**

Propagation of Space-Charge Waves in Infinite and Finite Electron Beams.—P. Parzen. (*Elect. Commun.*, June 1953, Vol. 30, No. 2, pp. 134–138.) Small-signal theory is developed for the case of beams of infinite lateral extent in planar diodes. The theory yields results in agreement with those of Llewellyn & Peterson (2578 of 1944) but can more readily be extended to deal with finite beams and with the effects of thermal velocities. Its application is not restricted to diodes. The analysis is relevant to the properties of travelling-wave valves.

621.385.017.72 **3451**

The Vapotron.—G. Ashdown. (*Electronic Engng*, Sept. 1953, Vol. 25, No. 307, pp. 378–379.) A detailed account of the vaporization method of cooling high-power transmitting valves has previously been given by Beurtheret (542 of 1952).

621.385.029.6 **3452**

Effect of Thermal-Velocity Spread on the Noise Figure in Traveling-Wave Tubes.—P. Parzen. (*Elect. Commun.*, June 1953, Vol. 30, No. 2, pp. 139–154.) Reprint. See 2934 of 1952.

621.385.029.63/.64 **3453**

On the Theory of the Helix-Type Travelling-Wave Valve.—E. I. Vasil'ev & V. M. Lopukhin. (*Zh. tekh. Fiz.*, Nov. 1952, Vol. 22, No. 11, pp. 1838–1842.) A theoretical investigation of the effect of the velocity scatter of the beam electrons on the range of the existence and on the value of the complex roots of the dispersion equation for a travelling-wave valve using a helix.

621.385.029.64 **3454**

Traveling-Wave Oscillator Tunes Electronically.—H. R. Johnson & J. R. Whinnery. (*Electronics*, Aug. 1953, Vol. 26, No. 8, pp. 177–179.) Details are given of the construction and operating characteristics of an electrically short valve with an output > 100 mW at 3 kMc/s. By variation of the helix voltage, a 4·5% change of frequency is obtainable. The electrical length is 14 λ. External feedback through a filter eliminates undesired oscillation modes.

621.385.032.216 : 537.581 : 537.311.32 **3455**

Relationship between Thermionic Emission and Electrical Conductivity of Oxide-Coated Cathodes.—S. Narita. (*J. phys. Soc. Japan*, May/June 1953, Vol. 8, No. 3, pp. 331–338.) (Ba, Sr)CO_3 and sintered $BaCO_3$ cathodes were investigated. A new emission and conduction mechanism is proposed for oxide-coated cathodes.

621.385.832 : 621.396.662 **3456**

A New Type of Tuning Indicator for Battery or Mains Receivers.—H. P. White. (*Mullard tech. Commun.*, July 1953, Vol. 1, No. 4, pp. 104–110.) Full details are given of the new Mullard DM70 tuning indicator. It has a simple electrode structure and a 1·4-V filament.

621.387 **3457**

The Characteristics of some Large-Current Glow-Discharge Tubes.—F. A. Benson. (*Electronic Engng*, Aug. 1953, Vol. 25, No. 306, p. 321.) Short-time characteristics of Type-CV1199 tubes, designed for the current range 30–180 mA, are reported.

621.396.615.141.2 **3458**

Theory of the Multisegment Magnetron.—V. I. Kalinin & T. P. Ryazonova. (*Zh. tekh. Fiz.*, Oct. 1952, Vol. 22, No. 10, pp. 1592–1598.) The results obtained by Slutskin (1839 of 1948) are further developed to cover the case when the modulation of the density of the tangential electron stream is taken into account. The discussion is based on the consideration of an electrical circuit equivalent to a multisegment magnetron (Fig. 1) and a general formula (12) is derived for the mean energy exchange between the electron streams and the slots during one cycle. Several particular cases are considered in detail; the theoretical conclusions are in good agreement with experimental results.

