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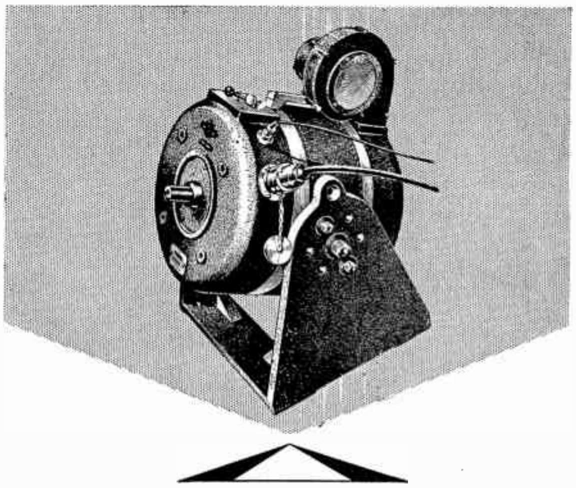

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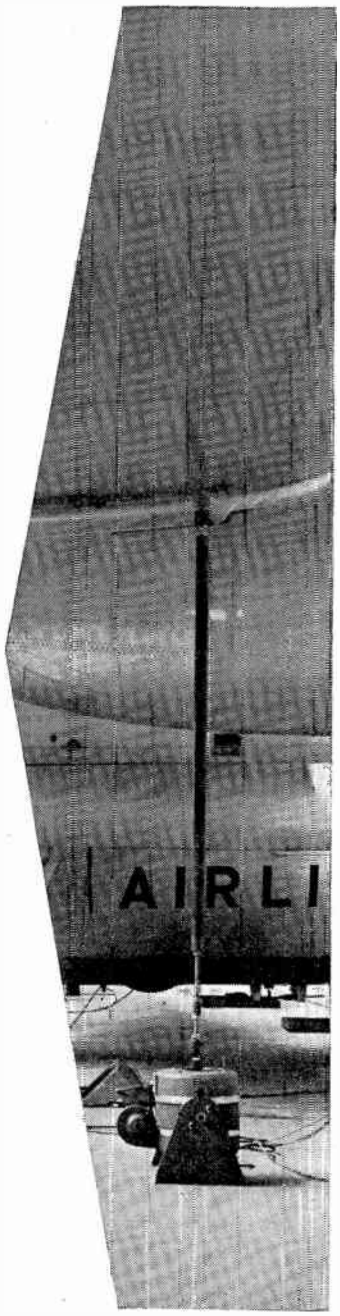

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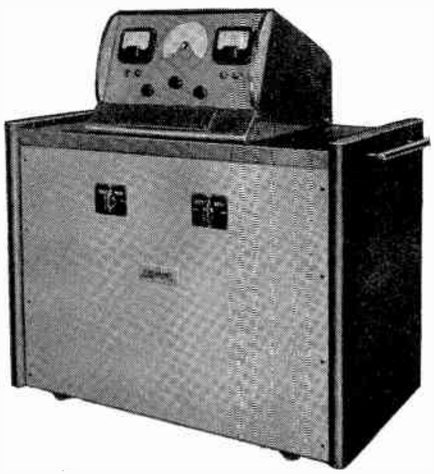

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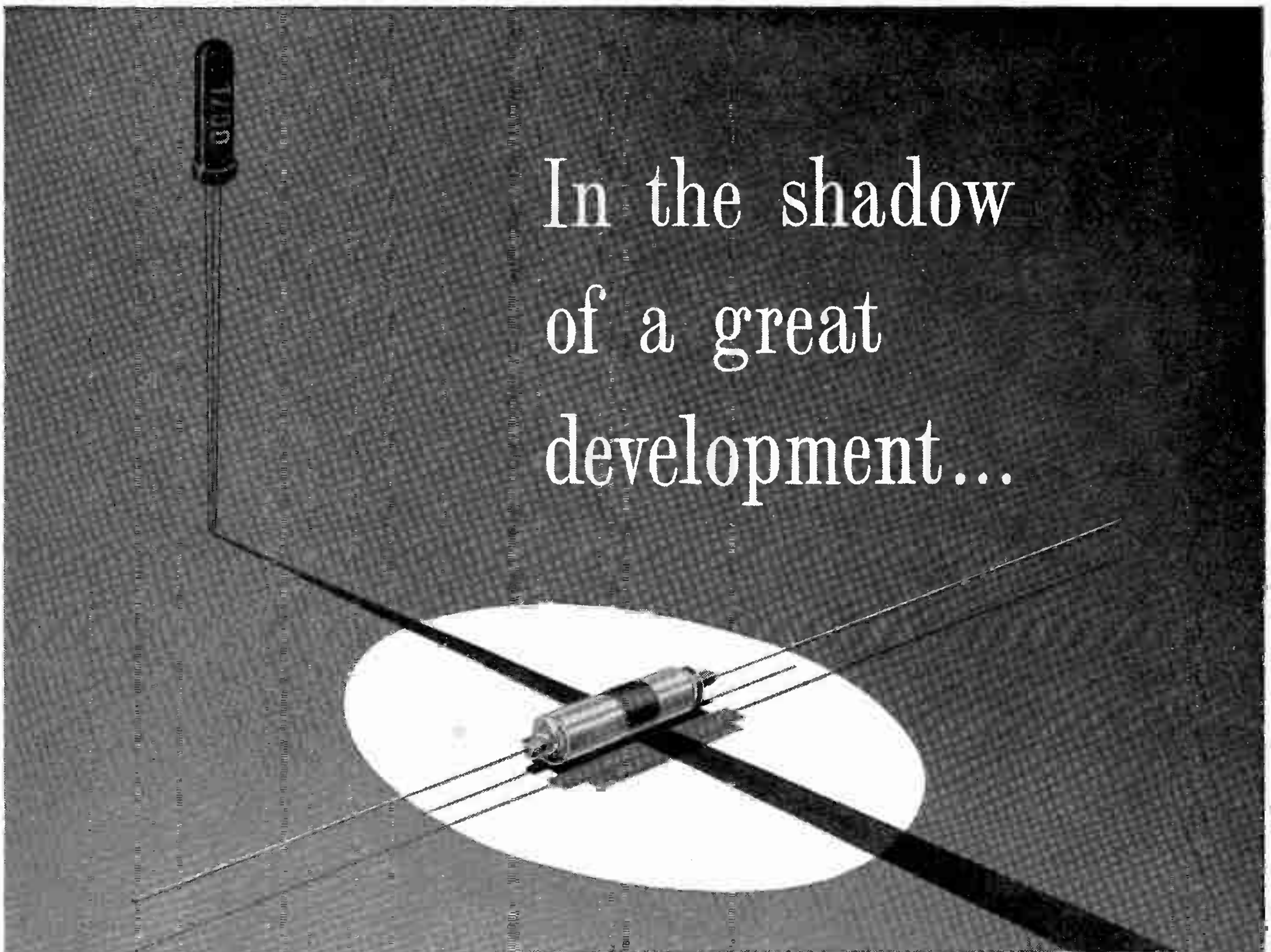

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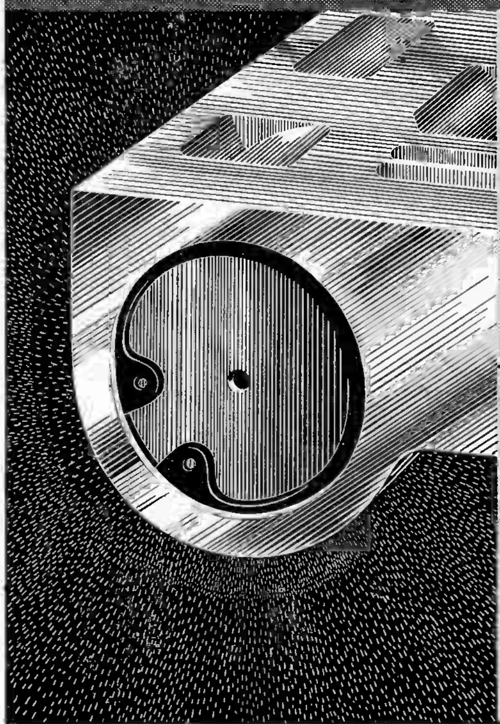

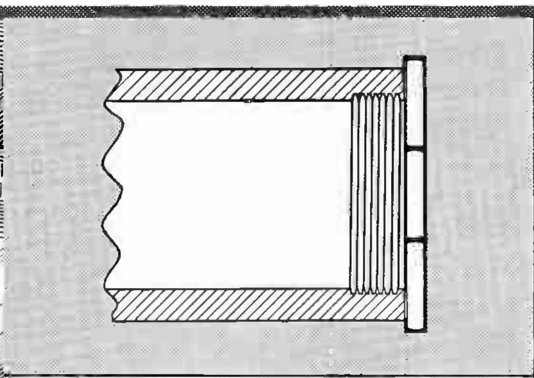

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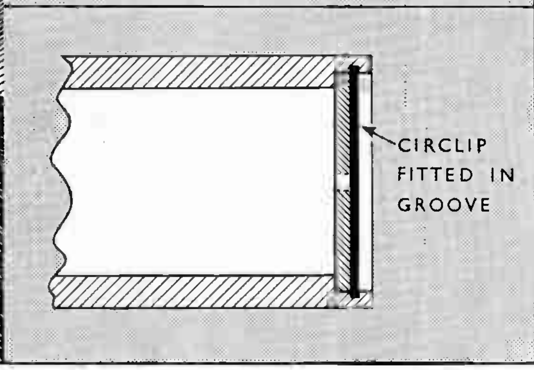

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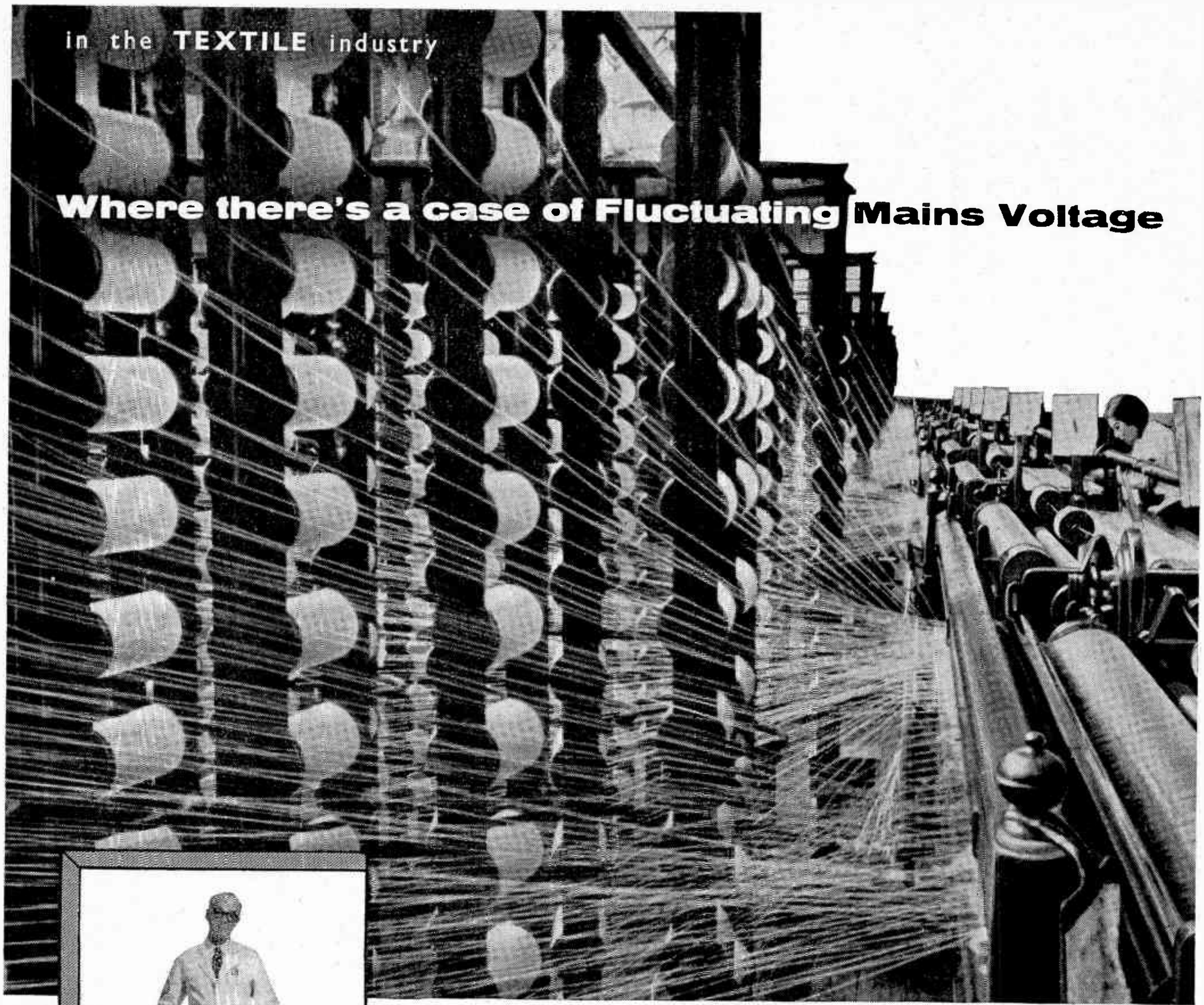

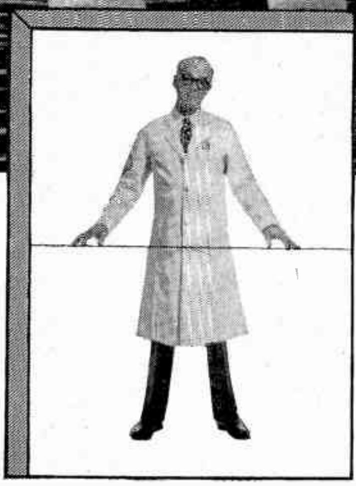

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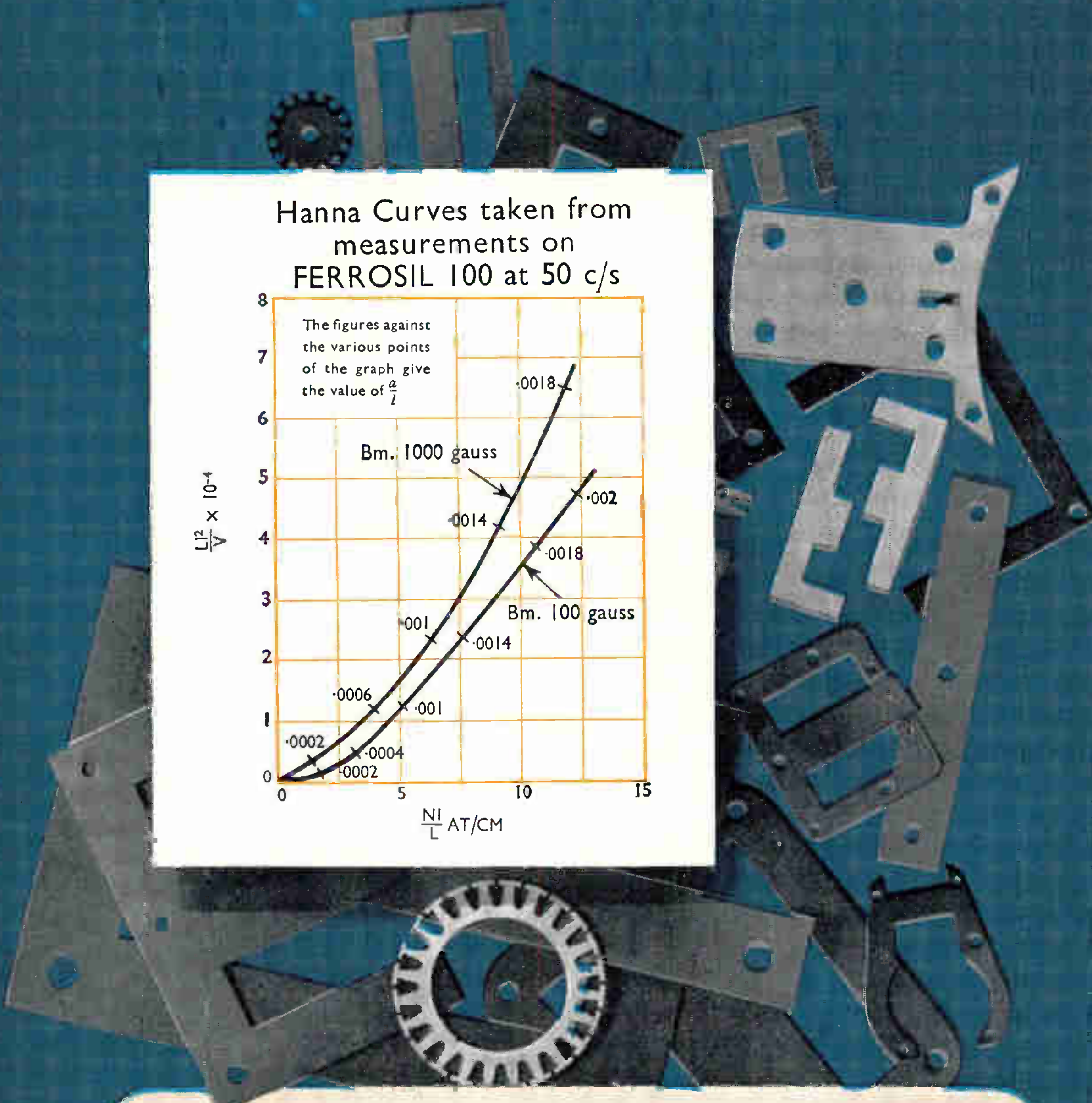

