WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

APRIL 1956

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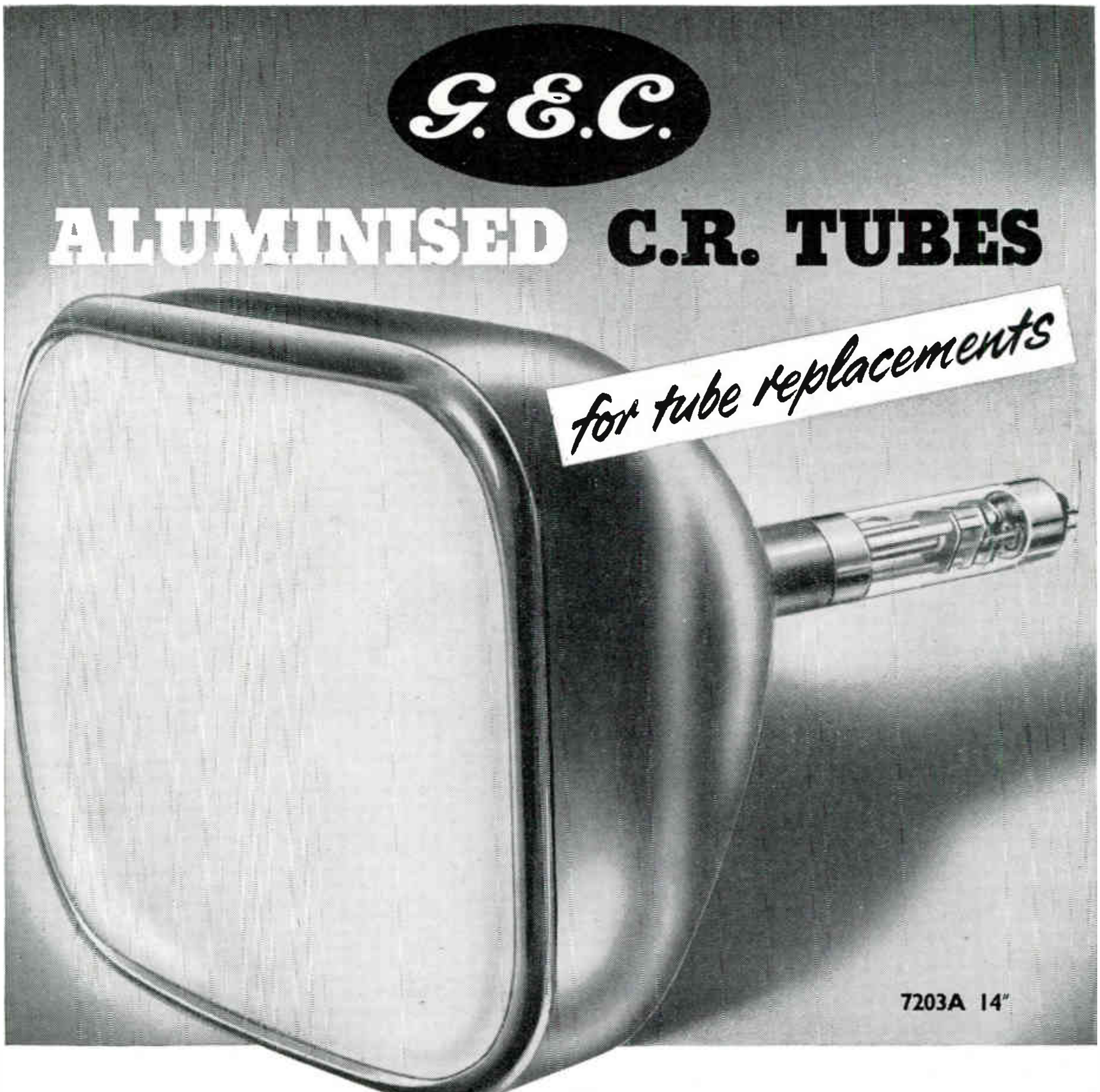

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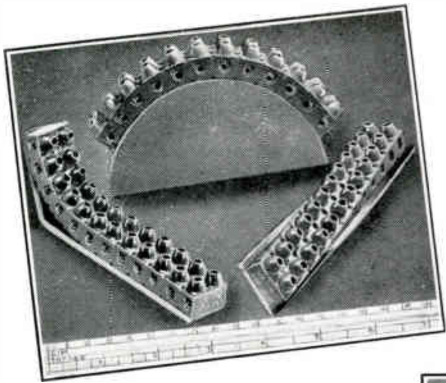

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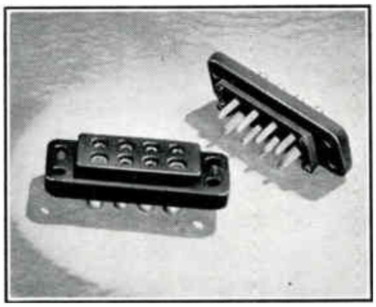

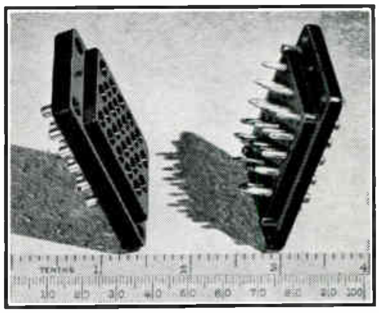

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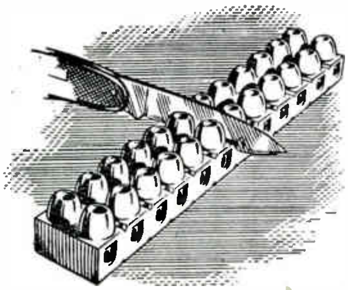

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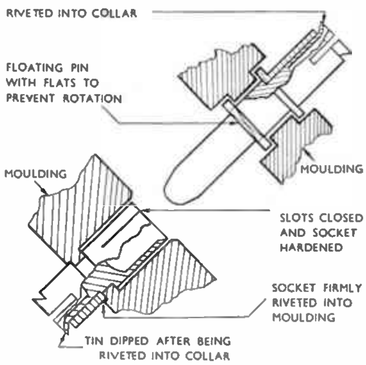

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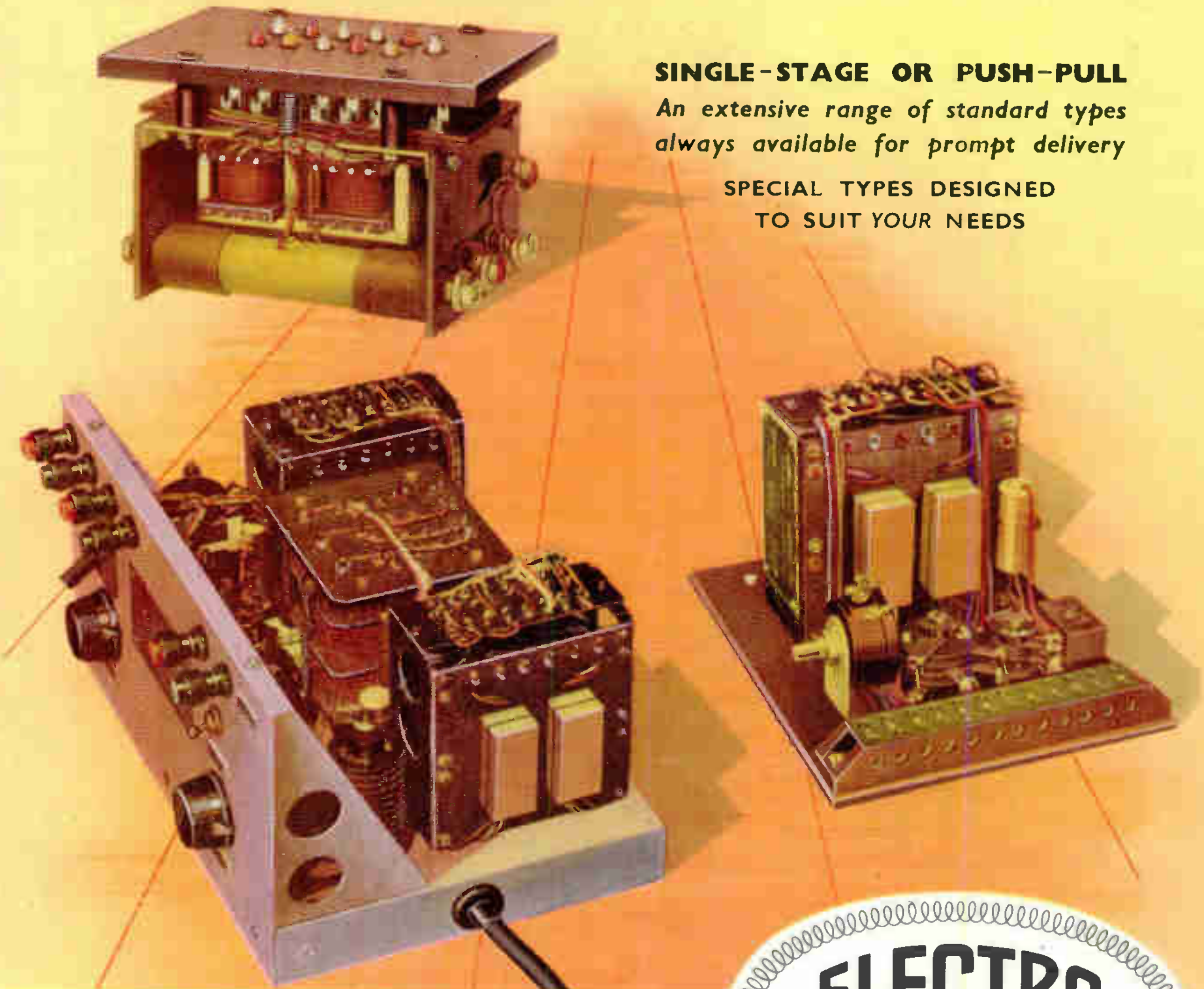

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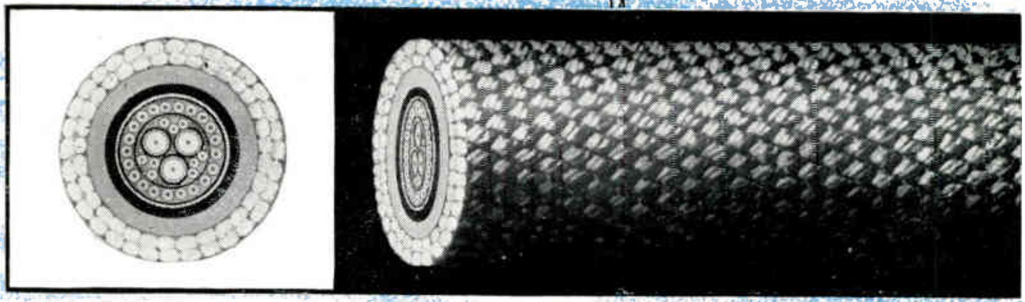

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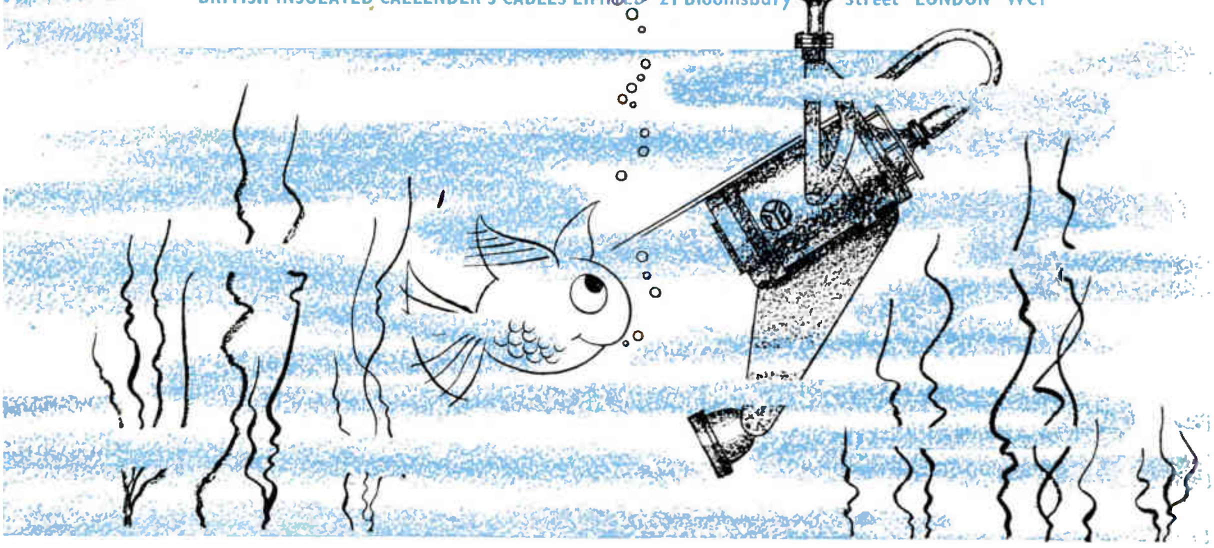

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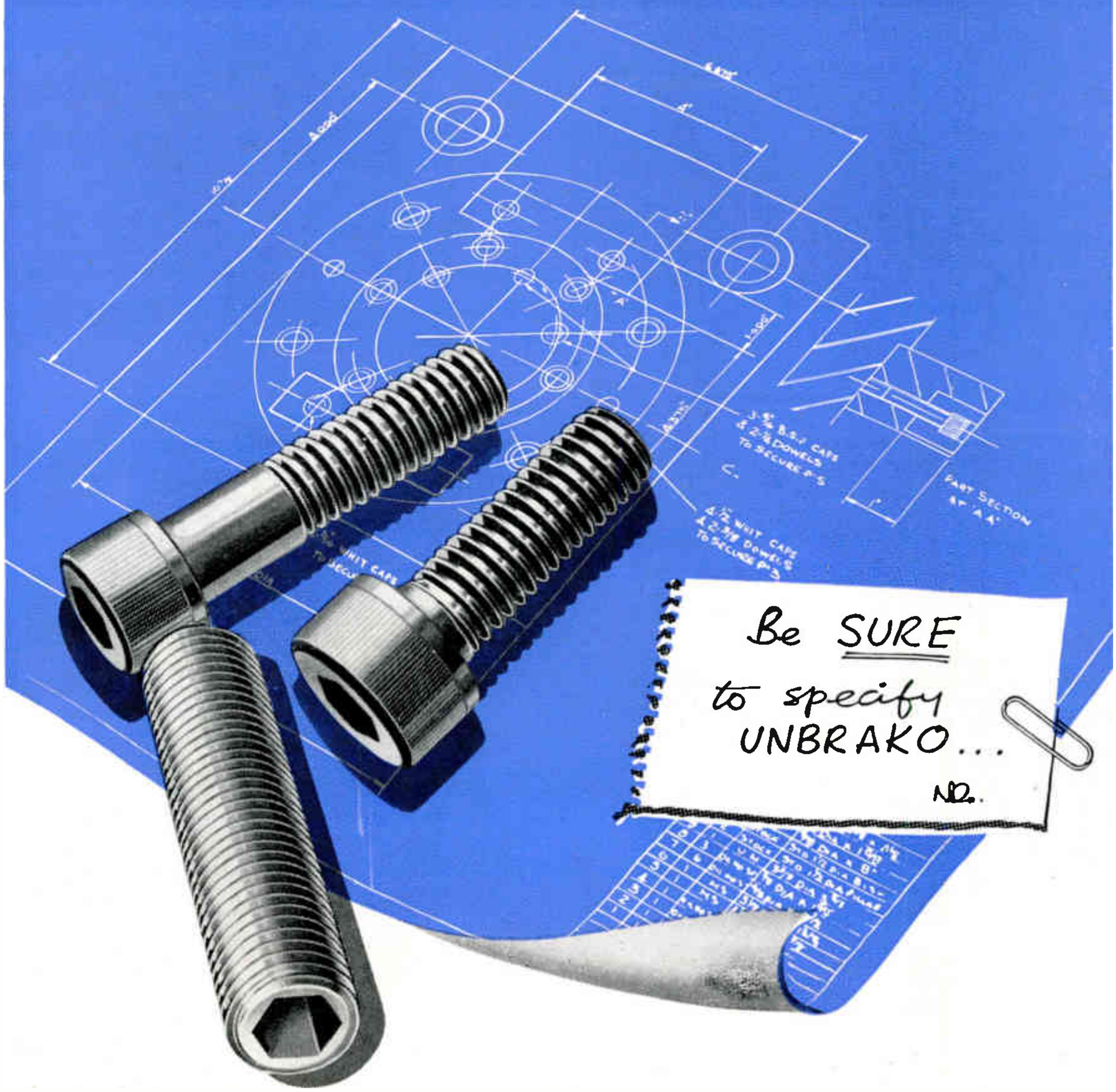

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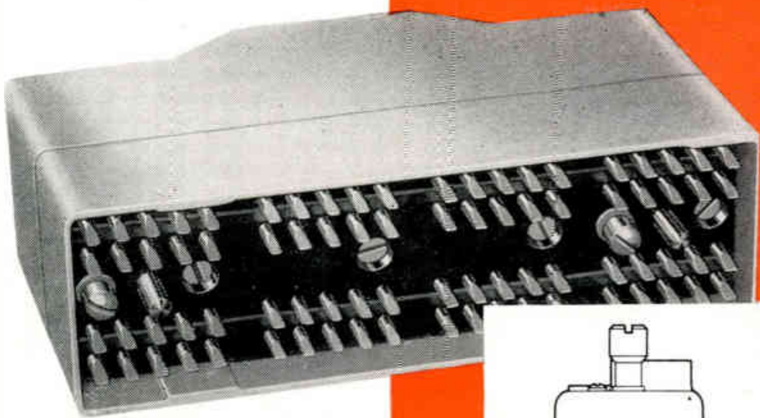

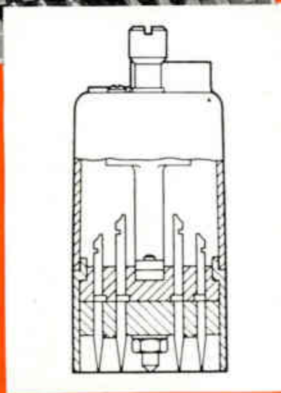

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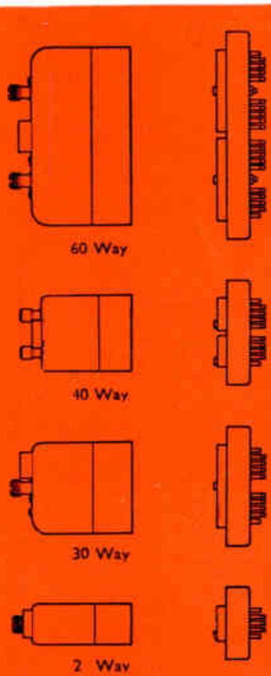

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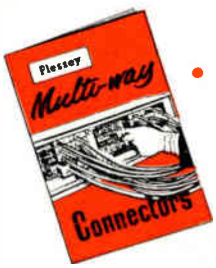

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WIRELESS ENGINEER

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New Rejector Circuit

ONE of the most difficult problems confronting the designer of a television receiver is that of securing adequate rejection of the sound signal without appreciably affecting the response of the video channel. In British television, and with the standard intermediate frequencies of 34·65 Mc/s for vision and 38·15 Mc/s for sound, the requirement is that the response of the vision channel should be at least 40 dB less at 38·15 Mc/s than it is at 37·65 Mc/s. Also, the response at this frequency, which is the upper edge of the video band, should not be down by more than a few decibels compared with that at lower frequencies.

The requisite sharp cut-off can only be obtained by the use of traps, absorbers, acceptors or rejectors, to quote the various names by which these circuits are known. Even then, very high Q in the trap circuit is desirable. Sometimes, bridged-T forms of trap are used; in these a phase balance is possible, so that the equivalent of a circuit of infinite Q can be obtained. Very high rejection is then possible, but the rejector bandwidth becomes exceedingly narrow.

An appreciable departure from the ideal conditions at the edge of the video band can occur in monochrome television without anything very drastic happening to the picture. The conditions for a colour picture are a good deal more stringent, however, for the phase of the colour components can be upset very easily by such defects.

A new form of sound-channel rejector was shown in the circuit of an experimental colour television receiver which has been recently described[1] and it has also been seen in some American colour-receiver circuit diagrams[2]. The overall response curve of the first set is reproduced from the first paper in Fig. 1 and it can be seen that the cut-off at the edge of the pass-band is almost incredibly sharp. Moreover, beyond the point of maximum attenuation, the rise in response is unusually small. These characteristics are largely due to a new form of rejector circuit, of which two are used in this receiver.

No published description of this rejector has yet appeared and it is understood to be due to

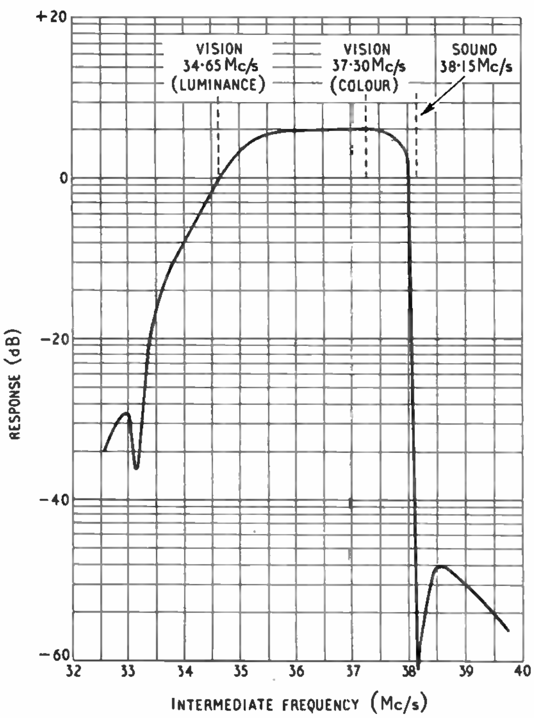

Fig. 1.

R.C.A. It is termed the bifilar-T trap and the essential part of the circuit is shown in Fig. 2. The trap proper is the conventional circuit L_1C_1 but it is connected into circuit via a centre-tapped coil of total inductance $4L$ having a resistance R across one of its windings. The two halves of this coil are bifilar wound to obtain maximum coupling between them.

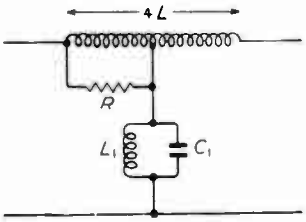

Fig. 2 (left).

The network is a π of parallel-tuned circuits in which the losses of the circuit in the series arm are effectively zero. In addition to this, however, the series arm is peculiar in having both negative inductance and capacitance. Instead of being inductive below resonance and capacitive above like an ordinary tuned circuit, the circuit is inductive above resonance and capacitive below.

This is the property of the bifilar-T circuit which distinguishes it from other rejectors, like the bridged-T, which are also capable of providing infinite attenuation at a particular frequency.

In Fig. 4, each shunt arm has a frequency of parallel resonance given by $\omega^2 = 1/L_1C_1$. The series arm has a frequency of parallel resonance given by $\omega^2 = (1 - L_1/L)/L_1C_1$ and which is, therefore, lower. Considering the series arm and the right-hand shunt arm as a pair of parallel-resonance circuits in series, there is a series-resonance frequency given by $\omega^2 = (1 + L_1/L)/L_1C_1$ and is higher than either of the other two frequencies. Normally, with circuits having positive inductance and capacitance, this series-resonance frequency always lies between the two parallel-resonance frequencies.

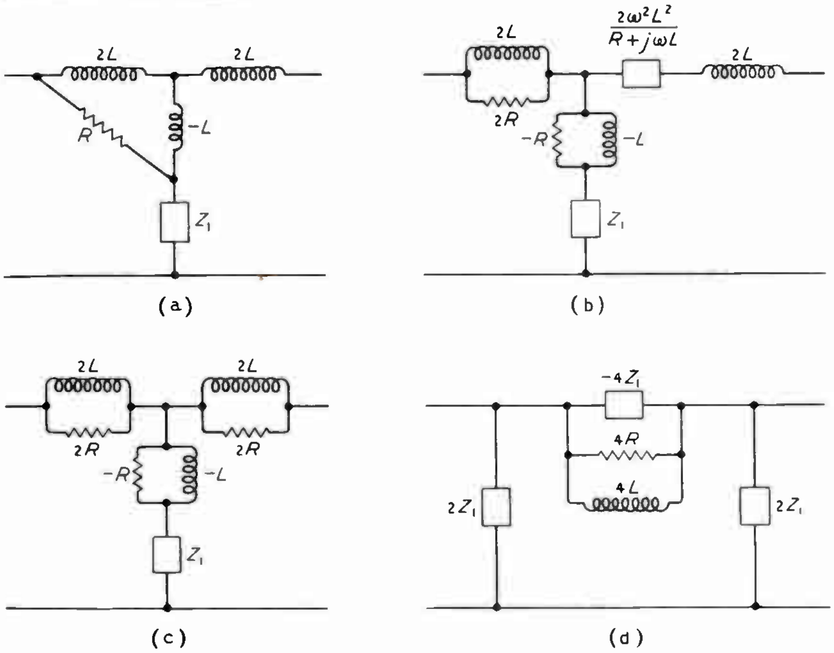

Fig. 3 (left).

Assuming that the coupling is, in fact, 100% and capacitance and resistance effects are negligible, an equivalent circuit has the form of Fig. 3(a) where Z_1 represents L_1C_1. By the application of the star-delta theorem to $2L$, R and $-L$, the circuit can be transformed to (b) and thence, by combining the elements of the right-hand series arm, to (c). The further use of the star-delta theorem enables the form (d) to be obtained.

If Z_1 is a circuit comprising L_1, C_1 and R_1 all in parallel, Fig. 3(d) can be re-drawn as in Fig. 4 and it is plain that the impedance of the series arm becomes infinite when

$$R = R_1$$

and

$$\omega^2 \frac{LL_1}{L - L_1} C_1 = 1$$

so that it is necessary to have $L > L_1$.

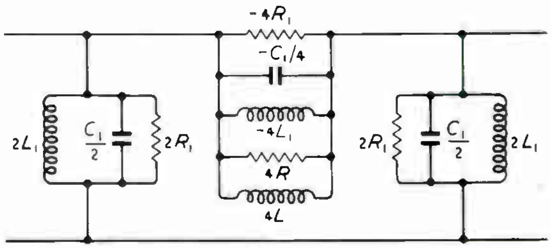

Fig. 4.

In the practical use of the bifilar-T circuit, other impedances are connected to the input and output terminals. The effective parallel-resonant frequencies of the shunt arms are, therefore, modified by such elements. The bifilar-T circuit of Fig. 2 may, for example, be used to couple together a pair of parallel-resonant circuits. In that case, the equivalent circuit for the whole network takes the form of Fig. 5 where $L' < 2L_1 > L''$, $C' > C_1/2 < C''$ and $R' < 2R_1 > R''$. Alternatively, one or both of the terminating impedances may be simple shunt RC combinations, in which case the circuit remains unchanged but the inductance values are different.

In most practical cases, therefore, the complete circuit of an intervalve coupling using a bifilar-T trap has the form of Fig. 5 and so is a pair of tuned circuits with 'top-end' coupling by a lossless trap. The difference between this circuit and others capable of infinite rejection is that the equivalent trap has negative inductance and capacitance.

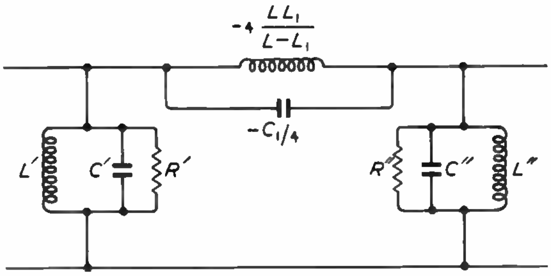

Fig. 5.

A complete and exact analysis is too complex to undertake here but we feel it desirable to draw attention to some of the unusual features of this circuit. The equivalent circuits given here are not, of course, exact, because we have ignored the leakage inductance in the bifilar coil and also its self-capacitance. This is likely to be large, because of the bifilar winding.

W. T. C.

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1 "Experimental Colour Receiver", by H. A. Fairhurst, *Wireless World*, March 1956, p. 112.

2 "Circuitry Report on the Admiral Color TV Receiver", by John Schumacher, *Service*, March 1956, p. 14.

WAVEGUIDE FILTERS

Narrow Band with Quarter-Wave Transformers

By M. H. N. Potok, B.Sc., A.M.I.E.E., A.M.Brit., I.R.E.

(Department of Electrical Engineering, The Royal Technical College, Glasgow)

SUMMARY.—High-Q, low insertion-loss waveguide filters can be made comparatively easily by inserting into the guide a number of suitably spaced, quarter-wave long, dielectric sections.

1. Introduction

THE most common methods of obtaining high-Q waveguide transmission filters are either to couple the output to a high-Q cavity, or to introduce into the guide two inductive slits or apertures just less than half guide wavelength apart. The first method necessitates high precision and careful finish, while the second is very difficult to tune over any but a very narrow band. Both methods result in a high insertion loss[1].

Recent years have seen the development of reflection-type optical filters based on the principle of Fabry Perot interferometer. The method can be applied at microwave frequencies and Culshaw[2] reported some excellent results using for the two reflectors blocks consisting of sheets of polystyrene. The blocks were a quarter-wave thick and were separated by a quarter-wave in air from one another, their inner faces being kept a multiple of half a wavelength apart. His arrangement is reported to lead to a Q of 60,000 when the two blocks are a considerable distance apart but by its very nature it results in considerable insertion loss and is somewhat cumbersome to operate.

MS accepted by the Editor, June 1955

Very good results, in some way better than those obtained in air, can be got by placing the quarter-wave transformers in a waveguide. The following paragraphs develop a simple theory of such a filter based on the impedance and reflection concepts common in waveguide technique.

2. Analysis of Quarter-Wave Transformer Filters

Let λ_g and λ_0 be the waveguide and unbounded space wavelengths respectively when the medium is air, and λ_{gd} and λ_{0d} refer to these wavelengths when the medium is a dielectric of negligible loss and a dielectric constant κ times larger than that of air.

The wave impedance of a guide filled with air is

$$Z_a = \frac{120\pi\lambda_g}{\lambda_0} = \frac{120\pi}{\sqrt{1 - \lambda_0^2/\lambda_c^2}}$$

while when the guide is filled with the dielectric

$$Z_d = \frac{120\pi\lambda_{gd}}{\lambda_{0d}} = \frac{120\pi}{\sqrt{\kappa - \lambda_0^2/\lambda_c^2}}$$

Since, as will be shown, the two types of transformers should have as different impedances as possible, one of them should have air as the medium since this gives the largest possible wave impedance, while the other should have large permittivity κ.

For simplicity all the quantities referring to the air transformer will have a suffix 'a' while those referring to the dielectric will have suffix 'd'. Let the ratio $\frac{Z_d}{Z_a}$ be denoted by $p = \frac{\sqrt{1 - \lambda_0^2/\lambda_c^2}}{\sqrt{\kappa - \lambda_0^2/\lambda_c^2}}$

As an example, for polystyrene ($\kappa = 2{\cdot}56$) when $\lambda_0 = 3{\cdot}14$ cm and is propagated in the H_{10} mode through a standard X-band guide (0·9 in. × 0·4 in.); then $p = 0{\cdot}5$.

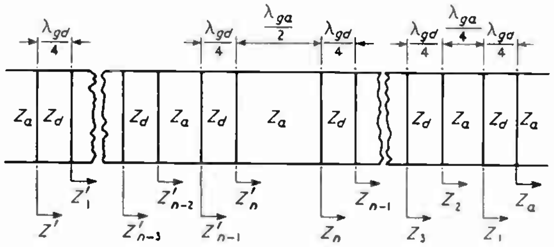

Fig. 1. *Arrangement of filter.*

Taking n quarter-wave transformers in a row so that the first is filled with a dielectric, second with air, etc., as in Fig. 1, the input impedance to the nth section (which is filled with dielectric), when the guide is matched at the output, is $Z_n = Z_d^{n+1}/Z_a^n = Z_a p^{n+1}$. The same impedance appears at the input to the preceding air-filled half-wave transformer (i.e., $Z_n' = Z_n$), hence, if p is small and n large, the half-wave transformer is an efficient cavity.

The frequency-selective nature of this arrangement can best be studied by assuming that λ_0 has been increased by a very small amount $\Delta\lambda$ to λ_1, then the total phase shift in a quarter-wave transformer (referred to λ_0) filled with dielectric d will be

$$\beta_{d1} l_d = \frac{2\pi\sqrt{\kappa\lambda_c^2 - \lambda_1^2}}{\lambda_c\lambda_1} \cdot \frac{\lambda_c\lambda_0}{4\sqrt{\kappa\lambda_c^2 - \lambda_0^2}}$$

$$\approx \frac{\pi}{2} \cdot \frac{\lambda_0}{\lambda_1} \cdot \left[1 - \frac{\lambda_0 \cdot \Delta\lambda}{(\kappa\lambda_c^2 - \lambda_0^2)}\right]$$

and since $(\kappa\lambda_c^2 - \lambda_0^2)$ is very likely to be of the same order of magnitude as λ_0^2 and for Qs in excess of 100, $\Delta\lambda$ will be less than $\lambda_0/100$

$$\beta_{d1} l_d = \frac{\pi}{2} \cdot \frac{\lambda_0}{\lambda_1} \text{ within } 1\%$$

and similarly for air

$$\beta_{a1} l_a = \frac{\pi}{2} \cdot \frac{\lambda_0}{\lambda_1}$$

Hence $\tan \beta_{d1} l_d \approx \tan \beta_{a1} l_a \approx \frac{1}{\pi/2 - \pi/2.\lambda_0/\lambda_1}$

$$= \frac{2\lambda_1}{\pi\Delta\lambda} \approx \frac{2\lambda_0}{\pi\Delta\lambda}$$

Since this is likely to be a large quantity if $\Delta\lambda$ is small, it is more convenient to work with $x = \frac{\pi\Delta\lambda}{2\lambda_0}$ which is small, allowing one to neglect after inspection terms containing x^2 and higher powers of x.

Similarly one can analyse the effect of variation in λ_0 on wave impedances

$$Z_{d1} = \frac{120\pi_{dc}}{\sqrt{\kappa\lambda_c^2 - \lambda_0^2}} = \frac{120\pi\lambda_c}{\sqrt{\kappa\lambda_c^2 - \lambda_0^2 - 2\lambda_0\Delta\lambda - \Delta\lambda^2}}$$

$$\approx Z_d \frac{1}{\sqrt{1 - \frac{2\lambda_0\Delta\lambda}{\kappa\lambda_c^2 - \lambda_0^2}}} \approx Z_d\left(1 + \frac{\lambda_0\Delta\lambda}{\kappa\lambda_c^2 - \lambda_0^2}\right)$$

$$\approx Z_d$$

In the same way, and provided $\Delta\lambda$ is very small $Z_{a1} = Z_a$. Thus the input impedance to the first dielectric transformer is

$$Z_1 = Z_d \frac{Z_a x + jZ_d}{Z_d x + jZ_a} \text{ at } x_1,$$

and by repeated application of this formula and neglecting terms containing x^2

$$Z_n = Z_d \frac{Z_a x(Z_a^n - Z_d^n) + jZ_d^n(Z_a - Z_d)}{Z_d x(Z_a^n - Z_d^n) + jZ_a^n(Z_a - Z_d)}$$

$$= Z_a \frac{px(1 - p^n) + jp^{n+1}(1 - p)}{px(1 - p^n) + j(1 - p)}$$

which simplifies to

$$Z_a\left(p^{n+1} - j\frac{px}{1 - p}\right) \text{ if } p^n \ll 1.$$

Using the same procedure as that outlined in connection with the quarter-wave transformer the phase shift across a half-wave transformer at λ_1 becomes $= \pi\lambda_0/\lambda_1$ and the tangent of this phase shift becomes $= -2x$.