621.396.615.142 **3459**

Debunching of Electron Beams constrained by Strong Magnetic Fields.—M. Chodorow, E. L. Ginzton & E. J. Nalos. (*Proc. Inst. Radio Engrs*, Aug. 1953, Vol. 41, No. 8, pp. 999–1003.) Feenberg's solution of the equations of electron motion in v.m. valves is arbitrarily extrapolated to cover large values of the bunching parameter. Experiments to test the validity of this procedure were made on a conventional type of two-cavity klystron. With increasing value of the debunching parameter, the output r.f. current decreased and the input voltage required to produce maximum output current increased. Voltages at the cavity gaps were well below the beam voltage in all experiments. A phase reversal of output current at a particular value of beam radius was observed. Experimental results are in reasonable accordance with theory, when this is corrected for nonuniformity of the beam cross-section. The extrapolation may be considered valid for values of the debunching parameter $< 3{\cdot}5$ and of the bunching parameter < 4.

MISCELLANEOUS

001.891 : 621.396 **3460**

Radio Research 1952. [Book Review]—Publishers: H.M. Stationery Office, London, 1953, 51 pp., 2s. (*Elect. J.*, 24th July 1953, Vol. 151, No. 4, p. 262.) Reports of the work carried out during 1952 at the Radio Research Station, Slough, at overseas stations, and at British universities collaborating with D.S.I.R.

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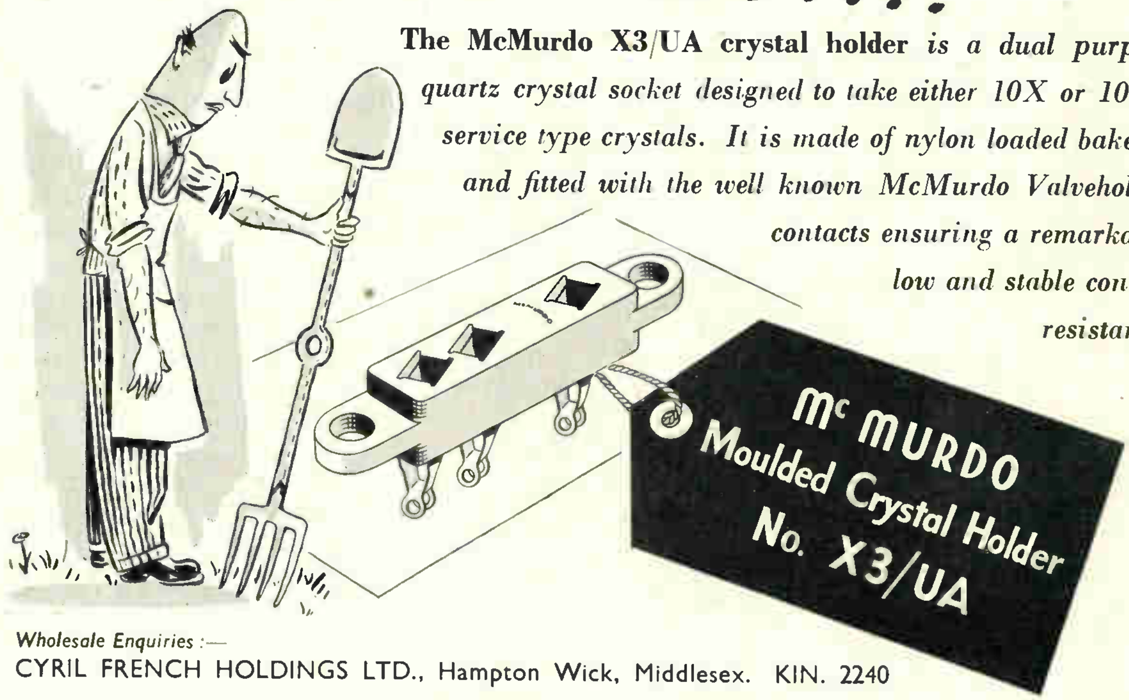