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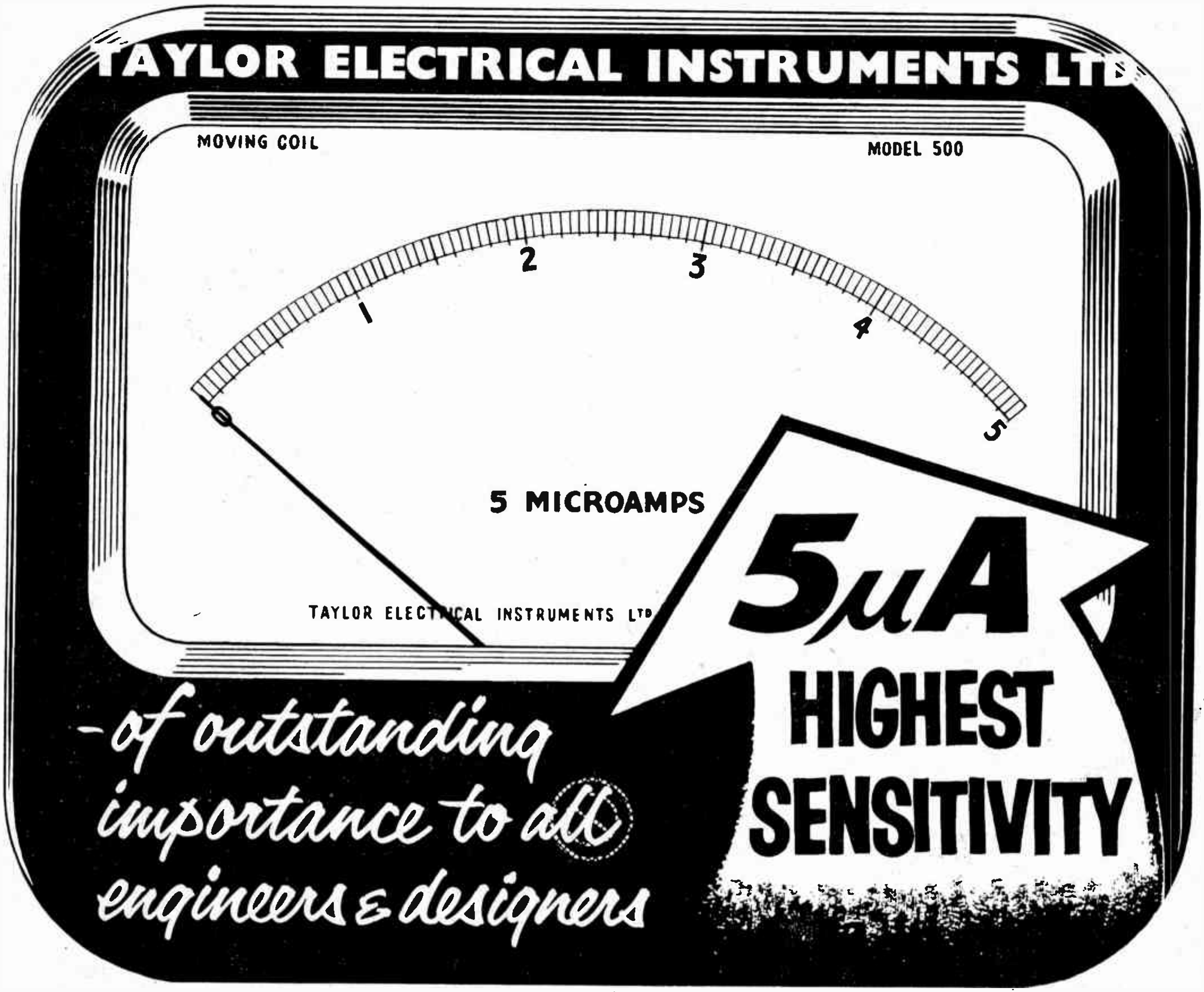
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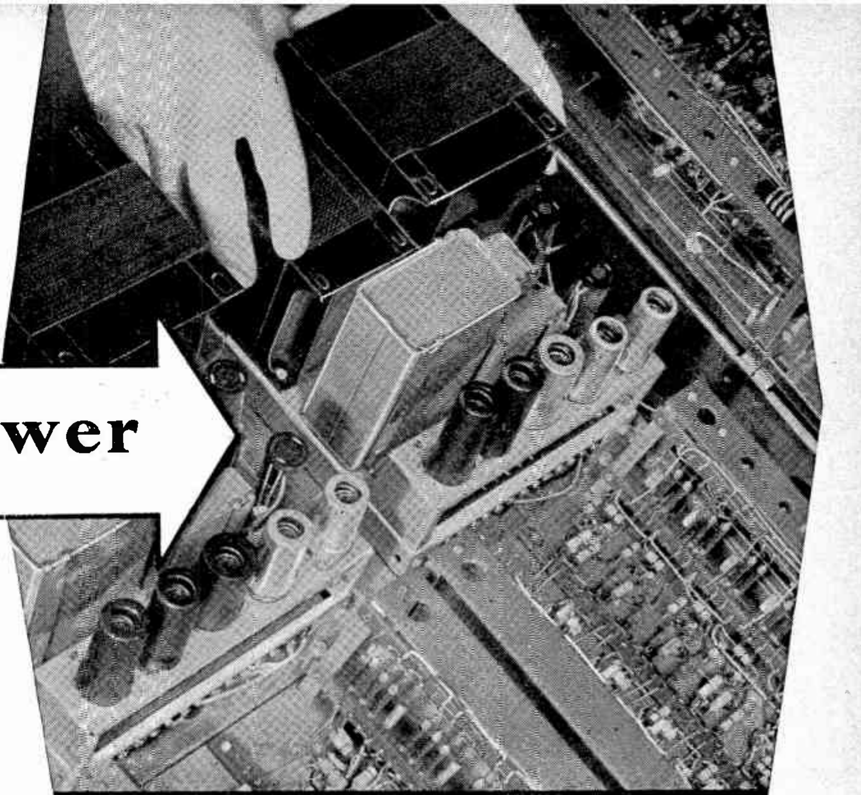

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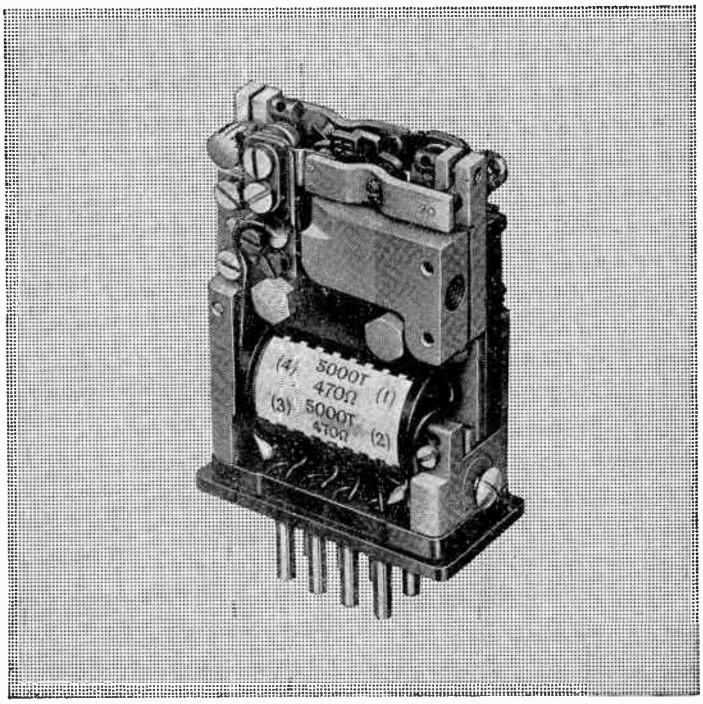
For further details send for leaflet F.3526 which gives information of operating characteristics and the available range of coils.



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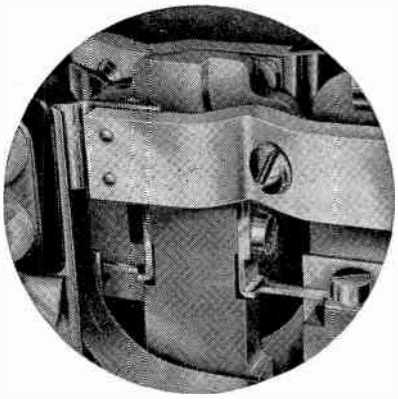

Close-up of armature suspension —magnet removed

* *In the centre-stable form, the armature of the relay remains in a central position between the side-contacts when the relay is unenergized, and moves to one or other side-contact when current is applied to the coil.*

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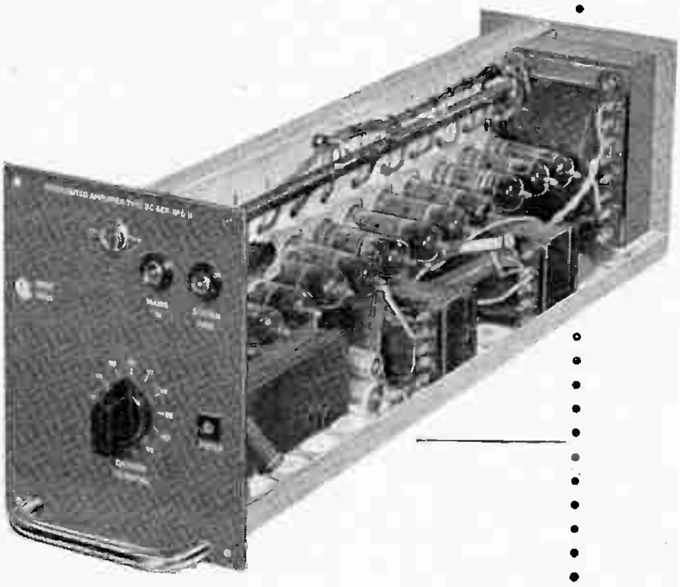

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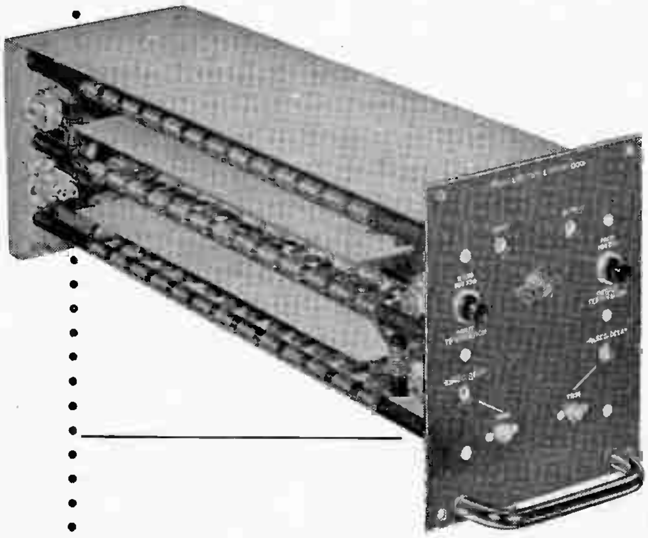

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CONTENTS

VOLUME 34 NUMBER 8 AUGUST 1957

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VOLUME 34 NUMBER 8
AUGUST 1957

ELECTRONIC & RADIO ENGINEER

incorporating WIRELESS ENGINEER

Radio on Show

TOWARDS the end of this month (28th August–7th September) the National Radio Exhibition will open at Earls Court. It is an exhibition designed primarily for the non-technical public and is concerned mainly with domestic sound and television receivers.

Hidden behind the façade of cabinets, the discerning will find much of technical interest if they refuse to let themselves be distracted by the popular sideshows of the B.B.C. and I.T.A. Domestic apparatus does not have the complexity of the radar set or of the computer, nor does it have to withstand the extremes of temperature and vibration of military equipment. It does, however, have to perform reliably in the hands of completely non-technical people; it must be attractive in appearance and low in price and, at the same time, be suited to quantity production.

All these things are not easy to achieve in combination and a good deal more sound engineering goes into modern domestic apparatus than is always realized. Because the exhibition is for the public the technicalities of the apparatus are not always evident and the visitor who is interested in them does not always find it easy to get behind externals. Our report of the exhibition, which will be in the October issue, will therefore be concerned mainly with the engineering aspects of the apparatus on view.

The National Radio Exhibition is not the only show in our field this month. Almost coincident with it is the Engineering, Marine, Welding and Nuclear Energy Exhibition (29th August–12th September) at Olympia. This is, of course, in itself neither an electronics nor a radio show. However, electronics enters into most branches of engineering nowadays and radio and radar are important ancillaries to all forms of shipping. It may thus be expected that there will be a good many exhibits of interest to us.

Colour Television Transmission

PRACTICAL ASPECTS OF THE TWO-SUB-CARRIER SYSTEM

By K. Teer*

SUMMARY. *In recent years a colour-television transmission system using two sub-carriers for the transmission of the chrominance information has been developed.*

Practical results were demonstrated to Study Group XI of the C.C.I.R. in 1955 and 1956. The principles and evolution of the system have been described and discussed in previous articles. In this article the technical aspects are considered in more detail and data are given about modifications and improvements which have since been introduced.

In the last three years a colour-television system has been developed which is based on the use of two sub-carriers combined with a luminance signal. The underlying principles of multiplex transmission for colour-television purposes and a description of this so-called two-sub-carrier system (t.s.c. system) have been published in previous articles.[1, 2] The practical results were demonstrated on the occasion of the meeting of the C.C.I.R. Study Group XI in Brussels in 1955 as well as during the visit of the C.C.I.R. Study Group XI to Holland in April 1956.

In this article the principles and experimental circuits are treated in more detail and subsequent improvements and modifications made are described. The description is split into the following sections:

1. Choice of the sub-carriers
2. Sub-carrier modulator
3. Pre-emphasis of sub-carrier sidebands
4. Cross-talk compensation
5. Demodulation
6. Matrixing
7. Automatic ratio control
8. Receiver diagram.

Of these, pre-emphasis, cross-talk compensation and automatic ratio control, in particular, have been improved in order (a) to reduce as much as possible the 'overswing' of the sub-carriers beyond the signal range while keeping the colour signal-to-noise ratio acceptable and (b) to safeguard the colour balance against variations in the amplitude characteristic of the transmission path. By means of pre-emphasis and cross-talk compensation random noise and luminance-signal cross-talk in the colour signals can be decreased; this permits lower sub-carrier amplitude levels. Relative sub-carrier amplitudes of 25% and 15% are considered to be suitable if pre-emphasis and cross-talk compensation are applied. (Percentages are referred to the black-to-white range of the luminance signal.)

For these values no overswing will occur beyond black level, whereas the maximum overswing beyond white level is 40%, as shown in Fig. 1. Although the latter value is only 7% above that in the N.T.S.C. system, it cannot be denied that at this point the t.s.c. system is at a disadvantage compared with the N.T.S.C. system, because maximum overswing conditions occur more frequently in it. This effect was noticed in practice.

The stringent requirements on the amplitude balance of the luminance signal and the sub-carriers seem to result in less difference from the N.T.S.C. system than was expected, for automatic gain control can be obtained fairly simply at both transmitter and receiver. Moreover, in recent papers[3] on requirements in the N.T.S.C. system, 2–3 dB is mentioned as being the maximum permissible sub-carrier amplitude deviation for that system also.

Of the remaining sections, 'Choice of the sub-carriers'

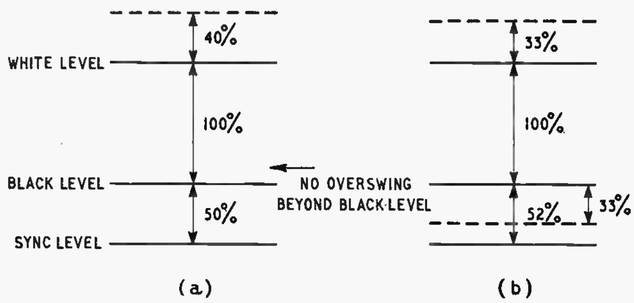


Fig. 1. *Maximum overswing in t.s.c. system (a) and N.T.S.C. system (b)*

gives a description of the experiments which led to the final choice of frequency and phase relations. 'Sub-carrier modulator' and 'Demodulation' deal with the typical problems which occur due to the large relative bandwidths of the sub-carriers. In 'Matrixing', the addition of mixed highs to small-band primary signals is discussed prior to a description of practical propositions. 'Automatic ratio control' gives some calculations concerning feedback gain control and the derivation of a reference voltage as well as practical solutions for the

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

comparison of colour-signal levels with luminance-signal level. Finally, in 'Receiver diagram' a complete diagram of the video part of a t.s.c. receiver is given.

Choice of Sub-Carriers

In order to find the frequency and phase relations for the sub-carriers which ensure minimum annoyance in the compatible picture, several experiments were carried out with combinations of two frequencies added to a television signal and having special relations to the scanning frequencies.

As described elsewhere,[2] not only the sub-carrier frequencies but also their difference frequency must show a pattern of low visibility, because this last frequency may easily occur in the picture in consequence of non-linearities in the transmission path, especially the non-linear cathode-ray tube characteristic. It is desirable, therefore, to choose an odd multiple of half the line frequency for this difference frequency. In order to establish such a difference the first idea was to choose sub-carrier frequencies which are an odd multiple of a quarter of the line frequency, because multiples of the line frequency have to be avoided. Experiments have shown, however, that for the lower sub-carrier frequency (about 3·6 Mc/s) an odd multiple of a quarter of the line frequency causes intolerable stroboscopic effects which appear as sloping lines moving in the horizontal direction.

As the second sub-carrier, with a corresponding pattern and thus with corresponding stroboscopic effects, gave much less annoyance than the first one, it was concluded that these effects decrease rather quickly with increasing sub-carrier frequency.

An experiment was made to find out whether this decrease in annoying effects with increasing frequency would permit the use of a second sub-carrier frequency which was a multiple of the line frequency. In this, the first sub-carrier was $(n + \frac{1}{2})f_l$ and the second was mf_l. Although in large areas of the picture the presence of the second sub-carrier was not very disturbing, it could not be tolerated owing to the interference of the vertical line structure, inherent in a multiple of the line frequency, which affected the fine picture detail. Annoying moiré effects occurred, especially in the vertical wedges of the test pattern.

An attempt was made to reduce these effects by introducing phase shifts in the second sub-carrier frequency at the beginning of each frame. Between even and odd frames a phase shift of 180° was introduced. This caused an interlacing of the dots on even and odd lines, which broke up the vertical line pattern (see Fig. 2). By means of this phase alternation the disturbing effects were eliminated and this pair of sub-carriers showed no effects that made their presence in the picture intolerable.