Thus if Z_n is the input impedance to the nth quarter-wave transformer, which is dielectric filled, the input impedance to the air-filled half-wave transformer preceding Z_n is Z_n'

$$Z_n' = Z_a \frac{Z_n - jZ_a 2x}{Z_a - jZ_n 2x}$$

$$= Z_a \frac{[p(1 - p^n) + 2(1 - p)]x + jp^{n+1}(1 - p)}{[p(1 - p^n) + 2p^{n+1}(1 - p)]x + j(1 - p)}$$

$$\approx Z_a\left(p^{n+1} - j\frac{2 - p}{1 - p}x\right)$$

If this section is now preceded by an identical set of quarter-wave transformers to that following the half-wave section, the input impedance to the whole arrangement will be at λ_1

$$Z' = Z_a \frac{Z_n' px(1 - p^n) + jZ_a p^{n+1}(1 - p)}{Z_a px(1 - p^n) + jZ_n'(1 - p)}$$

$$= Z_a \frac{xp^{n+2} + jp^{n+1}(1 - p)}{2x + jp^{n+1}(1 - p)}$$

Since the reflection coefficient $\rho = \frac{Z' - Z_a}{Z' + Z_a}$ and at half-power frequency $|\rho|^2 = \frac{1}{2}$ one obtains

$$|\rho|^2 = \frac{4x^2}{4x^2 + 4p^{2n+2}(1-p)^2} = \frac{1}{2}$$

hence

$$x^2 = p^{2n+2}(1-p)^2$$

therefore

$$x = p^{n+1}(1-p)$$

Since

$$x = \frac{\pi\Delta\lambda}{2\lambda_0} = \frac{\pi}{4Q}$$

i.e., $Q = \frac{\pi}{4p^{n+1}(1-p)}$

If one follows the previous example since

$$p = 0{\cdot}5,\ Q = \frac{\pi}{4 \times 0{\cdot}5^{n+1} \times 0{\cdot}5} = 2^n\pi$$

leading to Table 1.

TABLE 1

n	7	9	11	13	15
Q	400	1,600	6,400	25,600	102,400

A further substantial increase in Q can be obtained by making the centre section equal to an integral multiple of half-wavelengths. When the number of half-wavelengths is m, it can be shown by following the earlier procedure that

$$Q = \frac{\pi[m(1-p) + p]}{4p^{n+1}(1-p)}$$

hence if

$$p = 0{\cdot}5,\ Q = \frac{\pi(m+1)}{4p^{n+1}} = 2^{n-1}\pi(m+1)$$

Consider now the effect of an error in the lengths of the transformers such that the phase shift over each transformer is less than $\pi/2$ by a small angle x_1 so, that at λ_0 the input impedance

$$Z_n = Z_a\left(p^{n+1} - j\frac{px_1}{1-p}\right)$$

Reference to a Smith's Chart will make it clear that if the input impedance to the whole arrangement is to be $Z' = Z_a$ then

$$Z'_n = Z_a\left(p^{n+1} + j\frac{px_1}{1-p}\right)$$

hence the half-wave transformer section must be made longer than π by a small angle $\frac{2px_1}{1-p}$. Let now λ_0 increase to λ_1, then the phase shift across each quarter-wave transformer will be further reduced by a small angle x_2 while the phase shift across the half-wave transformer is reduced by $2x_2$.

Hence at λ_1

$$Z_n = Z_a\left[p^{n+1} - j\frac{p(x_1 + x_2)}{1-p}\right]$$

$$Z_n' = Z_a\frac{Z_n + jZ_a\left(2x_1\frac{p}{1-p} - 2x_2\right)}{Z_a + jZ_n\left(2x_1\frac{p}{1-p} - 2x_2\right)}$$

$$= Z_a\frac{p^{n+1} - j\frac{p(x_1+x_2)}{1-p} + j\frac{2}{1-p}(x_1p + x_2p - x_2)}{1 + j\left(p^{n+1} - j\frac{px_1 + px_2}{1-p}\right)\frac{2}{1-p}(x_1p + x_2p - x_2)}$$

$$\approx Z_a\left[p^{n+1} + j\frac{p(x_1+x_2) - 2x_2}{1-p}\right]\left[1 - j\frac{2p^{n+1}}{1-p}(px_1 + px_2 - x_2)\right]$$

$$\approx Z_a\left[p^{n+1} + j\frac{p(x_1+x_2) - 2x_2}{1-p}\right]$$

and

$$Z' = Z_a\frac{Z_n'p(x_1+x_2) + jZ_ap^{n+1}(1-p)}{Z_ap(x_1+x_2) + jZ_n'(1-p)}$$

$$= Z_a\frac{p^{n+2}(x_1+x_2) + jp^{n+1}(1-p)}{2x_2 + jp^{n+2}(1-p)}$$

$$\approx Z_a\frac{p^{2n+2}(1-p)^2 + jp^{n+1}(1-p)2x_2}{4x_2^2 + p^{2n+2}(1-p)^2}$$

which is seen to be of the same form as original Z', hence it follows that provided the error is very small there will be no reduction in Q.

The same argument can be used to prove that the arrangement can be tuned over a narrow range of frequencies by varying the length of the central half-wave transformer.

The above discussion assumes no losses in dielectric and waveguide. In practice, such a condition is unobtainable and the actual Q values will be considerably reduced. To check the results, some simple experiments were carried out.

3. Experimental Evidence

The quarter-wave transformers were made out of polystyrene sheet machined down to 0·214 in. corresponding to quarter-wavelength in polystyrene ($\kappa = 2{\cdot}56$) in a 0·9-in. × 0·4-in. guide when $\lambda_0 = 3{\cdot}14$ cm ($f = 9{,}550$ Mc/s). The sheet was cut up into small sections fitting closely into the guide, the sections being held together 0·425-in. apart ($\lambda_g/4$ in air-filled guide) by two very thin strips of polystyrene. The drawn-copper waveguide had a narrow slit in the top

B

face so that the separation between the two sets of transformers could be adjusted to the desired frequency. The graph of Fig. 2 gives the plot of log Q against the number of polystyrene sections used ($= n + 1$ in the previous notation) when the centre transformer is half-wave long. Since the theoretical formula assumes $Q \nless 100$ the values corresponding to four and six polystyrene sections have been calculated directly step by step. It is seen that the experimental curve is considerably below the theoretical and it appears to tend to a limit around $Q = 10{,}000$. Lack of suitable equipment did not allow the extension of the experiment beyond 14 sections. The overall length of the filter was $9\frac{1}{2}$ in.

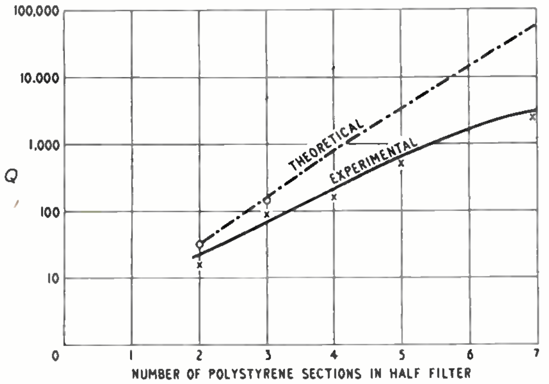

Fig. 2. *Theoretical (assuming no losses) and experimental relation between Q of filter and number of dielectric sections.*

In addition, measurements were taken with 14 polystyrene sections at wavelengths outside 3·14 cm giving the results of Table 2.

TABLE 2

f	9,407	9,547	9,602 Mc/s
Q	2,350	2,700	2,300

This shows that Q does tend to fall off below and above design-centre frequency, nevertheless the drop is not excessive. By doubling the length of the centre transformer from half to full wave-length the Q was found to increase from 2,700 to 3,400, an increase of 26% in place of the 50% estimated theoretically. The differences may be accounted by losses in polystyrene and guide walls. The insertion loss of the filter was less than 2 dB.

4. Conclusions

It has been shown that high-Q low-loss transmission filters can be obtained by the comparatively simple method of building up a chain of alternately high and low quarter-wave transformers. The arrangement is tunable over a narrow range of frequencies around the design centre.

The Q estimated on the basis of no losses is given by

$$Q = \frac{\pi}{4} \cdot \frac{m(1-p)+p}{p^{n+1}(1-p)}$$

where

p = ratio of dielectric to air wave impedances

$n + 1$ is the total number of dielectric sections used in the whole of the filter

m = number of half-wave air transformers in the centre of the filter

Since Q increases as p decreases, p must be as small as possible and the use of waveguide helps to reduce it appreciably.

If one considers losses in dielectrics and guide walls then the resultant Q values will be much reduced but they are still found to be large and in fact considerably larger than those obtainable with an identical set-up in air or in a coaxial cable of uniform structure (in air, the Q of a 14-polystyrene-section filter with half-wave transformer at centre and assuming no losses, is 1,550 compared with the experimentally obtained 2,700 in a waveguide).

Further work is contemplated; in particular, to study the effect of low-loss waveguides and higher permittivity low-loss dielectrics, such as quartz.

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CLUTTER ON RADAR DISPLAYS

Reduction by use of Logarithmic Receivers

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SUMMARY.—An analysis is developed of the action of an idealized logarithmic receiver followed by a differentiating circuit (high-pass filter), upon inherent receiver noise, sea-clutter and rain-clutter echoes. The analysis is extended to estimate the extent to which the performance of the practical logarithmic receiver may depart from that of the idealized receiver. Results are given of experiments with logarithmic receivers on both S and X band radars. The loss which occurs when a differentiating circuit follows a logarithmic receiver is stated, the cause examined and a method of minimizing the loss suggested. The important design parameters of a logarithmic receiver for clutter reduction are dealt with.

1. Introduction

THE investigations set forth in this paper have been carried out by the author, using both S and X band radars at different times in recent years. The objective has been to examine the effectiveness of logarithmic receivers, followed by differentiating circuits (high-pass filters), in reducing the intensity of sea and rain clutter interference on radar displays[2] to that of the normal inherent noise background. A theoretical treatment of the subject is given, which is followed by the collected results of the various experiments. Some of the effects met with, and the conditions which must be observed for a worth-while measure of success, were at first difficult to understand. As the work proceeded a reasonably clear picture of the action of the logarithmic receiver and differentiating circuit on noise and clutter did emerge, and enabled the correct parameters for these and the video amplifier feeding the display to be established.

Against sea clutter the most that can be expected from theoretical considerations of the logarithmic receiver and differentiating circuit is that it should be as effective as a so-called linear receiver followed by a perfectly-adjusted swept-gain or sensitivity-time control (s.t.c.). It should be noted here that the perfectly-adjusted swept-gain control is an ideal which cannot be realized in practice even with a highly-skilled operator. In the first place, the waveform of the control voltage, which must be applied to give perfect suppression of the clutter envelope to noise level, cannot be attained practically even for a single restricted arc of azimuth and, since the clutter envelope varies significantly with azimuth, only a compromise setting of the swept-gain control can be used for the whole three hundred and sixty degrees of azimuth. The logarithmic receiver may in theory, therefore, be expected to achieve automatically a little more than even a skilled operator can produce with the swept-gain control. Against rain clutter the swept-gain principle cannot be applied, and the only existing measure in Civil Marine Radars is to follow the linear receiver by a differentiating or fast time-constant circuit (f.t.c.). This has the effect of reducing the intensity of the clutter background and enabling large targets to be seen within it. It is expected that by similar use of a logarithmic receiver the rain clutter background would be rendered indistinguishable from the inherent noise.

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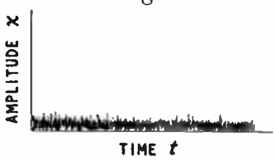


Fig. 1 (above). *Noise along a single range trace.*

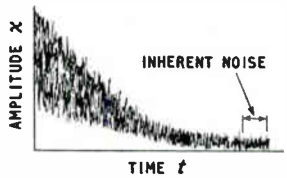


Fig. 2 (right). *Sea clutter along single range trace.*

2. Properties of Receiver Noise and of Sea Clutter

Fig. 1 represents receiver noise, and Fig. 2 sea clutter, along a single range trace of an A-scope display. The ordinate scale represents amplitude (x), and the abscissae range, which is proportional to time (t).

2.1. *Properties along a Single Range Trace*

The properties of noise and sea clutter will first be considered along any single range trace. Let the r.m.s. value of the amplitude x be represented by σ_0. Considering noise first, we have that the intensity is roughly constant (i.e., $\sigma_0 =$ constant); that the auto-correlation in amplitude extends over an interval approximately equal to $1/B$ where B is the receiver bandwidth; and that the amplitude x is distributed according to a Rayleigh probability distribution.

Taking now the case of sea clutter; there is a general decrease in intensity of clutter with increasing range (i.e., σ_0 falls); the auto-correlation extends over an interval T (where T = radar pulse length); and the amplitude x is not distributed according to a Rayleigh distribution.

2.2. Properties 'across' a Sequence of Range Traces at any Fixed Range

The properties are now considered at a fixed element of range from one range trace to the next and so on. This may be considered as moving 'across' or perpendicular to the range traces. In the case of noise, we have that the intensity is again constant (i.e., σ_0 = constant); there is no correlation from trace to trace; and that the amplitude x is distributed according to a Rayleigh probability distribution.

In the case of sea clutter, the intensity is roughly constant (i.e., σ_0 = constant); the correlation extends over a finite number of traces for a stationary aerial and may extend over the beam-width of an aerial for a rotating aerial; and the amplitude x is distributed roughly according to a Rayleigh probability distribution.

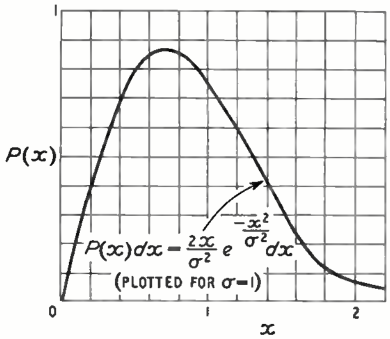


Fig. 3. *Curve of Rayleigh distribution of x plotted for scale factor $\sigma = 1$.*

2.3. Differences between Noise, Sea Clutter and Rain Clutter

From the foregoing, the important differences between noise and sea clutter are twofold. First, there is the variation in intensity of sea clutter along a range trace, which will result in saturation of a p.p.i. at short ranges, if the biasing and limiting levels are adjusted for long ranges where only noise is present. Secondly, the grain size of a sea-clutter background is usually coarser than that of a noise background, due to the trace-to-trace correlation of clutter, and any excess of T over $1/B$.

Given equality between T and $1/B$ the only observable difference between noise and rain clutter lies in the variation of intensity of rain clutter with range. This can take any form depending upon the variation of rainfall rate within the storm area.

The object of using a logarithmic receiver followed by a differentiating circuit is to eliminate this variation in intensity with range of both sea and rain clutter.

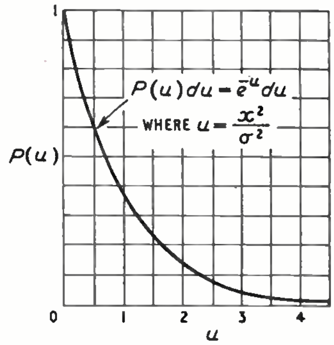


Fig. 4. *General curve representing Rayleigh distributions of x for all scale factors σ.*

3. Theory of the Logarithmic Receiver

3.1. The Idealized Logarithmic Receiver

It has been stated above that the amplitudes of inherent receiver noise, sea clutter, and rain clutter all obey more or less a Rayleigh probability distribution (after detection) when measured at a fixed range from trace to trace.

The Rayleigh probability distribution may be written as:

$$P(x)\, dx = (2x/\sigma^2) \exp. (-x^2/\sigma^2)\, dx, \quad x > 0 \qquad (1)$$

In this equation, x is the instantaneous amplitude and σ is the r.m.s. fluctuation about the mean (or scale factor).

By substituting a new variable:

$$u = x^2/\sigma^2, \quad u > 0 \quad .. \quad .. \quad .. \qquad (2)$$

the distribution can be transformed to:

$$P(u)\, du = \exp. (-u)\, du, \quad u > 0 \quad .. \quad .. \qquad (3)$$

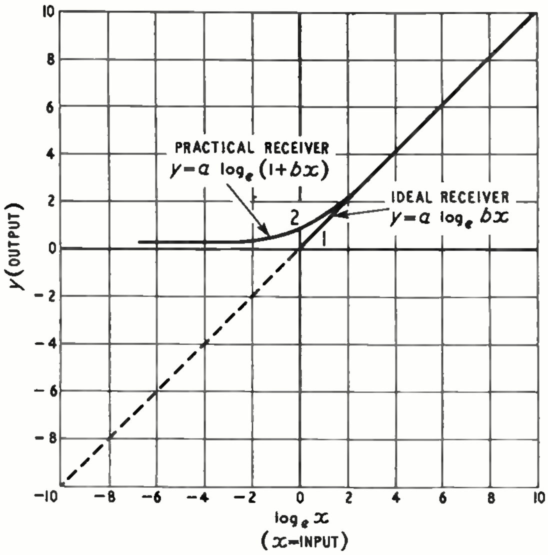


Fig. 5. *Curves representing laws for the ideal and practical logarithmic receivers. Constants 'a' and 'b' taken as unity for convenience in plotting.*

The original distribution (Equ. 1) is shown in Fig. 3 and the transformed distribution (Equ. 3) is shown in Fig. 4. It may be noted in passing that Equ. (3) is independent of σ, and Fig. 4 therefore represents all Rayleigh distributions regardless of the scale factor, σ.

We now consider an idealized logarithmic amplifier in which the input, x, is related to the output, y, by the law:

$$y = a \log_e(bx) \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

where a and b are constants of the amplifier. This law is shown graphically by curve (1) of Fig. 5, the constants a and b being assumed unity for convenience of plotting. It can be seen that as the input falls below unity ($x < 1$, $\log x < 0$) the output assumes negative values down to minus infinity (dotted part of characteristic).

Equ. (4) may be written as:

$$y = (a/2) \log_e(b^2x^2) \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

which, by substitution for x from Equ. (2), becomes:

$$y = (a/2) \{\log_e(b^2\sigma^2) + \log_e u\} \quad \ldots \quad (6)$$

If a random input signal of amplitude distribution $P(u)\,du$ (Equ. 3) is applied to the receiver input, the resulting probability distribution, $P(y)\,dy$, at the output of the amplifier will have the following parameters:

Mean value

$$= E(y) = \int_0^\infty (a/2)[\log b^2\sigma^2 + \log u]\, P(u)\, du \quad (7)$$

$$= (a/2) \left\{\log b^2\sigma^2 + \int_0^\infty (\log u)\, P(u)\, du\right\} \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

since $\int_0^\infty P(u)\,du = 1$ from probability definitions.

Similarly the mean-squared value of y, $E(y^2)$ is given by:

$$E(y^2) = \int_0^\infty (a^2/4)\left[\log b^2\sigma^2 + \log u\right]^2 P(u)\, du$$

$$= (a^2/4) \left\{(\log b^2\sigma^2)^2 + 2 \log b^2\sigma^2 \int_0^\infty (\log u) P(u)\, du + \int_0^\infty (\log u)^2\, P(u)\, du\right\} \quad \ldots \quad (9)$$

Hence the variance of y is given by:

$$E(y^2) - \{E(y)\}^2 = (a^2/4)\left\{\int_0^\infty (\log u\,)^2 P(u)\, du - \left[\int_0^\infty (\log u)\, P(u)\, du\right]^2\right\} \quad \ldots \quad \ldots \quad \ldots \quad (10)$$

Inserting the value of $P(u)\,du$, given for all Rayleigh distributions by Equ. (3) as exp. $(-u)\,du$, we have:

$$E(y^2) - \{E(y)\}^2 = (a^2/4)\left\{\int_0^\infty (\log u)^2 \exp.(-u)\, du - \left[\int_0^\infty (\log u) \exp.(-u)\, du\right]^2\right\} \quad \ldots \quad \ldots \quad (11)$$

The integrals can be found in tables of Laplace transforms, and Equ. (11) then reduces to:

$$E(y^2) - \{E(y)\}^2 = (a^2/4).(\pi^2/6) \quad \ldots \quad (12)$$

We have, therefore, the result that if any Rayleigh distribution is applied to the input of a receiver with an idealized logarithmic input–output law, the resulting r.m.s. fluctuation about the mean at the receiver output is constant, independent of the r.m.s. amplitude of the input fluctuations, and given by $(a\pi)/(24)^{\frac{1}{2}}$ where a is the slope of the logarithmic law. Fig. 6 illustrates this constant r.m.s. fluctuation of the clutter and noise signals about a varying mean level at the output of the logarithmic receiver, and in Fig. 7 the mean level has been removed by a high-pass filter, leaving the clutter and noise fluctuations at a constant r.m.s. amplitude across the display trace.

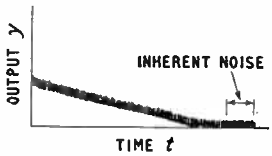


Fig. 6 (left). *Clutter output from logarithmic receiver before passage through high-pass filter.*

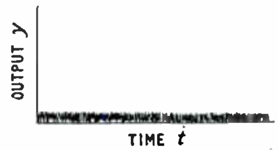


Fig. 7 (right). *Clutter output after removal of mean level by high-pass filter.*

3.2. *Practical Logarithmic Receiver*

A receiver with an idealized logarithmic characteristic (curve 1, Fig. 5) is clearly unattainable in practice; the practical logarithmic receiver having a law given by:

$$y = a \log_e (1 + bx) \quad \ldots \quad \ldots \quad (13)$$

shown as curve 2 in Fig. 5.

As the value of the input x becomes small, the law of the practical receiver begins to depart significantly from that of the ideal receiver; the output y of the practical receiver being zero for zero input, and the corresponding output of the ideal receiver being minus infinity. The above analysis based upon the ideal logarithmic law will therefore involve some error, the lower limits of the integrals in Equ. (11) corresponding to outputs of minus infinity which are impossible with the practical receiver. Attempts at an exact analysis for the practical receiver using Equ. (13) rapidly lead to intractable mathematics. However, the error resulting from the use of the idealized law will depend upon the extent to which the negative-going outputs of the ideal receiver contribute to the result of Equ. (12).

This error can be estimated in the following way.

Equ. (4) gives negative values to y when $bx < 1$ and, to avoid this, we define a new variable Y as follows:

$$Y = a \log (bx) \text{ when } bx \geqslant 1$$
$$Y = 0 \qquad \text{ when } bx < 1 \qquad \ldots \quad (14)$$

Let $u = z$ when $bx = 1$. From Equ. (2) $u = x^2/\sigma^2$, therefore:

$$z = 1/(b^2\sigma^2) \text{ or } b^2\sigma^2 z = 1 \qquad \ldots \quad \ldots \quad (15)$$

Now

$$E(Y) = \int_z^\infty (a/2) \log (b^2\sigma^2 u)\, P(u)\, du$$
$$= E(y) - \int_0^z (a/2) \log (b^2\sigma^2 u)\, P(u)\, du \quad (16)$$

Similarly

$$E(Y^2) = E(y^2) - \int_0^z (a^2/4)(\log b^2\sigma^2 u)^2\, P(u)\, du \quad (17)$$

When z is sufficiently small $\exp(-u) \approx 1 - u$, $(0 \leqslant u \leqslant z)$. Also the upper limit of u in the integrals of Equs. (16) and (17) is given by

$$z = 1/(b^2\sigma^2) = 0{\cdot}01$$

(assuming that the r.m.s. fluctuation of noise extends over 20 dB of the receiver characteristic). We can thus put $\exp(-u) \approx 1{\cdot}0$ as a first approximation, and therefore $P(u)$ can be written as unity in Equs. (16) and (17). Doing this, and changing the variable to $v = b^2\sigma^2 u$, we get:

$$(a/2) \int_0^z \log (b^2\sigma^2 u)\, P(u)\, du$$
$$= (a/2b^2\sigma^2) \int_0^1 \log v\, dv$$
$$= (a/2b^2\sigma^2) \Big[v (\log v - 1) \Big]_0^1$$
$$= -(a/2b^2\sigma^2) \qquad \ldots \quad (18)$$

and

$$(a^2/4) \int_0^z (\log b^2\sigma^2 u)^2\, P(u)\, du$$
$$= (a^2/4b^2\sigma^2) \int_0^1 (\log v)^2\, dv$$
$$= (a^2/4b^2\sigma^2) \Big[v (\log v - 1)^2 + v \Big]_0^1$$
$$= (2a^2/4b^2\sigma^2) \qquad \ldots \quad (19)$$

Substituting these results in Equs. (16) and (17) we get:

$$\text{Variance of } Y = E(Y^2) - \{E(Y)\}^2$$
$$= E(y^2) - \{E(y)\}^2 - (a^2/2b^2\sigma^2) - (a/b^2\sigma^2)\, E(y) - (a/2b^2\sigma^2)^2 \qquad \ldots \quad \ldots \quad (20)$$

$$= (a^2/4)\, \{\pi^2/6 - (2/b^2\sigma^2)\, [1 + \log b^2\sigma^2 - \gamma + 1/(2b^2\sigma^2)]\} \quad (21)$$

the integrals for $E(y)$ and $E(y^2)$ being found in tables of Laplace transforms. (γ = Euler's constant $\approx 0{\cdot}577$.)

The correction factor to be subtracted from the multiplying constant, $(\pi^2/6)$, in Equ. (12) because of the inability of the practical receiver to allow negative outputs, is therefore given by:

$$(2/b^2\sigma^2)[1 - \gamma + 1/(2b^2\sigma^2) + \log b^2\sigma^2] \quad \ldots \quad (22)$$

If the receiver is logarithmic to about 20 dB below r.m.s. noise level, we can take $b\sigma = 10$ for noise only, and the correction term will be of magnitude 0·1.

For clutter signals which may be as great as 80 dB above r.m.s. noise, σ^2 is greater by a factor of 10^8 and the correction term then becomes negligible. It is thus concluded that, as the noise plus clutter varies over a range of 80 dB, the variance of the output Y will vary from $(a^2/4)[\pi^2/6 - 0{\cdot}1]$ to $(a^2/4).(\pi^2/6)$, that is about 1 part in $16{\cdot}5 \approx 0{\cdot}25$ dB. Hence the variance of Y can be taken as sensibly independent of σ.

This result applies to a receiver having an ideal logarithmic law from zero output upwards, and in which the input r.m.s. noise fluctuations extend over the first 20 dB of the logarithmic characteristic. The full part of curve 1 in Fig. 5 shows the characteristic, and the curve for the practical receiver (curve 2, Fig. 5) departs from the desired characteristic for low values of the input x. However, by designing the practical receiver so that the constant b (gain) is very large, all the noise fluctuations within minus 20 dB of the r.m.s. noise level (99%) can be sensibly lifted on to the ideal characteristic (curve 1), and the above result is then closely applicable.

Before leaving these analytical considerations, it is worth noting that Equ. (10) is quite general for the logarithmic receiver and that if we insert for $P(u)\, du$ any distribution function of u which is independent of σ the resulting variance will be a constant. In practice, this means that the logarithmic receiver would be equally effective against any randomly varying input signal in which the amplitude x fluctuates over *positive* values according to a type of distribution in which the probability of x lying between x and $x + dx$ is determined by a scale factor σ; i.e.,

$$P(x)\, dx = P^*(x/\sigma)\, dx/\sigma, \quad x > 0.$$

Such a distribution can be transformed by the substitution $u = x^2/\sigma^2$ to a distribution of the type $P(u)\, du$. As shown in Equs. (1) to (3), the Rayleigh distribution is one example of such a probability function; there may be others.

4. Experiments with Logarithmic Receivers

4.1. *Limits of the Logarithmic Input–Output Law, and their Achievement*

It has been shown in Section 3 that a logarithmic receiver followed by differentiation can in theory reduce the r.m.s. clutter level to the r.m.s. level of inherent noise, provided that the logarithmic law extends at its upper limit to the highest level of clutter fluctuations (usually not more than 80 dB above r.m.s. noise) and at its lower limit to the deep troughs of the clutter and noise fluctuations. The extent to which the receiver should remain logarithmic below r.m.s.

noise level is of cardinal importance, and it can be stated at once that if the input–output characteristic ceases to be logarithmic at the r.m.s. noise level, little or no reduction in the intensity of the clutter background on a p.p.i. will be obtained in comparison with a linear receiver. However, as the logarithmic law is progressively extended below r.m.s. noise level, the disparity in intensity between the clutter background and the noise background is at first rapidly reduced, and then more slowly, until the logarithmic characteristic has been extended to a depth of about 20 dB below r.m.s. noise, beyond which no noticeable improvement appears to be realized. This is understandable since the probability of any fluctuation of the Rayleigh distribution reaching a value less than 20 dB below the r.m.s. level is only about 1%.

It is essential therefore to maintain the logarithmic characteristic uniformly down to a level of approximately 20 dB below the r.m.s. noise level. It is inadvisable to maintain the logarithmic characteristic to levels lower than 20 dB below r.m.s. noise, since nothing perceptible is gained in clutter reduction, and the extent to which the receiver is logarithmic above r.m.s. noise level is likely to be reduced as a consequence. This could be serious if the upper limit of the clutter fluctuation then lay beyond the upper limit of the logarithmic range of the receiver.

Methods of achieving a logarithmic response are covered by references one, two and three. All the experiments here reported were conducted with receivers constructed by the method of reference three. Certain important aspects of the design parameters covered in Section 5 are intended to supplement the material of reference three.

4.2. *Results on an S-Band Naval Radar*

The pulse length of the radar set was 2 μsec and the receiver constructed for the experiments had a bandwidth of 1 Mc/s. The receiver incorporated the facility that it could be switched at will to be either logarithmic or linear. The receiver was followed by a simple resistance-capacitance differentiating circuit (high-pass filter) to remove the mean value of the clutter and noise fluctuations.

The so-called 'linear' receiver has in fact a linear relationship between input and output signal only over a small range of output voltage. For instance, with a receiver of 10-Mc/s bandwidth, the relationship remains roughly linear for output signals up to about 18 dB above the normal level of output noise, thence upwards to about 30 dB above noise the relationship is roughly logarithmic, after which the receiver saturates, all further amplitude discrimination between signals being lost.

The effect of following the linear receiver by a differentiating circuit is to remove the mean level of the clutter and noise fluctuations at its output, and this in itself has a marked effect in reducing the intensity of the clutter region on the p.p.i. The fact that the linear receiver has itself a limited logarithmic range lends advantage to this process and certain civil marine radars with linear receivers have optional differentiation (by operation of a switch) which goes a considerable way to reducing the intensity of rain-clutter

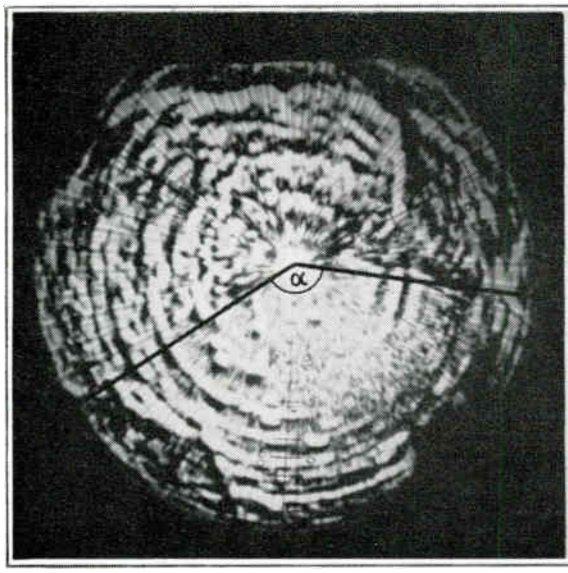


Fig. 8. *Linear receiver followed by differentiating time constant = T (2 μs).*

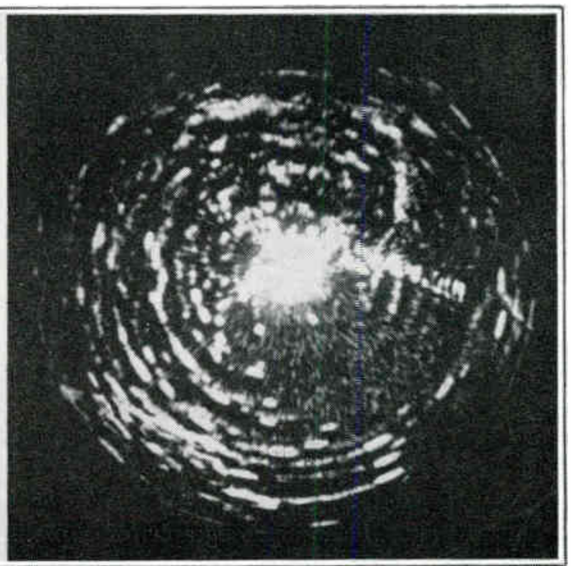


Fig. 9. *Conditions as Fig. 8, but with logarithmic receiver. Photographed ½ min. after Fig. 8.*

paints on the p.p.i., although it cannot eliminate them as is sensibly the case when a logarithmic receiver is followed by differentiation.

The comparison therefore with which this article is concerned, is of the linear receiver plus differentiation and the logarithmic receiver plus differentiation. Admittedly the results from the linear receiver (on sea clutter) could be immediately and immensely improved by the application of swept gain (s.t.c.), but this is a manual control requiring skilled adjustment, and the comparison it is desired to demonstrate is between two automatic devices.

Figs. 8 and 9 show examples of sea clutter effectively eliminates the rain clutter, so that those targets which produce echoes greater than that of the volume of rain illuminated at the same time, will paint upon a p.p.i. a background which is indistinguishable from the usual noise background.

Fig. 12 is interesting in that the radar aerial has been stopped, so that the beam is directed over the sea and photographs of an A-scan fed from the linear and logarithmic receivers have been taken. In Fig. 12(a) the first one-third of the trace to the right of the ground wave is occupied by sea clutter which gives place to rain clutter in the middle third, and then to inherent receiver

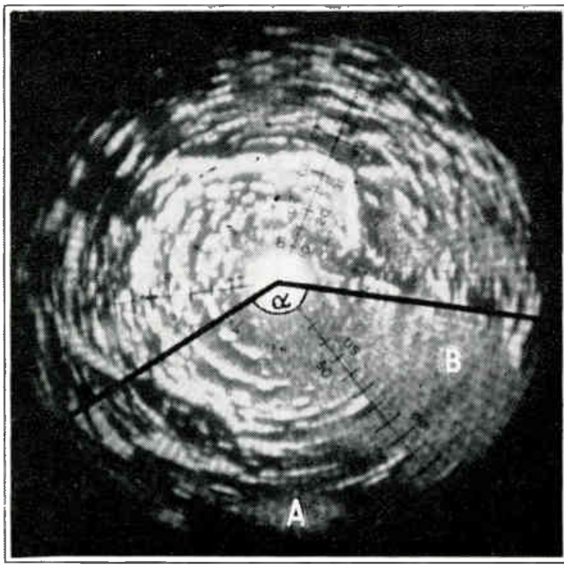


Fig. 10. *Linear receiver, 2 μs time-constant differentiation, sea sector = α. Rain clutter and sea clutter over most of the picture, heavy rain in areas A and B. Range* 15,000 *yards.*

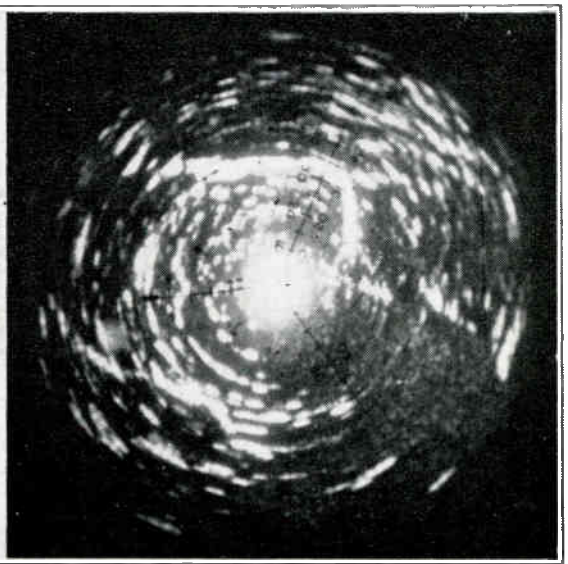


Fig. 11. *As Fig.* 10, *but logarithmic receiver in place of linear. Photograph taken ½ minute after Fig.* 10. *Figs.* 10 *and* 11 *taken several days after Figs.* 8 *and* 9.

photographed on the 12,000 yards range scale of the p.p.i. The sea aspect of the radar set was restricted to the angle α shown in Fig. 8 and it will be noted that clutter extends out to about 9,000 yards. It will be seen that the clutter background is largely eliminated by the combination of logarithmic receiver plus differentiation, but its presence can still be detected as a somewhat coarser type of noise background than is normal. This is mainly because of the trace-to-trace correlation of sea clutter, which as stated in Section 2 may extend over the beam-width of an aerial for a rotating aerial. The effect of this correlation is explained lower down.

Figs. 10 and 11 show photographs of a p.p.i. display with extensive rain clutter. There are heavy rain storms in the areas marked A and B. It will be noted that the logarithmic receiver noise. The structure of the rain clutter is not noticeably distinguishable from the inherent noise but the sea clutter has the appearance of a coarser structure due to the trace-to-trace correlation previously mentioned. It is clear from Fig. 12(b) that the r.m.s. amplitudes of the sea clutter, rain clutter, and inherent noise have been equalized by the logarithmic receiver followed by differentiation. The trace-to-trace correlation, however, will still make the sea-clutter region more prominent than noise on the p.p.i., because the clutter echoes tend to paint in one particular spot of the tube face for several successive p.p.i. traces, thus giving more energy to those particular spots than would inherent noise.