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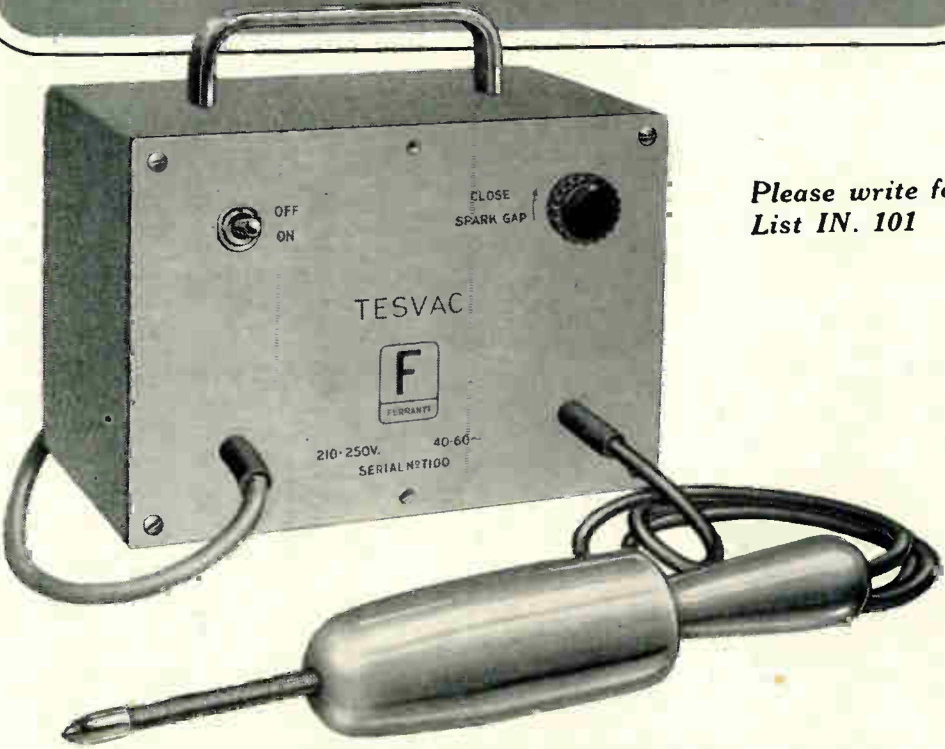

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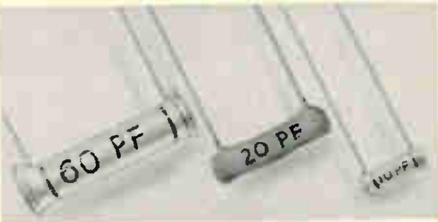

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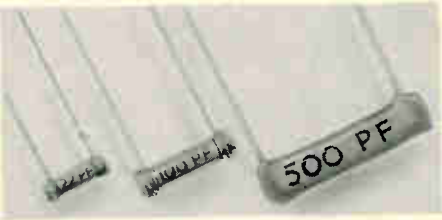

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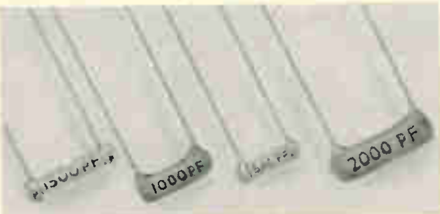

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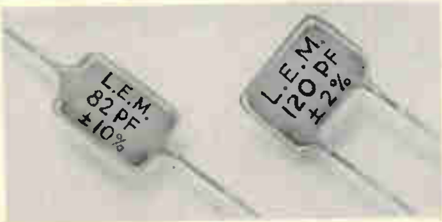

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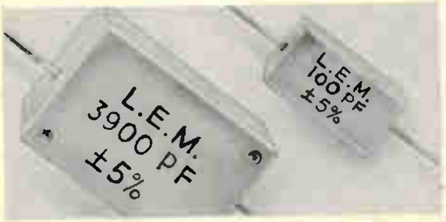

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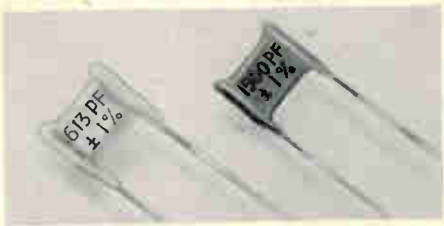