A further improvement was obtained in the first sub-carrier by introducing a phase alternation over 90° as described in Ref. 2. As this phase alternation causes a corresponding phase alternation in the difference frequency, the perceptibility in this pattern is also reduced. Apart from this reduction in visibility the introduction of the 90°-phase shift in the first sub-carrier gives another advantage which will be discussed in the section entitled 'Cross-talk compensation'.

The combination of an odd multiple of half the line frequency with 90°-phase alternation and a multiple of the line frequency with 180°-phase alternation was used in all experiments with the t.s.c. system.

Sub-Carrier Modulator

The sub-carriers are modulated by colour signals which in the t.s.c. system are the red and blue primary signals E_R and E_B. The modulation depth is 100%.

As is well known, modulation can be obtained by means of non-linear devices or valves with multiplicative action between two of their electrodes. If these devices are used in single-ended circuits, the sub-carrier and colour signals will not be suppressed. Separation of the colour signal E_c and the modulated sub-carrier $E_c \cos \omega t$ by filtering is also not readily possible, owing to

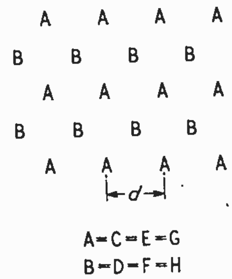


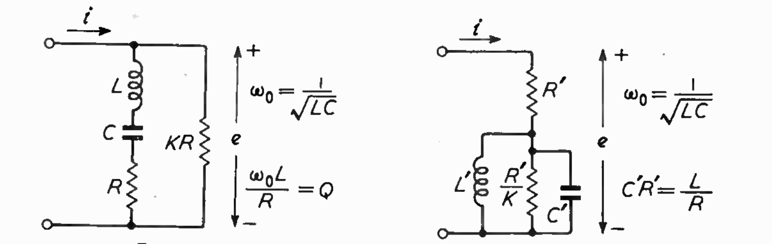
Fig. 2. *The dot pattern of a multiple of the line frequency with a phase alternation of 180° at the beginning of each frame*

the partial overlap of their frequency bands. For this reason modulator circuits are often used in which two modulators work in push-pull so that video signal and sub-carrier cancel on the common output impedance. For this purpose video signals and sub-carriers of opposite polarity are necessary at the input-terminals. The circuit has to be carefully designed in order to ensure proper balancing. For the video signal especially it is rather difficult to obtain and maintain good suppression over the whole band. Nevertheless, it is possible to build suitable modulators of this kind.

If pre-emphasis of the sub-carrier sidebands is used, as has been proposed in the t.s.c. system in order to improve the signal-to-noise ratio for the colour signal at the receiver, it is still more difficult to procure sufficient cancelling of the video signal.

In this case it is advisable to look for another solution of the problem. This can be found by modulating the colour signal first on a higher frequency than the sub-carrier in such a way that the video signal can be eliminated by a high-pass filter. By means of frequency conversion in a second modulator the carrier is converted into the right sub-carrier frequency.

Of course, sub-carrier elimination is still necessary in the first modulator, but the critical video-signal balance can be omitted. The second modulator can be of very simple design. Suppression of sub-carrier and colour signal presents no problems here because the carrier (which is the difference between the carrier of the first modulator and the sub-carrier) as well as the modulating signal (which is the output signal of the first modulator) can be easily eliminated by filters. This method was used in the experimental t.s.c. system. The colour system was first used to modulate the third harmonic of the sub-carrier frequency, and then the appropriate frequency conversion was effected by mixing with the second harmonic.



$$e = i\,\frac{K}{1+K}\,R\left[\frac{j\frac{x}{Q} + 1 - x^2}{j\frac{x}{Q} + \frac{1-x^2}{1+K}}\right] \qquad \text{(a)}$$

$$e = \frac{1+K}{K}\,R'\left[\frac{j\frac{x}{Q} + \frac{1-x^2}{1+K}}{j\frac{x}{Q} + 1 - x^2}\right] \qquad \text{(b)}$$

Fig. 3. *A simple pre-emphasis and de-emphasis circuit*

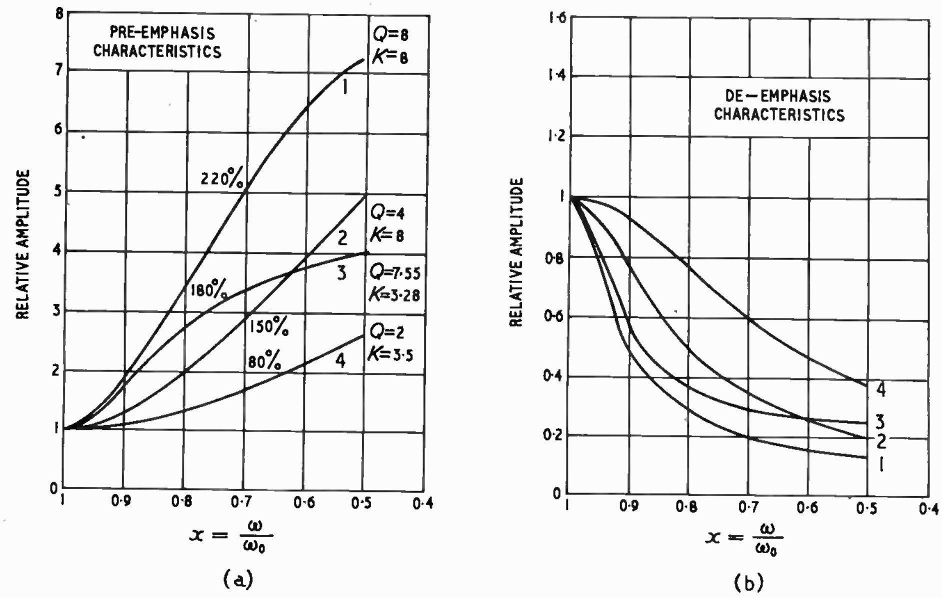


Fig. 4. *Characteristics of the circuit of Fig.* 3 *for several values of Q and K;* (*a*) *for pre-emphasis and* (*b*) *for de-emphasis*

Pre-Emphasis of Sub-Carrier Sidebands

The choice of the relative sub-carrier amplitude in the composite signal has to be a compromise between compatibility and colour-picture quality. If the amplitude is increased, the annoyance caused by the sub-carrier in monochrome reception will increase and, if it is decreased, the colour noise and luminance-signal cross-talk in colour reception will increase. In order to determine a suitable value it is necessary to study both annoying effects: the presence of the sub-carrier in monochrome reception and the noise in colour signals. In either case, the influence of all undesired components over the whole frequency spectrum of the modulated sub-carrier has to be considered. In this section we shall consider whether an improvement in the t.s.c. system can be made, especially in the signal-to-noise ratio for colour signals at the receiver, by transmitting the colour-sub-carrier sidebands with a higher relative amplitude level than the sub-carrier frequency itself. This may be of use because the annoyance of the modulated sub-carrier in the monochrome picture is mainly caused by the sub-carrier frequency and much less by the sub-carrier sidebands. An emphasis of the sidebands, therefore, results in only a slight increase of annoyance, whereas a substantial gain in signal-to-noise ratio at the receiver can be expected.

At the transmitter, pre-emphasis has to be applied to the modulated sub-carrier before the addition to the luminance signal, and in the receiver the pre-distortion introduced has to be eliminated by a corresponding 'de-distortion' (de-emphasis).

Of course, it would need extensive work to obtain all the necessary data. However, owing to some practical restrictions the number of possibilities is limited.

In the first place, the receiver de-emphasis circuit must be of simple design and well-reproducible type.

In the second place, peaking of sidebands may not go so far as to allow the transients in the sub-carrier envelope to reach far beyond the signal range of the transmitter.

In Fig. 3 simple pre-emphasis and corresponding de-emphasis circuits are shown. Both sidebands are peaked by the pre-emphasis circuit. This is not strictly necessary, as the sub-carrier is transmitted with only the lower sideband. However, this type of emphasis contributes to the necessary band-pass filter action which is needed for separating the modulated sub-carrier from the combined signal.

In Fig. 4 the amplitude characteristics of the pre-emphasis and de-emphasis circuits of Fig. 3 are represented for several values of the parameters K and Q. If $Q/(1+K)$ is chosen sufficiently small, the series resistance R'/K in the receiver circuit can be omitted, because its influence in the region of the sidebands can then be ignored. If $Q/(1+K) < 0{\cdot}45$ the deviation at the lower limit for the 3·6-Mc/s sub-carrier sidebands is less than 20% (curves 2 and 4).

For all curves of Fig. 4 is also given the percentage of overshoot that will occur in the response to a carrier transient. These figures hold for a bandwidth of the colour signal which is half the carrier frequency.

These values seem to be very high, but it should be noted that an overshoot of 100% of the red sub-carrier amplitude is only 10% of the maximum picture carrier level if sub-carrier amplitudes as given in the introduction (25% and 15%) are applied.

The permissible amount of overshoot depends on many things including the duration of the overshoot and possible measures at transmitter and receiver to reduce overmodulation distortion. For instance, it would be possible at the receiver to boost the picture-carrier frequency before the chrominance detector in order to reduce modulation depth during sub-carrier transients. However, the utmost limit is reached at 150% to 200%.

Once the practical restrictions have been formulated we must consider the influence of the pre-emphasis on noise and luminance-signal cross-talk in the colour signal.

We shall first direct our attention to the random noise which is assumed to be added in the transmission path and to be of equal energy distribution.

The distribution of the noise spectrum will be changed at the receiver by the de-emphasis circuit and the filter network in the colour-signal channel. Moreover, it has to be taken into account that the visual sensitivity for noise decreases with increasing frequency. This effect can be taken into consideration by the introduction of a so-called 'weight function' giving the relative sensitivity for noise as a function of frequency.

If $F_1(\omega)$ represents this weight function, $F_2(\omega)$ the filter characteristic and $F_3(\omega)$ the de-emphasis curve, converted to the video band, the weighted r.m.s. value of noise will be proportional to

$$\sqrt{\int_0^{\omega_B} F_1(\omega)F_2^2(\omega)F_3^2(\omega)d\omega}$$

If the de-emphasis is not applied this value will be

$$\sqrt{\int_0^{\omega_B} F_1(\omega)F_2^2(\omega)d\omega}$$

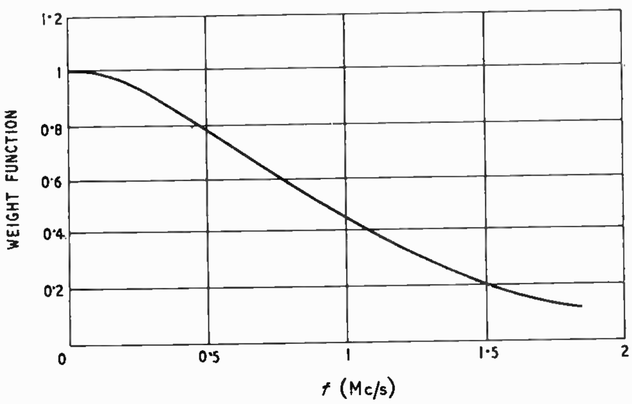


Fig. 5. *Weight-function for random noise in a 625-line television system*

So the improvement in signal-to-noise ratio is given by

$$\frac{\sqrt{\int_0^{\omega_B} F_1(\omega)F_2^2(\omega)d\omega}}{\sqrt{\int_0^{\omega_B} F_1(\omega)F_2^2(\omega)F_3^2(\omega)d\omega}}$$

For some de-emphasis characteristics this value has been calculated, applying a weight function given in Fig. 5. This function was derived as the mean value of curves published by several authors,[4, 5, 6] and a filter characteristic of the form shown in Fig. 6 was assumed to be present. The gain in signal-to-noise ratio for the characteristics of Fig. 4 is:

Curve No.	1	1·69
,, ,,	2	1·39
,, ,,	3	1·58
,, ,,	4	1·18

For No. 2, which is the characteristic applied in the experimental system, the gain was also estimated by observing the picture, and a satisfactory correspondence was found (a gain of 1·5).

It is more difficult to give a corresponding figure of merit for the decrease in luminance signal cross-talk as has been done for the random noise. The spectrum components of the cross-talk signal are correlated and

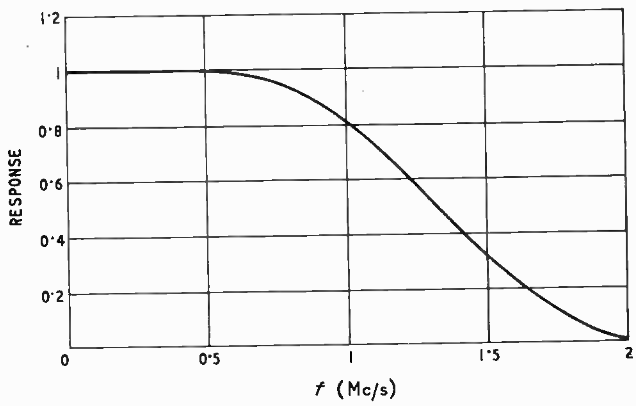


Fig. 6. *Low-pass filter characteristic as applied in signal-to-noise considerations for a 'pre- and de-emphasized' sub-carrier*

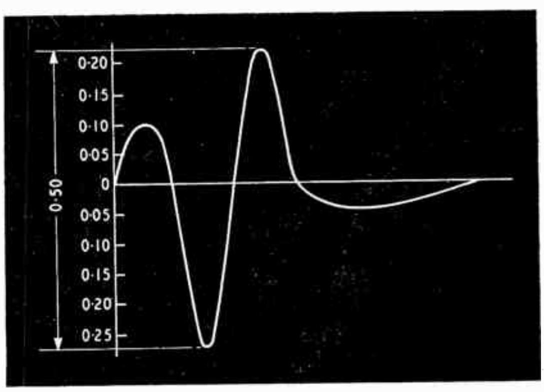


Fig. 7. *Transient response of a sub-carrier band-pass filter*

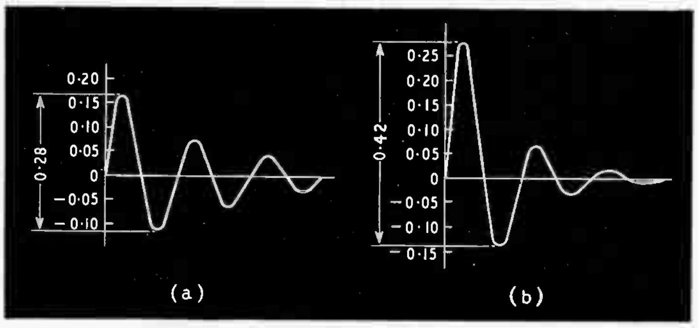


Fig. 8. *Transient response of the de-emphasis circuits of Fig. 4 (b): (a) circuit 2, (b) circuit 4*

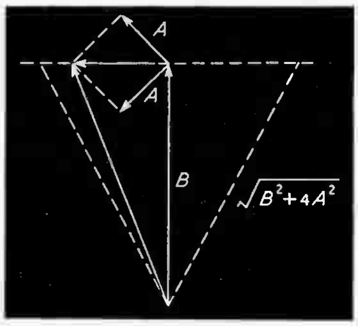


Fig. 9. *Vector diagram for 'cross-talk compensated' sub-carrier*

this annoying effect depends on picture content. Nevertheless, we will try to get some information by considering the cross-talk of a transient in the luminance signal. We shall do this only for the red signal, because the blue sub-carrier region is higher in frequency and the cross-talk is much less important.

In Fig. 7 is given the transient response of the band-pass filter which has to be employed in the sub-carrier channel if no pre-emphasis is used. This response was obtained experimentally using a flat-staggered triple which was 3 dB down at 1·8 Mc/s and 4·2 Mc/s.

If pre- and de-emphasis are introduced in the transmission of the sub-carrier, the cross-talk of the luminance signal is reduced by the de-emphasis circuit. In Fig. 8 the results are given for curves 2 and 4 of Fig. 4(b) if only the de-emphasis circuit is present in the sub-carrier channel and no further band-filter is used. The maximum excursion appears to be decreased to 56% and 84% of that of Fig. 7.

Of course, the annoyance of this cross-talk effect depends not only on this excursion but also on the increment of the response. It was found by observing a television picture made up of the red colour signal that for equal annoyance the relative sub-carrier level could be decreased to 50% and 65% respectively applying the de-emphasis curves 2 and 4 respectively.

Summing up the foregoing considerations and experiments, it may be concluded that the application of pre-emphasis in the t.s.c. system reduces the noise as well as the cross-talk in the colour signals. The gain in signal-to-noise ratio is not very high; but what is more important is the cross-talk reduction. In addition to these improvements, a simplification of the sub-carrier channel is obtained, because the band-pass filter is entirely or partially replaced by the simple de-emphasis circuit. In the experimental system, a pre-emphasis has been introduced for the red sub-carrier according to curve 2 of Fig. 4. Band-pass filters in the receiver for the red sub-carrier could be omitted. Provisionally, no pre-emphasis was applied to the blue sub-carrier. For reasons of receiver simplicity it is also useful there.

Cross-Talk Compensation

The reduction of cross-talk in the colour signal as just described is based on an attenuation of the relative value of the luminance signal in the pass-band of the sub-carrier. This attenuation varies with frequency and is maximum for the outer regions of the pass-band and zero for the sub-carrier frequency. Near the sub-carrier, therefore, the cross-talk is not decreased.

Now there is another method for cross-talk diminution which is especially effective near the sub-carrier frequency. This method is here called 'cross-talk compensation', because the cross-talk effects are eliminated by introducing extra cross-talk effects of opposite sign.

If a luminance-signal component

$$A \cos (\mu t + \psi)$$

is present near the sub-carrier

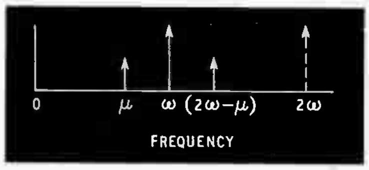
$$B \cos (\omega t + \theta)$$

a cross-talk component proportional to

$$A \cos [(\omega t + \phi) - (\mu t + \psi)] = A \cos [(\omega - \mu)t + (\phi - \psi)]$$

will be found in the colour signal.