Figs. 12(c) and 12(d) are similar photographs to Figs. 12(a) and 12(b) but sea clutter and inherent noise only are present. It is of interest to note in the inherent noise region the rather more well-defined level of the noise peaks on the logarithmic receiver due to the compressing action of the logarithmic characteristic.

4.3. *Results on an X-Band Civil Marine Radar*

A logarithmic receiver of similar type was constructed and installed in a Decca Type 12 Civil Marine Radar. The pulse length of the radar set was 0·2 μsec and the bandwidth of the receiver was made 10 Mc/s. This receiver also had the facility of being instantly convertible to a linear receiver by operation of a switch.

Figs. 13 to 16 show two pairs of p.p.i. photographs of sea clutter using the linear receiver followed by differentiation, and the logarithmic receiver followed by differentiation. The captions show the differentiation time constants employed in each case. These photographs were taken from a radar on the coast at Southsea (the lower half of each picture represents the sea aspect) during a defence exercise in which many small craft were involved. It is not often that sea clutter conditions happen to coincide with the presence of numbers of small craft close in shore as occurred in this instance. The range scale in use for these pictures was 3·0 nautical miles, and sea clutter extends to a little beyond one nautical mile. Fig. 16(a) gives a key to some of the echoes in Figs. 15 and 16. Comparison of Figs. 14 and 16 with 13 and 15 shows the great reduction in intensity of sea clutter when the logarithmic receiver is used. The complete elimination of sea clutter, however, which would be expected from the theoretical treatment of Section 3, is not obtained. While this is largely because of the trace-to-trace correlation effect, the elimination of clutter on the X-band radar seemed rather less complete than on the S-band radar, and it looks as if some additional factor is reducing the effectiveness of the logarithmic receiver on the X-band equipment. The anti-clutter operation of the logarithmic receiver is, as shown in Section 3, effective for an amplitude probability distribution of the clutter of Rayleigh form. Now the pulse length and beam-width of the S-band radar were 2 μsec and 5° respectively compared with 0·2 μsec and 1·7° for the X-band radar. The area of sea illuminated by the S-band pulse at a given range is therefore some 30 times that of the X-band pulse at a similar range. If the sea surface is considered as being made up of a number of scatterers, upon which the energy of the radar beam impinges, there will, on an area basis, be roughly 30 times as many scatterers illuminated within a range element of the beam of the S-band radar as are embraced by the equivalent range element of the X-band radar. It was stated in Section 2 that the amplitude probability distribution of sea clutter at any given range 'across' a sequence of radar traces is more or less a Rayleigh distribution, a fact which has been established for low-definition radars of pulse length and beam-width similar to the S-band radar under consideration here. However, in the case of the high-definition X-band radar, where the number of scatterers illuminated by the beam within the pulse length is so much less, it is possible that the amplitude probability distribution may depart significantly from the Rayleigh

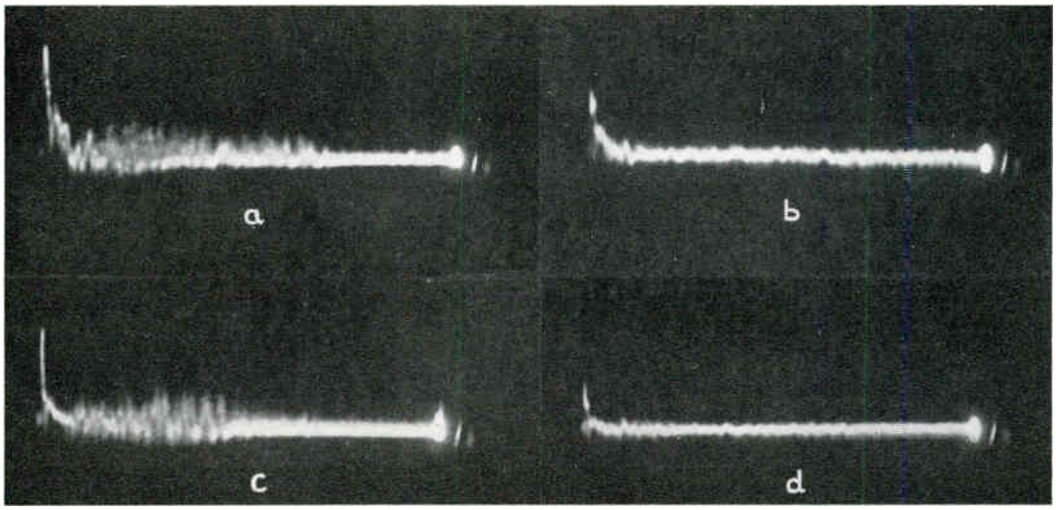


Fig. 12. *Photographs of clutter on 'A' type display; range scale* 15,000 *yards.* (*a*) *Linear receiver, differentiating time constant* 2*T*, *sea clutter extending into rain clutter, and then noise;* (*b*) *as* (*a*) *but with logarithmic receiver, rain clutter had spread to full range before exposure;* (*c*) *linear receiver, differentiating time constant T, sea clutter and noise only;* (*d*) *as* (*c*) *but with logarithmic receiver in place of linear.*

form. This might reduce the effectiveness of the logarithmic receiver on high-definition radars, but is of course in no way an argument against such radars. For the beam-widths and pulse lengths under consideration here, the signal-to-clutter ratio (for a given radar) is to a first approximation inversely proportional to the product of beam-width and pulse length, so that the advantage in improved signal-to-clutter ratio by going to the higher definition is obvious.

4.4. *Differentiation Losses and Effects*

The purpose of the differentiating circuit is to remove the mean level of the clutter, and it may be assumed that a time constant of several times the pulse length could safely be used to do this. In practice, however, it is as well to restrict the differentiating time constant to about one and a half times the pulse length, since longer time constants result in some extension of the ground-wave region at the centre of the p.p.i. The use

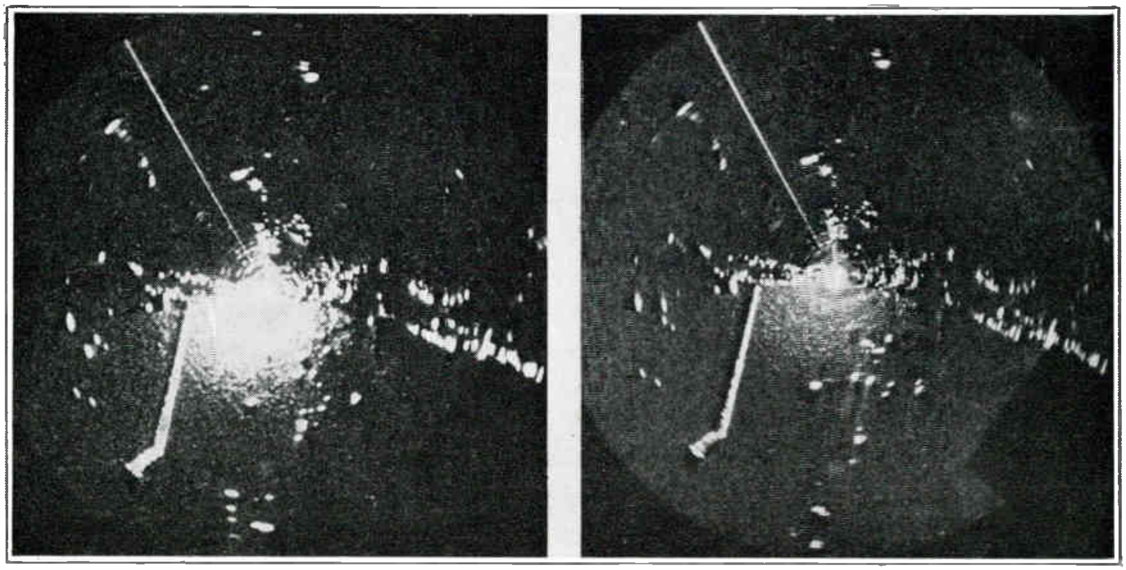


Fig. 13 (left) *for linear receiver and* Fig. 14 (right) *for logarithmic receiver, followed by differentiation; forward and backward time constants* 1·5 *T and* 0·5 *T respectively.*

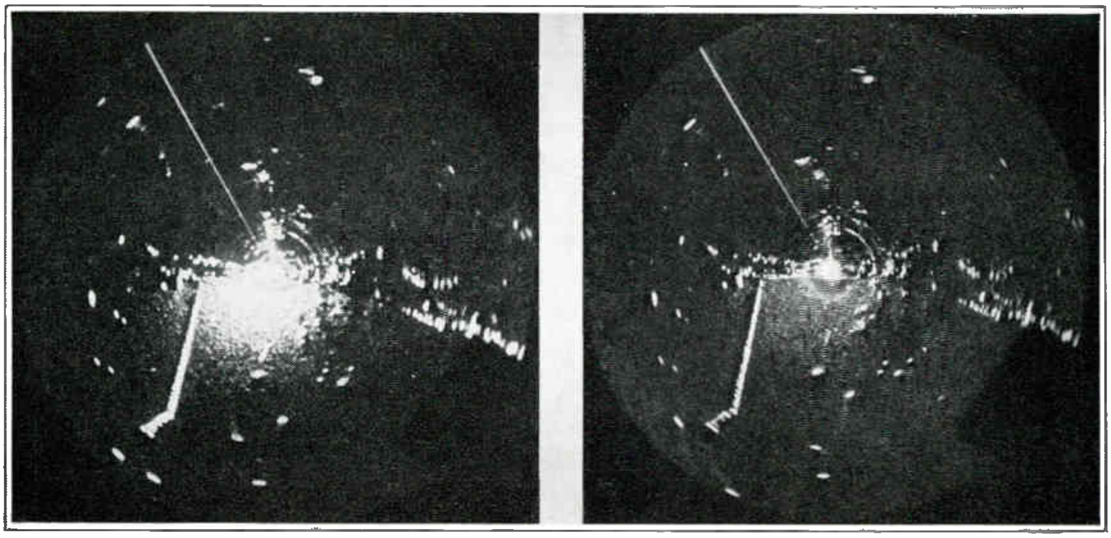


Fig. 15 (left) *for linear receiver and* Fig. 16 (right) *for logarithmic receiver, followed by differentiation; forward and backward time constants* 1·5 *T and* 0·5 *T respectively.*

of a time constant of this order gives rise to shadows immediately after each echo, due to the time taken for the differentiating capacitor to discharge after the echo pulse has ceased. It is usual, therefore, to connect a germanium diode across the differentiating resistance in such a way that it presents a very low resistance to discharge currents from the capacitor, so that discharge takes place much more quickly than charge, the shadows after echoes thereby being rendered negligible in extent. In this manner, recovery times which are a small fraction of the pulse length can be obtained.

When a linear receiver is followed by such an RC differentiating circuit, with a forward time constant of $1{\cdot}5T$ and backward time constant of $0{\cdot}2T$, there appears to be no reduction of signal-to-noise ratio on small signals when compared with the undifferentiated picture. When the same differentiating circuit follows a logarithmic receiver, however, there is a reduction in signal-to-noise ratio of not less than 3 dB, which is of course a considerable price to pay for any advantage offered by the logarithmic receiver.

Fig. 17 (a, b and c) show the photographs of an A-scope trace in which a small echo superimposed upon inherent noise can be seen. The A scope is showing the output of the linear receiver. In Fig. 17(a) the output is not differentiated, but in (b) the output is passed through an RC differentiating network in which the forward time constant is $1{\cdot}5T$ and the backward time constant also $1{\cdot}5T$. In (c) the output is passed through an RC differentiating network in which the forward time constant is $1{\cdot}5T$ and the backward time constant $0{\cdot}2T$. There is no noticeable loss in signal-to-noise ratio between

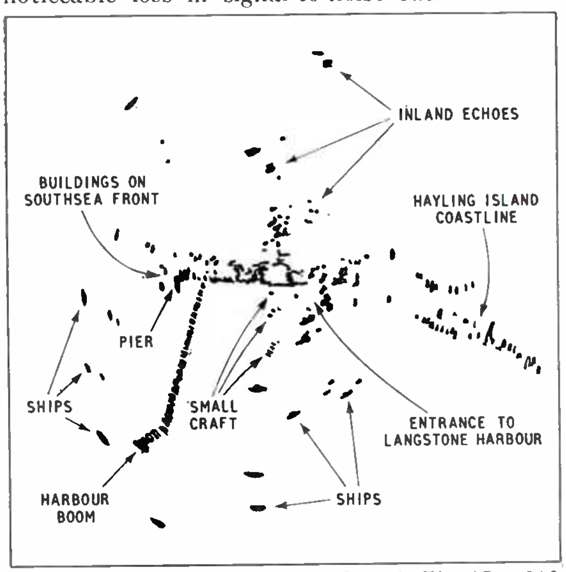


Fig. 16(a). *Key to some of the echoes shown in Figs.* 15 *and* 16.

the result shown in the differentiated outputs (b) and (c) and the undifferentiated output (a).

Fig. 17 (d, e and f) show photographs of the same A-scope trace but fed from the logarithmic

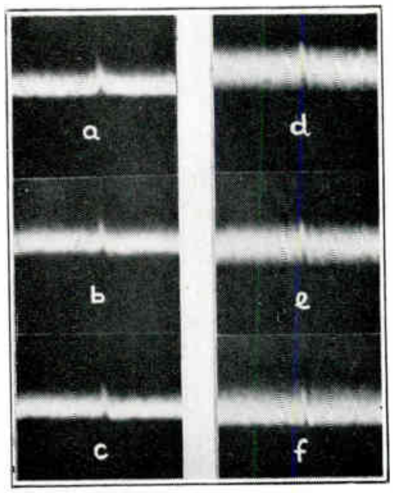


Fig. 17. *Effects of differentiation on weak signals with linear and logarithmic receivers.* (*a*) *Linear receiver output, not differentiated;* (*b*) *same output as* (*a*) *differentiated by forward time constant* $1{\cdot}5\,T$, *backward time constant* $1{\cdot}5\,T$; (*c*) *same output as* (*a*) *differentiated by forward time constant* $1{\cdot}5\,T$, *backward time constant* $0{\cdot}2\,T$; (*d*) *as* (*a*) *but with logarithmic receiver;* (*e*) *as* (*b*) *but with logarithmic receiver;* (*f*) *as* (*c*) *but with logarithmic receiver* (T = *radar pulse length*).

receiver. The sequence is identical with the previous one in respect of the parameters of the differentiating circuit. It will be noticed in (d) that the small echo appears significantly smaller than it does in (a) but this is merely the result of the compressing action of the logarithmic law of the receiver and represents no loss, since the character of the noise is also altered by compression. The signal-to-noise ratio of (d) can therefore be restored to that of (a) by a suitable expansion process (clipping or antilog law) in the video amplifier which follows the logarithmic receiver. Fig. 17(e) does possibly show a small loss in signal-to-noise ratio when compared with the undifferentiated output presented in (d) and, from observations of many p.p.i. pictures, it is thought that a very small loss does occur in differentiation of a logarithmic output even when the backward time constant is kept equal to the forward time constant. It is in the step from (e) to (f), however, where the backward time constant has been reduced to $0{\cdot}2T$, that the real loss occurs which is estimated to be at least 3 dB. Careful observation of (f) will show that the noise has become more feathery than in (d) and (e) and, by holding the edge of a sheet of white paper along the upper level of the noise, first in (d) and (e) and then in (f), the reduction of the extent by which the small echo protrudes above the noise in (f) as compared with (d) and (e) will be immediately obvious.

This effect was rather baffling until the difference in character between linear and logarithmic noise was considered. Fig. 18(a) represents an attempt to show a twice-noise signal at the output of a linear receiver, presented on an A-scope

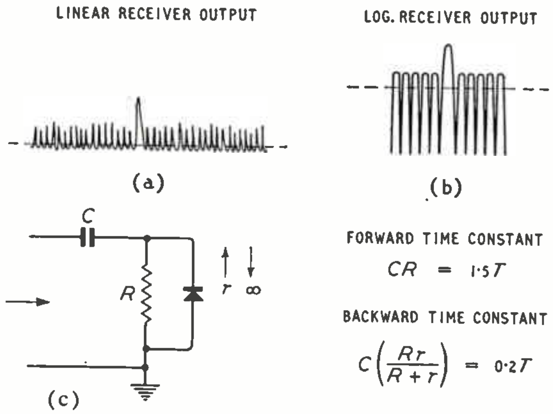


Fig. 18. *Representing small signals at the output of linear and logarithmic receivers and the differentiating circuit to which the outputs are applied.*

trace. Fig. 18(b) shows the same twice-noise signal at the output of a logarithmic receiver; note that the amplitude ratio is no longer 2:1 because of the logarithmic law. To get the same absolute amplitude of signal above noise from the logarithmic receiver as from the linear receiver, much more noise must be accepted from the output of the logarithmic receiver than from the linear receiver. Moreover, whereas the r.m.s. level of the linear noise is near the bottom of the trace, the r.m.s. level of the logarithmic noise is near the top of the trace, the upper excursions of the noise being very much flattened and the lower excursions made very peaky.

In considering the effect of the differentiating circuit of Fig. 18(c) on the outputs of (a) and (b), it should be noted that the negative-going peaks of (a) are small in comparison with the negative-going peaks of (b) and that the differentiating time constant to negative-going peaks is $0{\cdot}2T$ (T = pulse length) while that to positive-going peaks is $1{\cdot}5T$. The positive-going peaks are not significantly differentiated by the long time constant of $1{\cdot}5T$, but the negative-going peaks are differentiated by the much shorter time constant of $0{\cdot}2T$, and the result of this differentiation is to give rise to sharp positive-going peaks which are superimposed on the existing positive-going peaks thus raising the noise level and reducing the signal-to-noise ratio. This effect is negligible on the linear noise of Fig. 18(a) because the noise peaks are only of equal amplitude to the signal but, in Fig. 18(b), where the noise peaks are four times the signal amplitude, the reversed positive-going peaks which result from the short time constant of differentiation ($0{\cdot}2T$) do increase the noise level significantly, and reduce the signal-to-noise ratio by not less than 3 dB. The only way to minimize this effect appears to be to keep the reverse or backward time constant of differentiation equal to the forward time constant, and to accept the shadows which result after the echoes. Where shadows of the order of two pulse lengths cannot be tolerated the reverse time constant may, as a compromise, be reduced to about one-half the pulse length in which case the loss in signal-to-noise ratio is less than 3 dB.

There seems no advantage in using differentiating time constants of less than $1{\cdot}5T$, and with such time constants the author has found no advantage in using the delay-line type of differentiating network as opposed to a simple *RC* combination.

It is a useful facility to be able to switch the differentiation out of circuit on all ranges of the display, in the absence of clutter, but this is especially desirable on long-range scales of display where the spot size of the c.r.t. is much greater than the pulse length. In this case, echoes on the p.p.i. from coastlines tend to lose considerably in boldness and thickness as a result of differentiation, and it should not be used on long-range scales except in the presence of rain.

4.5. *Restoration of Signal/Noise Ratio in Video Amplifiers*

Considering again the illustrations of Fig. 18(a) and (b) it is seen that, for equal signal amplitudes above noise, the logarithmic receiver will have at its output 'peak' noise amplitudes which are roughly four times as great as the linear receiver. We are here writing in rather loose terms when we talk of 'peak' noise amplitudes since, for the Rayleigh distribution, there is no peak amplitude, all amplitudes up to infinity being present. In reading an A scope, however, the eye does assign a rough peak amplitude to the noise present on the trace and it is to this amplitude we refer.

This disparity in signal-to-noise ratio between the linear and logarithmic outputs is the natural result of using a receiver with a logarithmic law holding uniformly down to noise fluctuations 20 dB below the r.m.s. level. Nothing can be done to remove this disparity *before* the differentiating circuit, for the receiver would then no longer be logarithmic down to 20 dB below r.m.s. noise, and differentiation would not reduce the intensity of the clutter fluctuations to the constant level of noise. After differentiation, however, it is desirable to clip the base away from the logarithmic noise in the course of video amplification, so as to restore the signal-to-noise ratio of Fig. 18(a). Alternatively, a video amplifier having an anti-log characteristic may be used.

5. Design Notes

5.1. *Noise and Signal Output Amplitudes for Logarithmic Receivers*

For the type of logarithmic amplifier described in reference three, the output voltage increases uniformly with input signal at a rate which is usually not far off 1·0 V per 20 dB. Assuming therefore that the receiver remains logarithmic for signals 80 dB above r.m.s. noise and 20 dB below r.m.s. noise, the maximum signals will have an amplitude of 4 V above r.m.s. noise and the r.m.s. noise level at the output will be something above 1·0 V, usually about 1·3 to 1·5 V. A signal 6 dB above r.m.s. noise will have an amplitude of 0·3 V above noise. A similar signal at the input of a linear receiver would give an amplitude of 0·3 V at the output with 0·3 V of r.m.s. noise present. Assuming, therefore, no loss on differentiation of the logarithmic output, the same video amplifier gain could be used for these two receivers. Base clipping of the logarithmic noise would of course be necessary early in the video amplifier, since something like 40 V of signal is usually necessary to brighten fully an intensity-modulated display, and the video-output valve would have to pass on about 200 V of r.m.s. noise to do this unless clipping were resorted to.

5.2. *Saturation of Final Stages by Noise*

In this type of logarithmic amplifier, the r.m.s. level of noise will usually be saturating the final valve and almost saturating the penultimate valve when the receiver is so working that it is logarithmic to 20 dB below r.m.s. noise level. The valves cannot work at normal voltages in this condition without being seriously overrun, and it is necessary to reduce the screen volts to some safe level (about 140 V) in order to keep the dissipation within normal limits.

To preserve the constancy of the logarithmic law over the whole working range of inputs the fully saturated contributions of each stage to the total output should be equal and the stage gains equal. Alternatively, if the gain of any stage is down by say 50% the saturated contribution of that stage to the output should also be artificially reduced by 50% to compensate.

If all the valves in the amplifier are run at sub-normal screen voltages in order to safeguard the last two valves, the logarithmic law will remain constant over the working range but the total gain of the amplifier will be somewhat reduced by the process, and it may become necessary to add one stage more than would

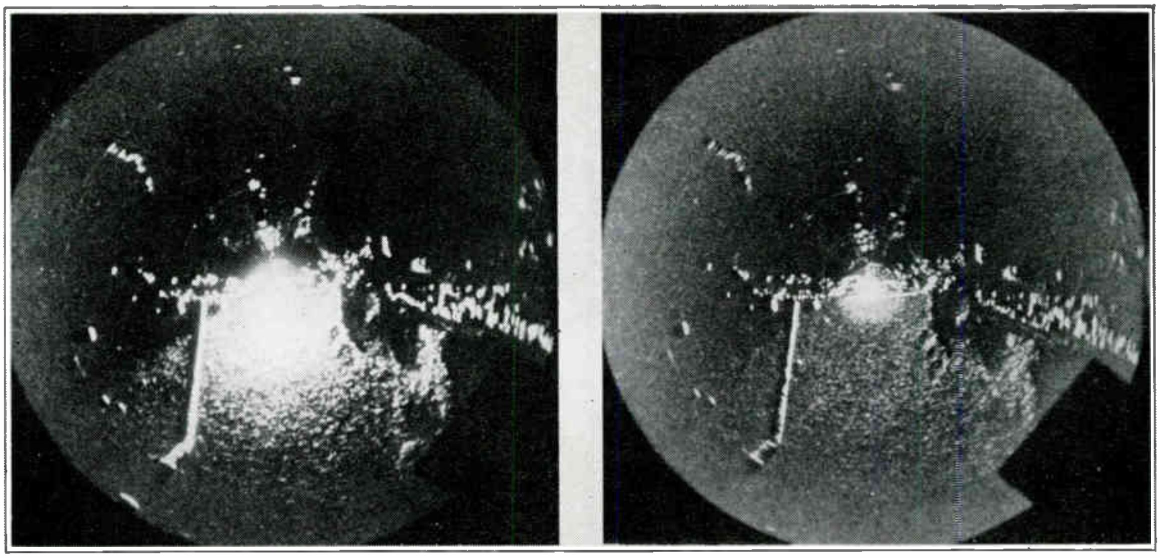


Fig. 19 (left) *for linear and* Fig. 20 (right) *for logarithmic receivers followed by differentiation. Forward time constant = backward time constant = 1·5 T. These pictures were taken with the receiver 'pulsed on' during the brightening pulse of the p.p.i. only. This limits valve dissipation to about one quarter of that of continuous-running conditions and ensures that the last two stages of the logarithmic receiver, which must always be saturated with receiver noise, do not exceed rated dissipation. Targets at 0·5 and 1 mile (7 o'clock) are dan buoys carrying corner reflectors. An X-band civil marine radar (3-miles range scale) was used.*

otherwise be needed. The output of the receiver per decibel of input will also be lowered because the saturated output of each stage will be reduced.

If this condition cannot be accepted the last two valves only may be run at reduced screen voltages; the effect will be to reduce the saturated output of each of these stages, and to reduce their gain very slightly. The reduction of saturated output can be compensated for, and the logarithmic law thus maintained constant, by feeding a suitably larger fraction of the output of these valves into the delay line.

A more elegant solution is to supply the screens with their full rated voltage, but to pulse them on only during the brightening pulse of the cathode-ray tube. Many radar equipments only utilize one-third of the available repetition period even with their longest ranges, and the dissipation can thus be reduced by a factor of three to one without difficulty. A pulse somewhat earlier than the triggering pulse should be used in order to give the screens time to settle down before the incidence of the ground wave. This system works satisfactorily and Figs. 19 to 24 are photographs of the display of a civil marine radar, working from a pulsed receiver under both linear and logarithmic conditions. The performance against sea and rain clutter is indicated in the captions.

5.3. *Tolerances of the Logarithmic Law*

In Section 3 it was shown that, after differentiation, the r.m.s. fluctuations at the output of the logarithmic receiver are given by:—

$$\text{r.m.s. fluctuations} = \frac{a\pi}{\sqrt{24}} \approx 0{\cdot}6a$$

where a is the slope of the logarithmic law $y = a \log_e bx$. If this slope varies over the range, then the r.m.s. fluctuations at the output will not be constant but will depend upon the r.m.s. amplitude of the input signal, since the input signal will decide the operative region of the input–output law of the receiver.

As a p.p.i. display limits at between twice and three times noise, clearly once the display is set up for a given noise level the variations of r.m.s. amplitude of noise and clutter along the trace must not be allowed to vary within anything like a ratio of two to one. In other words the variation of slope of the logarithmic law must be kept well within the limits two to one. In practice, a variation of $\pm$ 20% seems quite acceptable.

6. Trace-to-Trace Correlation

Even with the logarithmic receiver working against an input having a nearly perfect Rayleigh distribution, the trace-to-trace correlation still causes the clutter background to be brighter and coarser than the noise. From time to time suggestions have been put forward for destroying this correlation by sending out successive pulses on different frequencies. Certainly a worth-while improvement in signal-to-clutter ratio could result from destroying the correlation, and the suggestions seem worthy of experiment. A drawback is that either the local oscillator of the

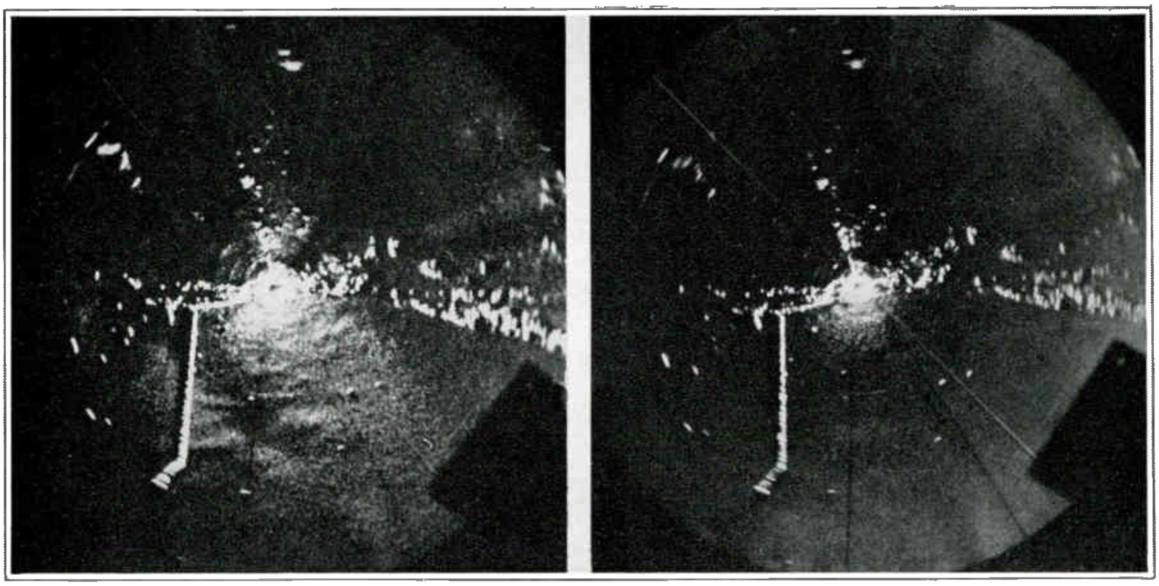
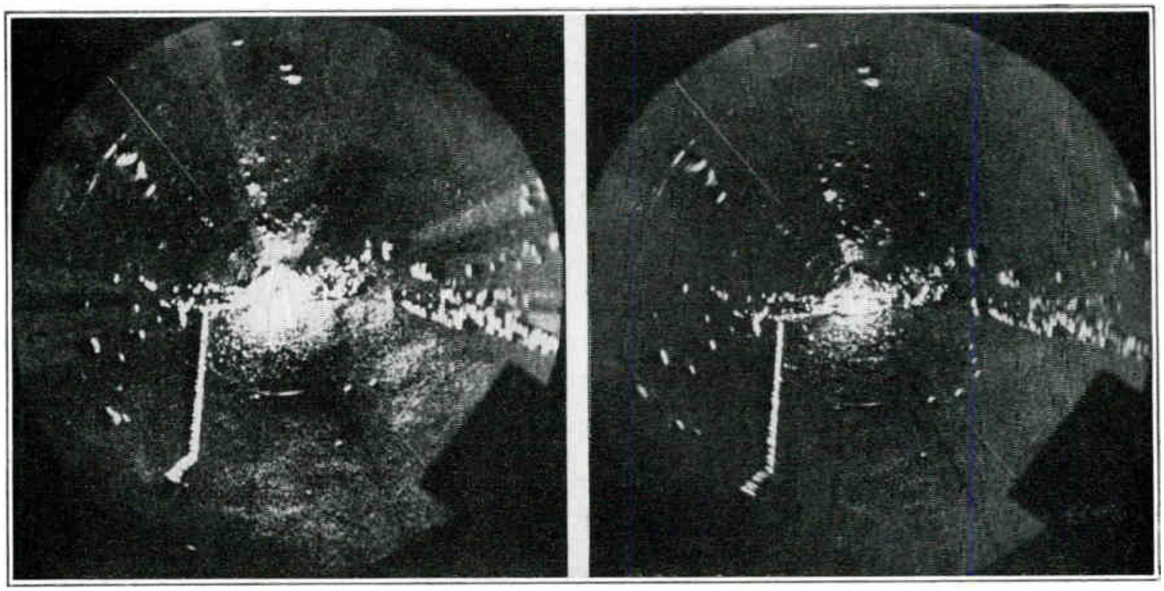


Fig. 21. *A pair of photographs of rain clutter using linear and logarithmic receivers on an X-band civil marine radar. The range scale in use is 3 miles; sea clutter extends to about* 0·5 *n. miles.* (Left) *For linear receiver and* Fig. 22 (right) *for logarithmic receiver, followed by differentiation. Forward time constant = backward time constant =* 1·5 *T in each case*

receiver must shift frequency correspondingly with the frequency change of the transmitted pulse or a very wide-band amplifier (100 Mc/s) must be used. Because of its increased noise, such an amplifier would sacrifice most of the improvement obtained, since the signal-to-noise ratio (for weak signals) cannot be adequately restored by restricting the subsequent video bandwidth.



Figs. 23 and 24. *Rain clutter on 3-miles range scale of X-band civil marine radar. The three targets at* 4 *o'clock* (1·5 *miles) in Fig.* 24 *are barely discernible in Fig.* 23. Fig. 23 (left) *for linear receiver and* Fig. 24 (right) *for logarithmic receiver, both followed by differentiation. Forward time constant = backward time constant = 1·5 T in each case.*

7. Conclusions

(1) The logarithmic receiver followed by differentiation will reduce the r.m.s. fluctuations (intensity) of rain clutter and sea clutter more or less to the intensity of the inherent receiver noise, but only provided that it remains logarithmic to 20 dB below r.m.s. level.

(2) With a p.p.i. display the performance against rain clutter is better than against sea clutter because of the pulse-to-pulse correlation of sea clutter.

(3) The performance against sea clutter on a low discrimination radar (5° beam, 2 μsec pulse) is somewhat better than on a high discrimination radar (1·7° beam, 0.2 μsec pulse), possibly because the amplitude probability distribution of the sea clutter on the low-definition equipment conforms more closely to the Rayleigh form than it does with the high-definition equipment.

(4) There is a loss of some 3 dB in signal-to-noise ratio when a receiver logarithmic to 20 dB below r.m.s. noise is followed by a differentiating circuit which uses a short recovery time constant to avoid shadows. For example, a differentiating circuit having a forward time constant of 1·5 times the pulse length and a recovery time constant of 0·2 times the pulse length shows this effect. This loss does *not* occur when the same differentiating circuit follows a linear receiver and is inherent in the lower signal-to-noise amplitude at the output of the logarithmic receiver. The loss can be mainly avoided on the logarithmic receiver by using a backward time constant equal to the forward time constant and tolerating the resulting shadows after echoes.

(5) Differentiation produces considerable display losses on the long-range scale particularly on coastlines. It should only be used on this scale in the presence of rain.

(6) Clipping or an anti-logarithmic characteristic is essential in the video amplifier which follows the differentiating circuit of the logarithmic receiver in order to restore the output signal-to-noise ratio to that of a linear receiver.

(7) For the type of logarithmic receiver considered in this paper, saturation of the final two stages by noise must be accepted if the receiver is to be logarithmic to 20 dB below r.m.s. noise. The dissipation of the valves can be kept within safe limits either by using a sub-normal screen voltage or by using a normal voltage and pulsing the screens on only during the brightening pulse of the p.p.i.

(8) The slope of the input–output law of the logarithmic receiver in volts per decibel should be kept constant within limits of ± 20%.

(9) A worth-while further improvement in the performance of the logarithmic receiver against sea clutter would result if the pulse-to-pulse correlation could be destroyed by, say, pulse-to-pulse frequency variation.

Acknowledgments

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THE IMPEDANCE CONCEPT

Part 1—*Its Relation to Stability and Feedback*

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SUMMARY.—Much useful information about the properties of a network can be obtained by means of its impedance or transfer function, which can be determined by repeated application of simple algebraic laws in terms of a variable p closely associated with time differentiation. These laws and operational calculus, by means of which results obtained in the 'p-world' can be transferred to the time world when absolutely necessary, are discussed simply *ab initio* and in the barest possible outline. Stability and damping, however, can be determined by algebraic processes within the 'p-world'; these processes include the determination of partial fractions and the conditions for an algebraic equation to be free from roots with positive real parts, which are discussed in detail, starting from first principles. We deduce the conditions for a system to have at least a specified damping rate, and for the stability of a three-stage *RC* amplifier.

In Part 2 the effect of adding terms in p^4 and p^5 on the roots of a given cubic in p is discussed more fully. It is assumed that the time constants associated with the amplifier are not under the designer's control; the effect of adding a 'step circuit' on the gain and the maximum feedback obtainable are therefore fully considered.

1. Introduction

RECENT developments in the whole field of servo and feedback-amplifier design make it doubtful whether the historical approach to the subject can any longer be regarded as the natural and most useful one. Our present approach is based upon the 'impedance concept' which is now well known. Impedances and the associated transfer functions are expressed, usually algebraically, in terms of a variable $p = \alpha + j\omega$ which is closely associated with time differentiation. Under steady-state conditions, we can regard α as zero, so that p takes the familiar form $j\omega$. We prefer, however, to work throughout in the 'p-world', where much relevant information can be obtained, usually by purely algebraic processes; we seldom have to carry out a 'translation' into the time or frequency worlds. In the 'p-world' we do not have to distinguish between steady-state and transient conditions. The fundamental properties of the 'p-world' and their relevance to practical considerations are discussed in a simple and elementary manner by Carter[1]. Operational calculus is the process by which the frontier between the 'p-world' and the familiar world of time and space is crossed in both directions. A comprehensive 'grammar' and 'dictionary' for 'translation' from 'p-language' to time language and vice versa (without proofs) is provided by McLachlan and Humbert[2]. Certain properties of algebraic equations (in the p-world) are also relevant; many of these are listed in reference 1 (pp. 115–116), for equations of degree not exceeding 4. We have included only sufficient reference, without proof, to these results for clarity. Many of the ideas we here set forth are not new, but their power to clarify fundamental thought about the whole subject of feedback and servo design has perhaps been inadequately exploited. Various difficulties inherent in the historical approach are irrelevant to our present approach.