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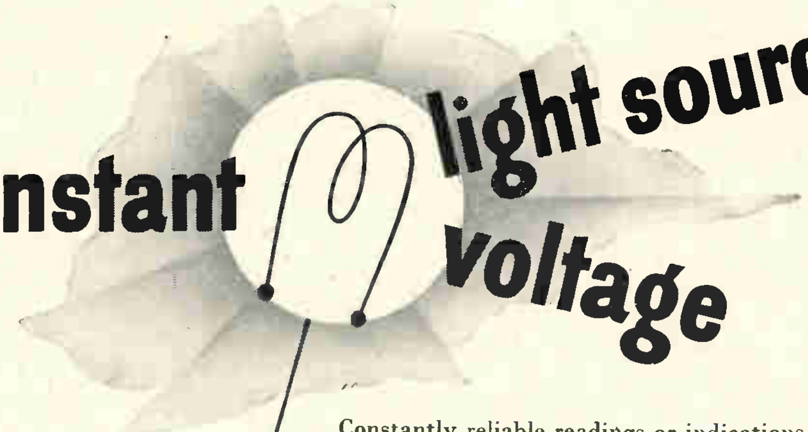

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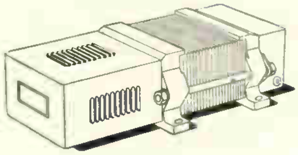

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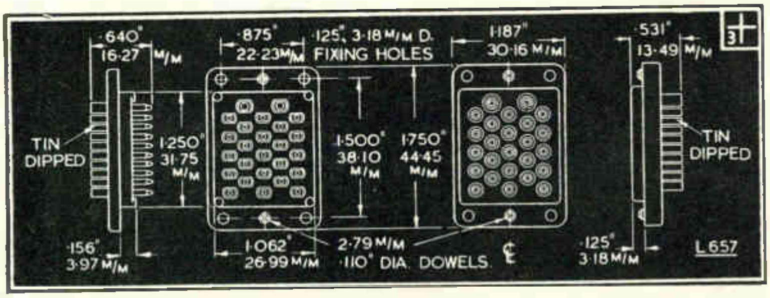
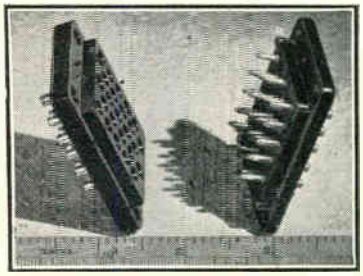


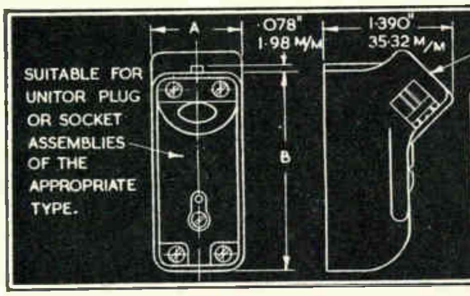


LIST NO.	PINS
L.653/P & S	4
L.654/P & S	8
L.655/P & S	12
L.656/P & S	18
L.657/P & S	25

UNITORS

These unitors (pronounced " unite-ors ") are the result of a Government development contract carried out with the closest collaboration on all sides, and the range is now very popular.

Like any other inter-unit connector, their main purpose is to facilitate the uncoupling of a sub-assembly for servicing or replacement in event of failure. The body is moulded from a moisture and tracking-resistant nylon-filled phenolic material. There is no metal surround, a four-hole fixing flange being incorporated in the moulding, together with two locating dowels. In the case of the four-in-line model two-hole fixing is provided with no dowels.



L No.	A INS.	A M/M	B INS.	B M/M
L 653/C2	·562	14·29	1·594	40·48
L 654/C2	·703	17·86	1·615	41·02
L 655/C2	·968	24·61	1·632	41·45
L 656/C2	·968	24·61	2·125	53·98
L 657/C2	1·312	33·34	1·881	47·78

Air Registration Board approval has been granted on all types, with the exception of the 4-pin model on which approval is expected in the very near future. Joint Service Approval has been granted on all types.