Fig. 10. *Frequency spectrum for 'cross-talk compensated' sub-carrier*



Such a component would also be found if a luminance-signal component

$$A \cos [(\omega t + \phi) + \{(\omega t + \phi) - (\mu t + \psi)\}] = A \cos [(2\omega t + 2\phi) - (\mu t + \psi)]$$

were present.

Thus the cross-talk effect of $A \cos (\mu t + \psi)$ can be eliminated by introducing an extra interference

$$- A \cos [2\omega t + 2\phi - (\mu t + \psi)],$$

because after demodulation both interfering frequencies give rise to cross-talk effects of opposite sign.

The cancellation will be complete if synchronous detection is used. If envelope detection is used—as in the t.s.c. system—the cancellation is partial. This can be seen from Fig. 9, in which sub-carrier- and luminance-signal components are represented as vectors. The envelope is

$$\sqrt{B^2 + 4A^2 \cos^2 \{(\omega - \mu)t + (\phi - \psi)\}}$$

Hence its value varies between B and $\sqrt{B^2 + 4A^2}$. If $4A^2/B^2$ is sufficiently small

$$\sqrt{B^2 + 4A^2} \approx B\,(1 + 2A^2/B^2)$$

If only the original component $A \cos(\mu t + \phi)$ is present the envelope varies between $B + A$ and $B - A$. The modulation is therefore decreased by a factor A/B, which is a considerable reduction if the cross-talk components are sufficiently small with respect to the sub-carrier amplitude.

The cross-talk compensating signal

$$A \cos\{2\omega t + 2\phi - (\mu t + \phi)\}$$

can be obtained by modulating the second harmonic of the sub-carrier frequency with the luminance signal and by filtering off the lower sidebands in the region of the sub-carrier frequency (Fig. 10). It will be clear from the foregoing that the phase of this second harmonic has to be matched to the sub-carrier phase.

The cross-talk compensating signal of proper amplitude and phase is added to the combined signal before transmission. Its compensating action will be maintained during transmission only if there is no asymmetrical influence on the transmission characteristic for the modulated sub-carrier from transmitter up to sub-carrier demodulation. As the sub-carrier is single-sideband modulated and therefore passes a vestigial sideband filter in the receiver, the compensation is only possible for a rather small region on both sides of the sub-carrier. Consequently, at the transmitter, only a small band of compensating signal has to be added.

Of course, the cross-talk compensation will cause in principle a disturbing visual effect in the monochrome picture. However, this effect is small for two reasons. In the first place only a small band of signal components is introduced. In the second place, due to the phase alteration of 90° in the red sub-carrier, a 180°-phase alternation will occur in the second harmonic of this sub-carrier, which means that the disturbing effect is of opposite sign in successive fields.

The latter circumstance does not hold for the blue sub-carrier but, in general, means of reducing the interference level are more important for the red signal. Interference by the blue signal is far less perceptible.

The experiments were carried out by means of an arrangement shown in Fig. 11. The luminance signal modulates the second harmonic of the red sub-carrier. The phase of this frequency can be varied over 360°. From the modulator output signal a small frequency range near the sub-carrier frequency (about 0·2 Mc/s on both sides) is filtered out and added to the luminance signal, which is combined with the modulated sub-carriers and synchronizing pulses in order to form the composite signal. The amount of compensating signal can be adjusted and it can also be switched off. In order to correct for the phase shift applied to the compensating signal by the modulator filter, a corresponding delay of the luminance signal in the composite signal is introduced. The effect of cross-talk compensation was studied by observing the red receiver picture of the t.s.c. system under normal conditions and when cross-talk compensation was introduced. Optimum cross-talk compensation was adjusted by varying the amplitude of the compensating signal and the phase of the second harmonic. An attempt was made to get the same cross-talk level for both cases by changing the luminance-signal level; it was found that for this arrangement the decrease in cross-talk effects in the red picture was equivalent to a decrease of luminance-signal level by a factor of two.

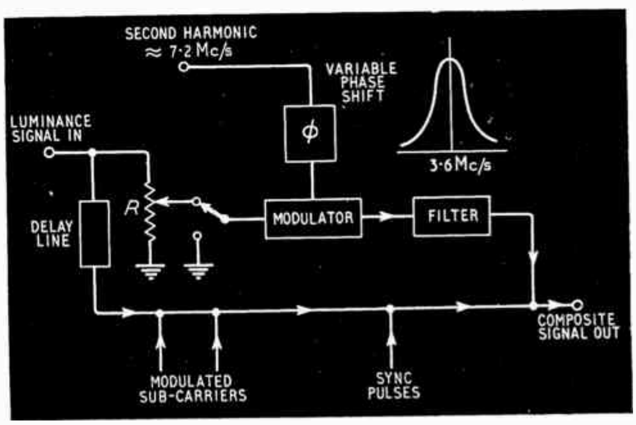


Fig. 11. *Arrangement for the addition of cross-talk compensating signal components*

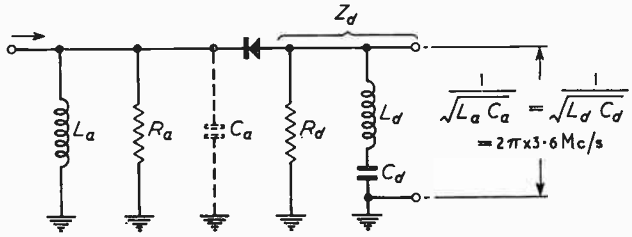


Fig. 12. *Detector circuit for the sub-carrier used in the t.s.c. system*

The question may arise as to what the effect of cross-talk compensation will be in the N.T.S.C. system. This has already been discussed by Loughlin, who first mentioned the possibility of cross-talk compensation.[5]

In the N.T.S.C. system only a limited reduction of cross-talk can be obtained, because the colour signals are modulated on sub-carriers having the same frequency and differing 90° in phase. The required phase of the second harmonic for compensation in one colour signal therefore differs 180° from that required for the other signal. Consequently if cross-talk cancellation is established for one signal the cross-talk effect increases in the other colour signal.

In general, it can be shown that in the N.T.S.C. system one of the colour components (which may be a colour-difference signal, the hue or the saturation) can be freed from luminance cross-talk, at the cost of a corresponding increase of cross-talk in the complementary colour information.

Perhaps the properties of human visual perception will allow an improvement of a picture-quality amelioration by a modified 'cross-talk distribution'.

However, there is a difficulty in the spurious pattern of the cross-talk compensating components. As a phase alternation of 90° is not introduced in the N.T.S.C. sub-carrier, the second harmonic is a multiple of the line frequency and so the added signal components will be fairly clearly visible in the luminance signal picture.

Demodulation

In the t.s.c. system, sub-carriers are demodulated by non-synchronous detectors. This type of detection is a common technique in communication engineering and is used in all radio and television receivers, but the problem here is somewhat more complicated because the relative bandwidth of the modulated carrier is not very small. The bandwidth of the red signal, for instance, is about half the sub-carrier frequency.

As in the literature only few data are present about the optimum design of detection circuits for single-sideband modulated sub-carriers with 100% modulation depth and large relative bandwidth, an investigation of this subject seems desirable for the t.s.c. system. However, the necessity for receiver simplicity limits the application of complicated networks and, as fairly good results were obtained by the simple detection circuit of Fig. 12, in all experiments with the t.s.c. system this circuit was used.

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(*To be concluded*)

Transducer Indicator System

ANALYSIS OF AN INDUCTIVE DISPLACEMENT INDICATOR

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

In the first article[1] to appear in this journal under its present title, a brief description was given of a displacement indicator consisting of an inductance transducer and a moving-coil dynamometer indicating instrument. The essential features of this device are illustrated in Fig. 1. In effect, this is an electromagnetic servo-mechanism which uses a.c. throughout. It comprises two units, a transmitter or sensing unit in which the small mechanical displacements to be measured modify the inductive arms of a bridge network, and a dynamometer indicator. Among the advantages which have been claimed for this system are higher sensitivity and better stability than can be achieved with the more common d.c. systems[2]. Performance has been stated to be substantially independent of variations of supply voltage and frequency. So far as the writer is aware, no theoretical analysis of the system has yet been published. Without such investigation, one's understanding of the behaviour of the arrangement can be only superficial and, of course, it is then not possible to assess the validity of these claims. The ingenuity of the system, no less than the importance it seems likely to acquire in telemetry, have prompted the following quantitative examination of its principles. Before commencing this, it should perhaps be mentioned that the same type of indicator has been employed with a transducer in which the mechanical displacement varies the mutual inductances of a pair of differentially-coupled transformers. Details of the theory of this system have been published[3], but these throw no light on the performance of the device now to be considered. The two systems are, in fact, fundamentally dissimilar.

From the circuit diagram of Fig. 1 it is seen that separate secondary windings of a mains transformer energize the yoke of the dynamometer indicator, and feed the bridge network consisting of the resistors R_1 and R_2 and the inductors L_3 and L_4. Across the other diagonal of the bridge is connected the moving coil of the indicator instrument. The latter is not spring-controlled, but is fed through ligaments which impose negligible mechanical torque. It is therefore free to take up a position in which the electromechanical torque is zero. The indicator coil is threaded by a fraction of the alternating flux carried by the yoke, and this induces

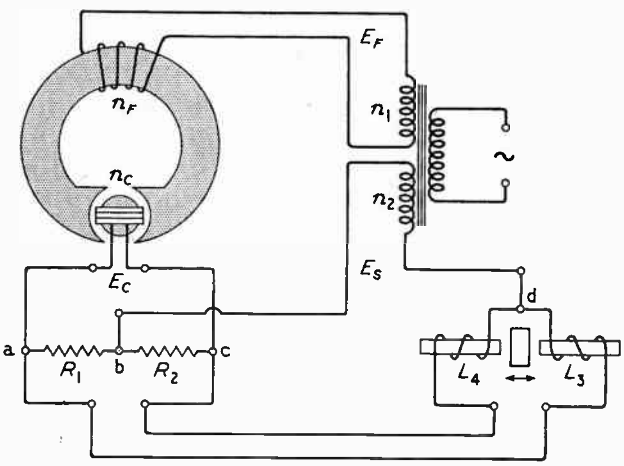

Fig. 1. *Transducer indicator system*

in it a voltage E_c which depends on its orientation. The current in the coil will therefore be a function of its position, the voltage E_s applied to the bridge, and the values of the bridge components. The mechanical displacements which it is desired to indicate are communicated through a stylus shaft to a high-permeability disc armature which moves in the air gap between the two coils. Displacement therefore modifies their self-inductances L_3 and L_4, and possibly their mutual inductance M as well.

It is, of course, well known that, averaged over a complete cycle, the torque experienced by a coil situated like that in the indicator instrument will not be zero if it carries a current having a component in phase with the alternating flux threading it. Since this current depends, in part, on the orientation of the coil, the latter, being without restraint, will seek the equilibrium position in which the current through the coil has no component in phase with the flux in the gap. The analytical problem is therefore to relate this equilibrium position to the values of the bridge parameters.

The Analytical Problem

Let us suppose that the flux in the yoke of the indicator instrument is represented by $\Phi \cos \omega t$. The flux in the indicator gap is then $\frac{\Phi}{\kappa} \cos \omega t$, where κ is the leakage factor. If ψ is the angle which the plane of the indicator coil makes with the flux direction, the flux threading the coil will be $\frac{\Phi}{\kappa} \cos \omega t \sin \psi$. This will induce in its n_c turns a voltage E_c proportional to $\frac{n_c}{\kappa} \Phi \sin \omega t \sin \psi$. The yoke flux induces a voltage proportional to $n_F \Phi \sin \omega t$ in the n_F turns of the field winding ; this is equal and opposite to the applied voltage E_F. Now $n_2 E_F = n_1 E_s$ where n_1 and n_2 are the numbers of turns on the secondaries of the supply transformer. The loads connected across the indicator and field coils will, of course, vary as the transducer setting is altered, but, from elementary transformer theory, this can be shown to have no effect on the flux linking them. Accordingly, E_s and E_c will remain in quadrature with the flux variations for all settings of the transducer. Furthermore E_s will be proportional to $\frac{n_2 n_F}{n_1} \Phi \sin \omega t$. Thus,

$$E_c = -\eta \sin \psi \, E_s \quad \cdots \quad (1)$$

where η is an instrumental constant proportional to the factor $\left[\frac{n_c n_1}{n_F n_2}\right]$. The problem may now be stated as follows. Determine the value of ψ for which the current in the indicator coil has no component in quadrature with E_c — and hence in phase with Φ — when the bridge network shown in Fig. 2 is energized with voltages E_s and $E_c = -\eta \sin \psi \, E_s$ of the same phase.

The prospect of analysing an unbalanced bridge network energized across both diagonals may, at first sight, seem formidable. The problem is, however, tedious rather than difficult. Appendix 1 derives the following expression for I_c in terms of E_s, E_c, and the values of the bridge parameters R_1, R_2, L_3, L_4, and M

$$I_c = \frac{E_s j\omega \left[R_1 (L_4 - M) - R_2 (L_3 - M)\right]}{R_1 R_2 j\omega (L_3 + L_4 - 2M) - \omega^2 (R_1 + R_2)(L_3 L_4 - M^2)}$$
$$+ E_c \frac{\left[R_1 R_2 - \omega^2 (L_3 L_4 - M^2) + j\omega (R_2 L_3 + R_1 L_4)\right]}{R_1 R_2 j\omega [L_3 + L_4 - 2M) - \omega^2 (R_1 + R_2)(L_3 L_4 - M^2)} \quad \cdots \quad (2)$$

The component of I_c in quadrature with E_s and E_c is, of course, the imaginary part of the expression on the right-hand side of Equ. (2). If the latter is rationalized I'_c, the imaginary part of I_c, is given by the relation,

$$I'_c = \frac{E_s}{f(z)} \omega \{[R_1(L_4 - M) - R_2(L_3 - M)] [\omega^2 (R_1 + R_2)(L_3 L_4 - M^2)]\}$$
$$+ \frac{E_c}{f(z)} \omega \{\omega^2 (L_3 L_4 - M^2)[R_1^2 L_4 + R_2^2 L_3 + 2 R_1 R_2 M] - (R_1 R_2)^2 (L_3 + L_4 - 2M)\} \quad \cdots \quad (3)$$

where

$$-f(z) = [R_1 R_2 \omega (L_3 + L_4 - 2M)]^2 + [\omega^2 (R_1 + R_2)(L_3 L_4 - 2M)]^2$$

The imaginary component I'_c of the current in the indicator coil is thus zero when

$$\frac{E_c}{E_s} = -\frac{[R_1(L_4 - M) - R_2(L_3 - M)][\omega^2 (R_1 + R_2)(L_3 L_4 - M^2)]}{\omega^2 (L_3 L_4 - M^2)[R_1^2 L_4 + R_2^2 L_3 + 2 R_1 R_2 M] - (R_1 R_2)^2 (L_3 + L_4 - 2M)}$$
$$= -\frac{(R_1 + R_2)[R_1(L_4 - M) - R_2(L_3 - M)]}{[R_1^2 L_4 + R_2^2 L_3 + 2 R_1 R_2 M] - \frac{(R_1 R_2)^2 (L_3 + L_4 - 2M)}{\omega^2 (L_3 L_4 - M^2)}} \quad \cdots \quad (4)$$

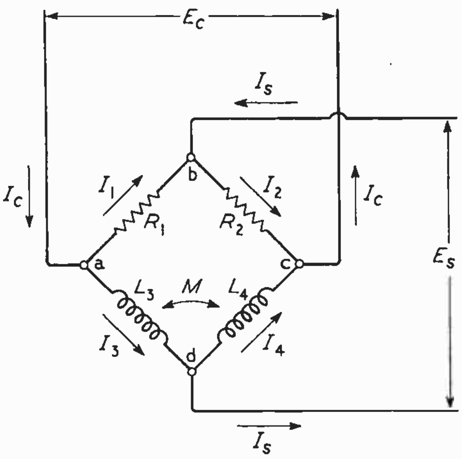

Fig. 2. *Equivalent bridge network*

Owing to the presence of magnetic materials in the system, I_c will inevitably contain harmonics of the supply frequency ; moreover, the harmonic content may vary with the setting of the transducer. It is therefore important that the condition specified by Equ. (4)

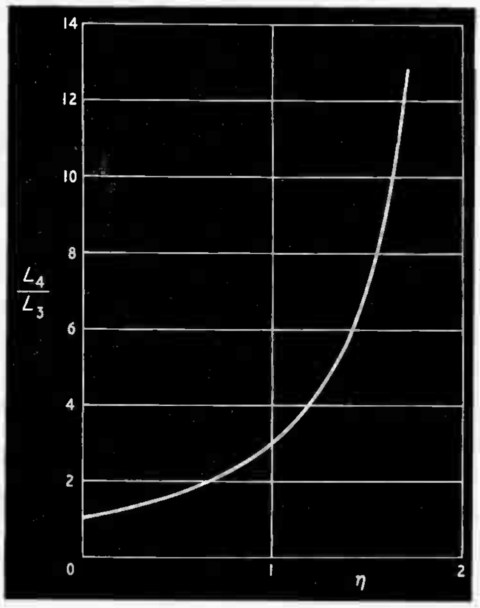

Fig. 3. *Value of ratio L_4/L_3 to give deflection of* $\pm 90°$

should be independent of frequency. This requires that

$$\frac{(R_1R_2)^2(L_3+L_4-2M)}{\omega^2(L_3L_4-M^2)} \ll [R_1{}^2L_4+R_2{}^2L_3+2R_1R_2M].$$

If this is so, then since, according to Equ. (1), $E_c = -\eta E_s \sin\psi$, the indicator will take up the position for which

$$\eta \sin\psi = \frac{(R_1+R_2)[R_1(L_4-M)-R_2(L_3-M)]}{R_1{}^2L_4+R_2{}^2L_3+2R_1R_2M} \qquad (5)$$

The Response of the System

The implications of Equ. (5) become more evident if we consider the special case in which both resistive arms are equal to R. In this instance, the torque on the indicator coil will disappear when

$$\eta \sin\psi = \frac{2(L_4-L_3)}{L_3+L_4+2M} \quad .. \quad .. \quad .. \qquad (6)$$

and the condition that the system should not be frequency sensitive is that R^2 should be negligible in comparison with $\omega^2 (L_3 L_4 - M^2)$.