We include, with proof in the Appendix, a general expression of the condition for a circuit with an algebraic characteristic equation of degree n to be stable. This result is given by Guillemin[3] and is also discussed by Morris and Head[4] from the purely algebraic standpoint. The pioneer in this field, however, was Routh[5]. Frazer and Duncan[6] also obtained, as early as 1928, determinantal conditions of stability which are somewhat unwieldy for a high value of n. We deduce the condition for a circuit to have

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damping at least as rapid as $e^{-\alpha t}$, and apply these results to the case of a three-stage RC amplifier to which we wish to add suitable elements to obtain greater feedback without producing instability and without excessive loss of gain.

2. Fundamental Impedance Laws and Properties of the '*p*-World'

We assume the following laws for networks with lumped elements:

(a) The impedances of a resistance R ohms, an inductance L henrys and a capacitance C farads are respectively R, pL and $1/pC$ ohms.

(b) Two impedances Z_1 and Z_2 in series are equivalent to a single impedance $Z_1 + Z_2$.

(c) Two impedances Z_3 and Z_4 in parallel are equivalent to a single impedance $Z_3Z_4/(Z_3 + Z_4)$.

(d) Within the p-world, p may be manipulated algebraically.

By repeated application of (a), (b), (c) and (d) the impedance Z of any network with lumped elements can be found. If the input voltage as a function of time is $V_1(t)$, the input current is $i_1(t)$ and the output voltage is $V_2(t)$, then operational calculus (as we shall see in Table 1) enables us to determine a unique corresponding function of p for each; we shall call these functions respectively V_1, i_1 and V_2. We then have a generalization of Ohm's Law for a network equivalent to a single impedance Z:

(e) $V_1 = i_1 Z$.

Knowing V_1 and Z we can therefore deduce i_1, and hence $i_1(t)$ by operational calculus. We have seen that Z is a function of p associated with the network; V_2/V_1 is another such function, called the transfer function. Our final fundamental law is

(f) If the transfer functions of two networks are T_1 and T_2, the transfer function of the single network formed by connecting these two networks in cascade is $T_1 T_2$, provided that the two networks are separated by a valve so that there is no interaction.

The very brief Table 1 of time functions $V_1(t)$ and the associated 'operationally equivalent' functions of p will be sufficient for our present purpose. In all cases $V_1(t)$ is to be taken as zero when $t < 0$, so that all functions tabulated have discontinuities, or discontinuous derivatives, when $t = 0$.

If $V_1(t)$ corresponds to V_1, then

(g) $\dfrac{d^r V_1(t)}{dt^r}$ corresponds to $p^r V_1 - \sum_{s=0}^{r-1} p^{r-s}\left[\dfrac{d^s V_1(t)}{dt^s}\right]_{t=0}$

(h) $\lim_{t \to 0} V_1(t) = \lim_{p \to \infty} V_1$.

(i) $\lim_{t \to \infty} V_1(t) = \lim_{p \to 0} V_1$.

For a 'dead' circuit the second term in (g) is absent. (h) and (i) frequently give us sufficient information about $V_1(t)$ and its derivatives if we know V_1, so that it is not necessary to determine $V_1(t)$ explicitly by means of Table 1.

3. Application to a Series *LCR* Circuit

Let us apply these assumptions to the series LCR circuit of Fig. 1 considered in detail by Carter[1] (p. 28).

By assumptions (a) and (b), the total impedance is

$$pL + R + \frac{1}{pC} \quad (1)$$

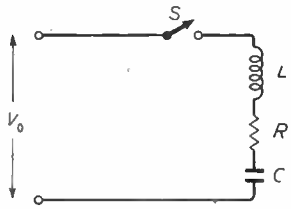


Fig. 1. *A series LCR circuit.*

If therefore the switch S is closed at $t = 0$, the input voltage is by Table 1 operationally equivalent to V_0, and by assumption (e) we have

$$i = V_0 \Big/ \left\{pL + R + \frac{1}{pC}\right\} \quad \cdots \quad (2)$$

Carter considers the capacitor voltage v instead of i; applying assumptions (a) and (e) to the capacitor only, we have

$$v = \frac{1}{pC} \cdot i$$

Hence eliminating i and using assumption (d)

$$v = \frac{V_0}{LC} \cdot \frac{1}{p^2 + \frac{R}{L}p + \frac{1}{LC}} \quad \cdots \cdots \quad (3)$$

as found by Carter. Now suppose that the zeros of the denominator in (3) are $-\alpha$ and $-\beta$ so that

TABLE 1*

Time Functions $V_1(t)$ and Operational Equivalents V_1

$V_1(t)$ $[t \geqslant 0]$	a	$t^n/n!$	e^{at}	$e^{at} - 1$	$\sin at$	$\cos at$
V_1	a	$1/p^n$	$p/(p - a)$	$a/(p - a)$	$pa/(p^2 + a^2)$	$p^2/(p^2 + a^2)$

$V_1(t)$ $[t \geqslant 0]$	$e^{-at} \cos bt$	$e^{-at} \sin bt$	te^{-at}
V_1	$p(p + a)/\{(p + a)^2 + b^2\}$	$pb/\{(p + a)^2 + b^2\}$	$p/(p + a)^2$

*We have here used the Heaviside system in preference to the Laplace.

C

$$\alpha + \beta = R/L; \ \alpha\beta = 1/LC \quad .. \quad .. \quad (4)$$

Then in order to apply Table 1, we must first express the right-hand side of (3) in partial fractions. When (as usually happens in circuit problems) the numerator of the fraction is of lower degree than the denominator, and the zeros of the denominator are real and distinct, this can be done easily, for

$$v = V_0 \frac{\alpha\beta}{(p + \alpha)(p + \beta)}$$

$$= V_0 \left[\frac{\alpha\beta}{(p + \alpha)(\beta - \alpha)} + \frac{\alpha\beta}{(\alpha - \beta)(p + \beta)} \right] \quad (5)$$

the rule being that to obtain the partial fraction with denominator $(p + \alpha)$, p is replaced by $-\alpha$ in the numerator and denominator of v everywhere except in the factor $(p + \alpha)$ in the denominator, and similarly for the partial fraction with denominator $(p + \beta)$ we replace p by $-\beta$ everywhere in the numerator and denominator of v except in the factor $p + \beta$.

By Table 1, we then deduce

$$v(t) = V_0 \left\{ 1 - \frac{\alpha e^{-\beta t} - \beta e^{-\alpha t}}{\alpha - \beta} \right\} \quad .. \quad (6)$$

If α, β are complex, they will be conjugate so that we can write from (4)

$$\alpha = \gamma + j\omega; \quad \beta = \gamma - j\omega \quad .. \quad .. \quad (7)$$

where $\gamma = R/2L, \quad \omega^2 = (1/LC) - (R^2/4L^2) \quad (8)$

In this case (6) is still valid, but is more conveniently expressed in terms of γ, ω in the form

$$v(t) = V_0 \{1 - e^{-\gamma t}[\cos \omega t + (\gamma/\omega) \sin \omega t]\} \quad .. \quad .. \quad .. \quad (9)$$

The borderline case between (6) and (9) occurs when $\alpha = \beta = \gamma = R/2L$ and $\omega = 0$. Now if ωt is sufficiently small, $\cos \omega t$ can be replaced by 1 and $\sin \omega t$ by ωt in (9); this suggests that in the borderline case we should expect to find

$$v(t) = V_0 \{1 - e^{-\gamma t} [1 + \gamma t]\} \quad .. \quad .. \quad (10)$$

and this result is in fact correct[1].

From (8) and (9) we see that the damping rate γ is proportional to R provided that there is oscillation ($R^2 < 4L/C$). If, however, $R^2 > 4L/C$, (6) applies and the effective damping rate is the lesser of α, β, that is

$$\alpha = (R/2L) - \{(R^2/4L^2) - (1/LC)\}^{\frac{1}{2}} \quad .. \quad (11)$$

(11) decreases as R increases [above $(4L/C)^{\frac{1}{2}}$]. For large values of R we have

$$\{(R^2/4L^2) - (1/LC)\}^{\frac{1}{2}} = (R/2L)\{1 - (4L/CR^2)\}^{\frac{1}{2}}$$
$$= (R/2L)\{1 - (2L/CR^2) + \ldots\} \quad (12)$$

so that α in (11) is nearly equal to $1/RC$. If, therefore, R is gradually increased from zero while L and C remain constant, the damping rate will be proportional to R until $R^2 = 4L/C$ (critical damping), when it will reach its maximum value.* When $R^2 > 4L/C$ the effective damping rate, now given by (11), decreases steadily and tends to zero as R tends to infinity. This apparent anomaly for large values of R is explained by further examination of (6), in which α is now small compared to β. Neglecting α/β, (6) reduces to

$$v(t) \approx V_0 \{1 - e^{-\alpha t}\} \approx V_0 \alpha t \text{ for small } \alpha t \quad (13)$$

so that the smallness of α simply means that $V(t)$ is initially proportional to α and t, and takes a long time to grow to its final value V_0.

We notice that assumptions (g), (h) and (i) give us directly from (3) the initial values of $v(t)$ and its first two derivatives just after the switch is closed, and the final value of $v(t)$; thus $v(0) = 0$, $dv/dt = 0$ and $d^2v/dt^2 = V_0/LC$ when $t = 0$, while the final value of $v(t)$ is V_0 and all its derivatives are zero. Assumption (h) ceases to be meaningful if the numerator of V_1 has higher degree than the denominator.

This indicates that we may be able to obtain adequate information about the circuit of Fig. 1 from (3); the full information contained in equations (6) or (9) may not be required, but the nature of the zeros of the denominator of (3) decides the type of behaviour to be expected. In (6) and (9), $v(t)$ contains one or more exponential terms with a negative exponent, because the denominator of (3) has either negative real zeros [$-\alpha$ and $-(R/L) + \alpha$, where α is given by (11)] or complex zeros with negative real parts [$-\gamma \pm j\omega$ where γ, ω are given by (8)] or equal negative real zeros $-R/2L$. Had the denominator of (3) possessed a zero with a positive real part, the associated circuit would have been unstable, because then (6) or (9) would contain one or more terms which would grow exponentially. In general, a passive network cannot be unstable, so that the existence of a zero with positive real part would indicate the presence of a generator. For a quadratic expression

$$a_0 + a_1 p + a_2 p^2$$

there cannot be zeros with positive real parts if a_0, a_1 and a_2 all have the same sign, but this is not a sufficient restriction on the coefficients of an expression of higher degree. This fact becomes important when we wish to determine the maximum feedback obtainable from a feedback amplifier having several stages or to determine what elements should be added to such an amplifier to increase the feedback obtainable. The fact that (3) has a denominator quadratic in p with positive coefficients is sufficient to show that $v(t)$ either consists of two exponential terms with negative exponents or a damped 'cisoidal' oscillation. 'Cisoidal' is a term used by Campbell

*In general, however, critical damping is not quite the same as maximum damping. See, for example, Carter[1] (p. 70).

and Foster[10] to mean 'of the form $e^{j\omega t}$'; the term is convenient when we are mainly concerned with ω, and are not greatly concerned with distinguishing between the sine-term and the cosine-term. More generally, we are likely to obtain in the 'p-world' an equation similar to (3) but involving a denominator D in p which may be of any degree. All we usually need to know is whether the quantity corresponding to $V(t)$ involves exponentially growing or exponentially decaying terms, or 'cisoidal' oscillations growing exponentially or damped exponentially. This can be determined from the nature of the zeros of D. In the next section, therefore, we consider the conditions for a polynomial in p of the third or higher degree to have no zeros with positive real parts. This does involve what at first sight appears to be pure complex-variable algebra, but it is in fact algebra put to its proper use as a labour-saving device. It enables us to be lazy in an enlightened way, so that we do not have to use Table 1 repeatedly to obtain detailed information which has no practical significance. We also deduce easily the conditions for a polynomial in p to have no zeros whose real parts are greater than a given quantity $-\alpha_0$. Such a polynomial is associated with time functions, 'cisoidal' or exponential, damped at least as rapidly as $e^{-\alpha_0 t}$.

4. Conditions of Stability

We wish to determine the conditions under which the polynomial

$$P_n = a_0 + a_1 p + a_2 p^2 + \ldots + a_n p^n \quad (14)$$

shall have no zeros with positive real parts. If in (14) we replace p by $j\omega$, P_n can be rewritten in the form

$$P_n = (a_0 - a_2\omega^2 + a_4\omega^4 - \ldots) + j\omega(a_1 - a_3\omega^2 + a_5\omega^4 - \ldots) \quad (15)$$

$$= E_n + j\,O_n \text{ for brevity} \quad (16)$$

Now suppose that $\omega_1^2, \omega_2^2 \ldots \omega_s^2$ are the values of ω^2 which make E_n zero while $\Omega_1^2, \Omega_2^2 \ldots \Omega_t^2$ are the non-zero values of ω^2 which make O_n zero. Then if n is odd, s and t will both be $\frac{1}{2}(n-1)$, whereas if n is even, $s = \frac{1}{2}n$ and $t = \frac{1}{2}n - 1$. The necessary and sufficient conditions for P_n to be free from zeros with positive real parts are then as follows:

(a) the coefficients of P_n must all have the same sign.

(b) $\omega_1^2, \omega_2^2 \ldots \omega_s^2$ and $\Omega_1^2, \Omega_2^2 \ldots \Omega_t^2$ must all be real and positive; we can therefore regard them as arranged in ascending order.

(c) $0 < \omega_1 < \Omega_1 < \omega_2 < \Omega_2 < \ldots$

This result is proved in the Appendix for the case $n = 4$, and the way in which the proof can be extended to higher values of n is there indicated. The possibility that the ω and/or Ω terms might not be distinct is also briefly considered, but this case is of mathematical rather than practical importance.

We can also easily obtain the condition that all the zeros shall have negative real parts less than $-\alpha$; if P_n satisfies this condition, the associated circuit will only give rise to oscillations or exponential decay damped at least as rapidly as $e^{-\alpha t}$. To obtain this condition, put $q = p + \alpha$ in (14) and let Q_n be the resulting polynomial in q. Then Q_n must be free from zeros in q with positive real parts, and we can apply the criteria just obtained to Q_n instead of to P_n. If

$$Q_n = b_0 + b_1 q + b_2 q^2 + \ldots + b_n q^n \quad (17)$$

then

$$\left.\begin{aligned}
b_n &= a_n \\
b_{n-1} &= a_{n-1} - \binom{n}{1}\alpha a_n \\
b_{n-2} &= a_{n-2} - \binom{n-1}{1}\alpha a_{n-1} + \binom{n}{2}\alpha^2 a_n \\
b_{n-3} &= a_{n-3} - \binom{n-2}{1}\alpha a_{n-2} + \binom{n-1}{2}\alpha^2 a_{n-1} - \binom{n}{3}\alpha^3 a_n \\
&-\,-\,-\,-\,-\,-\,-\,-\,-\,-\,-\,-\,-\,-\,-\,-\,-\,-\,- \\
b_{n-r} &= a_{n-r} - \binom{n-r+1}{1}\alpha a_{n-r+1} + \binom{n-r+2}{2}\alpha^2 a_{n-r+2} \ldots + (-1)^r \binom{n}{r}\alpha^r a_n \\
b_0 &= a_0 - a_1\alpha + a_2\alpha^2 + \ldots + (-1)^n \alpha^n a_n
\end{aligned}\right\} \quad \ldots (18)$$

and since α is given in advance, the b terms are easily calculated from (18). The symbol $\binom{m}{k}$ in (18) denotes the binomial coefficient $m!/\{k!\,(n-k)!\}$.

The result we have obtained is a perfectly general condition for a polynomial P_n in p to be free from zeros with positive real parts. This result was obtained in this form by Guillemin[3]. It is equivalent to results first obtained by Routh[5] and discussed fully by Frazer and Duncan[6]. Carter[1] (p. 116) lists the equivalent results for $n = 2, 3, 4$ where they are expressed simply in terms of the coefficients, but for higher values of n the results cannot be expressed simply and explicitly in terms of the coefficients.

We now apply these results to the case of a three-stage *RC* feedback amplifier.

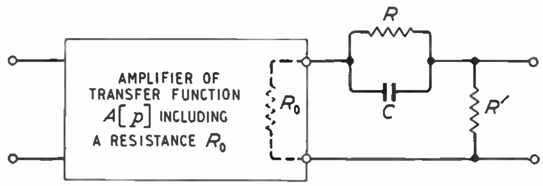


Fig. 2. *Three-stage amplifier with added step-circuit elements to improve feedback.*

5. Application to a Three-Stage *RC* Feedback Amplifier

In the general case of a feedback amplifier, if $A[p]$ is the transfer function of the forward path and $\beta[p]$ is the transfer function of the feedback path, the input voltage v is related to the output voltage V by the equation

$$V/v = A/(1 - A\beta) \quad \ldots \quad \ldots \quad \ldots \quad (19)$$

as first proved by Nyquist[8]. In the case of our three-stage valve amplifier, we shall suppose that

$$A = A_0/(1 + T_1p)(1 + T_2p)(1 + T_3p) \quad (20)$$

where A_0 is a constant and the time constants T_1, T_2 and T_3 are determined by the parameters of the three valves and are not under our control except in so far as some trading of bandwidth for gain is possible. We shall also suppose that the feedback is negative, and that β is a mere negative constant, say $-\lambda/A_0$. The equation determining the stability and damping of the amplifier is then

$$1 - A\beta = 0 \quad \ldots \quad \ldots \quad \ldots \quad (21)$$

This reduces to

$$(1 + \lambda) + (T_1 + T_2 + T_3)p + (T_2T_3 + T_3T_1 + T_1T_2)p^2 + T_1T_2T_3p^3 = 0 \quad \ldots \quad \ldots \quad \ldots \quad (22)$$

In the notation of the last section, the expressions we should have to consider for determining stability would be

$$\left.\begin{aligned} E_3 &= (1 + \lambda) - (T_2T_3 + T_3T_1 + T_1T_2)\omega^2 \\ O_3/\omega &= (T_1 + T_2 + T_3) - T_1T_2T_3\omega^2 \end{aligned}\right\} (23)$$

and for stability the zero of E_3 must be less than the zero of O_3, so that

$$(1 + \lambda)/(T_2T_3 + T_3T_1 + T_1T_2) < (T_1 + T_2 + T_3)/T_1T_2T_3$$

which reduces to

$$\lambda < (T_2/T_3) + (T_3/T_2) + (T_3/T_1) + (T_1/T_3) + (T_1/T_2) + (T_2/T_1) + 2 \quad \ldots \quad \ldots \quad (24)$$

a result obtained by one of the authors[9] in 1938. If we now regard λ as a quantity which can be continuously varied from zero upwards, we see that for small values of λ, (22) has three real negative roots in p, and that the effective damping rate is determined by the least of the quantities $1/T_i$; as λ is increased the zero of O_3, which is given by (23), remains fixed, while the zero of E_3 increases steadily with λ. When λ reaches the maximum value determined by (24), the zeros of E_3 and O_3 are equal, so that (22) has a purely imaginary root. It can be shown that there is a unique value λ_0 of λ satisfying (24) such that (22) has equal roots; as λ approaches λ_0 from below the three roots of (22) are real but the two numerically smallest roots approach each other. For λ between λ_0 and the maximum value λ_1 possible without violating (24), (22) has a pair of complex roots with equal negative real parts; these real parts tend to zero as λ approaches λ_1 from below, so that the system then has little damping, and there is a close association between the rate of damping and the difference between the zeros of E_3 and O_3 in (23) even if it is difficult to find an explicit formula expressing this association.

In our second article we determine the effect on the zeros of a cubic P_3 in p if terms in p^4 and p^5 are added. A possible method of obtaining increased feedback is to add a 'step circuit' as in Fig. 2. When this is done, we have three parameters R, C and R' which are under our control, whereas T_1, T_2, T_3, and R_0 are not; R_0 is taken into account in determining T_1, T_2 and T_3 and is merely a resistance which already exists as part of the original amplifier to which it happens to be convenient to attach the step circuit. We therefore consider how the choice of values for R, C and R' (or rather for certain parameters determined by them) affects the degree of feedback obtainable and the overall gain.

APPENDIX

Conditions for a Polynomial to be Free from Zeros with Positive Real Parts

In Section 4 we stated the necessary and sufficient conditions for the polynomial P_n defined by (14) to be free from zeros with positive real parts. We now prove these results for the case $n = 4$ and indicate how the proof may be extended to higher values of n. We also discuss briefly the possibility that the quantities called ω_s and/or Ω_s in Section 4 might not be distinct. Consider, therefore, the polynomial

$$P_4 = a_0 + a_1p + a_2p^2 + a_3p^3 + a_4p^4 \quad \ldots \quad (A1)$$

and let

$$p = \alpha + j\omega = re^{j\theta} \quad \ldots \quad \ldots \quad \ldots \quad (A2)$$

$$\begin{aligned} P_4 = Re^{j\phi} &= (p - p_1)(p - p_2)(p - p_3)(p - p_4) \\ &= R_1e^{j\phi_1} . R_2e^{j\phi_2} . R_3e^{j\phi_3} . R_4e^{j\phi_4} \quad \ldots \quad (A3) \end{aligned}$$

where $p_k = \alpha_k + j\omega_k$ are the zeros of P_4, so that

$$r = (\alpha^2 + \omega^2)^{1/2}; \quad \theta = \tan^{-1}(\omega/\alpha) \quad \ldots \quad \ldots \quad (A4)$$

$$R_k = \{(\alpha - \alpha_k)^2 + (\omega - \omega_k)^2\}^{1/2};$$
$$\phi_k = \tan^{-1}\{(\omega - \omega_k)/(\alpha - \alpha_k)\}\ (k = 1, 2, 3, 4) \qquad (A5)$$
$$R = R_1R_2R_3R_4;\ \phi = \phi_1 + \phi_2 + \phi_3 + \phi_4 \quad .. \quad (A6)$$

It is shown in reference 7 that if p is made to follow out any closed curve C, the consequent change in ϕ, the argument of P_4 [see (A3)], will be 2π times the number of zeros of P_4 contained within C. We shall give the argument fully, following reference 7, for the case when C is a quadrant of a circle of large radius r_0 bounded radially by the real and imaginary axes as in Fig. 3. If r_0 is sufficiently large, any zero of P_4 which has positive real and imaginary parts can be regarded as within C, and we can also regard the term a_4p^4 as the only important term of P_4 when $|p| = r_0$. As complex zeros of P_4 occur in conjugate pairs, any zero of P_4 in the first quadrant will be associated with a corresponding zero in the fourth quadrant which will also have a positive real part. If therefore the argument ϕ of P_4 undergoes no change when p describes the large quadrant C already specified, P_4 will be free from zeros with positive real parts as required, and if the argument ϕ of P_4 does change, it will change by $2N\pi$ where N is the number of zeros of P_4 with positive real and imaginary parts; P_4 will in this case also have N roots with positive real and negative imaginary part. If p is varied from zero following the closed path shown by arrows in Fig. 3, and r_0 is large, θ remains zero until $p = r_0$, when $\alpha = r_0$ and $\omega = 0$. Round the circular part of the path, r_0 remains constant while θ increases from 0 to $\pi/2$; round this part of the path α is $r_0 \cos\theta$, which decreases steadily, while ω is $r_0 \sin\theta$, which increases steadily, so that $\alpha^2 + \omega^2$ remains constant and equal to r_0^2. When $\theta = \pi/2$, α is zero and $\omega = r_0$. Thereafter, θ remains constant and equal to $\pi/2$, and $p = j\omega$; p thus falls from jr_0 to zero completing the contour.

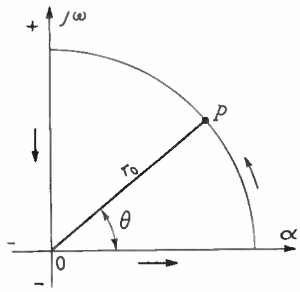
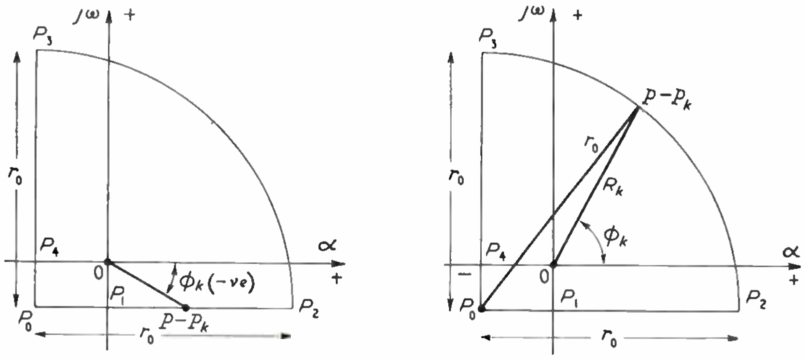


Fig. 3. *A closed contour for p.*



(a) (b)

Fig. 4(a). *Behaviour of $(p - p_k)$ when p moves from the origin along the positive real axis, and p_k has positive real and imaginary parts.* (b) *Behaviour of $(p - p_k)$ when p describes the circular part of the contour of Fig. 2 and p_k has positive real and imaginary parts.*

Now consider the behaviour of

$$p - p_k = R_k e^{j\phi_k}$$

as p describes the contour in Fig. 3. R_k and ϕ_k are given by (18), and we shall first suppose that α_k and ω_k are positive, so that the point $p = p_k$ lies within the contour of Fig. 3. The initial position of $p - p_k$ will be the point P_0 $(-\alpha_k, -\omega_k)$ which is in the third quadrant as in Fig. 4(a), and has an argument between $-\pi$ and $-\pi/2$. The locus of the point $(p - p_k)$ will be a quadrant, like that of Fig. 3, but having P_0 as centre instead of the origin.

Initially the argument of $(p - p_k)$ will be that of P_0 which we have taken as between $-\pi$ and $-\pi/2$. As p moves along the real α-axis as in Fig. 4(a), $(p - p_k)$ will move parallel to the real α-axis, and ϕ_k will become $-\pi/2$ at P_1 when $p = \alpha_k$ and ϕ_k will be small and negative when $p = r_0$ at P_2. While p describes the circular part of its contour, Fig. 4(b), since r_0 is large, p and $(p - p_k)$ will differ little, so ϕ_k will increase from a small and negative value to a value just over $\pi/2$ when p reaches the value jr_0 at P_3. As p descends the imaginary axis to the value $j\omega_k$ at P_4, the argument of $(p - p_k)$ increases steadily to π. Finally, when p completes the contour, the argument of $(p - p_k)$ has a value between π and $(3\pi/2)$ which is 2π more than the initial value. If now we suppose either of α_k and ω_k to be negative (or both as in Fig. 5), so that the point $p = p_k$ is no longer within the contour of Fig. 3, the point P_0 is in one of the other quadrants, and the quadrant $P_0P_2P_3$ will be found no longer to include the origin. A repetition of the discussion of the changes in ϕ_k shows that ϕ_k now returns to its initial value after p has described the contour of Fig. 3.

It follows that when p describes the contour of Fig. 3, each of ϕ_1, ϕ_2, ϕ_3 and ϕ_4 as defined by (A3) increases by either 0 or 2π, and therefore that the sum ϕ of these angles increases by 2π times the number of zeros of P_4 enclosed within the path of Fig. 3; for sufficiently large r_0, this means 2π times the number of zeros of P_4 lying in the first quadrant.

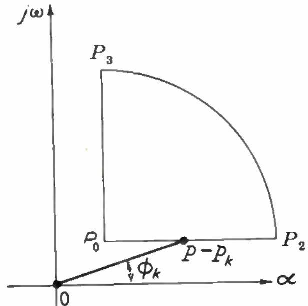


Fig. 5. *Behaviour of $(p - p_k)$ when p describes the contour of Fig. 3 and p_k has negative real and imaginary parts.*

Now consider the behaviour of the polynomial P_4 itself [see equation (A1)]. Initially the argument ϕ of P_4 is zero. We are interested only in the case where all the a terms are positive as it can easily be proved that P_4 must have positive real parts if one of the a terms (other than a_0, a_4) is zero or if they do not all have the same sign. It follows that ϕ remains zero as p moves along the real axis to $p = r_0$. While p describes the circular part of its contour, the predominant term of the polynomial P_4 is $a_4p^4 = a_4r_0^4e^{4j\theta}$ if r_0 is sufficiently large. Hence the argument ϕ of P_4 is approximately 4θ and increases from 0 to (approximately) 2π. p has now reached the value jr_0, so that ω has the large value r_0, and we have to consider what happens when p descends the imaginary axis and ω decreases to zero. At this stage $p = j\omega$ and P_4 reduces to

$$P_4 = (a_0 - a_2\omega^2 + a_4\omega^4) + j\omega(a_1 - a_3\omega^2) \quad .. \quad (A7)$$
$$= E_4 + jO_4 \text{ for brevity} \quad .. \quad .. \quad .. \quad (A8)$$

E_4 is formed from the terms of even degree in P_4, while O_4 is formed from those of odd degree. We first consider the case $a_2^2 < 4a_0a_4$, so that E_4 is positive for all values of ω. O_4 is negative for $\omega > (a_1/a_3)^{1/2}$ but positive for $\omega < (a_1/a_3)^{1/2}$. When ω has the large value r_0, O_4 is thus large and negative while E_4 is large and positive. The argument ϕ of P_4 is now given by

$$\tan\phi = \frac{O_4}{E_4} = \frac{\omega\,(a_1 - a_3\omega^2)}{a_0 - a_2\omega^2 + a_4\omega^4} \qquad (A9)$$

We have already seen that ϕ is approximately 2π when $\omega = r_0$; (A9) shows that $\tan\phi$ then has the small negative value$-\ a_3/(a_4r_0)$, so that ϕ is in fact slightly less than 2π.

From (A9) we see that ϕ cannot pass through an odd multiple of $\pi/2$ unless E_4 is zero and it cannot pass through an even multiple of $\pi/2$ unless O_4 is zero. The detailed implications of this for the case under consideration, when E_4 is always positive and ω steadily decreases, are as follows. E_4 remains positive and O_4 negative until ω reaches the value $(a_1/a_3)^{1/2}$; ϕ therefore remains between 2π and $3\pi/2$ returning to the value 2π when $\omega = (a_1/a_3)^{1/2}$. When $\omega < (a_1/a_3)^{1/2}$, E_4 and O_4 are both positive so that ϕ is between 2π and $5\pi/2$; for small values of ω, however, O_4 is small whereas E_4 is approximately a_0, so the final value of ϕ is 2π, and P_4 has one complex root with positive real part in the first quadrant, and therefore two complex roots with positive real parts altogether. The general behaviour of ϕ as p descends the imaginary axis is sketched roughly in Fig. 6.

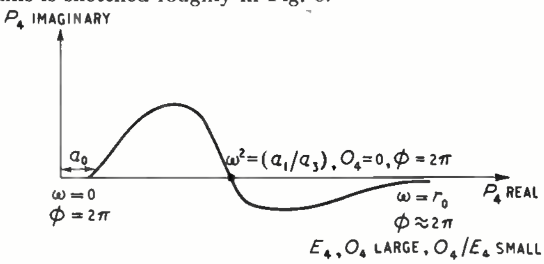


Fig. 6. *Rough sketch of behaviour of P_4 when p descends the imaginary axis and E_4 has no real zeros.*

Now suppose that $a_2^2 > 4a_0a_4$, so that there are two real positive values of ω for which E_4 is zero. If these values of ω and ω_1 and ω_2 $(\omega_2 > \omega_1 > 0)$, E_4 is positive for $\omega > \omega_2$, negative for $\omega_2 > \omega > \omega_1$ and positive for $\omega < \omega_1$, but ω_1 and ω_2 may have any relation to $(a_1/a_3)^{1/2}$. If $\omega_2 > (a_1/a_3)^{1/2} > \omega_1$, we have, as before, that the argument of ϕ of P_4 is slightly less than 2π when $\omega = r_0$, but now it descends to $3\pi/2$ when $\omega = \omega_2$ since for $\omega = \omega_2$, E_4 is zero while O_4 is negative. For $\omega_2 > \omega > (a_1/a_3)^{1/2}$ E_4 and O_4 are both negative, so that ϕ is between $3\pi/2$ and π, reaching the latter value when $\omega = (a_1/a_3)^{1/2}$. Similarly, ϕ is between π and $\pi/2$ for $(a_1/a_3)^{1/2} > \omega > \omega_1$, and ϕ becomes $\pi/2$ when $\omega = \omega_1$. Lastly, for $\omega < \omega_1$, ϕ lies between $\pi/2$ and 0 and ϕ clearly tends to zero with ω. This is sketched roughly in Fig. 7.

In this case, therefore, P_4 has no complex root in the first quadrant; it must therefore have no complex root in the fourth quadrant. If the a terms are all positive, P_4 clearly has no real positive zero, so that P_4 must be entirely free from roots with positive real parts.

It remains to consider the cases in which E_4 is zero for two real positive values ω_1, ω_2 of ω $(\omega_2 > \omega_1 > 0)$ but either $\omega_2 > \omega_1 > (a_1/a_3)^{1/2}$ or $(a_1/a_3)^{1/2} > \omega_2 > \omega_1$. By a similar argument to the foregoing we can show that P_4 must have at least one zero with positive real and positive imaginary part in either of these cases, and therefore at least one pair of conjugate complex roots with positive real parts.

In the general case the only differences are (i) the argument ϕ of P_n will increase by $\frac{1}{2}n\pi$ instead of by 2π as p describes the circular part of its path in Fig. 2,

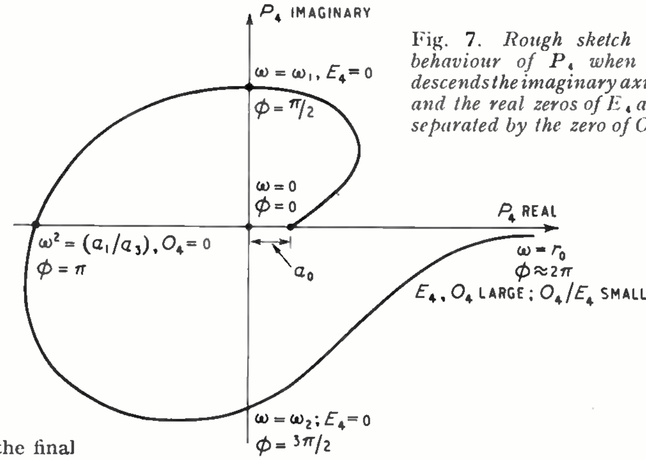


Fig. 7. *Rough sketch of behaviour of P_4 when p descends the imaginary axis, and the real zeros of E_4 are separated by the zero of O_4.*

(ii) E_n and O_n will have a number of roots, and there will therefore be more possible cases to consider. The result, however, is quite clear and definite, and involves the three conditions (a), (b) and (c) already given in Section 4.

We have disregarded the possibility that E_n or O_n in (16) might have equal zeros, for clearly the condition of separation (c) in Section 4 is then violated unless a repeated zero of E_n is also a zero of O_n or vice versa. In this case P_n has a purely imaginary zero, whereas the practical requirement is that all the zeros of P_n should have negative real parts.

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(*To be concluded*)

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.

Non-Linearity Distortion

SIR,—I was particularly interested in your January 1956 Editorial comments on non-linearity and its cancellation.

There is one point in connection with even-order harmonic distortion, however, namely, that the assymetry of the output wave produces a d.c. component. This is not transmitted by normal types of interstage coupling, so that a balance arrived at under steady-state conditions would not be maintained accurately under normal programme conditions, where the syllabic variation of the d.c. component would be passed (with a certain amount of attenuation and phase-shift). This in turn would cause a variation in the mean bias of the following stage.

Distortion cancellation is quite a practical proposition in a.m. radio transmitters employing low-level modulation. Where the r.f. driver is a screen-modulated pentode it is possible, by suitable adjustment of the grid drive, to produce a modulation characteristic which is the exact inverse of the linearity curve of the class B final stage, thereby reducing the distortion to a negligible amount.

The adjustment is easily and quickly made using an oscilloscope, and the balance remains exact under all conditions of syllabic variation of the modulation level.