U.K. Pat. 649739
SWISS Pat. 277799

A coupling comprises a block of plugs and a second block of sockets, arranged so as to be non-reversible. Round pins are employed, turned from high-grade brass, silver plated. The normal pins carry 3 amps, but each block has two large pins or sockets to carry 10 amps. All pins are fully floating, and the method of assembly is the subject of a patent.

Resilient sockets are of differentially hardened beryllium copper. All contacts, plugs and sockets are numbered on the face and reverse sides of the body.

The distance between flanges when plugged together is 0·281 in.

A cable cover for line connection with retainer for chassis mounting is available.

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Six G.B. tele-projectors, throwing a picture on to a screen 4 ft × 3 ft, suitable for viewing by up to 200 persons, are available at a special price of £160 each. The list price, new, is £225.

These were bought by a well-known engineering firm in order to show the Coronation ceremony to their overseas visitors, and have not been used for any other purpose. They are in excellent condition, are at present tuned for reception in the London area, but are easily adaptable. Demonstrations in London can be arranged on application.—Write to Box 1289, c/o *Wireless Engineer.*

Wireless Station Superintendent required by the **Nigeria Government** Posts and Telegraphs Department for one tour of 18 to 24 months in the first instance. Commencing salary according to experience in scale £864 rising to £1,392 a year. Gratuity £100/£150 a year. Outfit allowance £60. Free passages for Officer and wife and assistance towards cost of children's passages or their maintenance in the U.K. Liberal leave on full salary. Candidates must have had wide practical experience of modern radio techniques and equipment, in particular V.H.F. equipment, and preferably also V.H.F. multi-channel equipment.

Write to the Crown Agents, 4 Millbank, London, S.W.1. State age, name in block letters, full qualifications and experience and quote M2C/28927/WJ.

APPOINTMENTS—The engagement of persons answering these advertisements must be made through the local office of the Ministry of Labour and National Service, etc., if the applicant is a man aged 18-64 or a woman 18-59 inclusive, unless he or she or the employer is excepted from the provisions of The Notification of Vacancies Order, 1952.

The G.E.C. Stanmore Laboratories invite applications for the following vacancies for work on guided missile research. Good canteen facilities, sick fund and pension scheme. Holidays above average.

(i) **Senior Graduate** with 1st or 2nd class Honours Degree or equivalent qualifications required to work on the planning and assessment of complex trials on airborne radar. Previous experience in radar work highly desirable. (Ref. WE/LMKB. 1.)

(ii) **Graduate** aged about 30 with good degree or equivalent qualifications required for responsible work on technical flight trials of airborne radar. Previous experience of similar work and of flying is very desirable. There is also a junior post vacant for a man with similar qualifications. Both these posts (together with (i) above) involve some travelling within the United Kingdom. (Ref. WE LMKB. 2.)

Applications should be made in writing to the Staff Manager, G.E.C., Stanmore Laboratories, The Grove, Stanmore Common, Stanmore, Middlesex, stating age, qualifications and experience and quoting the appropriate reference.

Assistant Signals Officer required by the **Government of Nigeria** for the Aviation Department for one tour of 18 to 24 months in the first instance. Salary, etc., either (*a*) in scale £750 rising to £1,315 a year with prospect of permanency or (*b*) in scale £807 rising to £1,453 a year, on a temporary basis with gratuity at the rate of £100 a year. Outfit allowance £60. Free passages for officer and wife, and assistance towards cost of children's passages, or their maintenance in the United Kingdom. Liberal leave on full salary. Candidates must have a knowledge of the fundamental principles of electricity and magnetism and of radio engineering with experience in the maintenance of aeronautical radio transmitters and receivers, direction finders, test equipment and small petrol and diesel engine generator sets. Workshop experience and a knowledge of radar will be of advantage. Minimum examination qualifications are C. & G. Certificate in radio communications or technical electricity, or satisfactory pass in M.C.A. radio mechanics course.

Write to the Crown Agents, 4 Millbank, London, S.W.1. State age, name in block letters, full qualifications and experience and quote M2C/29637/WJ.