From a knowledge of the manner in which ψ varies with the values of the transducer inductive elements, it is now possible to ascertain the way in which displacement should vary these in order to give indicator deflections proportional to displacements. To simplify the illustration of this, we shall assume that there is no mutual coupling between the transducer windings. In this case, Equ. (6) reduces to,

$$\eta \sin\psi = \frac{2\,(L_4/L_3-1)}{(L_4/L_3+1)} \quad .. \quad .. \quad .. \qquad (7)$$

and the ratio L_4/L_3 required to give full-scale deflection of 90° is seen to be,

$$\frac{L_4}{L_3} = \frac{2+\eta}{2-\eta}$$

The value of the instrumental constant η therefore determines the sensitivity of the system. Fig. 3 shows how, as η is reduced, smaller values of the ratio L_4/L_3 are needed for full-scale deflection.

For any given value of η in Equ. (7) we may calculate the ratio L_4/L_3 corresponding to a series of indicator deflections between $\pm 90°$. Typical results are shown in Fig. 4. If now the horizontal scale of this diagram is regarded as representing the displacement δ of the transducer stylus, then each of the curves indicates the type of variation in the ratio L_4/L_3 required to ensure a linear relationship between transducer displacement and meter deflection for the selected value of the instrumental constant η. The latter, which can be adjusted by altering the ratio of turns on the two secondary windings of the supply transformer, determines the indicator deflection corresponding to the maximum displacement of the transducer stylus. The shapes of the curves in Fig. 4 suggest that it will usually be easier to secure a reasonably linear relation between δ and ψ if deflections are limited to, say, $\pm$ 60°.

Restoring Torque

When designing a system such as this, it is important to know the factors affecting the magnitude of the torque experienced by the indicator coil when it is displaced slightly from its equilibrium position. Such a displacement affects E_c but leaves E_s unaltered. In the simplified case in which there is no mutual coupling between the transducer windings and $R_1 = R_2 = R$, if $R^2/\omega^2 L_3L_4$ is very much smaller than unity,

$$I'_c = \frac{E_s}{f(z)}\,\omega\,[2\,\omega^2 R^2 L_3 L_4 (L_4-L_3)]$$
$$+\frac{E_c}{f(z)}\,\omega\,[\omega^2 R^2 L_3 L_4 (L_3+L_4)] \quad .. \qquad (8)$$

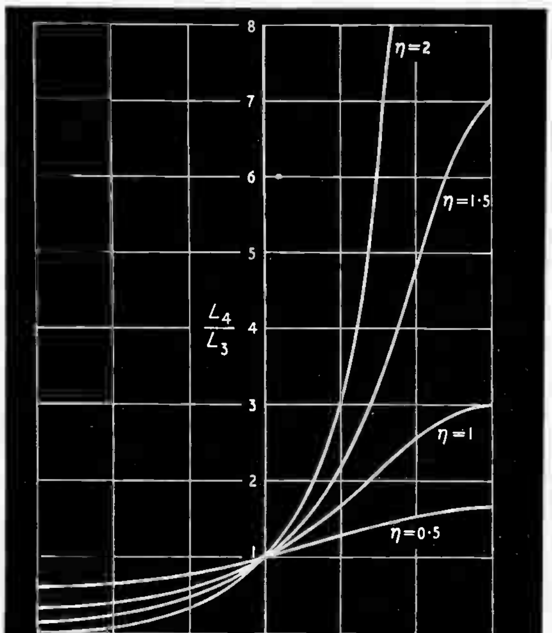

Fig. 4. *Variation of ratio L_4/L_3 with δ to give linear relation between transducer displacement δ and inductor deflection Ψ*

where $-f(z) = \omega^2 R^2 [R^2 (L_3 + L_4)^2 + 4\omega^2 (L_3 L_4)^2]$
$\approx 4\, \omega^4 R^2 (L_3 L_4)^2$

if R^2 is negligible in comparison with

$$\frac{\omega^2 L_3 L_4}{\frac{1}{4}(L_3/L_4 + 2 + L_4/L_3)}$$

Then, since $E_c = -\eta\, E_s \sin\psi$

$$I'_c = \frac{E_s \omega}{4\omega^2 L_3 L_4}\{2 (L_4 - L_3) - \eta\, (L_3 + L_4) \sin\psi\}$$

When the coil is in equilibrium in a position ψ, the term in the bracket is zero. Displacement of the indicator coil through a small angle $\Delta\psi$ increases I'_c from zero to

$$\Delta I'_c = -\frac{E_s\, \omega\, (L_3 + L_4)}{4\, \omega^2\, L_3\, L_4} \cos\psi\, \Delta\, \Psi$$

This will set up a restoring torque ΔT proportional to $\Phi\ \Delta I'_c$. The coefficient $\dfrac{\Delta T}{\Delta\psi}$ is the 'stiffness' of the system. Thus,

$$\frac{\Delta T}{\Delta\psi} \propto \Phi \frac{E_s\, \omega\, \eta\, (L_3 + L_4)}{4\, \omega^2\, L_3\, L_4} \cos\psi$$

Now Φ is proportional to $\dfrac{\eta_1}{\eta_F\, \eta_2}\, \dfrac{E_s}{\omega}$ and hence to $\dfrac{\eta}{\eta_c}\, \dfrac{E_s}{\omega}$, so that the variation in 'stiffness' through the scale range is,

$$\frac{\Delta T}{\Delta\psi} \propto \frac{\eta^2}{\eta_c} E_s^2 \frac{(L_3 + L_4)}{4\, \omega^2\, L_3\, L_4} \cos\psi \qquad \ldots \quad \ldots \quad (9)$$

The general inferences from this relation are not unexpected. As ψ approaches $\pm\, 90°$, $\Delta T/\Delta\psi$ falls to zero. Elsewhere the force tending to restore equilibrium may be increased by raising the supply voltage to the bridge network, or by making η larger. The latter method, however, reduces sensitivity. The value of the factor $(L_3 + L_4)/L_3\, L_4$ may be expected to increase slightly towards the extremities of the stylus displacement. Consequently, the coefficient $\Delta T/\Delta\psi$ is maintained at a somewhat higher value than if it followed the simple cosine relation. Its fall as Ψ approaches $\pm\, 90°$ will therefore be correspondingly steep.

Provided it does not vary with deflection, the effect of mutual coupling between the transducer windings can be inferred from the foregoing analysis. If, however, M is a function of δ, theoretical investigation becomes very complicated.

Alternative Connection of Transducer and Indicator

In Appendix 2, the solution of the network problem has been carried a stage further in order to obtain an expression for I_s, the supply current to the bridge. An interesting alternative to the circuit arrangement in Fig. 1 is that in which the points of application of the supply voltage E_s, and of the indicator coil are interchanged. In this instance, the voltage E_s is applied between a and c, and the indicator coil connected across b and d as shown in Fig. 5.

An expression for the current in the indicator coil is then given by replacing E_s by E_c, and I_s by I_c in Equ. 33 of Appendix 2. We then have,

$$I_c = E_s \frac{j\omega[R_1(L_4 - M) - R_2(L_3 - M)]}{j\omega R_1 R_2 (L_3 + L_4 - 2M) - \omega^2 (R_1 + R_2)(L_3 L_4 - M^2)}$$
$$+ E_c \frac{j\omega(R_1 + R_2)(L_3 + L_4 - 2M)}{j\omega R_1 R_2 (L_3 + L_4 - 2M) - \omega^2 (R_1 + R_2)(L_3 L_4 - M^2)} \qquad \ldots \quad \ldots \quad \ldots \quad (10)$$

The imaginary component I'_c of the current in the indicator coil is given by,

$$I'_c = \frac{E_s}{f(z)}\, \omega^2 [R_1(L_4 - M) - R_2(L_3 - M)]\, [(R_1 + R_2)\, (L_3 L_4 - M^2)]$$
$$+ \frac{E_c}{f(z)}\, \omega^2 (R_1 + R_2)^2 (L_3 + L_4 - 2M)(L_3 L_4 - M^2) \quad (11)$$

where

$$-f(z) = [R_1 R_2\, \omega(L_3 + L_4 - 2M)]^2 + [\omega^2\, (R_1 + R_2)\, (L_3\, L_4 - M^2)]^2$$

I'_c is zero when,

$$\frac{E_c}{E_s} = -\frac{R_1\, (L_4 - M) - R_2\, (L_3 - M)}{(R_1 + R_2)\, (L_3 + L_4 - 2M)} \qquad \ldots \quad (12)$$

This relation, unlike that shown in Equ. (4) is not a function of ω and therefore the condition which it represents is independent of frequency for all values of R_1 and R_2. As before, $E_c = -\eta\, E_s \sin\psi$ so that with the transducer connected in this way, the indicator coil will take up the position for which,

$$\eta \sin\psi = \frac{R_1\, (L_4 - M) - R_2\, (L_3 - M)}{(R_1 + R_2)\, (L_3 + L_4 - 2M)} \qquad \ldots \quad (13)$$

In the special case in which $R_1 = R_2 = R$, and there is no coupling between the inductive elements of the transducer, this relation reduces to

$$\eta \sin\psi = \frac{L_4 - L_3}{2\, (L_3 + L_4)} \qquad \ldots \quad \ldots \quad \ldots \quad (14)$$

The ratio L_4/L_3 required in order to obtain a $90°$ deflection is, in this instance,

$$\frac{L_4}{L_3} = \frac{1 + 2\eta}{1 - 2\eta}$$

The sensitivity of a system connected in this manner is, for any given value of η, evidently lower than that for the arrangement analysed earlier. Indeed, a value of

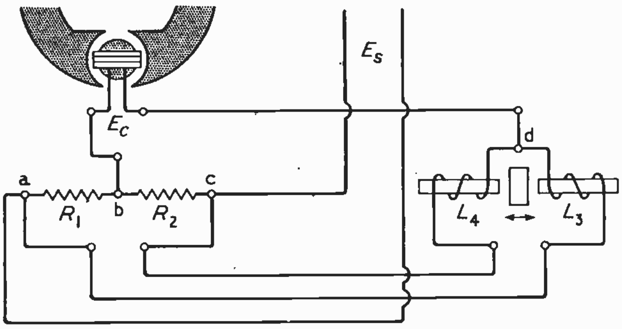

Fig. 5. *Alternative connection of transducer and indicator*

the ratio L_4/L_3 which, for a given value of the instrumental constant η provides a 90° deflection with the circuit shown in Fig. 1, will require a constant of $\eta/4$ to give the same deflection if the circuit is connected as in Fig. 5. Lower sensitivity is, however, compensated by a higher coefficient of 'stiffness', since for the system shown in Fig. 5

$$\frac{\Delta T}{\Delta \psi} \propto \frac{\eta^2}{\eta c} E_s^2 \frac{(L_3 + L_4)}{\omega^2 L_3 L_4} \cos \Psi \qquad \ldots \quad \ldots \quad (15)$$

Although the network analysis suggests that the equilibrium position of the indicator coil in the arrangement just described is independent of frequency, it would not be satisfactory in practice to rely on calibration at one frequency for operation at another. The constants of the magnetic materials, and in particular the transducer characteristics are, to some extent, functions of frequency. Nevertheless there may be substantial advantages in a system in which no inherent restrictions are placed on the choice of values for the resistive arms of the bridge. Such an arrangement may be designed and calibrated for use at any convenient frequency in the audio range. Moreover it may offer attractive possibilities for obtaining a particular type of indicator response from a given transducer characteristic by using unequal resistive arms in the bridge.

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[1] H. G. M. Spratt, "Transducer Characteristics", *Electronic and Radio Engineer*, 1957, Vol. 34, No. 1, p. 2.
[2] M. T. Mason, "Remote Indication and Control", *Elect. Rev., Lond.*, 1955, Vol. 157, No. 7, p. 295.
[3] J. E. Houldin, A. Matthews and G. T. Thompson, "The Principles of an Instrument for the Electrical Indication of Mechanical Quantities", *Instrument Practice*, 1952, Vol. 7, No. 1, p. 31.

APPENDIX I

If the inductive arms of the network shown in Fig. 2 have negligible resistance, then at a frequency $\omega/2\pi$, the currents in the four arms of the bridge satisfy the following relations

$$I_1 + I_3 = I_2 + I_4 \quad \ldots \quad (16)$$

$$E_c = I_1R_1 + I_2R_2 \quad \ldots \quad (17)$$

$$E_s = I_2R_2 - I_4 j\omega L_4 + I_3 j\omega M \quad \ldots \quad (18)$$

$$E_s = -I_1R_1 + I_3 j\omega L_3 - I_4 j\omega M \quad \ldots \quad (19)$$

The current I_c, for which we require an expression in terms of the voltages E_c and E_s and the values of the bridge components, is related to the currents in the bridge arms as follows,

$$I_c = I_1 + I_3 = I_2 + I_4$$

Substituting

$$\text{and} \quad \left.\begin{array}{l} I_1 = I_c - I_3 \\ I_2 = I_c - I_4 \end{array}\right\} \quad \ldots \quad (20)$$

in Equ. (17), (18) and (19), we obtain,

$$E_c = I_c(R_1 + R_2) - I_3R_1 - I_4R_2 \quad \ldots \quad (21)$$

$$E_s = I_c R_2 + I_3 j\omega M - I_4(R_2 + j\omega L_4) \quad \ldots \quad (22)$$

$$\text{and } E_s = -I_cR_1 + I_3(R_1 + j\omega L_3) - I_4 j\omega M \quad \ldots \quad (23)$$

From Equ. (21),

$$I_3 = -\frac{E_c}{R_1} + I_c\frac{(R_1 + R_2)}{R_1} - I_4\frac{R_2}{R_1} \quad \ldots \quad (24)$$

which, if substituted in (22) and (23) gives,

$$E_s R_1 + E_c j\omega M = I_c[R_1R_2 + (R_1 + R_2) j\omega M] - I_4[R_1R_2 + j\omega(R_2L_3 + R_1M)] \quad \ldots \quad (25)$$

and

$$E_sR_1 + E_c[R_1 + j\omega L_3] = I_c[R_1R_2 + (R_1 + R_2) j\omega L_3] - I_4[R_1R_2 + j\omega(R_2L_3 + R_1M)] \quad \ldots \quad (26)$$

It remains to eliminate I_4 between (25) and (26). This may be done by multiplying the former by $[R_1R_2 + j\omega(R_2L_3 + R_1M)]$ and the latter by $[R_1R_2 + j\omega(R_1L_4 + R_2M)]$. Subtracting the resulting equations we find

$$E_s j\omega[R_1(L_4-M) - R_2(L_3-M)] + E_c[R_1R_2 - \omega^2(L_3L_4-M^2) + j\omega(R_1L_4 + R_2L_3)] = I_c[R_1R_2 j\omega(L_3+L_4-2M) - \omega^2(R_1+R_2)(L_3L_4-M^2)] \quad (27)$$

whence,

$$I_c = E_s\frac{j\omega[R_1(L_4-M) - R_2(L_3-M)]}{R_1R_2 j\omega(L_3+L_4-2M) - \omega^2(R_1+R_2)(L_3L_4-M^2)} + E_c\frac{[R_1R_2 - \omega^2(L_3L_4-M^2) + j\omega(R_2L_3+R_1L_4)]}{R_1R_2 j\omega(L_3+L_4-2M) - \omega^2(R_1+R_2)(L_3L_4-M^2)} \quad \ldots \quad (28)$$

APPENDIX 2

We may obtain an expression for I_s in a similar fashion. The following alternative method is, however, more attractive. From Fig. 2 it is evident that

$$I_s = I_3 - I_4$$

Substituting for I_3 from Equ. (24) we have,

$$I_s R_1 = -E_c + I_c(R_1 + R_2) - I_4(R_1 + R_2) \quad \ldots \quad (29)$$

If we now multiply (29) by $[R_1R_2 + j\omega(R_1L_4 + R_2M)]$ and (25) by $(R_1 + R_2)$ then on subtracting the resulting equations we find,

$$E_s(R_1+R_2) - E_c[R_2 + j\omega(L_4-M)] = -I_c(R_1+R_2) j\omega(L_4-M) + I_s[R_1R_2 + j\omega(R_1L_4 + R_2M)] \quad \ldots \quad (30)$$

Similarly, on multiplying (29) by $[R_1R_2 + j\omega(R_2L_3 + R_1M)]$ and (26) by $(R_1 + R_2)$ we obtain on subtracting,

$$E_s(R_1+R_2) + E_c[R_1 + j\omega(L_3-M] = I_c(R_1+R_2) j\omega(L_3-M) + I_3[R_1R_2 + j\omega(R_2L_3 + R_1M)] \quad \ldots \quad (31)$$

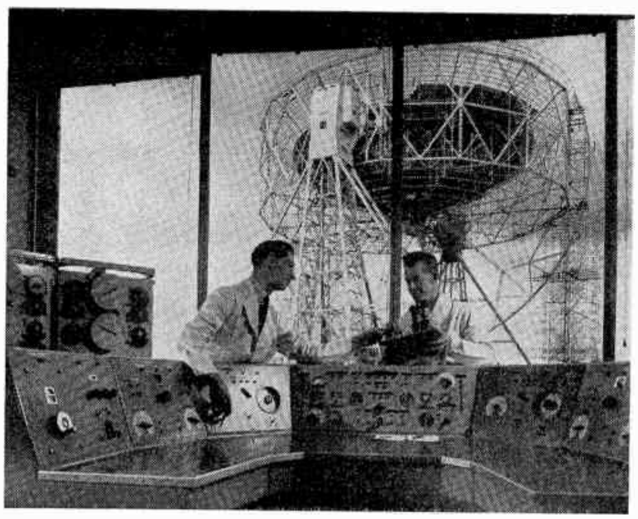
To eliminate I_c from (30) and (31) we multiply the former by $(L_3 - M)$ and the latter by $(L_4 - M)$ and add the resulting equations. This gives

$$E_s(R_1+R_2)[L_3+L_4-2M] + E_c[R_1(L_4-M) - R_2(L_3-M)] = I_s\{R_1R_2(L_3+L_4-2M) + j\omega[(R_1+R_2)(L_3L_4-M^2]\} \quad (32)$$

The current I_s is therefore given by,

$$I_s = \frac{E_s j\omega[(R_1+R_2)(L_3+L_4-2M)]}{R_1R_2 j\omega(L_3+L_4-2M) - \omega^2(R_1+R_2)(L_3L_4-M^2)} + \frac{E_c j\omega[R_1(L_4-M) - R_2(L_3-M)]}{R_1R_2 j\omega(L_3+L_4-2M) - \omega^2(R_1+R_2)(L_3L_4-M^2)} \quad (33)$$

RADIO TELESCOPE

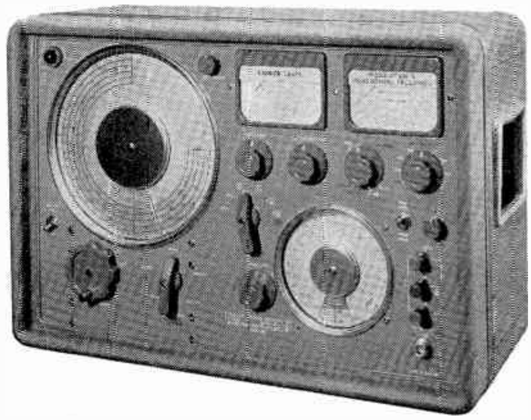


The control desk of the Jodrell Bank radio telescope with the telescope itself visible through the window. This structure is 360 ft in diameter and weighs 2,000 tons. The reflector weighs 750 tons

Institution of Electronics Exhibition

Held in Manchester from 11th to 20th July, this exhibition comprised both research and commercial sections. In addition, a convention was held in conjunction with it.