Wanganin, D. C. SUTHERLAND
New Zealand.
6th March 1956.

[In our January Editorial, we assumed a valve with a characteristic of

$$i_a = 0{\cdot}01 + 0{\cdot}0016\,V + 0{\cdot}00006\,V^2$$

and we let $V = -5 + 4 \sin \omega t$, giving

$$i_a = 0{\cdot}00398 + 0{\cdot}004 \sin \omega t - 0{\cdot}00048 \cos 2\omega t$$

We assumed that the a.c. component only was handed on to an identical stage by a coupling with a resistive impedance of 1 kΩ, so that for the second stage

$$V = -5 - 4 \sin \omega t + 0{\cdot}48 \cos 2\omega t$$

and we found that this gave for the anode current of the second stage

$$i_a = 0{\cdot}003987 - 0{\cdot}004115 \sin \omega t - 0{\cdot}0001152 \sin 3\omega t + 0{\cdot}000006912 \cos 4\omega t$$

Now if, in the first stage, the input were a sine wave of only 2-volts peak then, for this stage, $V = -5 + 2 \sin \omega t$ and the anode current is

$$i_a = 0{\cdot}00362 + 0{\cdot}002 \sin \omega t - 0{\cdot}00012 \cos 2\omega t$$

If now the input is suddenly changed from 2 volts to 4 volts, there is a sudden change of the mean anode current from 0·00362 A to 0·00398 A, which is 0·00036 A. This is handed on to the next stage as a change of mean grid voltage of 0·36 volt.

Immediately after the change, therefore, the second stage has

$$V = -5{\cdot}36 - 4 \sin \omega t + 0{\cdot}48 \cos 2\omega t$$

Inserting this in the equation for anode current gives

$$i_a = 0{\cdot}002674688 - 0{\cdot}003712 \sin \omega t - 0{\cdot}000020736 \cos 2\omega t - 0{\cdot}0001152 \sin 3\omega t + 0{\cdot}000006912 \cos 4\omega t$$

Comparing the two expressions, the amplitudes of the third and fourth harmonics are unchanged. The second harmonic is not zero and the fundamental is smaller. Instead of 2·8% third harmonic and 0·168% fourth, we have 0·56% second, 3·1% third and 0·186% fourth harmonic.

The distortion cancellation thus holds fully only under steady-state conditions and is momentarily upset during the transient which occurs on a change of amplitude. A d.c. coupling under the assumed conditions would give even worse results, but that is only because the conditions are not the right ones for d.c. coupling. The ideal is obviously a d.c. coupling and valve characteristics so chosen that, in combination, the overall static characteristic is linear. Nevertheless, even with a.c. coupling, the distortion under transient conditions, although greater than in the steady state, is still less than if there were no cancellation at all. W.T.C.]

Practical Transistor Circuits

SIR,—Some of the most interesting engineering possibilities of the transistor are associated with its small physical size: is it right then to conclude, as in your February Editorial, that a practical value for the circuit constant C_e is 35 μF?

This represents a component physically many times larger in volume than the small-sized transistors which are now or will presently be available. Similarly, one finds that practical designs for output transformers associated with transistors, having collector dissipation in the watt range, are considerably bigger than the transistors themselves.

SMALL TRANSISTOR
3 × 4 × 7mm
84 cu. mm

T.C.C CE 87B
50μF. 12VOLTS
11·5mm DIA. × 41·5mm LONG
4,300 cu. mm

The optimum design of a transistorized circuit will be indicated by the answers to these queries:

(a) Is it physically possible to produce 'large' electrolytic or tantalum capacitors of the order of 35 μF, for a few volts working level, in the lineal size range of the smallest transistors?

(b) Can we look forward to smaller transformers through the more effective use of materials of higher flux density and higher permeability than are now available commercially?

(c) Alternatively, shall we accept less gain per transistor stage, using a larger number of transistors, in order to have smaller magnitudes for the values of the coupling components?

Future practice may be represented by (c), just as with small pentode valves we are usually prepared to accept a stage amplification of × 100, although, as is well known, × 1000 or nearly the full valve amplification factor itself can be realized, at the expense of an inconveniently large anode load and high supply voltage.

Fortiphone Ltd., D. L. JOHNSTON
London, W.1.
16th February 1956.

AUTOMATION

The Proceedings of the Radio-Electronics-Television Manufacturers' Association Symposium on Automation is a record of the papers presented at the symposium held in the University of Pennsylvania, Philadelphia, Pennsylvania, U.S.A., on 26th and 27th September 1955. Copies cost $5 from Engineering Publishers, G.P.O. Box 1151, New York 1, N.Y., U.S.A.

NEW BOOKS

Services' Textbook of Radio; Vol. 3, Electronics

By J. Thomson, M.A., D.Sc., F.Inst.P., M.I.E.E. Edited by the Technical Staff of *Wireless World*. Pp. 259 + ix. H.M. Stationery Office, York House, Kingsway, London, W.C.2. Price 12s. 6d.

This book is one of a series of seven volumes which will cover the whole field of radio and radar. Although the third volume of the series, it is the first to be published.

"Electronics" deals mainly with the thermionic valve, but c.r. tubes, photo-electric devices and semiconductor 'valves' are treated also. Certain sections are marked for omission on a first reading and the level of treatment is then suited to the beginner; when read to include these marked parts, and still more with the appendixes, the treatment reaches a fairly advanced level.

***Wireless & Electrical Trader* Year Book 1956** (27th Edn)

Pp. 344. Trader Publishing Co. Ltd., Dorset House, Stamford Street, London, S.E.1. Price 12s. 6d.

Hi-Fi Loudspeakers and Enclosures

By Abraham B. Cohen. Pp. 360 + vii. John F. Rider Publisher Inc., 480 Canal Street, New York 13, N.Y., U.S.A. Price $4.60.

Introduction to Color TV (2nd Edn)

By M. Kaufman and H. Thomas. Pp. 156 + iv. John F. Rider Publisher Inc., 480 Canal Street, New York 13, N.Y., U.S.A. Price $2.70.

Multivibrators

Edited by Alexander Schure, Ph.D., Ed.D. Pp. 48 + iv. John F. Rider Publisher Inc., 480 Canal Street, New York 13, N.Y., U.S.A. Price 90 cents.

Color Television Standards

Edited by Donald G. Fink. Pp. 520 + xii. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price 64s.

This book comprises papers selected from the Proceedings of the Second National Television System Committee, whose deliberations from January 1950 to July 1953 resulted in the present American colour television standards.

Electronics

By A. W. Keen, M.I.R.E., A.M.I.E.E. Pp. 256. Ward Lock & Co. Ltd., 143 Piccadilly, London, W.1. Price 25s.

This is a semi-popular book for readers with "a sound knowledge of elementary electricity and magnetism". In spite of its title, it covers almost everything in which radio valves are employed, including radar.

High Vacuum Technique (3rd Edition, revised)

By J. Yarwood, M.Sc., F.Inst.P. Pp. 208 + viii. Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 25s.

PREMIUMS FOR TECHNICAL WRITING

The Radio Industry Council has awarded six premiums of 25 guineas each for articles published in the public technical press during 1955. They are to:—

A. W. M. Coombs, for his article "'Memory' Systems in Electronic Computers", which appeared in *Communications and Electronics* for March 1955.

R. A. Bracewell, for his article "An Infra-Red Radiation Pyrometer", which appeared in *Electronic Engineering* for June 1955.

Alan E. Crawford, for his article "Progress in High Power Ultrasonics", which appeared in *British Communications and Electronics* for August and September 1955.

A. H. Beck, T. M. Jackson and J. Lytollis, for their article "A Novel Gas-Gap Speech Switching Valve", which appeared in *Electronic Engineering* for January 1955.

The remaining two premiums have been awarded jointly to the authors of three articles which appeared in the *Post Office Electrical Engineers' Journal* for July and October 1955.

I. A. Ravenscroft and R. W. White, "A Frequency Modulator for Broad-Band Radio Relay Systems".

R. W. White and J. S. Whyte, "Equipment for Measurement of Inter-Channel Crosstalk and Noise on Broad-Band Multi-Channel Telephone Systems".

J. S. Whyte, "An Instrument for the Measurement and Display of V.H.F. Network Characteristics".

The awards were made at a luncheon held by the Public Relations Committee of the Radio Industry Council on 8th March 1956.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for February 1956

Date 1956 February	Frequency deviation from nominal: parts in 10^8	
	MSF 60 kc/s 1429-1530 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.
1	−0·1	+1
2	−0·2	0
3	−0·2	0
4	−0·1	0
5	−0·1	+2
6	−0·1	+2
7	−0·1	+1
8	−0·1	+1
9	−0·1	+2
10	−0·2	+2
11	−0·2	+4
12	−0·2	+2
13	−0·2	+2
14	−0·1	+2
15	−0·2	+2
16	−0·2	+2
17	−0·1	+2
18	−0·1	+3
19	−0·1	+4
20	−0·1	+4
21	−0·1	+5
22	−0·1	0
23	0·0	0
24	−0·1	0
25	−0·1	0
26	−0·1	−2
27	−0·1	−2
28	−0·1	−2
29	−0·2	−2

The values are based on astronomical data available on 1st March 1956.

ABSTRACTS and REFERENCES

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research and published by arrangement with that Department.

The abstracts are classified in accordance with the Universal Decimal Classification. They are arranged within broad subject sections in the order of the U.D.C. numbers, except that notices of book reviews are placed at the ends of the sections. U.D.C. numbers marked with a dagger (†) must be regarded as provisional. The abbreviations of journal titles conform generally with the style of the World List of Scientific Periodicals. An Author and Subject Index to the abstracts is published annually; it includes a selected list of journals abstracted, the abbreviations of their titles and their publishers' addresses.

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ACOUSTICS AND AUDIO FREQUENCIES

534.12 961

The Principal Frequencies of Vibrating Systems with Elliptic Boundaries.—S. D. Daymond. (*Quart. J. Mech. appl. Math.*, Sept. 1955, Vol. 8, Part 3, pp. 361–372.) Analysis involving Mathieu functions is presented.

534.121.1 962

On Upper and Lower Bounds of the Eigenvalues of a Free Plate.—Y. Nakata & H. Fujita. (*J. phys. Soc. Japan*, Sept. 1955, Vol. 10, No. 9, pp. 823–824.) A brief mathematical note on the vibration of a thin elastic plate.

534.2-8 963

New Methods for rendering Ultrasonic Waves Visible.—F. Hauer & G. Keck. (*Naturwissenschaften*, Nov. 1955, Vol. 42, No. 22, pp. 601–602.) Both travelling and standing waves in liquids can be made visible by using a specially prepared photographic plate; the sensitive material is dissolved at a rate depending on the strength of the ultrasonic field.

534.612.082.72-8 964

Capacitance Analyser of Sound Field.—V. A. Zverev, V. M. Bokov & I. E. Lur'e. (*Akust. Zh.*, July–Sept. 1955, Vol. 1, No. 3, pp. 218–220.) A circuit arrangement is described in which a capacitance change corresponding to the fractional change of the dielectric constant of water due to a sound field of frequency 85 kc/s is caused to modulate a r.f. signal; the change in the dielectric constant is of the order of 1 part in 10^6 for a pressure change of 10^5 bar.

534.833 : 534.75 965

Apparatus for the Protection of the Hearing from the Damaging Effects of Noise.—I. M. Polkovski. (*Akust. Zh.*, July–Sept. 1955, Vol. 1, No. 3, pp. 249–256.) Low-pass mechanical sound filters to be worn over the ears are described. Their attenuation is low up to 1 kc/s but increases rapidly to about 35 dB at 3·5 kc/s.

534.845 966

Measurement of the Absorption Coefficient of Acoustic Materials by the Reverberation-Chamber Method.—R. Lamoral. (*Ann. Télécommun.*, Oct. 1955, Vol. 10, No. 10, pp. 206–217.) Formulae for reverberation time developed by Sabine, Eyring and Millington are compared; values calculated from the two latter formulae may differ by as much as 25% for fairly 'dead' studios. Details are given of the test method and installation used by the R.T.F. organization; the volume of the reverberation chamber is 246 m³, and the excitation sound is a frequency-modulated tone. An electro-optical system for interrupting the sound without producing transients is described. Measurements are reported on panels of various materials either in one piece or divided up and variously arranged; material is most effective when located near the microphone. The results are embodied in graphs suitable for design purposes.

534.86 967

Conference on Electroacoustics, Kiev, 1st–5th July 1955.—(*Akust. Zh.*, July–Sept. 1955, Vol. 1, No. 3, pp. 294–296.) Brief notes on 17 papers presented at the conference.

534.86 : 534.76 968

Secondary Field in Stereophonic Two-Channel Transmission for Various Positions of the Sound Source in the Primary Field.—P. G. Tager. (*Akust. Zh.*, July–Sept. 1955, Vol. 1, No. 3, pp. 286–293.) The effect of varying one or more parameters in a system comprising two separate electroacoustic transmission channels on the stereophonic effect at various positions in an auditorium is considered theoretically, reverberation effects being neglected. The principal parameters are the differences of energy levels and time. These are taken into account in calculations by (*a*) the stereophonic-transmission factor $\beta\mu$ and (*b*) the delay factor τ, where β is a function of the directivity of the sound source, microphones and loudspeakers, the amplification factors of the two channels and the constructional factors of the microphones and loudspeakers, and μ is the ratio of the distances between the source and the microphones. Curves and tables are given to facilitate calculations.

534.861.1 969

Broadcast Studio Redesign.—L. L. Beranek. (*J. Soc. Mot. Pict. Telev. Engrs*, Oct. 1955, Vol. 64, No.

10, pp. 550–559.) "A review is made of psychoacoustic and audience-opinion information of recent and older sources from which criteria for studio design are drawn. New criteria are proposed for reverberation time for studios and auditoriums used for speech, music and general purposes. Examples of three types of existing studios that need revision are discussed."

621.395.61 970
A New Cardioid Microphone.—N. Friedman & C. Macpherson. (*Tele-Tech & Electronic Ind.*, Oct. 1955, Vol. 14, No. 10, pp. 70–72..133.) A cardioid polar diagram and flat frequency characteristic are achieved by an arrangement of entrance ports and acoustically coupled cavities which also substantially eliminates proximity effect and susceptibility to mechanical shock.

621.395.612.4 971
Design of a [pressure-type] **Ribbon Microphone.**—A. V. Rimski-Korsakov. (*Akust. Zh.*, July–Sept. 1955, Vol. 1, No. 3, pp. 257–263.) Design formulae are given.

AERIALS AND TRANSMISSION LINES

621.315.213 972
Polythene-Insulated Video-Pair Cables.—(*Elect. Commun.*, Sept. 1955, Vol. 32, No. 3, pp. 165–168.) Constructional features and electrical characteristics are described for cables with four, six and eight video pairs.

621.315.687 973
Electrical Resistance of Joints in Copper Cables.—K. Sagel. (*Nachrichtentech. Z.*, Oct. 1955, Vol. 8, No. 10, pp. 541–544.) Experimental results show that soldered joints (solder composition 40% Sn, 2·7% Sb, remainder Pb) have a lower resistance, and lower scatter of resistance values, than spliced or welded joints. Results are tabulated and microphotographs are reproduced.

621.372 974
A Method of launching Surface Waves.—J. D. Lawson. (*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, p. 111.) The principle used is to maintain the phase velocity of the wave constant throughout the launching system; tapered-horn arrangements are illustrated.

621.372.21 : 621.315.213 975
Strip Transmission Lines.—C. Bowness. (*Electronic Engng*, Jan. 1956, Vol. 28, No. 335, pp. 2–7.) Practical production techniques for strip transmission lines are investigated and the construction of typical component parts of a r.f. system is described. The system has advantages, especially as regards cost, in cases where some stray radiation can be tolerated.

621.372.22 976
A Transmission-Line Taper of Improved Design.—R. W. Klopfenstein. (*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, pp. 31–35.) Discussion of the design of a line with Dolph-Tchebycheff taper, which is optimum in the sense that for a given taper length the reflection coefficient is minimum throughout the pass band, while for a specified upper limit to the reflection coefficient the length of the taper is a minimum. Values of transcendental functions used in the design are tabulated.

621.372.22 : 621.314.2 977
A Tapered Strip Line for Pulse Transformer Service.—Primozich, Schatz & Woodford. (See 993.)

621.372.8 978
Circular and Rectangular Waveguides with Longitudinal Diaphragms.—E. G. Solov'ev. (*Zh. tekh. Fiz.*, April 1955, Vol. 25, No. 4, pp. 707–710.) The propagation of e.m. waves in waveguides with infinitely thin conducting diaphragms, equally spaced along the guide, is considered. The problem is solved by the method of matching the fields at the boundary between two regions. The electric field between the diaphragms is evaluated approximately. One of the dispersion equations obtained for the propagation constant is analysed for the case when the length of the wave in the waveguide is much greater than the space-period of the diaphragm structure.

621.372.8 : 537.226 979
Media rendered Artificially Anisotropic.—Ya. B. Fainberg & N. A. Khizhnyak. (*Zh. tekh. Fiz.*, April 1955, Vol. 25, No. 4, pp. 711–719.) An anisotropic medium can be obtained by arranging dielectric disks periodically along the axis of a waveguide, as suggested by Harvie (1744 of 1949). A mathematical discussion shows that the system is equivalent to a waveguide completely filled with an anisotropic dielectric, not only in respect of retarding properties but also in respect of the mean values of the electric and magnetic fields.

621.396.67 980
Distribution of Current along a Cylindrical Transmitting Aerial.—P. Poincelot. (*Ann. Télécommun.*, Sept. & Oct. 1955, Vol. 10, Nos. 9 & 10, pp. 186–194 & 219–228.) Detailed analysis. See 932 of 1955 and back references.

621.396.67 981
Method for Calculation of the [input] **Conductance of Cylindrical Aerials.**—G. Barzilai. (*Alta Frequenza*, Aug./Oct. 1955, Vol. 24, Nos. 4/5, pp. 339–355.) The method consists essentially in evaluating separately the input voltage and the radiated power for a given value of current in the vicinity of the maximum point. First and second-order calculations are made. For an abbreviated account in English see *Trans. Inst. Radio Engrs*, Jan. 1955, Vol. AP-3, No. 1, pp. 29–32.

621.396.67.012.12 982
Radiation Diagrams of Ring Arrays with Rotationally Symmetrical Horizontal Characteristics.—H. W. Fastert. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1955, Vol. 7, Nos. 9/10, pp. 157–165.) Long-wave aerial systems with sky-wave suppression are discussed. An examination is made of the number of radiators required in practice to give a constant-current system. Analysis for multi-ring systems is simplified by stipulating additional conditions, e.g. maximum gain.

621.396.677 983
Synthesis of Aerial Arrays.—A. Fernandez Huerta. (*Rev. Telecomunicación, Madrid*, Sept. 1955, Vol. 9, No. 41, pp. 3–13.) A theoretical study of linear uniformly spaced arrays. By resolving the characteristic polynomial into binomial expressions the radiation diagrams are found rapidly; broadside and end-on arrays are thus treated. Use of Tchebycheff polynomials to obtain optimum distribution and of small spacings to obtain superdirectivity are discussed.

621.396.677.3.012.12 984
Design of Line-Source Antennas for Narrow Beam Width and Low Side Lobes.—T. T. Taylor. (*Trans. Inst. Radio Engrs*, Jan. 1955, Vol. AP-3, No. 1, pp. 16–28.) The effect of nonuniform field distributions across the aerial aperture on the radiation pattern is discussed mathematically. The ideal compromise between beam width and side-lobe level is given by a space factor $\cos \pi\sqrt{u^2-A^2}$, where $u = (2a/\lambda)\cos\theta$, θ being the angle of radiation and a the half-length of the line source, and $\cosh \pi A$ is the side-lobe ratio. Practical approximations to this ideal pattern are discussed.

621.396.677.833 985

Double-Parabolic-Cylinder Pencil-Beam Antenna.—R. C. Spencer, F. S. Holt, H. M. Johanson & J. Sampson. (*Trans. Inst. Radio Engrs*, Jan. 1955, Vol. AP-3, No. 1, pp. 4–8.) Radiation from a point source on the focal line of one cylinder is reflected in turn from this cylinder and from a second one arranged so that its focal line coincides with the directrix of the first, resulting in the production of a parallel beam. Production advantages of such systems are stressed.

AUTOMATIC COMPUTERS

681.142 986

Electronic Computing Machines and their Uses.—J. H. Wilkinson. (*J. sci. Instrum.*, Nov. 1955, Vol. 32, No. 11, pp. 409–415.) Basic design of digital computers is described. Programming for suitable types of problem is discussed and illustrated by examples.

681.142 987

A Simple Electronic Fourier Synthesizer.—H. B. Mohanti & A. D. Booth. (*J. sci. Instrum.*, Nov. 1955, Vol. 32, No. 11, pp. 442–444.) An experimental model of a machine for two-dimensional Fourier synthesis is described which incorporates magnetic drum storage of the required sine and cosine functions.

681.142 988

Simplified Analog Computer.—V. B. Corey. (*Electronics*, Jan. 1956, Vol. 29, No. 1, pp. 128–131.) A differential analyser is described, based on ten identical operational amplifiers which may be used for integration, multiplication, etc., as required by the particular problem. The arrangement is very flexible and may be used with external circuitry for the generation of special functions required in synthesizing particular input analogues.

681.142 : 621.375.2.024 989

Some Aspects of the Design of a D.C. Amplifier for Use with a Slow Analogue Computer.—H. Fuchs. (*Electronic Engng*, Jan. 1956, Vol. 28, No. 335, pp. 22–25.) "The errors introduced as a result of an h.t. resistance common to several amplifiers is calculated, and a criterion for the accuracy of a linear computer stated."

681.142 : 621.375.43 990

Transistor Amplifiers for Use in a Digital Computer.—Simkins & Vogelsong. (See 1014.)

CIRCUITS AND CIRCUIT ELEMENTS

621.3.011.21 991

Contribution on the Calculation of the Total Impedance of Parallel-Connected Impedances.—W. Boesch. (*Rev. gén. Élect.*, Oct. 1955, Vol. 64, No. 10, pp. 517–520.) A method of calculation involving use of a special transparent rule and a simple nomogram is described and illustrated by a numerical example.

621.3.016.35 : 621-526 992

Stability Criteria for an Electrical or Mechanical System with Distributed Parameters.—A. S. Gladwin. (*Brit. J. appl. Phys.*, Nov. 1955, Vol. 6, No. 11, pp. 400–402.) For electrical or mechanical systems with distributed parameters, or with an element producing a finite time delay, to be stable all the roots of the appropriate characteristic equation must be negative or have negative real parts. If the equation is expressed in the form $C_0 + C_1z + (C_2 + C_3z) \tanh z = 0$, then for C_0 positive, C_1 and C_3 must be positive and C_2/C_1 must be greater than a certain critical value which is a function of C_0/C_3. An example is given of the application of these criteria to a servomechanism.

621.314.2 : 621.372.22 993

A Tapered Strip Line for Pulse Transformer Service.—F. G. Primozich, E. R. Schatz & J. B. Woodford. (*Elect. Engng, N.Y.*, Oct. 1955, Vol. 74, No. 10, p. 908.) Digest of paper in *Trans. Amer. Inst. Elect. Engrs*, Part 1, *Communication and Electronics*, May 1955, Vol. 74, pp. 158–161. A rolled-up construction is described for an exponential-line impedance transformer. The impedance is 25 Ω at one end of the 10-m line and 1 000 Ω at the other; pulses as long as 36 mμs can be handled.

621.316.825 + [621.314.632 : 546.289 994

The Turnover Phenomenon in Thermistors and in Point-Contact Germanium Rectifiers.—R. E. Burgess. (*Proc. phys. Soc.*, 1st Nov. 1955, Vol. 68, No. 431B, pp. 908–917.) The turnover power in thermistors increases as the square of the absolute ambient temperature. The similarity in form between the reverse characteristic of a point-contact Ge rectifier and the static characteristic of a thermistor suggests a thermal hypothesis for rectifier turnover; but in the latter case the turnover power decreases approximately linearly with ambient temperature. The type of function necessary to reproduce the rectifier characteristics at turnover is discussed; such a function cannot be derived from existing theories of rectification.

621.318.424 : 621.318.134 995

A Survey of the Application of Ferrites to Inductor Design.—R. S. Duncan & H. A. Stone, Jr. (*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, pp. 4–13.)

621.318.435 : 621.375.3 996

Influence of I.D.-O.D. [inside-diameter/outside-diameter] **Ratio on Magnetic Properties of Toroidal Cores.**—R. W. Roberts & R. I. Van Nice. (*Elect. Engng, N.Y.*, Oct. 1955, Vol. 74, No. 10, pp. 910–914.) The optimum design of cores for magnetic amplifiers is discussed. A theoretical result is obtained assuming a parallelogram-shaped hysteresis loop, and an experimental investigation is reported. The dependence of the optimum ratio on the magnetic properties of the core material is indicated.

621.318.57 + 621.314.7 + 621.385 997

Design of Electronic Devices for Production.—Thomson. (See 1249.)

621.319.4 : 621.387 998

Electrically Variable Gas-Dielectric Capacitor.—J. F. Gordon. (*Electronics*, Jan. 1956, Vol. 29, No. 1, pp. 158–160.) A neon tube placed between the plates of an air-dielectric capacitor gives a variation in dielectric constant proportional to the tube current. Maximum variation of capacitance is about 10%. The device may be used for frequency control of self-excited oscillators.

621.372 999

A Simplified Method of solving Linear and Nonlinear Systems.—R. Boxer & S. Thaler. (*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, pp. 89–101.) The method described uses the z transform [363 of 1955 (Ragazzini & Bergen)] directly to determine the responses of linear and nonlinear systems approximately. The solutions are given as time series representing the values of the response at equally spaced instants. Both constant and time-varying systems are considered.

621.372.413 1000

The Harmodotron—a Beam Harmonic Higher-Order-Mode Device for producing Millimeter and Submillimeter Waves.—P. D. Coleman & M. D. Sirkis. (*J. appl. Phys.*, Nov. 1955, Vol. 26, No. 11, pp. 1385–1386.) High-power oscillations are generated by exciting a single higher-order TM mode in a cylindrical cavity by means of a bunched beam of 1–1·5-MeV electrons.

621.372.542.2.01 **1001**

Simultaneous Approximation for the Amplitude and Delay Time of an Ideal Low-Pass Filter using the Flow Analogy.—J. Peters. (*Arch. elekt. Übertragung*, Oct. 1955, Vol. 9, No. 10, pp. 453–459.) Analogies between transmission phenomena in networks and current flow in a potential field are discussed [2370 of 1951 (Darlington)]. A distortion-free low-pass filter can be represented by two superposed fields, one with parallel flow, corresponding to the distortion-free aspect, and the other with radial flow, corresponding to the filter aspect. Such a representation involves a finite residual error which cannot be eliminated without introducing instability, but which can be reduced indefinitely by increasing the delay time.

621.372.543.2 : 538.652 **1002**

Electromechanical Filters for 100-kc/s Carrier and Sideband Selection.—R. W. George. (*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, pp. 14–18.) Torsional-type filters comprising $\lambda/2$ cylindrical resonators with $\lambda/4$ coupling necks are discussed. Descriptions are given of two filters with bandwidths of 50 c/s and 3·1 kc/s respectively. Using a particular Fe-Ni-Cr-Ti alloy known as Ni-Span-C, the temperature variation of frequency is $<$ 2 parts per million. Magnetostrictive transducers made of this alloy or of ferrite are used. Spurious modes of vibration are reduced by a balanced arrangement of the transducers and by mechanical damping in the end supports.

621.372.56 **1003**

A Push-Button Attenuator for Frequencies up to 10 Mc/s.—H. Paardekooper. (*Commun. News*, Oct. 1955, Vol. 16, No. 1, pp. 10–22.) Details are given of a resistance attenuator providing attenuation up to 99·5 dB in 1-dB steps with an interpolation step of 0·5 dB. Each section comprises a T-network with the shunt arm bridged by capacitance to compensate the inductance of the series arms.

621.373.4.029.65 **1004**

Cherenkov Radiation.—J. G. Linhart. (*Research, Lond.*, Oct. 1955, Vol. 8, No. 10, pp. 402–406.) A brief explanation is given of the mechanism of Čerenkov radiation, which may find application in mm-λ oscillators.

621.373.421 **1005**

Cathode-Follower Phase-Shift Oscillator.—J. C. Samuels. (*Elect. Commun.*, Sept. 1955, Vol. 32, No. 3, pp. 198–202.) Analysis indicates that better frequency stability is obtained by connecting the phase-shifting network between grid and cathode than by connecting it in the anode circuit. Calculations are made for three typical circuits.

621.373.431.1 **1006**

Investigations of the D.C. and A.C. Characteristics of Bi-stable Multivibrators.—K. Gosslau H. J. Harloff. (*Nachrichtentech. Z.*, Oct. 1955, Vol. 8, No. 10, pp. 521–530.) The characteristics are discussed theoretically with reference to the design of reliable counting and switching units. Design procedure is illustrated by two numerical examples, one of a binary counting unit with counting rate up to 1 Mc/s at supply voltages between 50 and 350 V, the other for a pulse-counting rate of 10 Mc/s at supply voltages between 60 and 300 V. A modification of the latter circuit using a Type-E88CC double-triode, instead of Type-E90CC, was used at counting rates up to 30 Mc/s.

621.374.3 : 621.318.57 **1007**

The Diode Pump Integrator.—J. B. Earnshaw. (*Electronic Engng*, Jan. 1956, Vol. 28, No. 335, pp. 26–30.) The design of a diode circuit for integrating a train of pulses is discussed and conditions are established for linear and nonlinear operation. An application to a frequency-sensitive relay with quick 'on-off' characteristics is described.

621.374.32 **1008**

A Logarithmic Voltage Quantizer.—E. M. Glaser & H. Blassbalg. (*Tele-Tech & Electronic Ind.*, Oct. 1955, Vol. 14, No. 10, pp. 73–75..128.) Details are given of an analogue-to-digital converter which provides an output pulse whose duration is proportional to the logarithm of the input voltage, the output pulse duration being determined by counting the number of fixed-frequency pulses occurring in the same interval. Range of input voltage is 3·3–100 V, with a threshold input pulse length of 0·5 μs.

621.374.32 **1009**

Analog-to-Digital Data Converter.—S. Rigby. (*Electronics*, Jan. 1956, Vol. 29, No. 1, pp. 152–155.) The required conversion is performed by arranging an oscillator generating pulses at a repetition rate proportional to the analogue voltage to feed a digital counter for a fixed period. The oscillator described has a range of four decades; stability is better than 0·1% of maximum pulse frequency.

621.375.221 **1010**

The Compensation of Wide-Band Amplifiers at High Frequencies.—P. Rohan & J. Weislitzer. (*Proc. Instn Radio Engrs, Aust.*, Oct. 1955, Vol. 16, No. 10, pp. 354–362.) Analysis is given for various coupling arrangements including *RC*, *RCL* and special two- and four-terminal networks.

621.375.23 : 621.3.018.78 **1011**

Distortion in Feedback Amplifiers.—R. W. Ketchledge. (*Bell Syst. tech. J.*, Nov. 1955, Vol. 34, No. 6, pp. 1265–1285.) Analysis is presented for frequency-dependent feedback. Consideration is restricted to cases where the distortion products are comparatively small and the nonlinear element can be described by a power series with only a few terms. Formulae are derived for a number of third-order products.

621.375.4 : 621.314.7 **1012**

Analysis of the Common-Base Transistor Circuit.—R. L. Pritchard. (*Electronic Engng*, Jan. 1956, Vol. 28, No. 335, p. 40.) Mathematical errors in a paper by Oakes (1600 of 1955) are pointed out.

621.375.4 : 621.314.7 **1013**

Transistor Circuit for Resonance-Coefficient Multiplier.—M. Soldi & M. Valeriani. (*Alta Frequenza*, Aug./Oct. 1955, Vol. 24, Nos. 4/5, pp. 375–389.) Use of transistors in *Q*-multiplier circuits [2678 of 1951 (Harris)] is discussed. The input admittance is calculated and the conditions for stabilization against self-oscillation are derived. Design and performance of experimental arrangements are described; composite transistors [927 of March (Pearlman)] may be used.

621.375.43 : 681.142 **1014**

Transistor Amplifiers for Use in a Digital Computer.—Q. W. Simkins & J. H. Vogelsong. (*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, pp. 43–55.) Pulse-regenerative amplifiers for a 3-Mc/s synchronous binary computer are based on use of external feedback, so that a negative-resistance transistor characteristic is not required. By using semi-gated feedback, allowance can be made for the slow recovery of Ge diodes incorporated in the circuit.

621.372 + 621.317].029.63/.64 1015

Schaltungstheorie und Messtechnik des Dezimeter- und Zentimeterwellengebietes. [Book Review]—A. Weissfloch. Publishers: Birkhäuser, Basel, 1954, 308 pp., Swiss Fr. 29.30. (*Bull. schweiz. elektrotech. Ver.*, 29th Oct. 1955, Vol. 46, No. 22, p. 1092.) Theory and applications of four-, six-, and eight-pole networks are covered. Circle-diagram representation is used to show transformation properties of four-pole networks.

GENERAL PHYSICS

535.325 : 538.566.2.029.6 1016

The Refractive Indices of Water Vapour, Air, Oxygen, Nitrogen and Argon at 72 kMc/s.—K. D. Froome. (*Proc. phys. Soc.*, 1st Nov. 1955, Vol. 68, No. 431B, pp. 833–835.) Experiments indicate that the refractive indices of water vapour, air and oxygen are modified at microwave frequencies in the manner to be expected from dipole theory; the values for nitrogen and argon should not vary with frequency, and this is confirmed.

537.122 1017

The Dielectric Theory of Electronic Interactions in Solids.—J. Hubbard. (*Proc. phys. Soc.*, 1st Nov. 1955, Vol. 68, No. 431A, pp. 976–986.) The long-range part of the Coulomb interaction of the electrons in solids is treated by regarding the solid as a homogeneous dielectric. Sustained oscillations are shown to be possible in metals and are identified with the plasma oscillations discussed by Bohm & Pines (1375 of 1954).

537.21 1018

The Capacity and Field of a Cylindrical Trough with a Plane Conductor in the Axial Plane of Symmetry.—H. J. Peake & N. Davy. (*Brit. J. appl. Phys.*, Nov. 1955, Vol. 6, No. 11, pp. 404–408.)

537.21 : 517.9 1019

Poisson's Partial Difference Equation.—H. Davies. (*Quart. J. Math.*, Sept. 1955, Vol. 6, No. 23, pp. 232–240.) The Green's functions and their asymptotic behaviour for Poisson's partial difference equation in two and three dimensions are discussed and the distribution of potential and current in a regular grid of wires is calculated. The method may also be applied to non-isotropic grids and can be extended to triangular and hexagonal grids in two dimensions and probably to more complicated patterns in three dimensions.

537.311.31 1020

Theory of Electrical Conductivity in Metals.—P. S. Zyryanov. (*Zh. eksp. teor. Fiz.*, Aug. 1955, Vol. 29, No. 2(8), pp. 193–200.) The fluctuations in the potential in the electron-ion plasma of a metal and the electrical resistance due to the scattering of electrons by these fluctuations are calculated.

537.311.31 1021

Number of States and the Magnetic Properties of an Electron Gas.—A. W. Sáenz & R. C. O'Rourke. (*Rev. mod. Phys.*, Oct. 1955, Vol. 27, No. 4, Part I, pp. 381–398.) A theoretical investigation of the thermodynamic properties of conduction electrons in metals.

537.311.33 : 538.614 1022

The Infrared Faraday Effect due to Free Carriers in a Semiconductor.—E. W. J. Mitchell. (*Proc. phys. Soc.*, 1st Nov. 1955, Vol. 68, No. 431B, pp. 973–974.)

537.52 : 546.11 1023

Growth of Pre-breakdown Ionization Currents in Hydrogen.—R. W. Crompton, J. Dutton & S. C. Haydon. (*Nature, Lond.*, 3rd Dec. 1955, Vol. 176, No. 4492, p. 1079.) Experiments are briefly reported, the results of which indicate that the mechanism of primary ionization is the same over a range of pressures from a few mm to 150 mm Hg.

537.52 : 621.317.729.2 1024

Theory of Langmuir Probes.—Yu. M. Kagan & V. I. Perel'. (*Zh. eksp. teor. Fiz.*, Aug. 1955, Vol. 29, No. 2(8), pp. 261–263.) A note contrasting results of Langmuir's theory with those of the authors' precise theory of the spherical probe at low pressures and negative probe potential (*C. R. Acad. Sci. U.R.S.S.*, 1954, Vol. 95, No. 4, pp. 765–768. In Russian).