An Opportunity occurs for a few Senior Engineers or Physicists (aged 30-35) with experience in light Electro-Mechanical Design, to join the expanding laboratory team of a well-known engineering company in the London area. Applicants should have a practical approach to problems and should have had experience in one or more of the following subjects:—Navigational Instruments, Communication Transmission Systems (with special reference to Interference Problems), Waveguide or Pulse Techniques. Salaries offered are in accordance with experience, and range from £800 to £1,100. A pension scheme is in operation. Full details of education and career should be sent in confidence, quoting ref. WE 772, to Box 1288, c/o *Wireless Engineer.*

Vacancies exist in the Valve Measurements and Applications Laboratory and in the Technical and Quality Control Department of the **Mullard Radio Valve Co., Ltd.**

Applicants should hold a degree or equivalent qualification in electrical engineering or physics. Some previous experience in general circuit work would be an advantage but is not essential. Candidates should, however, have a definite interest in circuit work as applied to tubes, valves and semi-conducting devices, as the work they will be required to assist in is concerned with the measurements of the more obscure characteristics of these articles, the investigation of, the possible applications and the solution of circuit problems. The salaries offered vary with age and experience. Applications in writing should be addressed to: Personnel Officer, **Mullard Radio Valve Co., Ltd.**, New Road, Mitcham Junction, Surrey, quoting ref. W1.A.

Senior Design Engineer. E. K. Cole, Ltd., Southend-on-Sea, require for Television Development Laboratory, Senior Engineer experienced R.F. side of television receiver design. Also Junior Engineer with laboratory experience. Write stating age, qualifications and experience to Personnel Manager.

Physicists and Development Engineers required for Research and Development work on Cold Cathode Tubes.
Applicants should preferably hold a degree in physics or engineering, or the equivalent, and have had at least two years' experience on Gas Discharge Tubes. Salary range £600 to £1,000 depending on qualifications and experience. Write giving full details to the Personnel Officer, Ericsson Telephones, Ltd., Beeston, Nottingham. (Reference LK/3.)

THE ENGLISH ELECTRIC CO., LTD.,

have vacancies
for the following staff in the Electronic Department at Luton:—

(*a*) **Senior Microwave Engineers** at a salary up to £1,000 p.a. Applicants should have had considerable design and/or engineering experience of microwave systems and associated equipment. Ref. 1160A.

(*b*) **Electronic Engineers** at a salary up to £750 p.a. Duties include the engineering of circuitry already developed and involve close liaison with and progressing of work through the drawing office and production department. Applicants with experience of engineering of radar and/or aircraft electronics for production will be especially welcome. Ref. 1071B.

(*c*) **Electronic Trials Engineers** at a salary of up to £750 p.a. Applicants should have experience of high frequency communication equipment and be prepared to accept responsibility for its installation and subsequent trial at locations in U.K. other than Luton. Ref. 456P.

(*d*) **Trials Radar Engineers** at a salary up to £750 p.a. Applicants should have had experience of operating development radar equipment in the field and have a sound basic knowledge of radar principles and circuits.
Vacancies also exist for Engineers in Australia. Ref. 1190.

(*e*) **Electronic Trials Assistants** at a salary up to £12 per week. These assistants will be formed into teams under the leadership of electronic trials engineers (see above) and will be required to assist in the installation and subsequent trial of H.F. communication or radar equipment. Ref. 456Q.

(*f*) **Electronic Laboratory Assistants** at a salary up to £10 per week.

Applicants should have either experience of:—

(i) The preparation of radar equipment for field trials for which a sound basic knowledge of radar circuitry is essential and special knowledge of radar equipment, AA No. 3, Mk. 7, would be advantageous.

or

(ii) The preparation of information for drawing office and production department from circuit diagrams. Ref. 1066D.

Housing accommodation is available to suitable applicants for posts (*c*) and (*e*) above.

Applications should be sent to Dept. C.P.S., 336/7 Strand, W.C.2, quoting appropriate reference.