Among the research exhibits were some examples of special-purpose equipment developed by the U.K.A.E.A. for use in connection with nuclear power reactors where, for various reasons, the conventional techniques are inapplicable. Fig. 1 shows the essentials of a device for indicating the level of liquid metals, strong acids or other highly-conducting liquids. It is based on the fact that a section of shorted coaxial line behaves like an inductance below the quarter-wave resonant frequency. A section of hollow line is immersed in the liquid so that, as the liquid level changes, so does the inductance. The latter forms part of a series-tuned circuit, the other reactance being the fixed capacitor C. A small r.f. voltage is applied to the circuit by way of a large shunt capacitor offering a very low reactance. Resonance is detected by rectifying the voltage which appears across C and displaying the resulting current on a microammeter. Analysis shows that, if the frequency is varied by means of capacitor C_0 in the oscillator tank circuit, then at resonance the length of the cable not immersed (x) is proportional to C_0, which can therefore be calibrated directly in units of length. In practice, most of the r.f. energy is dissipated in a cable-terminating resistor. The cable is therefore well matched, so that the level-detecting device may be remote from the oscillator. The accuracy of measurement is $\pm 1\%$.



*Marconi Instruments f.m./a.m. signal generator TF*1066/1

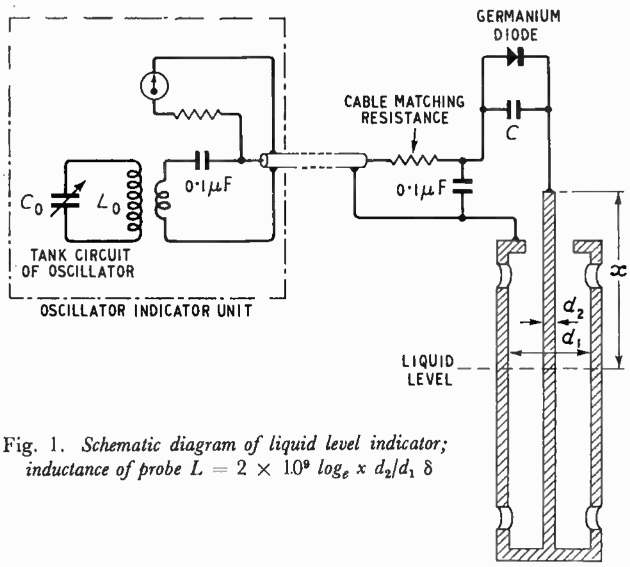


Fig. 1. *Schematic diagram of liquid level indicator; inductance of probe* $L = 2 \times 10^9 \log_e x\ d_2/d_1\ \delta$

An ion-current recorder of good accuracy ($\pm 2\%$) is given an accurate logarithmic characteristic by a novel method. The device employs the usual d.c. balancing arrangement whereby a servo loop is arranged to apply a voltage which nearly cancels the input voltage. This nulling voltage is accurately measurable, since it can be a known fraction of a large voltage. In the new equipment (Fig. 2), a servomechanism drives a slide-wire which forms part of a ladder attenuator. The attenuator is supplied at one end from a reference battery, the d.c. output at the slider forming the nulling voltage which opposes the input signal. The difference signal is converted to a.c. by a vibrating reed, and this a.c. output, suitably amplified, drives the servo motor. The pen of the recorder is coupled to the slider of the attenuator. The constants of the attenuator can easily be arranged so that the resulting record is logarithmic, a 3-decade scale being achieved in practice. The attenuator is also employed to control the level of signal applied to the main amplifier, so that the total gain round the feedback loop is constant. An additional advantage of the arrangement is that there are no rectifier plus smoothing circuits in the a.c. path, the a.c. output being applied directly to the servo motor. This reduces the lag in the servo loop.

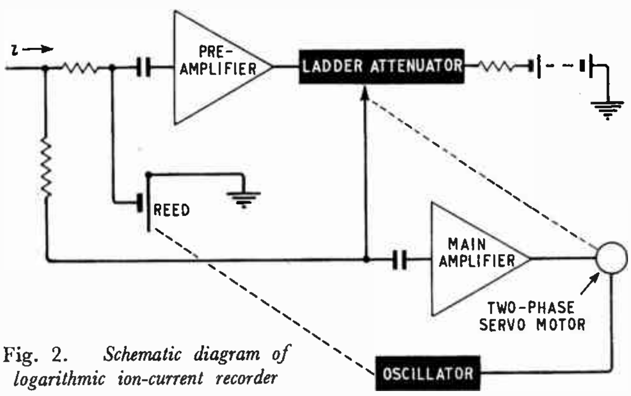


Fig. 2. *Schematic diagram of logarithmic ion-current recorder*

Another U.K.A.E.A. exhibit was a special reactor control-rod position indicator for use in sealed reactors. In this application, the measuring system must operate in a confined space at temperatures up to 300°C, and is inaccessible when the reactor is in service.

The control-rod is coupled mechanically to a ferromagnetic plunger which moves in a non-magnetic tube. A static coil system external to the tube enables changes in position to be detected. Coarse and fine measuring systems are incorporated. The plunger consists of alternate collars of magnetic and non-magnetic material. The coarse measuring system (Fig. 3) makes use of a type of differential transformer. The plunger is always in coil C_2, which is so proportioned that the amount of iron in its field is effectively constant; the resulting output voltage is used as a reference. The extent to which the plunger acts as a core to C_1 varies as the control rod moves. The ratio of the voltages in C_2 and C_1 is therefore an indication of rod position. In the fine measuring system (Fig. 4) three secondary windings are used. The voltage induced in each varies sinusoidally as an iron collar moves through it. The spacing of the windings is such that these three sinusoidal variations are 120° apart. A synchro and servo system gives an angular displacement which varies linearly with the control-rod position.

The application of semi-conductors to automatic train control was shown in an experimental system developed by British Railways Research Department. The purpose of the train-control system is to cause a train's brakes to be applied if it over-runs a 'caution' signal. The means of communicating information to the train about the state of a signal takes the form of two magnets laid between the rails. One is a permanent magnet and the other an electromagnet, which is linked with the signal and energized only when the latter is at 'clear'. (This arrangement 'fails safe'.) As a train passes the first magnet, an alarm and braking system is set, but does not operate immediately. If the second magnet is energized (line clear), the system is unset before it has time to operate. If not (signal at 'caution') the brakes are put on and the alarm sounded.

The proposed applications of semiconductors to the system take the form of a Hall effect probe for detecting the track magnets and transistor amplifiers for performing the other functions. The Hall effect device consists of a thin slab of indium antimonide. If a current is applied as shown in Fig. 5, the application of a magnetic field at right angles to the direction of current flow causes an output current to flow in a direction perpendicular to those of both the input current and the field. The device is quite sensitive and can detect the track magnets at a distance of several inches. Both d.c. and 3,000-c/s a.c. inputs are applied to the Hall effect device. The resultant d.c. output is used for switching transistor amplifiers on or off, while the a.c. output is convenient for amplification to the required power levels.

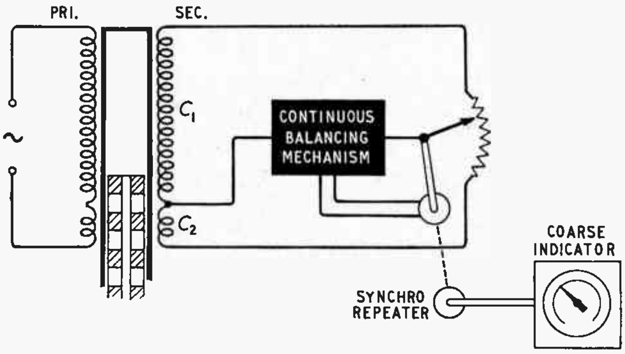


Fig. 3. *Coarse position-measuring system*

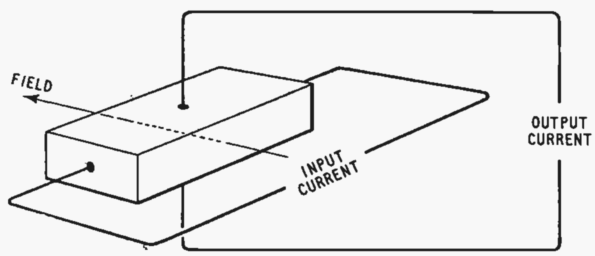


Fig. 5. *Hall effect device*

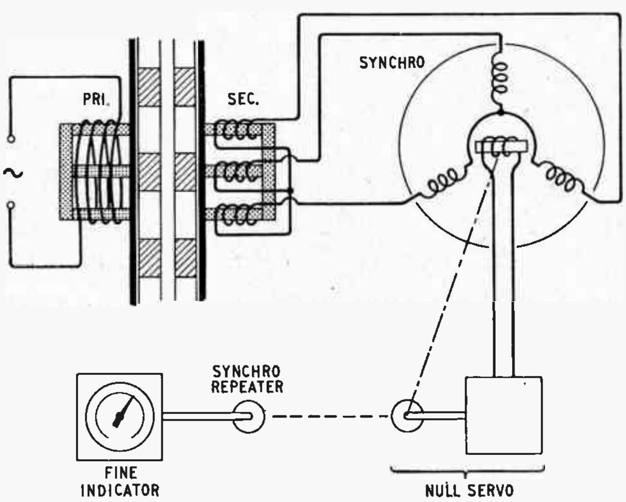
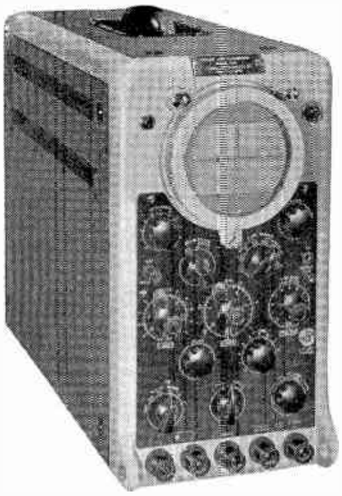
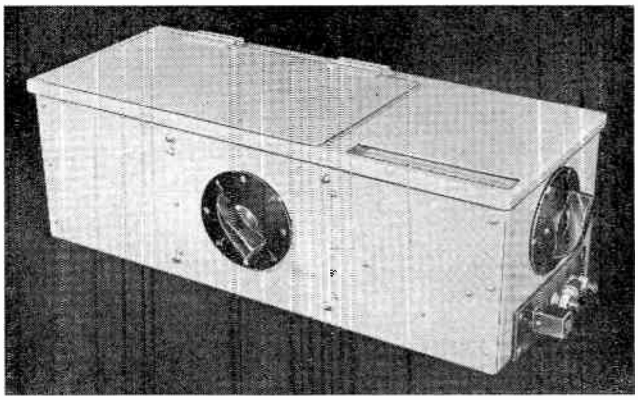


Fig. 4. *Fine position-measuring system*

British Railways also showed a number of devices for measuring the stresses and strains to which railway vehicles are subject. These included load cells used in determining the correct depth of ballast on the track, and buffer testing equipment. A compact six-decade dekatron counter, complete with power supply, occupies only $1\frac{3}{4}$ in. of panel space. Ten six-decade counters can thus be stacked on a standard 19-in. panel, and the counts recorded photographically.

Most of the commercial exhibits were of standard instruments. Avo showed some new coil-winding machines of large capacity. Vibrators with large power-handling capabilities (up to 2 kW) were shown by Goodmans. J. Langham Thompson had a Courtney-Pratt high-speed image-dissector camera embodying a Nipkow disc, and capable of picture rates up to 150,000 per second. A television receiver for schools broadcasts

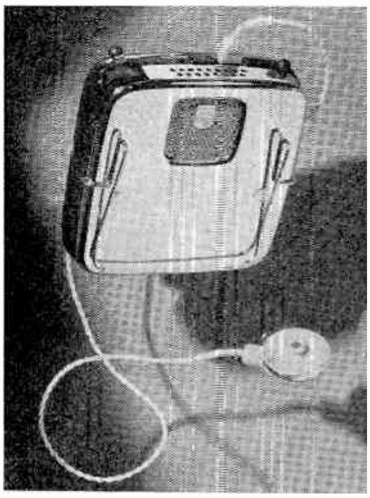




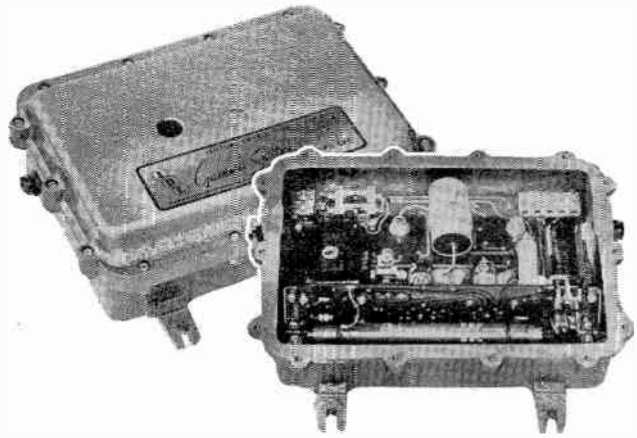
Cossor wide-band oscilloscope. The Y-amplifier response is flat from d.c. to 14 Mc/s

Dawe Instruments transistor oscillator

The Harrison colorimeter



Philips transistor hearing aid with adjustable li niter and telephone pickup coil



The Isotope Developments gamma switch employs a source of gamma-rays and a Geiger counter to detect objects, liquid levels, etc. The Geiger tube is mounted inside the equipment housing

(High-Definition Television) had a 27-in. aluminized tube. Transistors were used in some of the instruments; for instance, Marconi Instruments had a portable transmitter-receiver output test set, containing a transistor amplifier, and Dawe Instruments showed a bridge oscillator.

In a more conventional transistor application—hearing aids—Philips showed some very compact models. One of these is very versatile, in that it can be operated from one 1·5-V penlite cell, or two mercury cells (2·6 V) or three mercury cells (3·9 V), the amplification and output increasing with the battery voltage. The current consumption is 4–11 mA. A limiter with a variable threshold is included, as well as a 'listening coil' for picking up signals inductively from the earpiece of a telephone. Four transistors are employed in an *RC*-coupled circuit.

An interesting industrial device, the Hivolt 'Flockostat', makes use of high-voltage static charges to apply fibres to adhesive surfaces. The fibres and surface are given opposite charges. Under these circumstances the fibres are oriented vertically by the attractive force and, when the charge is increased, precipitated on to the adhesive in such a way as to form a 'pile'.

The Harrison colorimeter is a simple instrument for use in the textile, plastics and paper industries. It is claimed that the accuracy of hue and brightness measurement is equal to that of a trained human observer. A number of colour filters are used in conjunction with a standard light source and a large-area barrier-layer photocell. Measurements are taken by a null method with the aid of a mirror galvanometer.

A number of nucleonic instruments were shown by Isotope Developments, including scintillation counter heads, e.h.t. units, ratemeters and a simple 'radioactivity meter' (Geiger counter) for schools use. In the 'package monitor' gamma rays are used to detect unfilled containers on a conveyor belt and operate the necessary ejection mechanisms. A similar unit, the 'gamma switch', is available for industrial applications, where it is the nucleonic counterpart of the much used lamp and photocell type of detector. Since gamma rays are capable of penetrating metals and other dense substances, the device forms a ready means of detecting the level of liquids inside storage tanks, raw materials in hoppers, and so forth.

THERMODYNAMICS THROUGH THE LOOKING-GLASS

The expression 'negative temperature' is occurring more and more often in the literature and it is apt to puzzle the engineer as its most obvious meaning, a temperature below absolute zero, is unimaginable. The expression, which seems first to have been used by E. M. Purcell and R. V. Pound as long ago as 1951 and is now regularly applied to certain state- and spin-distributions, has recently been examined in more general terms in an important article by N. F. Ramsey ("Thermodynamics and Statistical Mechanics at Negative Absolute Temperatures", *Phys. Rev.*, Vol. 103, p. 1, July 1956). It seems a bit much, having just about completed the list of anti-particles, to top it up with anti-temperature as well for good measure. But this is not a new 'discovery', nor yet a mere name for the unusual state of things to which it applies; it is a means of extending the ideas of ordinary thermodynamics into a field which could never have been suspected when they were formulated.