537.525 : 621.385.833 1025

Electric 'Microdischarges' in a Dynamic Vacuum.—R. Arnal. (*Ann. Phys., Paris*, Sept./Oct. 1955, Vol. 10, pp. 830–873.) Account of an experimental and theoretical investigation of undesired discharges occurring as a result of surface imperfections in vacuum apparatus operated at high voltage. Experimental electrode systems similar to those in electron microscopes were used. The phenomena are distinguishable from field emission, and involve repeated secondary emission of electrons and H ions. For voltages applied in very short pulses the upper limit before onset of microdischarges is higher than for continuous voltages.

537.533 1026

Some Factors influencing Field Emission and the Fowler-Nordheim Law.—T. J. Lewis. (*Proc. phys. Soc.*, 1st Nov. 1955, Vol. 68, No. 431B, pp. 938–943.) "Departures from the normal Fowler-Nordheim law which have been observed at fields up to 7×10^7 V cm^{-1} have previously been ascribed to space-charge field distortion. It is shown that certain other effects due to failure of the image law, non-uniformity of work function and to possible surface irregularities could produce similar effects."

537.533 1027

Electron Emission from Metal Surfaces after Mechanical Working.—J. Lohff & H. Raether. (*Z. Phys.*, 1st Oct. 1955, Vol. 142, No. 3, pp. 310–320.) A fuller account of work described previously (2265 of 1955). Electron multiplication by secondary emission was used to facilitate the observations.

537.533 : 539.211 1028

A Study of the Structure of Abraded Metal Surfaces.—L. Grunberg & K. H. R. Wright. (*Proc. roy. Soc. A*, 8th Nov. 1955, Vol. 232, No. 1190, pp. 403–423.) A report is presented of an experimental investigation of the spontaneous and photoelectric electron emission from abraded surfaces in various atmospheres. All the metals investigated responded to radiation in the near ultraviolet; Al, Mg and Zn, which form 'excess metal' oxides, responded to wavelengths in the visible range also. From the observed rate of decay of emission it is inferred that both thermal and oxidation phenomena are involved. Previous theories that the emission depends on exothermal processes [2301 of 1953 (Kramer)] are considered untenable.

537.56 : 536.2 1029

Theory of Thermal Conductivity of a Plasma.—H. Schirmer. (*Z. Phys.*, 12th Sept. 1955, Vol. 142, No. 2, pp. 116–126.) The treatment is similar to that used previously for investigating the electrical conductivity of plasma (382 of February).

537.562 : 538.6 1030

Particle Transport, Electric Currents, and Pressure Balance in a Magnetically Immobilized Plasma.—P. Stehle. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, p. 443.) The problem discussed by Tonks (2595 of 1955) is treated without detailed analysis of particle trajectories; qualitatively similar results are obtained.

538.222 **1031**

A Note on the Paramagnetic Relaxation.—M. Yokota. (*J. phys. Soc. Japan*, Sept. 1955, Vol. 10, No. 9, pp. 762–768.) The theory given explains experimental results obtained by Gorter (*Paramagnetic Relaxation*, 1947) for the dependence of the spin-spin relaxation time on the static field strength.

538.3 **1032**

Electromagnetic Theory of Moving Matter: Part 2.—E. J. Post. (*Tijdschr. ned. Radiogenoot.*, Sept. 1955, Vol. 20, No. 5, pp. 307–321. In English.) Part 1: 3230 of 1955.

538.311 **1033**

Magnetic Field due to Circular Current.—R. Cazenave. (*Rev. gén. Élect.*, Oct. 1955, Vol. 64, No. 10, pp. 510–512.) Analysis is simplified by applying recent work on elliptic integrals.

538.561 : 537.533 **1034**

Production of Millimetre Waves by a Magnetic Undulator.—R. Combe & T. Frelot. (*C. R. Acad. Sci., Paris*, 28th Nov. 1955, Vol. 241, No. 22, pp. 1559–1560.) Measurements have been made on a device of a type described previously [1037 of 1954 (Combe & Feix) and back references] using an electron beam from a pulsed linear accelerator. With a mean beam intensity of 0·3 μA, the output power was about 1 mW, intermediate between the values corresponding to complete incoherence and complete coherence of the elementary vibrations. The spectrum extended from 4·8 to 8 mm with a maximum at 5·75 mm.

538.566 : 537.226.2 **1035**

Transmission of Electromagnetic Waves through Inhomogeneous Layers.—H. G. Haddenhorst. (*Z. angew. Phys.*, Oct. 1955, Vol. 7, No. 10, pp. 487–496.) Layers are considered in which the dielectric constant is graded from the value unity at each face to a maximum value at the mid-plane, the variation being either linear or exponential. Dielectrics are prepared satisfying this specification by forming ordinary homogeneous materials into a series of wedges or the like, thus producing a space-periodic structure. Reflection-factor measurements made in free space and in waveguides are reported. Variation of reflection factor with layer thickness is shown graphically for moltopren, paraffin and trolitul; variation with angle of incidence is shown for trolitul. The experimental results are compared with values of the reflection factor calculated from solutions of Maxwell's equations for the appropriate conditions. Such structures can give good transmission of microwaves over a wide frequency band.

538.566.029.6 **1036**

Generation of Submillimeter Waves.—H. Motz & K. B. Mallory. (*J. appl. Phys.*, Nov. 1955, Vol. 26, No. 11, p. 1384.) Brief preliminary note describing radiation observed unexpectedly when a bunched high-energy beam of electrons was passed through a simple rectangular waveguide.

538.569.4 : 535.33 **1037**

Pulse Techniques in Microwave Spectroscopy.—R. H. Dicke & R. H. Romer. (*Rev. sci. Instrum.*, Oct. 1955, Vol. 26, No. 10, pp. 915–928.) "Methods are described for exciting gases to states from which they emit coherent spontaneous radiation in the microwave frequency region. The excitation is produced by the application of short pulses of microwave power. The power subsequently radiated by the gas is calculated for several cases, and experimental methods used to detect the radiation are described. The problem of sensitivity is discussed and compared to the sensitivity obtainable in a continuous absorption experiment. The operation of a high resolution microwave spectrometer, which produces lines substantially narrower than the usual Doppler width, is described. The method used to stabilize the klystron used in these experiments, which is a modification of the Pound i.f. method, is given." See also 92 of January (Romer & Dicke).

538.569.4 : 539.152.2 **1038**

Spin Echo Serial Storage Memory.—A. G. Anderson, R. L. Garwin, E. L. Hahn, J. W. Horton, G. L. Tucker & R. M. Walker. (*J. appl. Phys.*, Nov. 1955, Vol. 26, No. 11, pp. 1324–1338.) Methods are discussed for storing information in the form of pulses in a nuclear-magnetic-resonance system, by use of the free-induction spin-echo technique [see also 1962 of 1955 (Fernbach & Proctor)]. The storage capacity in liquids is expressed in terms of the thermal noise of the detecting apparatus, the self-diffusion of the molecules, and the relaxation times. Undesired echoes arising from the interaction of input pulses are eliminated by f.m. and magnetic-field-modulation techniques. Glycerin and solutions of paramagnetic ions in water provide storage times of 10–50 ms, with a storage capacity of the order of 1 000 echoes. Larger capacities expected from liquids with long relaxation times are not realized owing to self-diffusion.

538.632 **1039**

The Hall Effect in Metals at High Frequencies.—B. Donovan. (*Proc. phys. Soc.*, 1st Nov. 1955, Vol. 68, No. 431A, pp. 1026–1032.) Theory is developed in terms of a 'surface Hall coefficient'; calculations are made for a metal with two overlapping energy bands, and the variations with frequency and with field strength are shown graphically for some special cases. At very low frequencies the Hall coefficient is constant; over a range of higher frequencies it increases as the square root of the frequency and then becomes constant again. The frequency variation of the phase angle between current and Hall e.m.f. is also investigated.

538.652 **1040**

Quantum Theory of Magnetostriction.—A. A. Gusev. (*Zh. eksp. teor. Fiz.*, Aug. 1955, Vol. 29, No. 2(8), pp. 181–192.) The theory of magnetostriction in hexagonal single crystals is developed on the basis of a polar model [*ibid.*, 1950, Vol. 19, pp. 256, 261 (Bogolyubov & Tyablikov)] by considering the magnetic and magnetoelastic interaction of electrons in the lattice. The energy spectrum of the crystal at low temperatures, the free energy and the temperature dependence of magnetostriction constants are calculated.

538.652 : 537.312.62 **1041**

Some Consequences of the Influence of Elastic Deformations on Superconductivity: Magnetostriction.—C. Grenier. (*C. R. Acad. Sci., Paris*, 7th Nov. 1955, Vol. 241, No. 19, pp. 1275–1277.) Analysis of conditions at the transition from the normal-conduction to the superconduction state in tin indicates that magnetostriction mainly in the direction of the quaternary axis is to be expected. Conditions for Hg are more complex.

548.0 : 53 **1042**

Effect of Defects on Lattice Vibrations.—E. W. Montroll & R. B. Potts. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 525–543.) "The theory of the effect of localized defects such as impurities, holes, and interstitials on the vibrations of crystal lattices is developed."

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16 **1043**

International Research in Radio Astronomy.—R. L. Smith-Rose. (*Nature, Lond.*, 10th Dec. 1955, Vol. 176, No. 4493, pp. 1110–1111.) Special Reports Nos. 3, 4 and 5 published by the International Scientific Radio Union, and entitled respectively 'Discrete Sources of Extraterrestrial Radio Noise', 'The Distribution of Brightness on the Solar Disk', and 'Interstellar Hydrogen' are noticed.

523.16 **1044**

Results of Observations of Discrete Emitters of Cosmic Radio Emission on Wavelength $\lambda = 3\cdot2$ cm.—N. L. Kaidanovski, N. S. Kardashev & I. S. Shklovski. (*C. R. Acad. Sci. U.R.S.S.*, 1st Oct. 1955, Vol. 104, No. 4, pp. 517–519. In Russian.) Results are given of a determination of the e.m. flux at 3·2-cm λ originating from the Omega nebula (NGC 6618), Orion nebula (NGC 1976), Cassiopeia A (supernova A.D. 369), Taurus A (NGC 1952, supernova A.D. 1054) and a source near the galactic centre Sagittarius X (possibly supernova A.D. 827). When these values are plotted together with observations at other radio wavelengths and, in the case of the Crab nebula, also at optical wavelengths, the results suggest that one mechanism may be responsible for radiation both at optical and radio wavelengths.

523.165 : 550.385 **1045**

The 27-Day Recurrence Tendency of Cosmic-Ray Intensity.—I. J. van Heerden & T. Thambyahpillai. (*Phil. Mag.*, Nov. 1955, Vol. 46, No. 382, pp. 1238–1251.) The 27-day-cycle variations of cosmic-ray intensity precede those of magnetic activity by about five days; it is tentatively suggested that the same solar disturbances are associated with both phenomena.

523.5 : 621.396.11 **1046**

Diurnal Variations in the Number of Shower Meteors detected by the Forward-Scattering of Radio Waves: Part 2—Experiment.—P. A. Forsyth, C. O. Hines & E. L. Vogan. (*Canad. J. Phys.*, Oct. 1955, Vol. 33, No. 10, pp. 600–606.) "The theory developed in Part 1 [738 of March (Hines)] is applied to derive the expected diurnal variations of the meteor signal rate for four showers as observed by means of a particular forward-scatter transmission path (Cedar Rapids—Ottawa). These results are then compared with the experimental signal rates. The good agreement obtained indicates that the approximations inherent in the theory are sufficiently accurate for practical purposes. The results also indicate that very few meteors, if any, are observed under conditions which do not satisfy the requirements for specular reflection."

523.746.5 **1047**

The Decreasing Phase of the Last Solar Cycle, as indicated by Measurements at the Astronomical Observatory at Rome.—T. Fortini. (*R. C. Accad. naz. Lincei*, Sept./Oct. & Nov. 1955, Vol. 19, Nos. 3/4 & 5, pp. 131–136 & 272–276.) Detailed observations are compared with those obtained at Zürich.

551.510.535 **1048**

Thermal Upward Flow in the Ionosphere.—Syun-ichi Akasofu. (*Sci. Rep. Tohoku Univ., 5th Ser., Geophys.*, June 1955, Vol. 6, No. 3, pp. 150–161.) The possibility of atmospheric waves due to heating effects of solar radiation is established; vertical thermal flow in the F_2 region may give rise to ion drift along the magnetic lines of force; this may explain anomalous diurnal variations in the F_2 layer. By applying the theory to observed ionospheric data for the North American zone estimates are obtained for the velocity of the ions and of the thermal flow.

551.510.535 **1049**

The Detection of the *Sq* Current System in Ionospheric Radio Sounding.—E. V. Appleton, A. J Lyon & A. G. Pritchard. (*J. atmos. terr. Phys.*, Oct. 1955, Vol. 7, Nos. 4/5, pp. 292–295.) Anomalies found in the application of Chapman's theory of ionosphere layer formation to E-layer phenomena are explicable if it is accepted that the *Sq* current system modifies the diurnal variation of electron density in the layer, most probably by some form of vertical-drift mechanism. Similar phenomena are found in the F_1 layer.

551.510.535 : 621.396.11 **1050**

The Night-Time Lower Ionosphere as deduced from a Theoretical and Experimental Investigation of Coupling Phenomena at 150 kc/s.—R. W. Parkinson. (*J. atmos. terr. Phys.*, Oct. 1955, Vol. 7, Nos. 4/5, pp. 203–234.) Experimental work over the period March 1953 to January 1954 is described and the results are compared with theoretical results given by Davids & Parkinson (1192 below). Cnrves show seasonal and diurnal variations of the night-time D layer in the coupling region. D–E-region models for the sunset period for July and November 1953 are proposed; the variation of electron distribution during the night is deduced from a recombination-coefficient model suggested by Mitra (*Scientific Report No.* 68, Ionospheric Research Laboratory, State College, Pennsylvania). All available experimental data tend to confirm the theoretical analysis.

551.510.535 : 621.396.11 **1051**

On the Variation of Ionospheric Absorption at Different Stations.—W. R. Piggott. (*J. atmos. terr. Phys.*, Oct. 1955, Vol. 7, Nos. 4/5, pp. 244–246.) "The possibility that the day-to-day changes in absorption at Slough, Swansea and Freiburg are identical is examined statistically. After allowing for known sources of error, it is considered that the residual differences in summer months are too small to be significant, but that it is probable that real differences occur in winter. In both cases the minimum probable correlation for the Swansea–Slough comparison is very high—about 0·92."

551.510.535 : 621.396.11 **1052**

Selective Annotated Bibliography on Ionospheric Propagation.—M. Rigby & M. L. Rice. (*Met. Abstracts Bibliography*, April 1955, Vol. 6, No. 4, pp. 489–547.) Includes some 300 references covering the period 1902–1955.

551.594.5 : 551.51 **1053**

Measurements of the Mean Lifetime of the Metastable ^1S-state of the Oxygen Atom in the Upper Atmosphere during Auroral Displays.—A. Omholt & L. Harang. (*J. atmos. terr. Phys.*, Oct. 1955, Vol. 7, Nos. 4/5, pp. 247–253.) The mean lifetime of the OI ^1S level is found to vary between 0·45 and 0·75 sec.

LOCATION AND AIDS TO NAVIGATION

621.396.93 + 621.396.96 **1054**

Contributions of the Siemens Company to Air-Safety Technique and Air-Navigation Electronics in the Years 1930–1945.—H. J. Zetzmann. (*Frequenz*, Oct. & Nov. 1955, Vol. 9, Nos. 10 & 11, pp. 351–360 & 386–395.) A detailed illustrated account covering radar equipment, altimeters, navigation aids and transmission technique.

621.396.96 **1055**

Shore-Based Radio Aids to Navigation on Centimetric Wavelength.—J. M. F. A. van Dijk, N. Schimmel & E. Goldbohm. (*Tijdschr. ned. Radiogenoot.*, Sept. 1955, Vol. 20, No. 5, pp. 271–279. In English.) "The problem of shore-based radar as a navigational aid for shipping is discussed from the system engineering point of view. A short review of the work carried out in the Netherlands, Belgium and Western Germany is given."

621.396.96 **1056**

A Ramark Beacon for Use with Marine Radars.—J. M. F. A. van Dijk, N. Schimmel & E. Goldbohm. (*Tijdschr. ned. Radiogenoot.*, Sept. 1955, Vol. 20, No. 5, pp. 281–290. In English.) Description of wide-band 3-cm-λ equipment with a range of 30 miles. A diversity system of transmitting aerials is used. The ramark signal can be eliminated from the ship's p.p.i. by switching on the fast-time-constant anti-clutter circuit provided in the marine radar. Results of trials are discussed.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 + 537.311.33 **1057**

Ultrashort Light and Voltage Pulses applied to Silver Halide Crystals by Turbine-Driven Mirror and Spark-Gap Switch.—J. H. Webb. (*J. appl. Phys.*, Nov. 1955, Vol. 26, No. 11, pp. 1309–1314.) Technique for studying electron mobility at room temperature is described.

535.215 **1058**

Photoelectric Yields in the Vacuum Ultraviolet.—W. C. Walker, N. Wainfan & G. L. Weissler. (*J. appl. Phys.*, Nov. 1955, Vol. 26, No. 11, pp. 1366–1371.) Measurements are reported on polycrystalline specimens of Ni, Cu, Pt, Au, W, Mo, Ag and Pd exposed to radiation of wavelengths 473–1 400 Å.

535.215 : 538.221 **1059**

Photoelectron Emission in a Ferromagnetic Material.—A. Z. Veksler. (*Zh. eksp. teor. Fiz.*, Aug. 1955, Vol. 29, No. 2(8), pp. 201–208.) Formulae are derived for the velocity distribution of photoelectrons and the temperature dependence of the photocurrent near the Curie point. The calculations are made on the basis of Vonsovski's *s-d* exchange model (*ibid.*, 1946, Vol. 16, p. 981). Results show that the photocurrent varies quadratically with magnetization.

535.215 : 546.482.21 **1060**

Infrared Quenching of Cadmium Sulfide.—S. H. Liebson. (*J. electrochem. Soc.*, Sept. 1955, Vol. 102, No. 9, pp. 529–533.) "Infrared quenching of CdS has been investigated as a function of applied voltage and both infrared and exciting light intensities. Infrared quenching of photoconductivity due to light of wavelengths shorter than that corresponding to the absorption edge increases as the voltage is increased. The quenching spectrum is found to shift with temperature in approximately the same amount as the corresponding shift of the absorption edge. An explanation of infrared quenching is offered based on infrared freed holes recombining with trapped electrons within the crystal."

535.215 : 546.482.21 **1061**

Nonstationary Processes in Photoconductors: Part 1—Slow Build-Up of Photoconduction in CdS Single Crystals as a Method for Analysis of Disturbed States.—K. W. Böer & H. Vogel. (*Ann. Phys., Lpz.*, 15th Oct. 1955, Vol. 17, No. 1, pp. 10–22.) Observed photoconductivity build-up curves obtained with weak illumination exhibit inflection points from whose positions the concentration of acceptor impurities can be determined.

535.3 **1062**

Optical Properties of Cadmium Sulfide and Zinc Sulfide from 0·6 Micron to 14 Microns.—J. F. Hall, Jr, & W. F. C. Ferguson. (*J. opt. Soc. Amer.*, Sept. 1955, Vol. 45, No. 9, pp. 714–718.)

535.33/.34 : 537.228.1 **1063**

[Optical] **Vibration Spectrum of Piezoelectric Crystals: Part 5—Lithium Sulphate and Potassium Sulphate.**—J. P. Mathieu, L. Couture & H. Poulet. (*J. Phys. Radium*, Oct. 1955, Vol. 16, No. 10, pp. 781–785.) A study was made of the Raman and infrared-absorption spectra of single crystals with various orientations.

535.37 **1064**

Oxidation States of Europium in the Alkaline Earth Oxide and Sulfide Phosphors.—P. M. Jaffe & E. Banks. (*J. electrochem. Soc.*, Sept. 1955, Vol. 102, No. 9, pp. 518–523.)

535.37 **1065**

Preliminary Studies of the Perovskite-Type Ternary Oxides as Luminophors.—S. Terol & R. Ward. (*J. electrochem. Soc.*, Sept. 1955, Vol. 102, No. 9, pp. 524–528.) Report of an investigation of phosphors with $LaAlO_3$ base and various activators.

535.37 **1066**

Yellow Luminescence caused by Titanium in Magnesium Silicate containing Fluoride.—H. Dziergwa & H. Panke. (*Z. Phys.*, 1st Oct. 1955, Vol. 142, No. 3, pp. 259–265.) Quantum yield, temperature dependence, absorption and afterglow are investigated for $MgF_2.MgO.SiO_2$ phosphors containing TiO_2 alone or with MnO.

535.37 : 546.412.84 **1067**

The Influence of the Crystal Structure on the Luminescence of Calcium Silicate (Mn, Pb).—H. Lange & G. Kressin. (*Z. Phys.*, 15th Oct. 1955, Vol. 142, No. 4, pp. 380–386.)

535.376 **1068**

Electroluminescence. Absolute Brightness Waves and Mean Lifetime of Excited Centres.—G. Destriau. (*J. Phys. Radium*, Oct. 1955, Vol. 16, No. 10, pp. 798–800.) The frequency variation of the alternating and steady components of brightness was studied for ZnS with various activators; the dependence of the relative amplitudes of the two components on the mean lifetime of the excited centres was demonstrated.

535.376 **1069**

Thermoluminescence Measurements of Electroluminescent ZnS:Mn Films.—R. E. Freund. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 760–761.) A typical glow curve is shown for a film which emitted light on application of an alternating or direct field; the measurements were made over the temperature range $-196°$ to $+25°C$.

535.376 : 537.534.9 **1070**

Deterioration of Luminescent Phosphors under Positive-Ion Bombardment.—J. R. Young. (*J. appl. Phys.*, Nov. 1955, Vol. 26, No. 11, pp. 1302–1306.) "The reduction of cathodoluminescence efficiency of phosphors has been observed after bombarding with H^+, H_2^+, He^+, Ne^+, N_2, and A^+ ions having energies from 1 to 25 keV. Detectable deterioration could be observed after a bombardment of less that 10^{-9} coul/mm² of ions (5×10^{11} ions/cm²). The deterioration could be annealed out at temperatures between 450° and 700°C. Results indicate that light 25-keV ions probably penetrate 0·1 μ to 0·2 μ into the phosphor." The phosphors studied were mainly ZnS:Ag.

535.376 : 546.472.21 **1071**

Edge Electroluminescence from ZnS Single Crystals.—R. W. Smith. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, p. 760.) Experimental results are presented which indicate that both free electrons and holes may contribute to the electroluminescence of ZnS.

537.226/.227 : 546.431.824-31 **1072**

Single-Crystal Neutron Analysis of Tetragonal $BaTiO_3$.—B. C. Frazer, H. R. Danner & R. Pepinsky. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 745–746.)

537.227/.228.1 **1073**

Properties of Piezoelectric Ceramics in the Solid-Solution Series Lead-Titanate/Lead-Zirconate/Lead-Oxide: Tin Oxide and Lead-Titanate/Lead-Hafnate.—B. Jaffe, R. S. Roth & S. Marzullo. (*J. Res. nat. Bur. Stand.*, Nov. 1955, Vol. 55, No. 5, pp. 239–254.) Results of a detailed investigation are reported. Ferroelectric ceramics with composition close to a morphotropic phase boundary exhibit high dielectric constant and relatively good piezoelectric properties. Where polymorphic inversions below the Curie temperature do not occur, these desirable properties are stable over a wide temperature range. The significance of the results for transducer materials is indicated.

537.227 : 546.431.824-31 **1074**

Observations on Etched Crystals of Barium Titanate.—D. S. Campbell. (*Phil. Mag.*, Nov. 1955, Vol. 46, No. 382, pp. 1261–1262.) Observations similar to those reported by Hooton & Merz (2990 of 1955) have been made using hydrochloric acid and phosphoric acid on crystals direct from the melt, with no polarizing treatment. Photomicrographs are reproduced.

537.228.2 : 546.431.824-31 **1075**

Temperature Dependence of Electrostrictive and Elastic Properties of Ceramic Barium Titanate.—N. A. Roi. (*Akust. Zh.*, July–Sept. 1955, Vol. 1, No. 3, pp. 264–271.) An experimental investigation over the temperature range 15°–150°C is reported. Results are presented graphically.

537.311.31 : 537.311.62 **1076**

Surface Resistance and Reactance of Metals at Infrared Frequencies.—J. R. Beattie & G. K. T. Conn. (*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, pp. 78–81.) Properties of Ag, Al, Cu and Ni surfaces have been studied over the frequency range 2·5–15 $\times 10^{13}$ c/s. Experimental results are compared with predictions from theory.

537.311.33 **1077**

Reflectivity of Several Crystals in the Far Infrared Region between 20 and 200 Microns.—H. Yoshinaga. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 753–754.) Results of measurements on Ge, Si, ZnS, InSb, PbS, PbSe and TlCl are shown graphically.

537.311.33 **1078**

Surface Potential and Surface Charge Distribution from Semiconductor Field Effect Measurements.—W. L. Brown. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 590–591.) "It is possible from measurements of the change in conductance of a semiconductor with application of an electric field normal to its surface to determine both the electrostatic potential of the surface and the distribution of charge in surface states. Such determinations depend upon the uniqueness of the minimum in conductance which can be observed in these experiments, and its independence of the surface state charge."

537.311.33 **1079**

Ambipolar Thermodiffusion of Electrons and Holes in Semiconductors.—P. J. Price. (*Phil. Mag.*, Nov. 1955, Vol. 46, No. 382, pp. 1252–1260.) If a temperature gradient is maintained in a semiconductor then, since the equilibrium concentrations of electrons and holes increase with temperature, there will be a concentration gradient of both in the direction of the temperature gradient and a diffusion of both down this gradient ('ambipolar thermodiffusion'). A formula is derived for the magnitude of this effect and is used to determine the electronic contribution to the thermal conductivity and to the Nernst effect.

537.311.33 **1080**

An Easy Derivation of the Hole Lifetime in an *n*-Type Semiconductor with Acceptor Traps.—F. W. G. Rose & D. J. Sandiford. (*Proc. phys. Soc.*, 1st Nov. 1955, Vol. 68, No. 431B, pp. 894–897.) The relations between the constants of recombination, generation and the mass-action laws are described and used for the calculation of hole lifetime in the case of steady-state injection, the concentrations of electrons and holes being assumed sufficiently low for Boltzmann statistics to apply. The result agrees with that given by Shockley & Read (420 of 1953). Lifetime at the commencement of injection is different from the steady-state value.

537.311.33 **1081**

The Use of a Modulated Light Spot in Semiconductor Measurements.—D. G. Avery & J. B. Gunn. (*Proc. phys. Soc.*, 1st Nov. 1955, Vol. 68, No. 431B, pp. 918–921.) Modifications to the theory of the 'travelling-light-spot' method for the measurement of minority-carrier lifetime τ, which become necessary when the light spot is intensity modulated at a frequency $\omega/2\pi$, are discussed and experimentally verified. A necessary condition for accurate determinations is $\omega^2\tau^2 \ll 1$. Measurement of the phase of the collector signal as well as its amplitude enables the diffusion constant, and hence the drift mobility, of minority carriers to be determined directly.

537.311.33 : 535.215 **1082**

Method for determining the Mobility of Minority Current Carriers injected by Light.—S. M. Ryvkin & R. V. Khar'yuzov. (*Zh. tekh. Fiz.*, April 1955, Vol. 25, No. 4, pp. 563–568.) A detailed description of the apparatus is given and results of measurements are presented, including a number of oscillograms. A comparison of the results obtained with those published in the literature indicates that the method can be applied with advantage to investigations of the photoelectric properties of semiconductors.

537.311.33 : 537.32 : 546.873.241 **1083**

Investigation of the Thermoelectric Properties of Bismuth Telluride.—R. M. Vlasova & L. S. Stil'bans. (*Zh. tekh. Fiz.*, April 1955, Vol. 25, No. 4, pp. 569–576.) Report of an extensive investigation of the dependence of the thermo-e.m.f., electrical conductivity, concentration and mobility of current carriers of Bi_2Te_3 on temperature and on departure from stoichiometric composition.

537.311.33 : 537.533 **1084**

The Effect of Field Emission on the Behaviour of Semiconductor Contacts.—R. W. Sillars. (*Proc. phys. Soc.*, 1st Nov. 1955, Vol. 68, No. 431B, pp. 881–893.) Quite a low potential across a semiconductor contact may be sufficient to produce field emission around the area of contact. This condition is investigated theoretically and experimentally for semiconductors with negligible surface barrier; the current due to field emission varies as the square or three-halves power of the applied voltage. Similar effects are found with metals which are not quite in contact.

537.311.33 : 546.28 **1085**
Lifetime of Electrons in *p*-Type Silicon.—G. Bemski. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 523–524.) Measurements of the temperature variation of the lifetime of electrons in the vicinity of a *p-n* junction give results consistent with theory developed by Hall (3453 of 1952) and Shockley & Read (420 of 1953), assuming the level of recombination centres to be 0·2 eV above the valence band.

537.311.33 : 546.28 **1086**
Broadening of Impurity Levels in Silicon.—M. Lax & E. Burstein. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 592–602.) Calculations of infrared absorption based on a hydrogen-type energy-level scheme are compared with experimental results on boron-doped Si. Results confirm the hypothesis that the broadening of absorption lines observed as temperature increases from that of liquid helium to that of liquid nitrogen is due to the interaction of trapped electrons with acoustic lattice vibrations.

537.311.33 : 546.28 **1087**
Trapping of Minority Carriers in Silicon: Part 2—*n*-Type Silicon.—J. R. Haynes & J. A. Hornbeck. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 606–615.) Complementary investigation to that described previously [2013 of 1955 (Hornbeck & Haynes)]. Evidence is obtained of hole trapping at two or more levels; the heights of the trap levels deduced from photoconductivity decay curves for a range of temperatures are consistent with accepted values of the energy gap.

537.311.33 : 546.28 : 535.37 **1088**
Visible Light from a Silicon *p-n* Junction.—R. Newman. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 700–703.) "When low-voltage silicon *p-n* junctions are biased in the reverse direction to breakdown, visible light is emitted from the junction region. The effects of surface treatment on the phenomenon are discussed. Two typical light output *vs* reverse current curves are shown. A typical spectral distribution curve in the 1·8–3·4 eV range is shown. The observations suggest that the light results from a radiative relaxation mechanism involving the high-energy carriers produced in the avalanche breakdown process."

537.311.33 : 546.28 : 621.396.822 **1089**
The Effect of Illumination on the Excess Noise of Silicon Filaments.—C. A. Hogarth & W. Bardsley. (*Proc. phys. Soc.*, 1st Nov. 1955, Vol. 68, No. 431B, pp. 968–970.) Experiments indicate that the excess noise in Si filaments due to trapping effects is reduced by irradiation with white light. This is interpreted as showing that the free minority carriers produced by the absorbed light keep the traps occupied and so reduce their effectiveness.

537.311.33 : 546.289 **1090**
Precipitation of Copper in Germanium.—R. A. Logan. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 615–617.) "The density of dislocations is shown to have a marked effect on the rate of anneal of copper in germanium. At 500°C samples containing high dislocation density ($\sim 10^6$/cm²) anneal in about 1 hour in contrast to material of low dislocation density ($\sim 10^4$/cm²) which requires about 24 hours. When copper-doped germanium is cooled from a high temperature in regions of high dislocation density, significant precipitation occurs in a cooling cycle of only a few seconds. In this case, in order to prevent precipitation the sample must be quenched from the high temperature in a time of the order of 0·1 second."

537.311.33 : 546.289 **1091**
Properties of Germanium Doped with Manganese.—H. H. Woodbury & W. W. Tyler. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 659–662.) "The temperature dependence of the electrical resistivity and Hall coefficient in *p*- and *n*-type manganese-doped germanium crystals indicates that manganese introduces two acceptor levels in germanium at 0·16 ± 0·01 eV from the valence band and 0·37 ± 0·02 eV from the conduction band. The distribution coefficient for manganese in germanium is $(1{\cdot}0 \pm 0{\cdot}2) \times 10^{-6}$. Comparison is made with other fourth-row metals (V, Fe, Co, and Ni) as impurities in germanium."

537.311.33 : 546.289 **1092**
Classical Theory of Cyclotron Resonance for Holes in Ge.—J. M. Luttinger & R. R. Goodman. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 673–674.)

537.311.33 : 546.289 **1093**
Observation of Quantum Effects in Cyclotron Resonance [in Ge].—R. C. Fletcher, W. A. Yager & F. R. Merritt. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 747–748.) Observations of new cyclotron resonance lines at very low temperatures are interpreted as confirming theoretical predictions by Luttinger & Kohn (2316 of 1955).

537.311.33 : 546.289 **1094**
Restoration of Resistivity and Lifetime in Heat Treated Germanium.—R. A. Logan & M. Schwartz. (*J. appl. Phys.*, Nov. 1955, Vol. 26, No. 11, pp. 1287–1289.) "Experiments are described in which minority carrier lifetime and resistivity of germanium are altered by the addition of copper by diffusion, under conditions where extraneous chemical contamination is minimized. This copper is then gettered by heating the samples in contact with liquid lead or gold, and the resistivity and lifetime are substantially restored. The gettering process is interpreted in terms of the low distribution coefficient of copper in the ternary system which copper and germanium form with the getter."

537.311.33 : 546.289 **1095**
The Influence of Pressure on Metal/Germanium Contacts.—J. Lees & S. Walton. (*Proc. phys. Soc.*, 1st Nov. 1955, Vol. 68, No. 431B, pp. 922–928.) Variation of the resistance of Ge with pressure is measured by comparing the areas of these contacts as given by electrical and by optical methods.

537.311.33 : 546.289 **1096**
Some Measurements connected with Carrier Deficit Phenomena in Germanium.—A. Many, D. Gerlich & E. Harnik. (*Proc. phys. Soc.*, 1st Nov. 1955, Vol. 68, No. 431B, pp. 970–972.) Carrier exclusion was studied in *n*-type Ge filaments composed of two sections, one containing a higher proportion of donor impurities than the other, separated by a sharp transition region. Results are in accordance with theory.

537.311.33 : 546.289 : 535.215 **1097**
Temperature Dependence of Photocurrent in Ge *p-n* Junctions.—R. Wiesner & E. Groschwitz. (*Z. angew. Phys.*, Oct. 1955, Vol. 7, No. 10, pp. 496–499.) Measurements were made of the photocurrent at temperatures between 10° and 60°C; results are consistent with the recombination mechanism proposed by Hall (3453

of 1952), the temperature dependence being determined mainly by the carrier diffusion length. The production of carriers by the illumination is nearly independent of temperature. With the specimens investigated, surface recombination contributed little to the temperature variation of the photocurrent, hence the problem could be treated with sufficient accuracy as unidimensional.

537.311.33 : 546.289 : 538.614 **1098**

Faraday Effect in Germanium at Room Temperature.—R. R. Rau & M. E. Caspari. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 632–639.) Room-temperature measurements were made of the Faraday rotation of plane-polarized microwaves traversing 4-mm-thick samples of Ge inserted in a circular waveguide; for a longitudinal magnetic field of 1 400 G the angle of rotation is about 3°. Values of 3 780 and 3 300 cm/s per V/cm were deduced for the electron and hole mobilities in *n*- and *p*-type specimens respectively, based on analysis using the Drude-Zener model. For small losses, the ellipticity of the polarization of the transmitted wave is proportional to the relaxation time.

537.311.33 : 546.289 : 548.0 **1099**

Some Aspects of Slip in Germanium.—R. G. Treuting. (*J. Metals, N.Y.*, Sept. 1955, Vol. 7, No. 9, Section 2, pp. 1027–1031.) Experiments are reported indicating that slip in Ge crystals subjected to tension at 600°C is in the < 100 > direction.

537.311.33 : 546.289 : 621.317.331 **1100**

Surface Preparation and Resistivity Measurements on Semiconductor Crystals.—R. Manfrino. (*Alta Frequenza*, Aug./Oct. 1955, Vol. 24, Nos. 4/5, pp. 390–420.) The main chemical and electrolytic etching processes suitable for Ge single crystals are described; micrographs illustrating results are reproduced. Apparatus for resistivity measurements by the four-probe method [1502 of 1954 (Valdes)] is described. A resistivity map obtained with a crystal surface of area 150 mm² is shown.

537.311.33 : 546.289 : 621.396.822 **1101**

Hall Effect Noise.—J. J. Brophy & N. Rostoker. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 754–756.) Hall-effect measurements were made on Ge 'bridge' specimens exhibiting excess noise with the usual $1/f$ spectrum. A magnetic-field-dependent noise voltage was observed; this is interpreted as indicative of carrier density fluctuations.