Belling & Lee, Ltd., Cambridge Arterial Road, Enfield, Middlesex, require research assistants in connection with work on electronic components, fuses, interference suppressors and television aerials. Applicants must be graduates of the I.E.E. or possess equivalent qualifications together with similar laboratory experience. Salary will be commensurate with previous experience: five-day week, contributory pension scheme. Applications must be detailed and concise, and will be treated as confidential.

THE BRITISH IRON & STEEL RESEARCH ASSOCIATION
Graduate Physicist of Electrical Engineering required by the above Association, for research and development in connection with industrial process measuring and control instruments. Experience in the development of automatic control mechanisms an advantage. Age up to 27. Starting salary up to £675 p.a. according to age, qualifications and experience. Location of work London. Post is superannuated under the F.S.S.U. Written applications only quoting "Instruments" to the Personnel Officer, B.I.S.R.A., 11 Park Lane, London, W.1.

Two Electronic Engineers or **Physicists** with interest in radio frequency techniques are required for the Nuclear Physics Section. One post will involve the development of radio frequency equipment for a particle accelerator. Experience of or interest in waveguide and line circuits is essential, and sufficient mathematical ability to apply electromagnetic theory to unusual structures would be useful. The other post is similar but would involve a more fundamental investigation of the behaviour of discharges in a radio frequency field. Apply in writing to Personnel Officer, Associated Electrical Industries Limited, Research Laboratory, Aldermaston Court, Aldermaston, Nr. Reading, Berks.

EXPERIENCED RADIO TESTERS AND INSPECTORS
Required for production of communication and radio apparatus. Also instrument makers, wirers and assemblers for factory test apparatus. Apply Personnel Manager, E. K. Cole, Ltd., Ekço Works, Malmesbury, Wilts.

Ferranti, Ltd., have immediate vacancies for:—

(*a*) Men with electrical engineering qualifications for the advanced testing of naval anti-aircraft Fire Control Equipment involving electronics and servo mechanisms either in factories or on board H.M. ships in home ports. Normal expenses plus a generous allowance when working out.

(*b*) Men with electrical engineering qualifications for production testing of Aircraft Gyroscopic Flight Instruments.

Salary in accordance with age and experience between £356 and £650 per annum. Previous experience though desirable, is not essential.

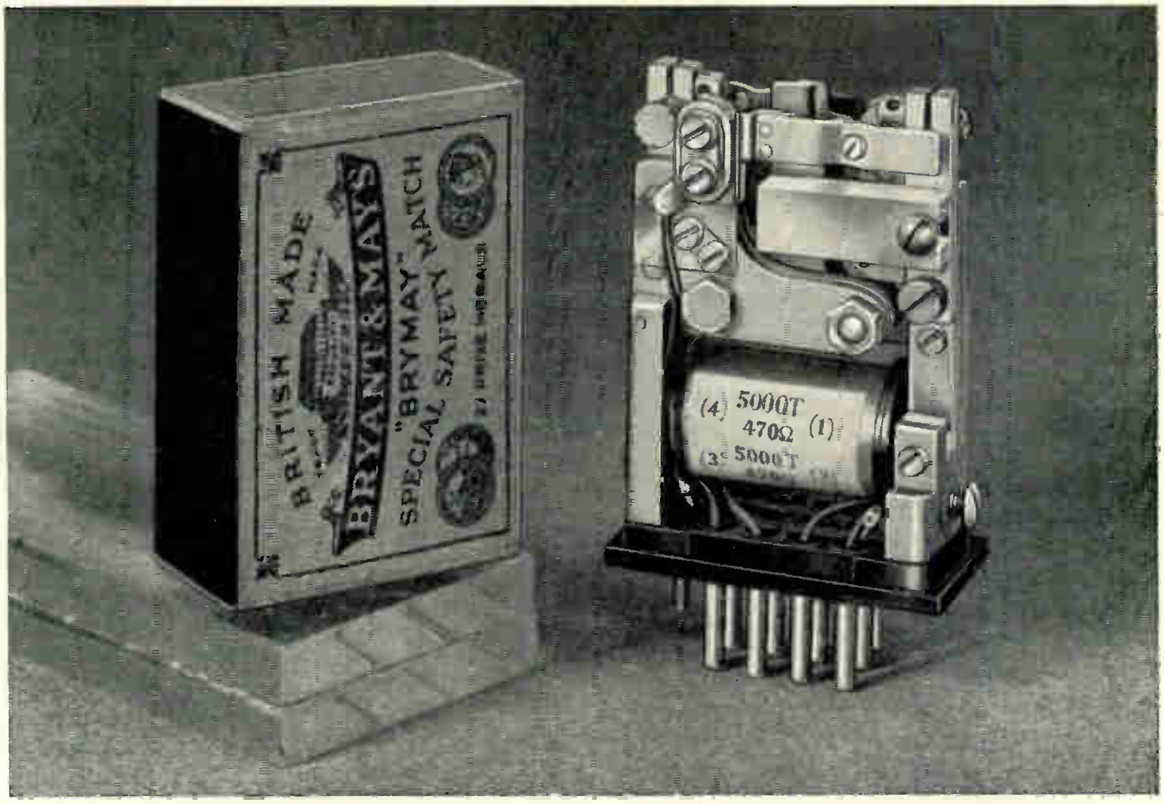
Permanent staff appointments with pension benefits. Forms of application from Mr. R. J. Hebbert, Staff Manager, Ferranti, Ltd., Hollinwood, Lancs. Please quote reference HGN (*a*) or (*b*).