Temperature as a Statistical Concept

In this context, temperature is used to denote something quite different from the ordinarily understood idea presented to us by a thermometer, and the basic concept of temperature is really that of a statistical factor. It is a coefficient which turns up almost as a pleasant surprise in the Maxwell-Boltzmann distribution of energy among the members of a system in equilibrium. Indeed, the climax of the usual derivation of this, which opens like Haydn's *Creation* with a Representation of Chaos (symbolized by a quantity H) and leads on to an exponential formula with a lower-case h (which isn't Planck's constant) in the index, comes when this h is suddenly identified with $1/2kT$. This distribution formula, which need not be given in full here, leads to the constantly-used result for the ratio between the numbers N_2 and N_1 of particles in a system in equilibrium at $T°$ Kelvin which possess energies W_2 and W_1. It is $N_2/N_1 = e^{-(W_2-W_1)/kT}$. Here the symbol 1 denotes the lower, and 2 the higher energy state; that is, $W_2 > W_1$. As to algebraic signs, N_1 and N_2 are necessarily positive numbers; W_1 and W_2 are always positive (but not *necessarily*, for negative energy states can occur in, for example, the Dirac theory of the electron); and T is in every ordinary case positive (but not *necessarily*, because a negative sign is simply extraordinarily unusual). If T is positive, than $N_2 < N_1$. An equilibrium distribution in which $N_2 > N_1$ is a rare achievement, a sort of mirror-image of the common state of affairs; and this is what is represented by a negative value of T. Fig. 1 shows N_2/N_1 plotted against T; and Fig. 2 the complete plot of T against N_2/N_1 for all imaginable values of N_2/N_1.

The Maxwell-Boltzmann calculation, though based on the idea of perfect-gas molecules interchanging energy by collisions, ends up only with energy and numbers, and is valid for any kind of process, provided (a) that the detailed balancing by which as many enter a given state as leave it during any interval persists, and (b) that classical considerations hold. In quantum-mechanics, the distribution formula is modified by a factor representing the number of possible wave-functions for each state; and if Fermi-Dirac statistics apply, the term $e^{-(W_2-W_1)/kT}$ is replaced by $\left(1 + e^{(W_2-W_1)/kT}\right)^{-1}$. Neither of these modifications, even if they happen to be relevant, affects the present discussion. For it appears that the ratio N_2/N_1 in any case works out to the Boltzmann exponential.

What Negative Temperature is Not

It seems worth while at this stage to clear up some points which might confuse the literally-minded person who is suspicious of statistics.

First, since the microwave experimental work in which the term originated is done at liquid helium temperatures, it would be easy for the idea to get abroad that some temperature below absolute zero had been reached by simple cooling, or that such a process could be regarded as conceivable. The statistical T becomes negative by passing through *infinity*, not by overshooting absolute zero. The temperature $-1°$K is infinitely distant from $+1°$K; and, in fact, $-10{,}000°$K is closer to $+1°$K than $-1°$K is. If this has aroused, rather than allayed, suspicions, so much the better; the title warned you that we were going into the looking-glass world.

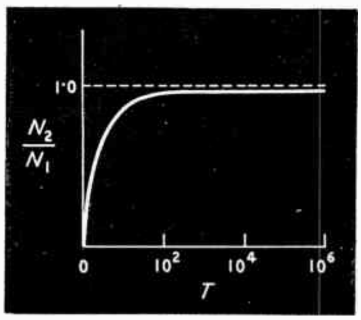


Fig. 1. *The graph of N_2/N_1 plotted against T for the Boltzmann equilibrium distribution. This is the only possible plot if T is the independent variable, changed at will. Here $(W_2 - W_1)/k$ has been taken as 1, a reasonable microwave value*

Next, as the degree K is already used in microwave work as an energy unit of single-quantum or microscopic status, and microwave values happen to be a little less than one degree as a rule, it may be well to emphasize that this has nothing at all to do with the matter. Indeed, it is a distortion of the proper idea of temperature, assigning to individual quanta what is more properly an attribute of averages. The relations used are:

$$kT = \tfrac{1}{2}mv^2 = (W_2 - W_1) = h\nu = eV,$$

where k is Boltzmann's constant, h this time really is Planck's constant, e is the electronic charge, and the other symbols have their usual meanings. In this way T°K can represent the kinetic energy of a particle, a difference in energy levels, or a quantum of radiation, as an alternative to using ergs, wave-numbers, or electron-volts. One degree K corresponds to about $1 \cdot 38 \times 10^{-13}$ erg; one eV is about $1 \cdot 6 \times 10^{-12}$ erg; so one eV can be expressed as 11,600°K, and so on. This is purely a numerical matter, and needs no interpretation. People who use this system are not suspicious of statistics, but merely callous towards the subject.

Finally, there is a relevant case of the statistical use of the term temperature, which comes grammatically into this paragraph by rights, since it illustrates what *positive* temperature *is*. In nuclear physics, the state of an excited nucleus is determined by observing the numbers N_i of neutrons it emits with energy W_i; from N_1 and N_2 with energies W_1 and W_2, using the Boltzmann formula $N_2/N_1 = e^{-(W_2 - W_1)/kT}$, a value T for the temperature of the nucleus is calculated. If this is about 10^{10} °K, or 10^3 eV, say, it doesn't mean that it is excessively thermonuclear; the figure describes the energy *distribution* between its constituent parts. A negative value for T in such a case is not inconceivable; but it just doesn't happen to happen.

Temperature and Thermodynamics

We ought now to try to see why ordinary distributions never demand a negative T. Referring to J. K. Roberts' "Heat and Thermodynamics", he says that equality of temperature between two communicating systems happens when no energy is being transferred from one to the other. If energy is being transferred, then there is a temperature difference. He calls this the "zeroth law of thermodynamics"; as a definition, it simply says that there is something, other than energy itself, which determines whether energy will flow or not, and giving that something a name.

The second law of thermodynamics states how temperature does its job if left to itself. When, unassisted by externally supplied energy, a quantity of energy dQ travels from a system at temperature T_1 to another at temperature T_2, then the expression $\left(\dfrac{dQ}{T_2} - \dfrac{dQ}{T_1}\right)$, which is called the change in entropy, must always be positive. That is, the naturally occurring process is for energy to flow from a higher temperature to a lower temperature, and to stop flowing when the temperatures are equal. But why does this happen, and what does this named but vaguely defined something have to do with it really?

Maxwell-Boltzmann statistics give the answer. When temperature has been equalized among the different parts of a system, then the most chaotic, and hence the most probable, distribution of energy amongst the individual particles is attained. Temperature is the factor which determines how the total energy available is parcelled out, if pure chaos reigns. For every temperature there is a definite distribution, and to every distribution there corresponds a definite temperature. The zeroth law of thermodynamics states that there is a one-one correspondence between distribution and temperature, while the second law as it is often expressed shows that the distribution-probability tends always to increase. The reason why this always happens in the ordinary way is that, with what might be called second-order chaos, there are huge numbers of particles and huge numbers of energy states, and anything else would be extremely *im*probable!

The point is, that the mirror-image trick can be done with a pair of Ns, or possibly several pairs of Ns taken separately, but to rearrange huge numbers of

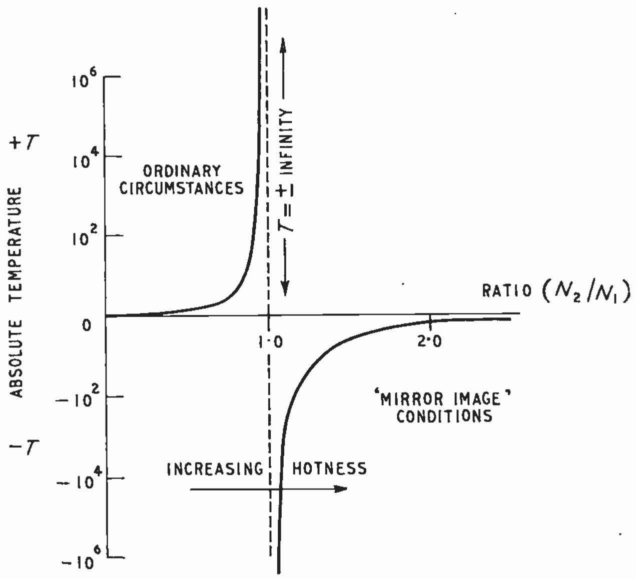


Fig. 2. *The graph of T against N_2/N_1. Both branches of the curve are possible if N_2/N_1 is regarded as the independent variable, which we can imagine to take any value we like*

Ns and Ws to simulate a complete Boltzmann distribution inverted is a statistically inconceivable process, and to preserve it in equilibrium for a finite time is a feat even further beyond the grasp of the imagination. With a two-state system, however, it is relatively simple both to invert the numbers, and to secure a transient equilibrium.

Two-State Systems

The next steps in the argument contain a number of assumptions. The first is that if we consider huge numbers of particles but only a *very few states*, perhaps only two states, then the relative distribution of particles between the states will be just the same as it would have been if there had been huge numbers of other states about. That is to say that, although this time $(N_1 + N_2)$ must add up to the total number N of particles, we *still* have $\dfrac{N_2}{N_1} = e^{\frac{-(W_2 - W_1)}{kT}}$, with the additional result, since $(N_1 + N_2) = N$, that $\dfrac{(N_1 - N_2)}{N} = \dfrac{(W_2 - W_1)}{2kT}$, approximately.

One feels intuitively that this is something like trying to apply the laws of *black-body* radiation to calculate relative intensities of the lines in a *line-spectrum*, and that it does need some justification! Here intuition has been over-apprehensive, for the procedure is quite correct in fact, and justified by quantum mechanics.

The second assumption is that if we start altering the relative numbers in the two states by some external process, and so change the temperature, we can preserve equilibrium at every stage in the process. (No equilibrium, no temperature!) This can be done if the process is *adiabatic*; that is, isolated thermally from the surroundings and proceeding slowly via a series of successive transient equilibria.

The third is that we can actually get a simple two-state (or few-state) system, and that we can disarrange its population distribution adiabatically. Such systems, and also the means of redistribution, are found in the (3, 3) ammonia resonance, for example, and also in the very limited range of spin-levels used in nuclear and paramagnetic resonance, the latter only being of interest for microwave purposes. And transitions between two chosen electron-spin levels occur with the emission or absorption of microwave radiation at the resonant frequency. Adiabatic conditions can be secured by cunning negotiation with or moderation of the spin-spin and spin-lattice relaxation times, but this is too long a story to go into now.

Supposing then that the electron-spins have *all* exchanged roles, so that there are N_1 in state W_2, and N_2 in state W_1.

Then, $\frac{N_2}{N_1} = e^{\frac{-(W_2-W_1)}{-kT}}$, and $\frac{N_1-N_2}{N} = \frac{(W_2-W_1)}{-2kT}$.

This inversion of the original distribution has simply reversed the sign of T!

The difference (N_1-N_2) is inversely proportional to T, so that if the difference is to be employed, then the numerical value of T, whether positive or negative, is best made as small as possible. At negative temperatures, however, the higher energy state is the more densely populated; so the negative temperature represents *a reservoir of usable energy*. Negative temperatures are thus 'hotter' than any normal positive temperature, and the smaller the size of T, the hotter they are. Unfortunately, it is impossible to obtain useful work by making energy travel from a system for which T is negative, however hot it is, to a system for which T is positive, however cold. Infinity intervenes!

Applications

It would be better if I discussed the applications of these ideas in a later article. The mathematics of "population inversion" and the method of "adiabatic fast passage" are described in the article by J. P. Wittke which I quoted last month. A theoretical paper by H. Motz, on "Negative Temperature Reservoir Amplifiers", considering doped silicon, appeared in the *Journal of Electronics* for May 1957, and an operating solid-state maser using paramagnetic salts, proposed by N. Bloembergen in 1956, has been described by H. E. D. Scovill, G. Feher, and H. Seidel (*Phys. Rev.*, 1957, Vol. 105, p. 762). I feel that it might be more useful to sort out the underlying principles than merely to quote the work; but I am not so sure that this is going to be easy.

N. F. Ramsey's Article

What I want to do now is to get back to the temperatures. It might appear that negative temperature is little more than a convenient way of describing a spin system in an unusual excited equilibrium condition. To some extent this may be so, but it opens a vast extrapolation of ordinary thermodynamics which may conceivably be applicable to the mirror-image world of anti-particles and anti-matter discussed in a recent *Times Science Review* by Prof. O. W. Frisch, as well as to the inverted distributions of ordinary states.

In his article, Ramsey goes into the matter very thoroughly, both as regards the assumptions made in describing this kind of system, and the conclusions reached about its properties. In the first place, supposing that we isolate a two-state system, and alter N_1 and N_2 adiabatically, establishing equilibrium at each successive stage, then the first thing that happens when N_2 is slightly increased is that T is *raised*; as N_2 approaches N_1, then T gets larger and larger, becoming infinite when $N_2 = N_1$. If N_2 is only slightly larger than N_1, then T has a large negative value, while if N_2 is much greater than N_1, then T is numerically small and negative. In particular, the complete inversion of N_1 to N_2, which just reverses the sign of T, takes place *via infinity*, not via absolute zero.

The second and third laws of thermodynamics are both obeyed. For the third law, the unattainability of absolute zero persists, from whichever side it is approached. For the second law, heat can be extracted from a negative-temperature reservoir to perform work, violating one of the usual formulations of the law; but the more general statement of the second law, due to Carathéodory, is not contravened.

There are a number of inversions of ordinary effects. Most resistances are negative at negative temperatures, so that a negative-temperature resistance amplifies instead of attenuating. For nuclear spin systems at negative temperatures, adiabatic demagnetization *raises* the temperature instead of cooling it. The realm of negative temperatures is shown to have its mirror-image counterparts of all ordinary thermodynamic observations, and Ramsey states that, though cases of this kind are necessarily infrequent, they should be considered in any really general discussion. I think I have probably done what I set out to do, in explaining what the term "negative temperature" means, though perhaps rather fuzzily. If you really want to go into things properly, I recommend you to Ramsey's article.

INTERNATIONAL CONFERENCE ON SOLID STATE PHYSICS

The Postal and Telecommunications Group of the Brussels Universal and International Exhibition is to hold an international conference on solid state phenomena in electronics and telecommunications at the University of Brussels from 2nd to 7th of June 1958.

Three major subjects will be treated at this conference: semiconductors, magnetic materials and photo-sensitive and luminescent materials. The complete proceedings of the conference will be published in a special volume. The official languages of the conference will be English, French and German.

Those interested in attending the conference should write to the Hon. General Secretary, Société Belge de Physique, Loverval, Belgium.

Artificial Transmission Lines

COIL DESIGN FOR THE NEGATIVE MUTUAL INDUCTANCE TYPE

By A. C. Hudson*

SUMMARY. *Some relations are derived between the dimensions of the coils for negative mutual inductance transmission lines and the constants of such lines.*

Artificial lines of the negative mutual inductance type were first described by Johnson and Shea[1] and are now frequently used,[2] for example, as the anode and grid lines of distributed-line amplifiers. An investigation has been made of the relations between the geometrical parameters of the coil and the constants of the line, and some equations derived, notably (19a) and (20a), which may be used for design.

The negative mutual inductance line comprises cascaded m-derived low-pass filter sections, with m typically between 1 and 1·5. A value of m greater than unity requires that the shunt inductance [see Fig. 1 (a)] be negative; in practice, this is achieved in effect by constructing the circuit of Fig. 1(c), where series-aiding mutual inductance exists between the two halves of L_T.

The required mutual inductance M may be obtained by employing either two adjacent solenoids separated by a suitable distance, or contiguous solenoids, as in Fig. 1(c), with a length-to-diameter ratio chosen to give the correct coefficient of coupling. The latter alternative would appear to be better from considerations of undesirable shunt capacitance across the coil. The argument for this is outlined briefly here. Section 11.2 of Reference 3 shows that the stray capacitance of a single-layer solenoid is directly proportional to coil diameter and much less strongly dependent on length. Thus it is desirable to use as small a diameter as possible. Since the inductance is fixed, this requires as large a pitch n as possible. The maximum pitch will be limited either by heat dissipation (for example, in the anode line of a distributed-line amplifier the coils must carry the anode current of all valves); or by mechanical weakness of the wire; and this maximum will apply equally to the spaced or contiguous design. Thus, in a typical design problem we have determined pitch, inductance, and coefficient of coupling. But under these conditions it is easily seen that the contiguous design will always have a smaller diameter than the spaced design, and hence will have smaller stray capacitance. The contiguous case is the one considered here, the coil being indicated in Fig. 1(c). It is assumed that coupling between successive full sections is negligible.

It is also possible to construct the coil assembly of a negative mutual inductance transmission line as one continuous tapped solenoid. This method is not considered here. Among the disadvantages of this method are the difficulty in adjustment of individual coils, and the fact that, for the case of a distributed-line amplifier, the use of a continuous solenoid arbitrarily equates coil length to the realizable spacing between valves, and hence will not, in general, provide an optimum design from the point of view of stray capacitance.

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Relations Between the Coil Parameters

Any one of the four parameters l/D, K, m and L_1/L_T will determine the others, and hence twelve equations may be derived by expressing each explicitly in terms of one of the others. The relations which are needed to derive these twelve equations are:

(i) The Esnault-Pelterie formula [see Section 10.1 (iii) of Reference 3]:

$$L = 0{\cdot}1008\frac{a^2N^2}{l + 0{\cdot}92a}\text{ (microhenrys)} \quad .. \quad (1)$$

which gives an approximation to the inductance of a single-layer solenoid. (This relation is applied in turn to the full coil L_T and to the half coil L_1.)

(ii) The condition that the networks of Figs. 1(a) and (b) must appear identical to the external circuit.

(iii) The exact relations $K = M/L_1$ and $L_T = L_1 + L_1 + 2M$.

The twelve equations are arranged in a manner indicating their derivation in Fig. 2. They are also collected in Table 1.

Coil Design

Fig. 3 is a chart showing relations which exist between various design parameters.

L_0, C_0, R_0 and ω_c are related by the usual filter theory relations which are summarized in Table 2.

In Fig. 3 each node represents an equation and a quantity is determined when it connects to a node for which the other quantities joined to that node are already determined. For example, if C_s and R_0 are given, then obviously $R_0\sqrt{C_s}$ is fixed. Now if m is also chosen, then the chart shows that B is determined.

The derivations of the important relations implied in Fig. 3 are given below.