537.311.33 : 546.3.171 **1102**

Production of Metal Nitrides by Glow Discharge, and some of their Properties.—W. Janeff. (*Z. Phys.*, 26th Oct. 1955, Vol. 142, No. 5, pp. 619–636.) The preparation of nitrides of Ag, Cu, Zn, Cd, Sb, Bi, Pb, Sn, Fe and Ni and measurements of their resistivities are described in detail. The temperature coefficient of resistance was also determined. No photoelectric effect was observed.

537.311.33 : 546.361.31 **1103**

Conductance Mechanisms and the Thermal Transition in Caesium Chloride.—W. W. Harpur & A. R. Ubbelohde. (*Proc. roy. Soc. A*, 8th Nov. 1955, Vol. 232, No. 1190, pp. 310–319.)

537.311.33 : 546.46.814 **1104**

Electrical and Optical Properties of Intermetallic Compounds: Part 4—Magnesium Stannide.—R. F. Blunt, H. P. R. Frederikse & W. R. Hosler. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 663–666.) Measurements over a temperature range indicate that the value of the energy gap is 0·33 eV at 0°K, and room-temperature mobilities are 320 and 260 cm/s per V/cm for electrons and holes respectively. Appreciable photoconductivity is observed at 85° and 5°K. Part 3: 1405 of 1955 (Blunt et al).

537.311.33 : 546.47-31 **1105**

The Electrical Conductivity at the Surface of Zinc Oxide Crystals.—G. Heiland. (*Z. Phys.*, 15th Oct. 1955, Vol. 142, No. 4, pp. 415–432.) Continuation of experiments reported previously (159 of 1955). The effects of exposure to an oxygen atmosphere and to light and electron irradiation were studied using single-crystal specimens. Both reversible and irreversible variations of conductivity are produced by irradiation at low temperature; this is interpreted as indicating that thin surface layers are effective in absorbing the radiation as well as in the oxygen interaction.

537.311.33 : 546.621.86 **1106**

Effect of Impurities on the Electrical Conduction Mechanism of AlSb.—A. R. Regel' & M. S. Sominski. (*Zh. tekh. Fiz.*, April 1955, Vol. 25, No. 4, pp. 768–770.) Experimental evidence shows that an admixture of Ge does not alter the conduction mechanism of AlSb while admixtures of Sn, Pb, As, Bi, Se and Te transform it into *n*-type. The rectifying properties of AlSb are also greatly affected by impurities. A theoretical interpretation of the results is presented.

537.311.33 : [546.682.86 + 546.681.86 **1107**

Nuclear Magnetic Resonance in Semiconductors: Part 1—Exchange Broadening in InSb and GaSb.—R. G. Shulman, J. M. Mays & D. W. McCall. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 692–699.) Experimental work is reported.

537.311.33 : 546.682.86 **1108**

Plasma Resonance in Crystals: Observations and Theory.—G. Dresselhaus, A. F. Kip & C. Kittel. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 618–625.) Observations are reported of magnetic resonance observed in *n*-type InSb at 9 and 24 kMc/s, at temperatures of 4° and 77°K. Theory presented previously in relation to resonance in Si and Ge (2995 of 1955) is extended to deal with this case. The influence of specimen shape and the possibility of detecting minority-carrier cyclotron resonance in the presence of a majority-carrier plasma are examined and eddy-current effects are discussed.

537.311.33 : 546.817.221 : 621.396.822 **1109**

Measurements of Current Noise in Lead Sulphide at Audio Frequencies.—D. Barber. (*Proc. phys. Soc.*, 1st Nov. 1955, Vol. 68, No. 431B, pp. 898–907.) The mean square noise voltage per unit bandwidth, $\overline{v_n^2}$, was found to vary as the square of the exciting current and to increase with decreasing frequency in a manner which varied from cell to cell, being directly dependent on cell resistance. Temperature and illumination affected $\overline{v_n^2}$ only through their effect on resistance.

537.311.33 : [546.873.231 + 546.873.241 **1110**

Diffusion of Tin and Antimony in the Semiconducting Compounds Bi_2Se_3 and Bi_2Te_3.—B. Boltaks. (*Zh. tekh. Fiz.*, April 1955, Vol. 25, No. 4, pp. 767–768.) Experimental data are presented graphically. The activation energy in the case of Sb is two or three times greater than in the case of Sn.

537.311.33 : 621.396.822 **1111**

Relative Influence of Majority and Minority Carriers on Excess Noise in Semiconductor Filaments.—L. Bess. (*J. appl. Phys.*, Nov. 1955, Vol. 26, No. 11, pp. 1377–1381.) "An experiment has been devised whereby excess noise is measured along various different directions in a germanium filament after the directions of current flow for the majority and minority carriers had been altered by a magnetic field. From this experiment it is possible to determine that whereas shot noise is caused by both majority and minority carrier fluctuation, $1/f$ noise is essentially produced only by majority carrier fluctuation."

537.323 : 546.56-1 **1112**

Effect of Point Imperfections on the Electrical Properties of Copper: Part 2—Thermoelectric Power.—F. J. Blatt. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 666–670.) Calculations are presented of the variation of thermoelectric power due to the presence of interstitials and vacancies and small concentrations of As. Part 1: 790 of March.

538.214 : 546.26-1 **1113**

Theory of the Magnetic Susceptibility of Graphite.—J. E. Hove. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 645–649.)

538.22 **1114**

Neutron Diffraction Study of the Magnetic Properties of the Series of Perovskite-Type Compounds [(1 − x) La, xCa] MnO₃.—E. O. Wollan & W. C. Koehler. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 545–563.) "This series of compounds exhibits ferromagnetic and antiferromagnetic properties which depend upon the relative trivalent and tetravalent manganese ion content. The samples are purely ferromagnetic over a relatively narrow range of composition ($x \sim 0{\cdot}35$) and show simultaneous occurrence of ferromagnetic and antiferromagnetic phases in the ranges ($0 < x < 0{\cdot}25$) and ($0{\cdot}40 < x < 0{\cdot}5$). Several types of antiferromagnetic structures at $x = 0$ and $x > 0{\cdot}5$ have also been determined." The results are in good agreement with Goodenough's predictions (1115 below).

538.22 **1115**

Theory of the Role of Covalence in the Perovskite-Type Manganites [La, M(II)] MnO₃.—J. B. Goodenough. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 564–573.) The structure of these manganites is studied using the concept of semicovalent exchange [3016 of 1955 (Goodenough & Loeb)]. Predicted properties are consistent with experimental results (e.g. 1114 above).

538.22 **1116**

Magnetic Moment Arrangements and Magnetocrystalline Deformations in Antiferromagnetic Compounds.—Yin-Yuan Li. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 627–631.)

538.22 **1117**

Considerations on Double Exchange.—P. W. Anderson & H. Hasegawa. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 675–681.) Calculations are presented of interactions between magnetic ions, of the type discussed by Zener (2441 of 1951).

538.22 : 621.318.134 **1118**

Properties of Lithium Ferrite FeLiO₂.—R. Collongues. (*C. R. Acad. Sci., Paris*, 28th Nov. 1955, Vol. 241, No. 22, pp. 1577–1580.) Structural changes observed at temperatures up to 670°C are described.

538.22 : 621.318.134 **1119**

Micrographic Study of the Cubic→Quadratic Transformation in Copper Ferrite.—I. Behar. (*C. R. Acad. Sci., Paris*, 28th Nov. 1955, Vol. 241, No. 22, pp. 1580–1581.)

538.22 : 621.318.134 **1120**

Influence of Substitutions on Quadratic Deformation in Copper Ferrite and Chromite.—C. Delorme. (*C. R. Acad. Sci., Paris*, 28th Nov. 1955, Vol. 241, No. 22, pp. 1588–1589.)

538.221 **1121**

A Simple Method for rendering Elementary Magnetic Domains Visible by use of Dry Powder.—W. Andrä & E. Schwabe. (*Ann. Phys., Lpz.*, 15th Oct. 1955, Vol. 17, No. 1, pp. 55–56.) The specimen is exposed to smoke from ignited iron pentacarbonyl.

538.221 **1122**

Solutions of the Equations of Ferrimagnetic Resonance.—B. Dreyfus. (*C. R. Acad. Sci., Paris*, 7th Nov. 1955, Vol. 241, No. 19, pp. 1270–1272.) A detailed study is made of the four solutions of the resonance equation derived previously (185 of January) in the light of the fact that only two resonances are observed experimentally and one of these is generally very weak [2335 of 1955 (McGuire)].

538.221 **1123**

Irreversible Magnetic After-Effect and its Influence on the Form of the Magnetization Curve.—O. Yamada. (*Z. Phys.*, 12th Sept. 1955, Vol. 142, No. 2, pp. 225–240.) Results of an experimental investigation on alnico (24% Co; 13·5% Ni; 8·5% Al; 3% Cu; 51% Fe) are reported and discussed.

538.221 **1124**

Results and Problems of the Quantitative Theory of Coercive Force.—R. Brenner. (*Z. angew. Phys.*, Oct. 1955, Vol. 7, No. 10, pp. 499–507.) A critical survey of published theoretical work on the nature of the coercive force in ferromagnetic materials, in relation to physical and practical aspects.

538.221 **1125**

Theory of Remagnetization of Thin Tapes.—H. Ekstein. (*J. appl. Phys.*, Nov. 1955, Vol. 26, No. 11, pp. 1342–1343.) Analysis based on domain-wall displacements is given for the reversal of magnetization of ferromagnetic tapes with rectangular hysteresis loops, subjected to an external field parallel to the surface.

538.221 **1126**

Influence of Pulsed Magnetic Fields on the Reversal of Magnetization in Square-Loop Metallic Tapes.—D. S. Rodbell & C. P. Bean. (*J. appl. Phys.*, Nov. 1955, Vol. 26, No. 11, pp. 1318–1323.) Report of experiments confirming results obtained previously [1706 of 1955 (Bean & Rodbell)] and illuminating the process of domain formation. The 'intrinsic mobility' of domain walls is evaluated; for a particular permalloy specimen the value is about 100 cm/s per oersted.

538.221 **1127**

Electrical Analog of the Eddy-Current-Limited Domain-Boundary Motion in Ferromagnetics.—G. Brouwer. (*J. appl. Phys.*, Nov. 1955, Vol. 26, No. 11, pp. 1297–1301.) When a ferromagnetic material is magnetized by domain-boundary displacement, the dispersion region of the permeability/frequency curve shifts towards lower frequencies with increasing domain size, in consequence of the inhomogeneous distribution of the permeability at the domain boundaries. This corresponds to the eddy-current phenomenon in laminated materials. Analysis based on equivalent networks is presented for some idealized cases. Successive stages in the magnetization process are illustrated by field maps with boundary movements indicated.

538.221 **1128**

Hysteresis Loops of Mixtures of Ferromagnetic Micropowders.—C. P. Bean. (*J. appl. Phys.*, Nov. 1955, Vol. 26, No. 11, pp. 1381–1383.) The mixtures considered are dilute compacts with < 1% of the volume occupied by the ferromagnetic powders, so that the total magnetization can be determined simply from the magnetization of the constituents. The latter comprise particles in three different size ranges, large (multi-domain), medium (single-domain) and small (single-domain 'super-paramagnetic'). Experimental results are discussed in relation to theories regarding the coercive force of such mixtures.

538.221 **1129**

The Remanent Magnetization of Haematite Powders.—E. P. Wohlfarth. (*Phil. Mag.*, Nov. 1955, Vol. 46, No. 382, pp. 1155–1164.) "Calculations are described of the remanent magnetization and the remanence after demagnetization as a function of field strength for powders of haematite, αFe_2O_3. The calculations are based on the assumptions that the powder particles are single domains and that their hysteretic properties are due to magnetocrystalline anisotropy with threefold symmetry in the basal crystal plane. The calculations are intended for eventual comparison with experiments now in progress."

538.221 **1130**

The Thermal Effects associated with the Magnetization of High-Coercivity Materials.—L. F. Bates & A. W. Simpson. (*Proc. phys. Soc.*, 1st Nov. 1955, Vol. 68, No. 431B, pp. 849–858.) Experiments with alloys of the Fe_2NiAl type, over a field range of 4 000 oersteds, give results in accordance with the theory advanced by Stoner & Rhodes (*Phil. Mag.*, May 1949, Vol. 40, No. 304, pp. 481–522).

538.221 + 549.514.51] : 534.22-8 **1131**

Variation of Elastic-Wave Velocity with Frequency in Fused Quartz and Armco Iron.—D. S. Hughes & J. M. Kennel. (*J. appl. Phys.*, Nov. 1955, Vol. 26, No. 11, pp. 1307–1309.) Using small cylindrical samples specially built up to transmit only torsional waves, the propagation velocity was observed to be constant within ± 1 m/s over the frequency range 0·6–3 Mc/s for fused quartz; for armco the velocity was constant within ± 2 m/s from 0·5 to 1·5 Mc/s and decreased by about 10 m/s between 1·5 and 3 Mc/s.

538.221 : 539.234 **1132**

Magnetic Domains in Thin Films of Nickel-Iron.—C. A. Fowler, Jr, & E. M. Fryer. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 746–747.) Photographs of the domain pattern on both faces of an evaporated film are reproduced.

538.221 : 621.318.132 **1133**

Relation of Magnetic Characteristics of Magnetically Soft Alloys to Thickness of Lamination.—E. I. Gurvich & E. Kondorski. (*C. R. Acad. Sci. U.R.S.S.*, 1st Oct. 1955, Vol. 104, No. 4, pp. 530–532. In Russian.) Experimental results indicate that the change in the frequency characteristics of the permeability and loss angle with change in the thickness of the specimen is probably due to a greater penetration, during the heat treatment, of impurities from the surface into the material in thinner specimens. Molybdenum permalloy and 50% permalloy were investigated.

538.221 : 621.318.134 **1134**

Ferrimagnetic Resonance in Single Crystals of Manganese Ferrite.—J. F. Dillon, Jr, S. Geschwind & V. Jaccarino. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 750–752.) Report of experiments at frequencies of 5·9, 9·3 and 24 kMc/s, and at temperatures up to 300°K. Resonance curves and line-width/temperature characteristics are shown.

538.221 : 621.318.134 **1135**

Transient Phenomena in a Ferrite.—S. Matz. (*Rev. gén. Élect.*, Oct. 1955, Vol. 64, No. 10, pp. 491–509.) Detailed account of an investigation, previously described briefly (3605 of 1954), of the complex-permeability variations of ferrites in a h.f. magnetic field. Results are discussed in relation to theory and measurements of magnetic after-effect.

538.221 : 621.318.134 **1136**

Low-Magnetic-Saturation Ferrites for Microwave Applications.—L. G. Van Uitert. (*J. appl. Phys.*, Nov. 1955, Vol. 26, No. 11, pp. 1289–1290.) Values of Curie temperature, coercive force, magnetic saturations at 25°C, and initial permeability at 0·1 Mc/s are given for six MgAlMn ferrites.

538.652 **1137**

Calculation of Magnetostriction Constants for Nickel.—G. C. Fletcher. (*Proc. phys. Soc.*, 1st Nov. 1955, Vol. 68, No. 431A, pp. 1066–1071.) Calculations for Ni at 0°K, based on the collective electron model, give values of the magnetic constants rather larger than the latest figures obtained experimentally at 120°K.

548.0 : 537.311.33 **1138**

Symmetry Properties of the Energy Bands of the Zinc Blende Structure.—R. H. Parmenter. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 573–579.) Analysis is based on group theory, first excluding and then including spin-orbit coupling. Character tables and compatibility tables are presented. The degeneracies and gradients of the various possible energy bands are studied at lines and points of symmetry in the Brillouin zone. The results are compared with those for the equivalent energy bands in crystals of diamond structure.

548.0 : 537.311.33 **1139**

Spin-Orbit Coupling Effects in Zinc Blende Structures.—G. Dresselhaus. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 580–586.) "Character tables for the 'group of the wave vector' at certain points of symmetry in the Brillouin zone are given. The additional degeneracies due to time reversal symmetry are indicated. The form of energy *vs* wave vector at these points of symmetry is derived."

548.5 **1140**

Single Crystals with Pure Surfaces.—E. Menzel, W. Stössel & M. Otter. (*Z. Phys.*, 1st Oct. 1955, Vol. 142, No. 3, pp. 241–244.) A method is outlined for producing spherical single crystals of metals by congelation from the molten state on to a plane base material in a vacuum.

548.5 : 669.046.54/.55 **1141**

Contribution to Mathematics of Zone Melting.—L. Burris, Jr, C. H. Stockman & I. G. Dillon. (*J. Metals, N.Y.*, Sept. 1955, Vol. 7, No. 9, Section 2, pp. 1017–1023.)

548.5 : 669.046.54/.55 : 537.311.33 **1142**

Temperature Gradient Zone Melting.—W. G. Pfann. (*J. Metals, N.Y.*, Sept. 1955, Vol. 7, No. 9, pp. 961–964.) "Under certain conditions, a molten zone can be made to move through a solid by impressing a stationary temperature gradient across the solid. This phenomenon can be utilized in fabricating semiconductive devices, growing single crystals, joining, boring fine holes in solids, measuring diffusivities in liquids, small scale alloying, and purification. Fundamentals and exemplary applications are outlined."

621.315.56 : 621.315.616 **1143**

Rubber as an Electrical Conductor in Engineering: Properties, General and Special Applications.—S. de Meij. (*Bull. schweiz. elektrotech. Ver.*, 29th Oct. 1955, Vol. 46, No. 22, pp. 1067–1069.) The electrical and mechanical properties of rubber containing furnace black are briefly reviewed. Applications mentioned include use in antistatic devices, cable sheaths, panel heating, noise-free cables and electromechanical transducers. Use of conducting rubber for semiconductor diodes and triodes is suggested, but no details are given.

621.315.614.64 **1144**

Acetylized Paper as an Insulating Material in Electrical Engineering.—W. Dieterle. (*Bull. schweiz. elektrotech. Ver.*, 29th Oct. 1955, Vol. 46, No. 22, pp. 1045–1065.) A comprehensive survey is presented of the electrical properties of paper treated with acetic acid. The treatment reduces the hygroscopicity. Impregnation with chlorinated oils of high dielectric constant further improves the insulating qualities. 26 graphs, 3 tables of properties of papers, fibres and oils, and 43 references are given.

621.315.616 **1145**

The Mechanical Engineering of Dielectrics.—B. Maxwell. (*Elect. Engng, N.Y.*, Oct. 1955, Vol. 74, No. 10, pp. 870–873.) The mechanical properties of polymeric plastics are discussed; the importance of their variation with time and with temperature is illustrated by results of measurements.

621.315.616 **1146**

Electric Strength of Irradiated Polythene.—K. H. Stark & C. G. Garton. (*Nature, Lond.*, 24th Dec. 1955, Vol. 176, No. 4495, pp. 1225–1226.) Measurements were made at temperatures up to 220°C on moulded disks recessed to a thickness of about 50 μ, irradiated with 4-MeV electrons. Results are compared with those obtained by Bird & Pelzer for unirradiated material (1696 of 1949).

621.318.2 : 621.385.029.6 **1147**

Permanent Magnets for Microwave Electron Valves.—de Bennetot. (See 1254.)

621.357 **1148**

Micromachining with Virtual Electrodes.—A. Uhlir, Jr. (*Rev. sci. Instrum.*, Oct. 1955, Vol. 26, No. 10, pp. 965–968.) An electrolytic etching method is described which has been used to drill very small holes in Ge, Mo, Fe, etc., and to plate small areas. The technique involves use of a glass tube drawn to a fine tip to localize the current.

537.311.33 **1149**

Poluprovodniki v Sovremennoi Fizike (Semiconductors in Contemporary Physics). [Book Review]—A. F. Ioffe. Publishers: Izd. A.N. S.S.S.R. (Acad. Sci. U.S.S.R.), Moscow-Leningrad, 1954, 355 pp., 17r.60k. (*Uspekhi fiz. Nauk*, Sept. 1955, Vol. 57, No. 1, pp. 165–169.) A monograph written for physicists, chemists and engineers. Chapters are included on (*a*) solid electrolytes, (*b*) metals, (*c*) electronic semiconductors, (*d*) quantum theory of semiconductors, (*e*) physical phenomena in semiconductors, and (*f*) experimental results.

MATHEMATICS

512 **1150**

Vectors, Matrices and Determinants, Tensors.—R. Fortet. (*Bull. Soc. franç. Élect.*, Oct. 1955, Vol. 5, No. 58, pp. 710–726.) An introductory article for electrical engineers.

517.522.2 **1151**

A Simple Method for the Numerical Calculation of Time Functions for Given Broken Rational Transform Functions.—R. Hofmann & W. Walcher. (*Arch. elekt. Übertragung*, Oct. 1955, Vol. 9, No. 10, pp. 475–478.) The method described is useful in analysis involving Laplace transforms. The required time function is presented in the form of a rapidly converging power series.

517.9 : 537.21 **1152**

Poisson's Partial Difference Equation.—Davies. (See 1019.)

MEASUREMENTS AND TEST GEAR

531.761 + 529.7 **1153**

Definition of the Second of Time.—G. M. Clemence. (*Nature, Lond.*, 24th Dec. 1955, Vol. 176, No. 4495, p. 1230.) Attention is directed to certain undesirable consequences involved in the adoption of a 'physical second' in place of the astronomical second [3686 of 1955 (Bullard)]. It is suggested that if a new physical unit is adopted it should be given a name other than 'second', to avoid confusion. See also 513 of February (Jones).

621.317.331 : 537.311.33 : 546.289 **1154**

Surface Preparation and Resistivity Measurements on Semiconductor Crystals.—Manfrino. (See 1100.)

621.317.335 : 621.315.61 **1155**

Measurement of ϵ' and tan δ of a Solid Dielectric at Centimetre Wavelengths in the Temperature Range from − 100° to + 100°C.—P. F. Veselovski. (*Zh. tekh. Fiz.*, April 1955, Vol. 25, No. 4, pp. 601–609.) A new method is described in which use is made of a rectangular resonator designed for oscillations of the H_{10} type. The resonator consists of two waveguide sections separated by a sheet of mica and closed by movable pistons at the ends. Results of measurements on polythene and other insulating materials are given in tables and curves.

621.317.335 : 621.317.755 **1156**

A Precision Resonance Method for measuring Dielectric Properties of Low-Loss Solid Materials in the Microwave Region.—S. Saito & K. Kurokawa. (*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, pp. 35–42.) A disk sample is inserted into a cylindrical cavity resonator; the dielectric constant is found from the difference between the lengths for resonance with and without the sample, and the loss factor is found from the difference between the Q factors. Good precision is obtained by using a differentiated resonance curve as a frequency marker on the oscilloscope.

621.317.335.3 **1157**

The Measurement of Dielectric Constants of Liquids at Decimetre Wavelengths.—T. Jäkel. (*Ann. Phys., Lpz.*, 15th Oct. 1955, Vol. 17, No. 1, pp. 42–54.) Detailed description of a coaxial-line method in which the evaluation of the results does not involve transcendental functions.

621.317.336 **1158**

Reflection and Impedance Measurements by means of a Long Transmission Line.—J. C. van den Hoogenband & J. Stolk. (*Philips tech. Rev.*, May 1955, Vol. 16, No. 11, pp. 309–320.) Description of a variable-frequency stationary-detector method using a line about 60 m long. Direct visual indication of matching is given by means of an oscilloscope. For quick measurements a wobbulator is used, for more accurate measurements a variable-frequency oscillator is preferred, in conjunction with a d.c. microammeter.

621.317.34 : 621.373.42.025.3 **1159**

Wide-Band Three-Phase *RC*-Generators for Complex Measurements of Two-Poles and Four-Poles.—G. Thirup. (*J. Brit. Instn Radio Engrs*, Dec. 1955, Vol. 15, No. 12, pp. 597–605.) "From a 3-phase *RC* oscillator two voltages are derived, the complex ratio of which can be varied. The complex ratio is independent of the frequency. The two voltages are used in a compensation circuit for measuring the parameters of two-poles and four-poles. Two equipments are described covering the frequency ranges 20 c/s–22 kc/s and 22 kc/s–10 Mc/s. The possible error is ± 0·5 dB and ± 2° in the frequency range 100 c/s–3 Mc/s. Outside this range the phase error may increase about 3 times while the amplitude error remains nearly the same."

621.317.715 **1160**

Very-High-Sensitivity Screened Galvanometer.—W. Meissner & R. Doll. (*Z. angew. Phys.*, Oct. 1955, Vol. 7, No. 10, pp. 461–468.)

621.317.722 **1161**

Design for a Sensitive Self-Recording Gold-Leaf Electroscope.—P. Goodman. (*J. sci. Instrum.*, Nov. 1955, Vol. 32, No. 11, pp. 439–440.) An electrode system is used such that the leaf deflection passes through 90° and the leaf then discharges; the output is in the form of voltage pulses whose frequency is proportional to current for low values and logarithmically related for higher values. Steady electron currents of the order of 10^{-14} A can be measured.

621.317.733 : 621.314.7 **1162**

A D.C. and A.C. Balance Detector with Automatic Protection from Overload.—C. Morton. (*J. sci. Instrum.*, Nov. 1955, Vol. 32, No. 11, pp. 437–439.) Two junction transistors in a balanced bridge circuit give a current gain of 50. The input resistance approaches infinity for input voltages > 150 mV, thus providing automatic protection. For a.c. working the application of an auxiliary voltage of about 4 V, of the same frequency as the bridge voltage, gives a sensitivity of 15 mA/V (r.m.s.).

621.317.75.029.3 **1163**

Audio Frequency Spectrum Analyzer.—E. F. Feldman. (*Tele-Tech & Electronic Ind.*, Oct. 1955, Vol. 14, No. 10, pp. 78–81 . . 135.) The heterodyne-type equipment described displays the Fourier analysis of complex waveforms, providing either logarithmic or linear frequency and amplitude indications. Points of the design are discussed, particularly the i.f. stages, the bandwidth of which is varied in synchronism with the log-frequency scan to maintain optimum resolution.

621.317.755 **1164**

Voltage Coincidence Oscilloscope.—R. J. D. Reeves. (*Wireless World*, Feb. 1956, Vol. 62, No. 2, pp. 85–87.) A method of presenting multichannel displays on a single-beam c.r.o. uses a voltage-coincidence circuit in place of a Y amplifier, with a normal timebase voltage on the X plates. A number of co-phased outputs are taken from an independent a.f. oscillator and one of them is applied to the Y plates to give a full-scale deflection; the others are used as reference signals in the voltage-coincidence circuit to explore the waveforms under examination, a bright spot appearing on the screen whenever the two signals are equal. The method is not useful for very fast displays, and care is necessary to avoid erroneous interpretation of the display.

621.317.755 : 621.3.015.3 **1165**

New Circuits for Recurrent Surge Oscillography.—D. D. Davis. (*Elect. Engng, N.Y.*, Oct. 1955, Vol. 74, No. 10, pp. 919–923.)

621.317.772 **1166**

Measuring Phase at R.F. and Video Frequencies.—Y. P. Yu. (*Electronics*, Jan. 1956, Vol. 29, No. 1, pp. 138–140.) A continuously variable delay line is used in conjunction with a balanced phase detector to indicate the phase difference between two input signals. The frequency range is 10 kc/s–20 Mc/s; a delay of 5×10^{-10} sec can be measured. Accuracy is within ± 0·1° of phase angle, or ± 1% of indicated time delay.

621.317.772.029.3 : 621.314.7 **1167**

Direct-Indicating Audio-Frequency Phase-Angle Meters using Transistors.—K. Homilius. (*Arch. tech. Messen*, Oct. 1955, No. 237, pp. 221–224.) Circuits are described based on the technique of converting the sinusoidal signals whose phase difference is unknown into constant-amplitude pulses. Use of both point-contact and junction transistors is indicated.

621.317.784 **1168**

Wattmeters.—H. D. Hawkes & A. H. M. Arnold. (*J. Instn elect. Engrs*, Nov. 1955, Vol. 1, No. 11, pp. 676–683.) A review of commercial instruments, including one torsion-head microwave type.

621.317 + 621.372].029.63/.64 **1169**

Schaltungstheorie und Messtechnik des Dezimeter- und Zentimeterwellengebietes. [Book Review]—Weissfloch. (See 1015.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.1-8 : 621.9 **1170**

Ultrasonic Machining of Brittle Materials.—M. S. Hartley. (*Electronics*, Jan. 1956, Vol. 29, No. 1, pp. 132–135.) High-speed cutting is achieved by means of a correctly shaped tool vibrating above the specimen at a frequency of 25 kc/s. The tool face and the specimen do not come into contact; an abrasive powder in a liquid suspension is fed between the two and cutting is done by impact of the powder on the specimen surface.

550.837 **1171**

Geophysical Prospection of Underground Water in the Desert by means of Electromagnetic Interference Fringes.—M. A. H. El-Said. (*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, pp. 24–30.) A variable-frequency and a variable-distance method are described. Interference patterns obtained at two places in the Egyptian deserts are shown; results are in good agreement with information regarding the water table obtained from boring.

621-52 : 621.9 **1172**

An Electronically Controlled Machine Tool.—(*Electronic Engng*, Jan. 1956, Vol. 28, No. 335, p. 37.) A fully automatic copy milling machine for the production of cams is described. Coded design information is fed on plastic tape to a control unit and information store which also contains provision for enabling smooth profile curves to be deduced from a relatively small number of stored marker points.

621-526(083.7) **1173**

I.R.E. Standards on Terminology for Feedback Control Systems, 1955.—(*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, pp. 107–109.) Standard 55 I.R.E. 26. S2.

621.365.5 : 621.385 **1174**

Brown Boveri Transmitting and Rectifier Tubes in Industrial Generators for R.F. Heating.—R. Hübner. (*Brown Boveri Rev.*, Sept. 1955, Vol. 42, No. 9, pp. 370–387.) Brief illustrated descriptions are given of induction and dielectric heating apparatus for various purposes.

621.374.3 : 621.316.86 : 77 **1175**

Precision Photographic Timer.—J. G. Thomason. (*Wireless World*, Feb. 1956, Vol. 62, No. 2, pp. 71–74.) An electronic timing circuit based on the Miller integrator is described. Compensation for variations in mains voltage is obtained by use of a feed resistor in which the current passed varies approximately as the fifth power of the applied voltage.

621.384.613 **1176**

Energy Stability of the 22-MeV Betatron at the University of Illinois.—B. M. Spicer & A. S. Penfold. (*Rev. sci. Instrum.*, Oct. 1955, Vol. 26, No. 10, pp. 952–953.) The greatest energy fluctuation observed over a period of eighteen months was 40 keV at 17 MeV.

621.385.83.032.2 **1177**

Some Anomalous Emission Effects in Low-Energy Electron Beams.—W. W. H. Clarke & L. Jacob. (*Proc. phys. Soc.*, 1st Nov. 1955, Vol. 68, No. 431B, pp. 805–816.) Cathode emission in a specially made electron gun, using accelerating voltages in the range 18–30 V, was found to vary in a periodic manner over the cathode surface, variations being greatest in the central region where the emission pattern was critically dependent on accelerating voltage and on cathode temperature. Variation of total emission was also periodic. The wavelength of the standing-wave pattern across the emitter was 0·01 mm; the electrons are assumed to oscillate in the field resulting from the effects of the initial velocities and the space charge.

621.385.833 **1178**

An Investigation of the Electron-Optical Properties of Straight Magnetic Gaps.—S. N. Baranovski, D. L. Kaminski & V. M. Kel'man. (*Zh. tekh. Fiz.*, April 1955, Vol. 25, No. 4, pp. 610–624.) The system considered consists of two magnetic pole pieces greatly extended in one direction; electrons are subjected to the action not of the field in the gap but of the 'dissipation field' in front of it. A theoretical as well as experimental investigation into the operation of the system is presented.

621.385.833 **1179**

The Scanning Electron Microscope and its Fields of Application.—K. C. A. Smith & C. W. Oatley. (*Brit. J. appl. Phys.*, Nov. 1955, Vol. 6, No. 11, pp. 391–399.) The advantages of the scanning electron microscope, due especially to the large angles of incidence and reflection which can be used, are discussed and examples are given of applications for which the conventional electron microscopes are unsuitable. Bombardment of the specimen is much less severe than with conventional types. A resolution of 200 Å has been attained and a value of < 100 Å may be attainable.

621.385.833 **1180**

A Possible Chromatic Correction System for Electron Lenses.—G. D. Archard. (*Proc. phys. Soc.*, 1st Nov. 1955, Vol. 68, No. 431B, pp. 817–829.) It is shown that a combination of two two-pole groups of electrodes can provide axial chromatic correction in one plane. Applications to practical correction systems are discussed and curves are given for the chromatic and focal properties of the corrective electrode group considered.

621.385.833 **1181**

Chromatic Aberration and Aperture Error of Cylindrical Electron-Optical Lenses.—M. Rheinfurth. (*Optik, Stuttgart*, 1955, Vol. 12, No. 9, pp. 411–416.) Expressions are derived for the two types of error, and it is shown that they cannot both be eliminated.

621.385.833 **1182**

Use of a Cylindrical Lens for reducing Image Distortion in Reflection Electron Microscopy.—C. Fert & B. Marty. (*C. R. Acad. Sci., Paris*, 21st Nov. 1955, Vol. 241, No. 21, pp. 1454–1456.)

621.385.833 : 535.767 **1183**

Theory and Technique of Electron-Microscope Stereograms.—(*Optik, Stuttgart*, 1954, Vol. 11, No. 12, pp. 562–577 & 1955, Vol. 12, No. 6, pp. 253–272.)

Part 1: 3657 of 1954.

Part 2: Errors in the Determination of Depth of Electron-Microscope Stereograms.—J. G. Helmcke & H. J. Orthmann.

Part 3: Methods for determining the Depth of Objects and Requirements for a Photogrammetric Apparatus.—J. G. Helmcke.

621.385.833 : 537.525 **1184**

Electric 'Microdischarges' in a Dynamic Vacuum.—Arnal. (See 1025.)

621.396.934 : 621.396.96 **1185**

Radar-Guided Missiles.—(*Wireless World*, Feb. 1956, Vol. 62, No. 2, pp. 67–70.) A brief survey of the principles of command-guidance, homing and beam-riding control systems used with ground-to-air missiles.

621.396.934 : 621.398 **1186**

Fuel Cut-Off Control for Guided Missiles.—G. L. Zomber & D. Macmillan. (*Electronics*, Jan. 1956, Vol. 29, No. 1, pp. 126–127.) The input circuit of the control relay comprises a parallel-T network tunable over the range 88–154 c/s, and is designed for use with a radar beacon system. The operating potential of a neon-tube voltage regulator in the switching circuit is stabilized by radiation from a radium pellet.

621.398 **1187**

The New Centralized Telecontrol System on 175 c/s adopted by Électricité de France.—F. Cahen & H. Prigent. (*Rev. gén. Élect.*, Oct. 1955, Vol. 64, No. 10, pp. 475–484.)

621.384.6 **1188**

High-Energy Accelerators. [Book Review]—M. S. Livingston. Publishers: Interscience, New York and London, 1954, 157 pp., 26s. (*Phil. Mag.*, Nov. 1955, Vol. 46, No. 382, p. 1264.) Recommended as a useful introduction to and survey of particle accelerators.

PROPAGATION OF WAVES

538.566.2 : 551.510.535 **1189**

Reflection of a Plane Electromagnetic Wave from a Stratified Ionized Gas.—P. Poincelot. (*C. R. Acad. Sci., Paris*, 7th Nov. 1955, Vol. 241, No. 19, pp.

1272–1275.) Continuation of analysis presented previously (3725 of 1955). The conditions for total reflection are indicated and expressions are derived for the phase shift of the reflected wave and the group propagation time in terms involving the apparent height of reflection. Results are discussed briefly in relation to ionosphere soundings.

621.396.11 **1190**

Measurements of the Propagation Time of Standard Time Signals.—M. Boella & C. Egidi. (*Alta Frequenza*, Aug./Oct. 1955, Vol. 24, Nos. 4/5, pp. 309–338.) Report of experiments made in May 1951, using transmissions in both directions between Turin, Italy, and Beltsville, U.S.A. Because the signals become distorted during transmission, an arbitrary criterion must be adopted for the moment at which the signal is considered to be received. Various criteria are compared on the basis of a statistical analysis of results. With a series of about 60 observations in 1 min, the accuracy of the determination is within about 0·1 ms, using photographic technique. On comparing the results with data obtained by ionospheric soundings at points near the propagation path, the most probable number of hops is deduced to be 4 or 5, the corresponding single-hop distance being 1 670 or 1 340 km respectively. Discrepancies between this result and the values usually assumed are discussed.