Ericsson Telephones, Ltd., have a number of vacancies in their Research Laboratories in connection with Electronic Switching and Computing. The posts will carry starting salaries between £600 and £1,000 according to age and experience.

(*a*) Electronic circuit designers of all grades. Applicants should have a degree, or equivalent, and several years' experience in the design of electronic circuits. (Reference LE/1.)

(*b*) Electronic equipment engineers with experience in the layout and mechanical design of electronic instruments and equipment in the radio, radar or communication fields. They will be required to co-operate with the circuit designers in the early stages and may ultimately be responsible for development to the production stage. (Reference LE/2.)

Applicants, who should be British born and between 25 and 35 years of age, should write, quoting the reference above and giving details of age, experience, academic or other training, and required starting salary, to the Personnel Officer, Ericsson Telephones, Ltd., Beeston, Nottingham.



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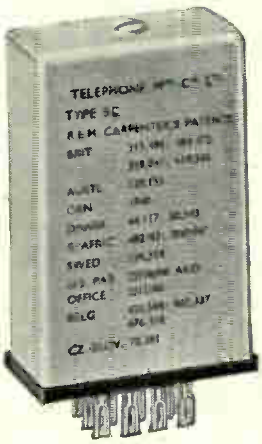

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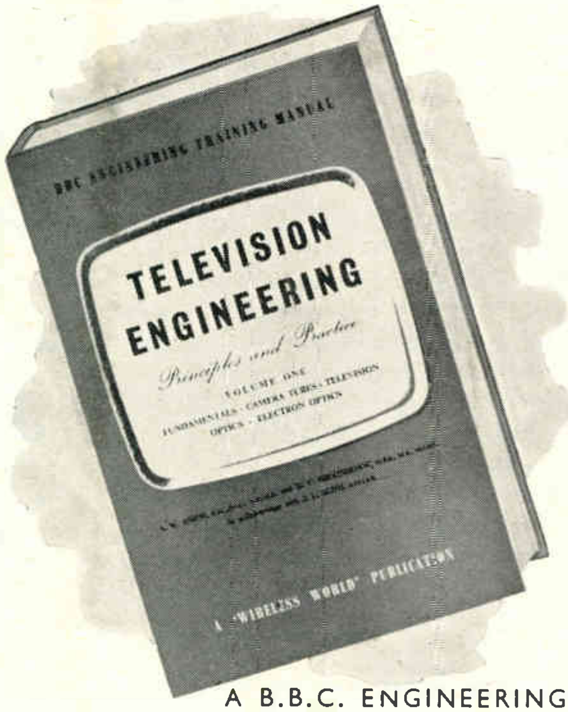

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