621.396.11 : 537.56 **1191**

Theory of Radio Reflections from Electron-Ion Clouds.—V. R. Eshleman. (*Trans. Inst. Radio Engrs*, Jan. 1955, Vol. AP-3, No. 1, pp. 32–39.) First-order approximations to the effects of size, shape and density on the reflecting properties of electron-ion clouds are obtained. For low-density clouds the reflection coefficients are determined by summation of the wavelets from individual ions; high-density clouds are treated as total reflectors at the critical-density radius.

621.396.11 : 551.510.535 **1192**

Wave Solutions for Critical and Near-Critical Coupling Conditions in the Ionosphere.—N. Davids & R. W. Parkinson. (*J. atmos. terr. Phys.*, Oct. 1955, Vol. 7, Nos. 4/5, pp. 173–202.) "Principal dielectric axis co-ordinates are used to put the coupled wave equations for ionospheric propagation into a form which is suitable for an analysis of critical coupling. This approach removes the singularity formerly associated with critical coupling. The analysis is carried out through the second order and compared with some recent data on polarization. The results compare favourably for several ionospheric models considered. A satisfactory analysis of split echoes at 150 kc/s is developed."

621.396.11 : 551.510.535 **1193**

Observations of the Effects of Ionospheric Storms over a North Atlantic Circuit.—T. W. Bennington. (*J. atmos. terr. Phys.*, Oct. 1955, Vol. 7, Nos. 4/5, pp. 235–243.) The median m.u.f. (i.e. the frequency up to which signals are propagated for 50% of a given period) for normal and disturbed conditions has been deduced from regular observations of signals from WWV, Washington, received at Tatsfield, England, over the period July–December 1953. The results show that disturbance of communications is most severe during the period 0700 to 1100 of any disturbed day. This agrees with the findings of Appleton & Piggott (2802 of 1952) from vertical-incidence data. The effect is also dependent on the normal value of m.u.f. for the time of day, being greater at night, particularly during winter when the normal m.u.f. is low.

621.396.11 : 551.510.535 **1194**

On the Variation of Ionospheric Absorption at Different Stations.—Piggott. (See 1051.)

621.396.11 : 551.510.535 **1195**

Selective Annotated Bibliography on Ionospheric Propagation.—M. Rigby & M. L. Rice. (*Met. Abstracts Bibliography*, April 1955, Vol. 6, No. 4, pp. 489–547.) Includes some 300 references covering the period 1902–1955.

621.396.11 : 551.594.6 **1196**

An Atmospherics Analyzer.—P. F. Smith. (*Trans. Inst. Radio Engrs*, Jan. 1955, Vol. AP-3, No. 1, pp. 9–12.) A system is described which records the received atmospheric waveform and simultaneously derives a plot of the time intervals between peaks of the waveform/time curve, facilitating reduction of data relating to range and ionospheric layer height.

621.396.11 : 621.396.6 **1197**

Tests on Long-Range Decimetre-Wavelength Multiplex Links.—R. Goublin. (*Bull. Soc. franç. Élect.*, Oct. 1955, Vol. 5, No. 58, pp. 700–709.) Transmissions on frequencies in the band 450–470 Mc/s over distances from 50 to 100 km show ground losses varying from 6 to 36 dB for angles of diffraction from 0 to 25 × 10^{-3} rad. Over paths of 215 and 250 km, one at grazing incidence to a forest terrain and the other over line of sight, fading is found to be worse in summer than in winter, especially over the forest, but communication is possible for practically all the time with a 20-W transmitter and parabolic aerials. F.m. terminal equipment for permanent communication links is described.

621.396.11 : 621.396.67.012.12 **1198**

The Dependence of Meteoric Forward-Scattering on Antenna Patterns and Orientations.—C. O. Hines, P. A. Forsyth, E. L. Vogan & R. E. Pugh. (*Canad. J. Phys.*, Oct. 1955, Vol. 33, No. 10, pp. 609–610.) Radio waves of frequencies above accepted m.u.f. may be propagated by scattering either from ionospheric irregularities [2581 of 1952 (Bailey et al.)] or from meteor trails [e.g. 2117 of 1953 (Villard et al.)]. In any given set of observations the predominant mechanism depends on the aerial radiation pattern; directing the radiation off the transmitter-receiver axis favours the meteor-trail-scattering mechanism.

621.396.11.029.53/.55 **1199**

A Theoretical Model for High-Frequency Back-scatter from the Sea Surface via the Ionosphere.—W. C. Hoffman. (*J. atmos. terr. Phys.*, Oct. 1955, Vol. 7, Nos. 4/5, pp. 278–284.) The average far-zone back-scattered power is computed for perfectly conducting doubly-trochoidal and doubly-sinusoidal surfaces (see also *Trans. Inst. Radio Engrs*, July 1955, Vol. AP-3, No. 3, pp. 96–100). Either assumption gives results of the correct order of magnitude; from the nature of the problem, this is the best that can be expected. The doubly-sinusoidal surface leads to much more convenient computations.

621.396.11.029.55 **1200**

Arrival Angle of H.F. Waves.—A. F. Wilkins & C. M. Minnis. (*Wireless Engr*, Feb. 1956, Vol. 33, No. 2, pp. 47–53.) "Apparatus is described for measuring the angle of elevation of h.f. waves arriving at the ground after transmission by way of the ionosphere. The system involves the comparison of the amplitudes of the signals received by two vertically-spaced horizontal loop aerials at heights above the ground of about 33 metres and 12·5 metres. The comparison is effected by amplifying the signal e.m.fs in a twin-channel receiver and applying the i.f. voltages to the deflecting plates of a cathode-ray oscilloscope. It is shown that, if a single ray is incident on the system, the e.m.fs produced in the two aerials are in phase irrespective of their height above ground, so that the corresponding oscilloscope pattern is a straight line, the slope of which is a simple function of the angle of elevation of the ray. Calibrations of the system over the band 10–20 Mc/s have shown that its behaviour is very close to that expected from the simple theory and that there is no noteworthy change in the calibration with azimuth. Methods of observation for the general case of the arriving signal consisting of more than one ray are considered."

621.396.11.029.55 : 551.510.535 **1201**
Azimuthal Fluctuations of the Direction of Arrival of Short Radio Waves.—W. Kronjäger & K. Vogt. (*Nachrichtentech. Z.*, Oct. 1955, Vol. 8, No. 10, pp. 537–540.) The directions of arrival of signals from Tangiers, Rio de Janiero, Buenos Aires, Osaka and Tokio, were measured at Eschborn. The frequencies were between 13 and 21 Mc/s and were in all cases below the m.u.f. The distribution of deviations was Gaussian; maximum fluctuations were ± 3° for 80% of the observations and ± 5° for 98%. Median directions of arrival for the first three stations coincided with the great-circle directions, but those for Osaka and Tokio deviated by 2° and 3° respectively, to the south. The aerial system comprised two rhombic aerials set at an angle of 9° to each other. Results are tabulated and presented graphically.

RECEPTION

621.396.82 : 551.594.6 **1202**
Atmospheric Noise Interference to Broadcasting in the 3-Mc/s Band at Poona.—S. V. C. Aiya & K. R. Phadke. (*J. atmos. terr. Phys.*, Oct. 1955, Vol. 7, Nos. 4/5, pp. 254–277.) Using an objective method previously described [257 of 1955 (Aiya)] systematic measurements were made daily from 1800 to 2300 I.S.T. throughout 1953; noise data required for the design of broadcast services are deduced. The results are compared with estimates of noise taken from *Nat. Bur. Stand. Circulars*, No. 462 and with experimental results given in *Radio Research Board Special Report* No. 26 [842 of 1954 (Horner)]; both of these tend to be low. An attempt is made to estimate noise levels from known thunderstorm data; reasonable agreement with measured values is obtained.

STATIONS AND COMMUNICATION SYSTEMS

621.376.5 **1203**
Calculation of the Spectra of Modulated Pulse Trains.—M. Sánchez & F. Popert. (*Arch. elekt. Übertragung*, Oct. 1955, Vol. 9, No. 10, pp. 441–452.) The theory and method of representation developed by Bennett (see *Modulation Theory*, Black, 1953, p. 266) are applied to determine the spectra of pulse trains subjected to various types of modulation of practical importance.

621.39 : 621.376.5 **1204**
A Multichannel Pulse Communication System with Automatically Variable Number of Pulses.—R. Filipowsky. (*Öst. Z. Telegr. Teleph. Funk Fernsehtech.*, Sept./Oct. & Nov./Dec. 1955, Vol. 9, Nos. 9/10 & 11/12, pp. 113–129 & 147–158.) The operation of a p.c.m. system with asynchronous sampling is discussed in detail. In this system sampling is performed on the statistical basis of the amount of information in the different channels; this leads to improved use of overall channel capacity. Circuits for selecting the information-carrying channels are described. See also 562 of February.

621.39 : 621.376.5 : 621.396.822 **1205**
Noise Characteristics of Pulse-Slope-Modulation.—J. Das. (*Electronic Engng*, Jan. 1956, Vol. 28, No. 335, pp. 16–21.) The calculated signal/noise ratio at the output of a p.s.m. audio amplifier is 27·4 dB for an input peak ratio of 6 dB with random noise only. This is confirmed experimentally. The ratio improves with increased bandwidth and for a signal containing impulse noise.

621.39.018.756 **1206**
A Type of Signal Practically Limited in Time and in Frequency.—J. A. Ville & J. Bouzitat. (*Câbles & Transm.*, Oct. 1955, Vol. 9, No. 4, pp. 293–303.) The incompatibility of perfectly limited spectrum and finite duration in pulses is demonstrated; an expression is given for the spectrum of a finite-duration signal which can be considered limited for practical purposes. If as an approximation the spectrum of a nonlimited signal is represented by Shannon's series, the error introduced can be considered as a modulation; any signal can thus be represented as a sequence of modulated signals. This method is applied to various signal types, particularly the raised-cosine signal; the limited-spectrum signal is represented by a succession of raised-cosine signals.

621.396 : 621.376.5 **1207**
Pulse Multiplex Radiotelephone System in Greece.—J. J. Muller. (*Onde élect.*, Aug./Sept. 1955, Vol. 35, Nos. 341/342, pp. 711–713.) An account of the network covering nearly 2 000 km, which was put into service in 1954. There are 34 stations, with separations ranging from 14 to 120 km, at altitudes ranging from 20 to 1 800 m. Some of the links handle telegraphy and radio-program transmission as well as telephony. Fading effects are reduced by using diversity reception. Carrier frequencies are between 1·7 and 2·3 kMc/s. Particular aspects of the system are dealt with in the following separate papers:

High-Quality Radio Links over Sea Paths in Greece.—R. Cabessa (pp. 714–727).

Synchronization of the Greek Radio-Link System.—G. X. Potier (pp. 728–732).

Operational Reliability of the Greek Radio-Link System.—R. Basard (pp. 733–738).

621.396.41 : 621.396.65 **1208**
A 4-Channel Carrier Telephone System for the Argentine V.H.F. Radio Links.—Y. Boers. (*Commun. News*, Oct. 1955, Vol. 16, No. 1, pp. 30–34.) Description of the Type-STR 128 system. Each channel has a width of 6 kc/s; the a.f. band used is 300 c/s–3·4 kc/s, while signalling is effected on 4·3 kc/s. The four bands are transmitted in the inverted position, so that the total modulation frequency range is transformed from 300 c/s –22·3 kc/s to 1·7–23·7 kc/s; this reduces the modulation frequency spectrum by 2½ octaves, considerably reducing the risk of intermodulation distortion.

621.396.41 : 621.396.65 : 621.376.4 **1209**
V.H.F. Phase-Modulated Transmitter-Receiver for 4-Channel Telephony.—J. M. M. Veldstra. (*Commun. News*, Oct. 1955, Vol. 16, No. 1, pp. 22–29.) Account of Type-SFR 329 radio equipment for use in conjunction with the channelling equipment described by Boers (1208 above).

621.396.44 + 621.397.242 **1210**
Distribution Systems.—J. Kason. (*Wireless World*, Feb. 1956, Vol. 62, No. 2, pp. 88–90.) A system for relaying sound and television broadcasts, picked up on a central receiving aerial, to small blocks of flats etc., is

described. Signals are distributed at transmitted frequencies through channel amplifiers; suitable filter networks incorporated in mixing and splitter units enable subscribers to receive any programme at will.

621.396.6 : 621.396.11 **1211**

Tests on Long-Range Decimetre-Wavelength Multiplex Links.—Goublin. (See 1197.)

621.396.65.029.64 : 621.397.26 **1212**

New Microwave Repeater System using Traveling-Wave Tubes.—N. Sawazaki & T. Honma. (*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, pp. 19–24.) A system is described using travelling-wave valves as wide-band amplifiers in a reflex arrangement permitting the same valve to serve at r.f. and i.f. See also 2783 of 1954 (Nomura et al.).

621.396.712 : 621.396.65 **1213**

F.M. for B.F.N.—J. D. Parker. (*Wireless World*, Feb. 1956, Vol. 62, No. 2, pp. 81–84.) The British Forces broadcasting network in Germany brought into service in 1956 constitutes a radio relay system in which the intermediate stations function as broadcasting stations, re-radiating the received program on different frequencies in the range 87·6–99·3 Mc/s. F.m. is used and the stations are automatically operated.

621.396.721 : 621.396.6 **1214**

Two-Metre Transmitter-Converter.—G. P. Anderson. (*Wireless World*, Feb. 1956, Vol. 62, No. 2, pp. 60–66.) A portable 6-W crystal-controlled transmitter working in the band 144–146 Mc/s and a receiver-converter are described for use with the equipment for 3·75 and 7·5 Mc/s of which details were given earlier (530 of 1954).

SUBSIDIARY APPARATUS

621-526 : 621.3.016.35 **1215**

Stability Criteria for an Electrical or Mechanical System with Distributed Parameters.—Gladwin. (See 992.)

621.311.62 : 621.314.5 **1216**

Advances in Vibratory-Power-Supply Techniques.—L. W. D. Sharp. (*Brit. Commun. Electronics*, Oct. 1955, Vol. 2, No. 10, pp. 54–58.) A brief review of British commercial vibrators. A table of the comparative performance of three typical vibrator units is given.

621.314.63 : 546.289 **1217**

The EW54 Germanium Junction Rectifier.—R. D. Knott & J. I. Missen. (*G.E.C. J.*, Oct. 1955, Vol. 22, No. 4, pp. 197–208.) The junction is produced by alloying a controlled quantity of In with a Ge crystal about 6 mm square and 0·4 mm thick with resistivity about 10 Ω. cm. The manufacture, electrical characteristics, and methods of assessing the performance of the rectifier are described. Applications are grouped into (*a*) use with resistive load, (*b*) use with battery load, and (*c*) pulse and miscellaneous. By virtue of its high efficiency this rectifier can be used for high powers with comparatively simple cooling systems.

621.314.63 : 546.824-3 **1218**

High-Temperature Area-Type Titanium-Dioxide Rectifiers.—H. C. Gorton, T. S. Shilliday & F. K. Eggleston. (*Elect. Engng, N.Y.*, Oct. 1955, Vol. 74, No. 10, pp. 904–907.) Modifications in the methods of production, including the deliberate introduction of impurities, have led to improvements over the performance reported previously [2755 of 1955 (Shilliday & Peet)] in respect of maximum operating temperature, reverse/forward resistance ratio, and reproducibility of characteristics. Satisfactory operation at temperatures above 250°C for hundreds of hours is reported.

621.352.32 **1219**

Behaviour of MnO_2 as Depolarizer [in Leclanché-type cells].—J. Brenet, A. Grund & A. M. Moussard. (*Rev gén. Élect.*, Oct. 1955, Vol. 64, No. 10, pp. 513–516.)

621.355 : 621.316.722.1 **1220**

Low-Voltage Stabilization.—(*Wireless World*, Feb. 1956, Vol. 62, No. 2, p. 70.) A form of Ni-Cd secondary cell is described having applications to the cathode biasing of valves. It provides a stable voltage of about 1·5 V which is practically independent of the current passed. The impedance is $< 1\ \Omega$ and is independent of frequency.

TELEVISION AND PHOTOTELEGRAPHY

621.397.26 : 621.396.65.029.64 **1221**

New Microwave Repeater System using Traveling-Wave Tubes.—Sawazaki & Honma. (See 1212.)

621.397.335 **1222**

Television Transmission without Synchronization Levels.—H. J. Griese. (*Nachrichtentech. Z.*, Oct. 1955, Vol. 8, No. 10, pp. 552–555.) A system is described in which the picture brightness levels occupy the whole modulation range and synchronization depends on coincidence between a local pulse and a stepped pulse transmitted during the horizontal blanking interval. The principal advantages are the improved signal/noise ratio and the removal of difficulties connected with holding the various levels constant.

621.397.5 **1223**

Phonevision—an Effective Method for Subscription Television.—A. L. C. Webb & A. Ellett. (*Proc. Instn Radio Engrs, Aust.*, Oct. 1955, Vol. 16, No. 10, pp. 341–353.)

621.397.5 : 535.623 **1224**

Compatible Colour Television.—J. Haantjes & K. Teer. (*Wireless Engr*, Jan. & Feb. 1956, Vol. 33, Nos. 1 & 2, pp. 3–9 & 39–46.) A system is described using two distinct subcarriers for the red and blue signals; the bandwidth for the red signal is 2 Mc/s and that for the blue is 1 Mc/s, and both lie within the normal-width luminance band. At the receiver the colour signals are fed to separate demodulators which include appropriate band-pass filters, and an equivalent green signal is formed locally. Receivers for this system, in contrast to those for the N.T.S.C. system, do not require subcarrier generators or synchronization of the colour-signal detectors. Important aspects of the two systems are tabulated for comparison.

621.397.61 **1225**

A 16-mm Projector for Operation with a Television Film Chain on a Partial-Storage Basis.—E. C. Fritts. (*J. Soc. Mot. Pict. Telev. Engrs*, Oct. 1955, Vol. 64, No. 10, pp. 576–577.) Apparatus described previously (268 of 1954) was modified for use with vidicon tubes. A 2–3 system separates the pull-down by the required integral number of television fields while maintaining the usual $2\frac{1}{2} : 1$ ratio between television fields and motion-picture frames.

621.397.61 **1226**

The Development and Design of an Underwater Television Camera.—D. R. Coleman, D. Allanson & B. A. Horlock. (*J. Brit. Instn Radio Engrs*, Dec. 1955, Vol. 15, No. 12, pp. 625–636. Discussion, pp. 636–639.) The camera described uses an image orthicon operating at 625 lines 50 frames or 525 lines 60 frames. The mechanical design of the casing is discussed and the handling arrangements for both diver-controlled and deep-sea cameras are described. Problems of underwater illumination are examined and a new type of remotely focused lamp is mentioned.

621.397.611 : 778.5 **1227**

Critical Considerations on Optical Compensation by means of Rotating Prisms in Television Film Scanners.—H. Grabke. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1955, Vol. 7, Nos. 9/10, pp. 166–170.) A prism with at least 24 sides and a built-in glass corrector is necessary to obtain satisfactory pictures. The coupling between film and prism is discussed.

621.397.611.2 **1228**

The 'Scenioscope', a New Television Camera Tube.—P. Schagen. (*Tijdschr. ned. Radiogenoot.*, Sept. 1955, Vol. 20, No. 5, pp. 291–305. In English.) Description of a tube in which the target is of slightly conducting glass, permitting charge to pass from the signal plate to the scanned surface. Very good picture quality can be achieved with an illumination of a few hundred lux; recognizable though 'noisy' pictures have been obtained with illuminations as low as 15 lux. See also 3762 of 1955.

621.397. 621.2 : 621.385.832: 535.65 **1229**

Two Photoelectric Colorimeters for Television Picture Tubes.—R. S. Hunter. (*J. electrochem. Soc.*, Sept. 1955, Vol. 102, No. 9, pp. 512–517.) The instruments described are intended primarily for measurements on monochrome picture tubes. One uses a single photocell and three filters on a disk, the other uses three photocells each with appropriate filter.

621.397.7 **1230**

A New N.W.D.R. Television [outside-broadcast] **Transmission Van.**—H. Krause & H. Käding. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1955, Vol. 7, Nos. 9/10, pp. 174–183.) The van is about 11 m long and weighs about 14 tons and carries complete equipment for three image-orthicon channels and 12 microphone channels. A detailed illustrated description is given.

621.397.7 : 535.623 **1231**

Integration of Color Equipment in a Television Station.—P. B. Laeser. (*J. Soc. Mot. Pict. Telev. Engrs*, Oct. 1955, Vol. 64, No. 10, pp. 537–541.) Problems involved in the installation at WTMJ-TV, Milwaukee, are discussed.

621.397.7 : 535.623 **1232**

CBS Television Color Studio 72.—R. B. Monroe. (*J. Soc. Mot. Pict. Telev. Engrs*, Oct. 1955, Vol. 64, No. 10, pp. 542–549.) Illustrated description of an installation in a converted theatre in New York.

621.397.8 : 535.8 **1233**

Influence of the Optical System on the Television Picture.—F. Below & H. Grabke. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1955, Vol. 7, Nos. 9/10, pp. 171–173.) Results of preliminary measurements indicate that local variations of picture quality can be assessed with sufficient accuracy from observations of the reproduction of black-to-white steps.

621.397.8 : 621.396.822 **1234**

The Visibility of Noise in Television.—(*B.B.C. Engng Div. Monographs*, Oct. 1955, No. 3, pp. 1–22.) Comprises the following parts:—

Part 1—The Visibility of Noise over the Grey Scale.—R. D. A. Maurice (pp. 5–12).

Part 2—The Visibility of Noise as a Function of Frequency.—M. Gilbert (pp. 13–16).

Part 3—Photographic Records of the Effect of Random Noise on a Television Picture.—G. F. Newell & J. G. Spencer (pp. 17–22).

TRANSMISSION

621.396.61 : 621.385 : 621.316.9 **1235**

Gas Tubes protect High-Power Transmitters.—W. N. Parker & M. V. Hoover. (*Electronics*, Jan. 1956, Vol. 29, No. 1, pp. 144–147.) Fault-detection and protective devices for transmitter valves are described.

VALVES AND THERMIONICS

621.314.63 **1236**

On the Transient Behaviour of Semiconductor Rectifiers.—B. R. Gossick. (*J. appl. Phys.*, Nov. 1955, Vol. 26, No. 11, pp. 1356–1365.) Theory and practical details of a pulse method of determining the characteristics of a *p-n* junction rectifier are discussed. See also 887 of 1954.

621.314.63.002.2 : 546.289 **1237**

Manufacture of Germanium Power Diodes.—(*Elect. Commun.*, Sept. 1955, Vol. 32, No. 3, pp. 146–164.) Various stages in the production of diffused-junction diodes are illustrated by photographs.

621.314.632 : 546.289] + 621.316.825 **1238**

The Turnover Phenomenon in Thermistors and in Point-Contact Germanium Rectifiers.—Burgess. (See 994.)

621.314.7 **1239**

Transistor Characteristics for Circuit Designers.—S. Schwartz. (*Electronics*, Jan. 1956, Vol. 29, No. 1, pp. 161–174.) Comprehensive tables giving the characteristics and operating parameters of 218 types of transistor made in the United States.

621.314.7 **1240**

The Field Effect Transistor.—G. C. Dacey & I. M. Ross. (*Bell Syst. tech. J.*, Nov. 1955, Vol. 34, No. 6, pp. 1149–1189.) Theory presented previously (3445 of 1953) is recapitulated and extended. Considerations of the field dependence of cut-off frequency f and transconductance on the one hand, and power dissipation on the other, lead to the conclusion that the best operating point is the critical field E_c above which carrier mobility is proportional to $E^{-0\cdot5}$. The performance for this case is discussed in detail and the results are summarized in design nomograms. f is found to be inversely proportional to the 'pinch-off' voltage, but the latter cannot reasonably be reduced below 0·5 V, corresponding to a maximum value of 1 kMc/s for f. The manufacture and performance of various types, operating with field values in some cases below E_c and in others above, are described. Transistors can be designed for high-power operation at low frequencies and can be used to replace pentodes.

621.314.7 **1241**

Gain of Amplifying Transistors.—R. Monelli. (*Alta Frequenza*, Aug./Oct. 1955, Vol. 24, Nos. 4/5, pp. 356–374.) Equivalent circuits are presented for point-contact and junction transistors. Expressions are derived for the gain, considering the transistor as an active quadripole. Values of gain and input resistance are tabulated for various commercially available transistors with various connections.

621.314.7 **1242**

Accelerated Power Aging with Lithium-Doped Point-Contact Transistors.—L. E. Miller & J. H. Forster. (*Trans. Inst. Radio Engrs*, July 1955, Vol. ED-2, No. 3, pp. 4–6.) Use of lithium-doped collector points facilitates observation of changes in transistor parameters during artificial aging tests; changes in the effective donor concentration in the formed region are responsible for the enhanced effects. The results indicate that changes associated with aging are not entirely due to changes in surface conduction.

621.314.7 **1243**

Negative-Resistance Regions in the Collector Characteristics of the Point-Contact Transistor.—L. E. Miller. (*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, pp. 65–72.) Anomalies of three different types have been observed in the *V*/*I* characteristics of various transistors. Typical curves are shown and the mechanisms responsible for the anomalies are discussed; the observed negative resistance is a property of the transistor itself and is attributable to a variation in the collection efficiency of the reverse-biased collector contact.

621.314.7 **1244**

The Dependence of Transistor Parameters on the Distribution of Base Layer Resistivity.—J. L. Moll & I. M. Ross. (*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, pp. 72–78.) Expressions are derived for the emitter efficiency, the transverse sheet resistance, the transit time and the cut-off frequency for uniform, linear, exponential and complementary-error-function distributions of impurities in the base of a junction transistor. For equivalent parameters the nonuniform distributions permit the use of wider base layers, but require greater maximum impurity concentrations and higher current densities.

621.314.7 : 537.531.9 **1245**

Dependence of Barrier-Layer Photoeffect on X Radiation.—J. Braunbeck & J. Zakovsky. (*Naturwissenschaften*, Nov. 1955, Vol. 42, No. 22, pp. 602–603.) Experiments on commercial junction transistors exposed to X rays are briefly reported; results are given for the Type-OC70 as typical. Curves show the variation of emitter and collector current (*a*) with wavelength and angle of irradiation and (*b*) with temperature.

621.314.7 : 621.396.822 **1246**

Shot Noise in *p-n-p* Transistors.—G. H. Hanson. (*J. appl. Phys.*, Nov. 1955, Vol. 26, No. 11, pp. 1388–1389.) Measurements on several types of transistor at frequencies up to 800 kc/s are reported. The frequency dependence of noise at high frequencies is in accordance with a formula derived by van der Ziel (600 of February).

621.38 : 001.4 **1247**

Russian Vacuum-Tube Terminology.—G. F. Schultz. (*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, p. 112.) A short list of basic terms is given together with the English translations.

621.383.4 : 621.396.822 **1248**

Reduction of Noise in Photoconductive Cells.—R. E. Burgess. (*Brit. J. appl. Phys.*, Nov. 1955, Vol. 6, No. 11, pp. 385–387.) "The factors determining the signal/noise ratio of photoconductive cells are briefly discussed and attention is drawn to the importance of the noise generated at the electrode contacts when current is flowing through the cell. Means are described for reducing or eliminating this component of noise by means of potential probes near the electrodes combined with suitable external circuits."

621.385 + 621.314.7 + 621.318.57 **1249**

Design of Electronic Devices for Production.—J. Thomson. (*Nature, Lond.*, 3rd Dec. 1955, Vol. 176, No. 4492, pp. 1059–1060.) Report of a summer school held at the Services Electronics Research Laboratories in September 1955. Various aspects of valve design were discussed, with emphasis on u.h.f. types; transistors and magnetic switching devices were also discussed.

621.385.029.6 **1250**

Thermal Motion of Electrons in a Two-Beam Amplifier.—M. I. Rodak. (*Zh. tekh. Fiz.*, April 1955, Vol. 25, No. 4, pp. 644–648.) The theory of a two-beam electron-wave valve is further developed by taking into account the effect of the thermal motion of electrons on amplification. The electron velocity distribution function is assumed to be Maxwellian. It is shown that this velocity distribution decreases the frequency range and the amplification. A sufficiently large velocity range makes amplification impossible.

621.385.029.6 **1251**

A Developmental Wide-Band, 100-Watt, 20-dB, S-Band Traveling-Wave Amplifier utilizing Periodic Permanent Magnets.—W. W. Siekanowicz & F. Sterzer. (*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, pp. 55–61.) Details are given of design, construction and performance. A permanent-magnet focusing system weighing only 2·8 lb has been developed.

621.385.029.6 **1252**

Transverse-Field Traveling-Wave Tubes with Periodic Electrostatic Focusing.—R. Adler, O. M. Kromhout & P. A. Clavier. (*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, pp. 82–89.) Valves are described in which the wave-propagating structure constitutes a smooth balanced transmission line for r.f. signals and serves at the same time as a space-periodic focusing field. Experimental structures for the 500–900-Mc/s band give wave-propagation velocities 1%–2% that of light, with characteristic impedances of about 500 Ω. Analysis based on transverse electron waves indicates that in order to obtain gain the electron stream must travel faster than the circuit wave by a substantial margin; gain is proportional to the square root of beam current. Experimental results are in good agreement with prediction from theory; noise figures as low as 6 dB have been obtained.

621.385.029.6 **1253**

Coupled Helices for Use in Traveling-Wave Tubes.—G. Wade & N. Rynn. (*Trans. Inst. Radio Engrs*, July 1955, Vol. ED-2, No. 3, pp. 15–24.) Theory is presented for propagation along coaxial helices in the presence of an axial electron beam. Curves show the propagation constants as functions of coupling coefficient and beam velocity, the helix voltages as functions of the physical dimensions, and a parameter related to the frequency response for different input and output couplers. The optimum length for coupling between the helices is greater in the presence of the beam, and the frequency for maximum transmission is lower. Preliminary work on a coupled-helix attenuator is mentioned.

621.385.029.6 : 621.318.2 **1254**

Permanent Magnets for Microwave Electron Valves.—M. de Bennetot. (*Onde élect.*, Aug./Sept. 1955, Vol. 35, Nos. 341/342, pp. 747–763.) Magnets are considered both for simple focusing, as in travelling-wave valves and multicavity klystrons, and for crossed-field systems, as in magnetrons. 'Tubular' constructions are discussed for which the main flux is from regions other than the pole pieces; uniformity of magnet composition is particularly important in such cases. Weights of focusing magnets range from 1·5 to 4·5 kg. See also 2991 of 1954.

621.385.029.6 : 621.374.4 **1255**

Multi-beam Velocity-Type Frequency Multiplier.—Y. Matsuo. (*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, pp. 101–106.) Experiments on a klystron with two nearly parallel beams are described; owing to the reduction of mutual-electron-repulsion effects, much higher output is obtainable than with a single-beam klystron. The device is expected to be useful as a mm-λ generator.

621.385.029.6 : 621.396.822 **1256**

Spurious Modulation of Electron Beams.—C. C. Cutler. (*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, pp. 61–64.) Noise at frequencies up to about 5 Mc/s in valves with high-intensity beams is found to be due to positive ions and secondary electrons. Oscillograms of various types of resulting modulation are shown. The effects are eliminated by improving the outgassing and evacuating and if necessary by deflecting the secondary electrons out of the beam.

621.385.032.2 **1257**

A Design of Triode Electron Gun.—K. C. Ho. (*Trans. Inst. Radio Engrs*, July 1955, Vol. ED-2, No. 3, pp. 10–14.) Analysis is given for a gun with two successive apertured-disk anodes, for producing a pencil beam. Focusing details for a 50-μA and 250-μA beam are treated as examples.

621.385.032.216 **1258**

New Type of Diffusion Cathode.—A. H. Beck, A. D. Brisbane, A. B. Cutting & G. King. (*Elect. Commun.*, Sept. 1955, Vol. 32, No. 3, pp. 172–178.) Reprinted from *Le Vide*, Nov. 1954, Vol. 9, No. 54, pp. 302–309. See also 1841 of 1955.

621.385.032.216 **1259**

Theoretical Basis for measuring the Saturation Emission of Highly Emitting Cathodes under Space-Charge-Limited Conditions.—C. R. Crowell. (*J. appl. Phys.*, Nov. 1955, Vol. 26, No. 11, pp. 1353–1356.) Analysis is presented confirming that the current at the inflection point of the I/V characteristic of a diode gives a sensitive indication of the saturation emission of an oxide cathode. See also 2787 of 1955 (Hopkins & Shrivastava).

621.385.032.216 **1260**

Measurements of Retarding Potential on the Hollow Cathode.—A. W. Bright & J. S. Thorp. (*Nature, Lond.*, 3rd Dec. 1955, Vol. 176, No. 4492, pp. 1079–1080.) I/V characteristics of diodes with two different types of hollow cathode are compared with characteristics of planar diodes. The results appear to indicate that the electron temperature for the hollow cathode does not differ greatly from that for a normal oxide cathode at the same cathode temperature, thus supporting the hypothesis that the greater part of the emission comes from a region near the edge of the hole.

621.385.032.216 : 537.583 **1261**

Schottky Effect for SrO Films on Molybdenum.—G. A. Haas & E. A. Coomes. (*Phys. Rev.*, 15th Oct. 1955, Vol. 100, No. 2, pp. 640–641.) Thermionic-emission measurements on Mo filaments with very thin films of SrO are discussed briefly in relation to the theory of the surface potential barrier.

621.385.832 : 681.142 **1262**

Digital Memory in Barrier-Grid Storage Tubes.—M. E. Hines, M. Chruney & J. A. McCarthy. (*Bell Syst. tech. J.*, Nov. 1955, Vol. 34, No. 6, pp. 1241–1264.) A description is given of the operation of a c.r. tube in which the same beam is used for writing and reading signals in the form of charges on a dielectric plate held between the barrier grid and a backplate; storage capacity and probability of error due to amplifier noise are particularly discussed. Experimental tubes have been produced with a capacity of 16 000 information bits, with reading and writing times of about 1 μs.

621.387 **1263**

Glow-Discharge Tubes.—F. A. Benson & L. J. Bental. (*Wireless Engr*, Feb. 1956, Vol. 33, No. 2, pp. 33–38.) Continuing the investigation reported previously [3422 of 1954 (Benson & Mayo)], an examination has been made of the influence of cathode material, gas filling, gas pressure and geometrical dimensions on the magnitude and duration of the initial drift and on the running-voltage/temperature characteristics; specially made tubes were used. Results are shown graphically; they indicate that gas pressure and tube dimensions have little influence, while the cathode material and the nature of the gas filling do affect the characteristic.

621.387 **1264**

Improving Gas-Diode Voltage Characteristics.—(*Tele-Tech & Electronic Ind.*, Oct. 1955, Vol. 14, No. 10, pp. 67, 123.) Note on a method developed at the National Bureau of Standards. By applying a pulsed voltage through a common resistor to a large number of cold-cathode gas diodes connected in parallel the individual operating characteristics are rapidly equalized and stabilized.

621.387 : 621.318.57 **1265**

Traveling-Glow-Discharge Counter Tube.—E. W. Cowan. (*Rev. sci. Instrum.*, Oct. 1955, Vol. 26, No. 10, pp. 988–989.) Brief description of a tube using wire electrodes of special curved form.

MISCELLANEOUS

016.3 : 621.3 **1266**

Index to Convention Record of the I.R.E., Vol. 3, 1955.—(*Proc. Inst. Radio Engrs*, Jan. 1956, Vol. 44, No. 1, pp. 149 et seq.)

621.37/.39(083.74) **1267**

Standard Functional Divisions for Electronic Equipment.—R. S. Shultz & W. R. Waltz. (*Elect. Engng, N.Y.*, Oct. 1955, Vol. 74, No. 10, pp. 896–900.) Preliminary report of a survey of existing U.S. Navy electronic equipment undertaken with a view to further standardization. Edge-punched cards used in the analysis are shown.

621.3 **1268**

Radio and Television Engineers' Reference Book. [Book Review]—E. Molloy & W. E. Pannett. Publishers: Newnes, 1600 pp., 70s. (*J. Instn elect. Engrs*, Nov. 1955, Vol. 1, No. 11, pp. 717–718.) ". . . well worthy of a place in the library of every radio engineer."

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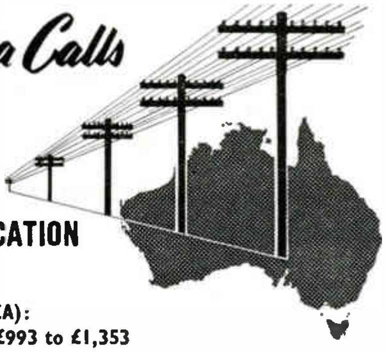

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