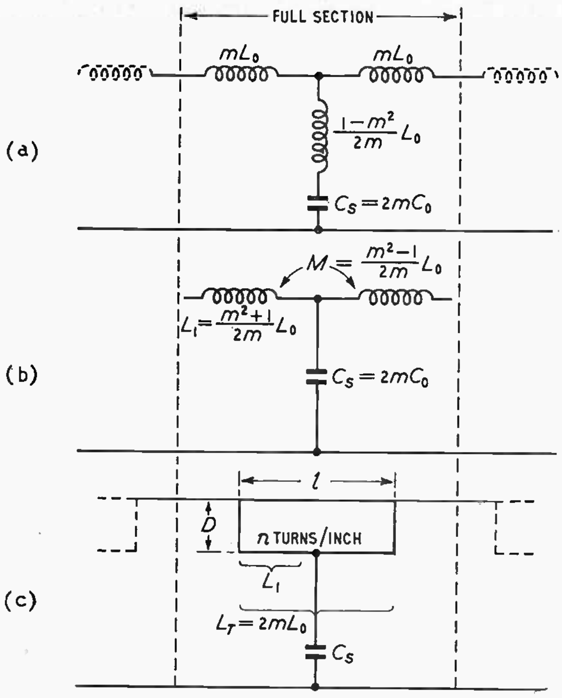

Fig. 1. *Three equivalent representations of the coil; (a) parameters as given by filter theory, (b) exact equivalent, (c) coil geometry*

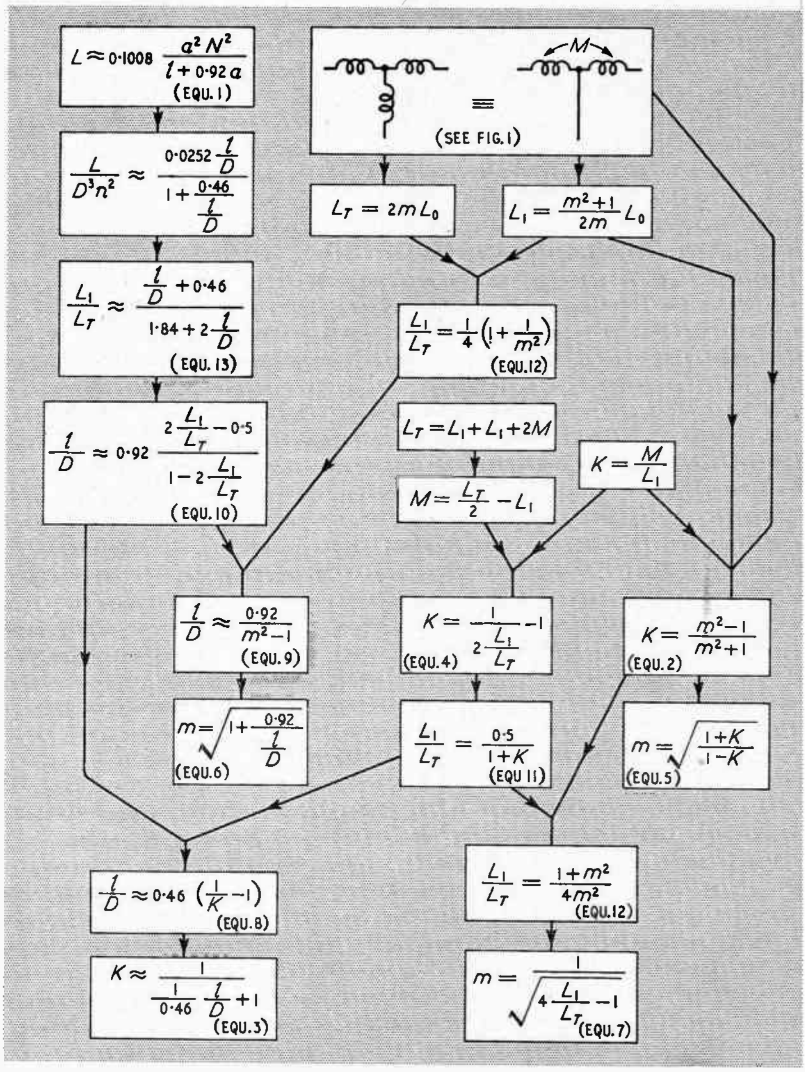

Fig. 2. *Derivation of the twelve coil equations*

The most useful relations implied in the chart of Fig. 3 are Equs. (19) and (20) relating $R_0\sqrt{C_s}$ and $\omega_c\sqrt{C_s}$ to m and the coil-form parameters; they are derived as follows.

TABLE I

	Equ. (2) $K = \dfrac{m^2-1}{m^2+1}$	Equ. (3) $K \approx \dfrac{1}{1+\dfrac{1}{0{\cdot}46}\dfrac{l}{D}}$	Equ. (4) $K = \dfrac{1}{2L_1/L_T} - 1$
Equ. (5) $m = \sqrt{\dfrac{1+K}{1-K}}$		Equ. (6) $m \approx \sqrt{\dfrac{0{\cdot}92}{l/D}+1}$	(Equ. (7) $m = 1\Big/\sqrt{4\dfrac{L_1}{L_T}-1}$
Equ. (8) $\dfrac{l}{D} \approx 0{\cdot}46\left(\dfrac{1}{K}-1\right)$	Equ. (9) $\dfrac{l}{D} \approx \dfrac{0{\cdot}92}{m^2-1}$		Equ. (10) $\dfrac{l}{D} \approx 0{\cdot}92\dfrac{2L_1/L_T-0{\cdot}5}{1-2L_1/L_T}$
Equ. (11) $\dfrac{L_1}{L_T} = \dfrac{0{\cdot}5}{1+K}$	Equ. (12) $\dfrac{L_1}{L_T} = \dfrac{1+m^2}{4m^2}$	Equ. (13) $\dfrac{L_1}{L_T} \approx \dfrac{\dfrac{l}{D}+0{\cdot}46}{1{\cdot}84+2\dfrac{l}{D}}$	

The Esnault-Pelterie relation, Equ. (1), may be arranged as follows:

$$\frac{L}{D^3n^2} = \frac{0{\cdot}0252\,(l/D)}{1+0{\cdot}46D/l}$$

Applying this to the entire coil L_T, with inductance $2mL_0$, and substituting $(0{\cdot}92)/(m^2-1)$ for l/D [see Equ. (9)], gives:

$$L_0 = \frac{2B^2}{m(m^4-1)} \quad \ldots \quad (18)$$

Combining this in turn with Equs. (15) and (14), using also the relation $C_s = 2mC_0$ gives,

$$R_0\sqrt{C_s} = \frac{2B}{\sqrt{m^4-1}} \quad \ldots \quad (19)$$

$$\omega_c\sqrt{C_s} = \frac{m\sqrt{m^4-1}}{B} \quad \ldots \quad (20)$$

Equs. (18) and (19) may be put in a slightly more convenient form by replacing B with its definition, giving,

$$D^3n^2 = 21{\cdot}55R_0^2C_s\,(m^4-1) \quad \ldots \quad (19a)$$

and $$D^3n^2 = \frac{86{\cdot}2m^2\,(m^4-1)}{\omega_c^2C_s} \quad \ldots \quad (20a)$$

TABLE 2

Equ. (14)	Equ. (15)	Equ. (16)	Equ. (17)
	$R_0 = \sqrt{L_0/C_0}$	$R_0 = \dfrac{1}{\omega_c C_0}$	$R_0 = \omega_c L_0$
$\omega_c = \dfrac{1}{\sqrt{L_0 C_0}}$		$\omega_c = \dfrac{1}{R_0 C_0}$	$\omega_c = R_0/L_0$
$L_0 = \dfrac{1}{\omega_c^2 C_0}$	$L_0 = C_0 R_0^2$		$L_0 = R_0/\omega_c$
$C_0 = \dfrac{1}{\omega_c^2 L_0}$	$C_0 = L_0/R_0^2$	$C_0 = \dfrac{1}{\omega_c R_0}$	

The coil design may start from these relations or, more conveniently, from either one of them and the basic m-derived filter relation

$$m = (1/2)R_0 C_s \omega_c \quad .. \quad .. \quad .. \quad .. \quad (21)$$
$$= \frac{R_0}{\text{Reactance of } C_s/2 \text{ at } f = f_c}$$

The specification of a negative mutual inductance transmission line when substituted in these relations will be sufficient to determine B, and hence the coil pitch and diameter may be chosen, since these two variables affect the design only through B.

Briefly, the coil design procedure could be as follows:

(1) State the wanted values of R_0, C_s and ω_c.
(2) Calculate m from Equ. (21).
(3) Decide if m has a practical value (see References 1 and 2); if not, one or more of R_0, C_s or ω_c must be altered.
(4) Find D^3n^2 from Equ. (19a) or Equ. (20a).
(5) Choose n and find D, or vice versa.
(6) Find l from Equ. (9), or N from Equ. (22) of the Appendix.

Conclusion

The chief result is to show how the coil-form parameter D^3n^2 is related to the parameters which are initially given in the design of a negative mutual inductance line.

No attempt is made to discuss fully the design of such lines—reference may be made to References 1, 2 and 4; while the theory of distributed amplification is discussed in Reference 2 by Ginzton et al. and Reference 5 by Percival.

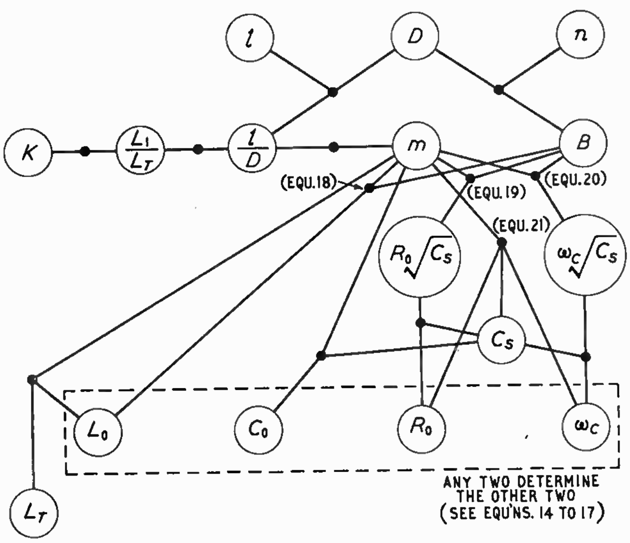

Fig. 3. *Chart of design equations*

Acknowledgment

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LIST OF SYMBOLS

R_0 = Characteristic impedance of transmission line in ohms.
ω_c = $2\pi f_c$.
f_c = Cut-off frequency of artificial line in cycles per second.
C_0 = Half-section shunt capacitance of constant-k artificial line in farads.
L_0 = Half-section inductance of constant-k artificial line in henrys.
M = Mutual inductance between the two halves of a full coil, in henrys.
m = Zobel's parameter in filter theory.
f = Frequency in cycles per second.
l = Length of a full coil, in inches.
D = Diameter of a full coil, in inches.
a = Radius of a full coil, in inches.
N = Number of turns.
n = Pitch, in turns per inch.
K = Coefficient of coupling between two halves of a full coil.
L_1 = Inductance of a half coil.
L_T = Inductance of a full coil.
C_s = Total shunt capacitance of a full m-derived section of artificial line, in farads.
B = $\sqrt{0{\cdot}0116 D^3 n^2}$, a property of the coil form.

NOTE: Microhenrys and microfarads are used in some cases in place of henrys and farads.

APPENDIX

Obtaining a Given Inductance on a Given Form

It is sometimes convenient to establish properties of a coil of given pitch and diameter, either because a coil form is already in existence, or because certain pitches and diameters are more convenient to fabricate, or for reasons of standardization.

The inductance of a single-layer solenoid of length at least one-quarter of its diameter is given by the Esnault-Pelterie relation

$$L = 0{\cdot}1008 \frac{a^2 N^2}{l + 0{\cdot}92a} \text{ microhenrys} \quad .. \quad .. \quad .. \quad (1)$$

[Section 10.1 (iii) of Ref. 3.]

Pitch may be introduced explicitly in this equation by writing N/n for l and the resulting relation solved for N, giving

$$N = L/k_1[1 + \sqrt{1 + (k_2/L)}] \quad .. \quad .. \quad .. \quad (22)$$

where $k_1 = (D^2 n)/(19{\cdot}84)$

and $k_2 = (D^3 n^2)/(21{\cdot}56)$.

Thus if a coil form of given pitch and diameter is to be used frequently, k_1 and k_2 may be calculated once, and Equ. (22) used for coil design. In practical work there is no need to measure the resulting inductance. Wheeler's formula could also be used, but it is slightly less accurate in the regions of interest for distributed-line amplifier design.

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V.H.F. Broadcasting

REDUCTION OF IMPULSIVE INTERFERENCE IN F.M. RECEPTION

By R. D. A. Maurice, Ing.-Dr., Ing.E.S.E., A.M.I.E.E.*

"*Put not yourself into amazement how these things should be: all difficulties are but easy when they are known*".
(*W. Shakespeare, "Measure for Measure", Act IV, Scene 2.*)

Recourse to very-high-frequency sound broadcasting seems to be the only way of ensuring listeners in the United Kingdom of a reasonable freedom from interference, a freedom which has been denied to many on account of the failure of the nations of Europe to evolve and maintain a satisfactory plan of channel allocation in the medium and long-wave bands.

The principal advantage in using the very high frequencies is the absense of long-distance interference. This enables a broadcasting authority to serve a more well-defined area than would be possible on longer wavelengths.

The main disadvantage of the use of v.h.f. is the susceptibility of receivers to man-made interference or impulse noise, and this aspect is discussed here at some length.

Relative freedom from impulsive interference may be bought at a price, the cost being reckoned in bandwidth. Shannon's formula for channel capacity measured in 'bits' per second shows that decibels of signal-to-noise ratio are equivalid with kilocycles per second of bandwidth. Wide-band frequency modulation is a way of increasing the bandwidth required by the signal or the signal spectrum width, and it thus leads to a gain in signal-to-noise ratio over that obtainable with amplitude modulation.

The major part of the article is concerned with showing how this gain in signal-to-noise ratio is achieved.

An example serves to show what field strengths are required to protect a given reception from motor-car-ignition interference.

The f.m./a.m. controversy has at last come to a close, probably due, in this country, to the Government's adoption of the Television Advisory Committee's report which recommended f.m. as the system of modulation to be used for a very-high-frequency sound broadcast service.

The B.B.C. ran two experimental transmissions using a.m. and f.m. respectively from two transmitters and one wide-band aerial at Wrotham, in Kent. A number of receivers were tried and, in particular, forty f.m./a.m. comparison sets[1] were made by a commercial firm and circulated to a large number of members of B.B.C. staff and others. These were designed to a very complete and stringent performance specification[2] and could be used to compare most features of the three modulation systems under investigation at the time. These were first, narrow-band a.m.; secondly, wide-band a.m. with an impulse limiter; and thirdly, wide-band f.m.

If in what follows there seems to be undue stress on f.m. performance rather than on the behaviour of the other two systems, this is because f.m. is now of immediate interest in view of the existence of virtually nation-wide coverage.

The F.M./A.M. Improvement for Random Fluctuation Noise

Normally, an f.m. receiver may be assumed to be quite insensitive to amplitude fluctuations but by means of its discriminator it will be directly subject to phase or frequency fluctuations. The action of a discriminator of the type usually employed in f.m. receivers is to convert changes of frequency directly into changes of amplitude; these are then detected in the usual manner by a conventional envelope detector or demodulator. Now to change a frequency fluctuation into one of amplitude one may use either the reactance of an inductor or the susceptance of a capacitor, the operation being described in the frequency or spectrum or

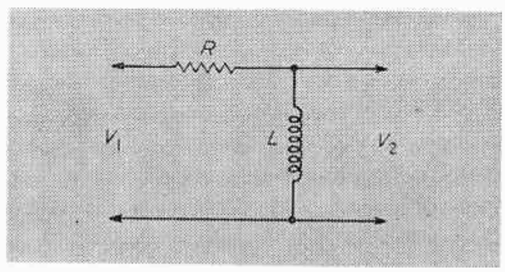

Fig. 1. *An elementary differentiator.* *If* $R/L = \omega_c$, $\left|\dfrac{V_2}{V_1}\right| = \dfrac{\omega/\omega_c}{\sqrt{1 + \omega^2/\omega^2_c}}$.

If $\omega/\omega_c \ll 1$, $\left|\dfrac{V_2}{V_1}\right| = \omega/\omega_c$

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impedance world by a multiplication by $j\omega$ or p. These symbols transformed into the time world represent a differentiation. Thus a discriminator is, over the linear portion of its characteristic, an electrical differentiator and nothing more. An elementary low-pass differentiator is shown in Fig. 1. The coefficient of the differentiation[3] is the time-constant of the circuit $1/\omega_c$.

Now consider two equally sensitive receivers with identical r.f. circuits except that the a.m. receiver has an r.f. bandwidth of $\pm f_a$ (say $\pm$ 10 kc/s) while the f.m. receiver has an r.f. bandwidth of $\pm \Delta f$ (say $\pm$ 75 kc/s). Suppose the densities of the spectra of random fluctuation noise voltage in the two receivers, Fig. 2, be N volts per cycle per second. The a.m. receiver mean noise power would be

$$W_{AM} \propto \frac{1}{2f_a}\int_{-f_a}^{f_a} N^2\,df = N^2 \quad \ldots \quad (1)$$

Now the discriminator of the f.m. receiver will, as already stated, differentiate this noise voltage with a time-constant equal to the reciprocal of the half-bandwidth of the r.f. or discriminator circuits. Thus the noise-voltage spectrum emerging from the discriminator will be the product of the uniform r.f. spectrum in a bandwidth $2\,\Delta f$, N V/c/s, and the ratio $\pm\,\omega/\omega_c$ where $\omega_c/2\pi$ is the half-bandwidth of the entire r.f. portion of the receiver and $\omega/2\pi$ is the departure of frequency from the mid-band value f_0. The f.m. receiver noise spectrum before detection and passage through the audio circuits is thus $N\omega/\omega_c = Nf/\Delta f$ with $-\Delta f \approx f \approx \Delta f$. The mean power of this noise restricted to a bandwidth $\pm f_a$ c/s around f_0 (the audio portion of the spectrum) is

$$W_{FM} \propto \frac{1}{2f_a}\int_{-f_a}^{f_a} (N/\Delta f)^2 f^2 df = N^2 f^2{}_a/(\sqrt{3}\Delta f)^2 \quad (2)$$

The ratio of the a.m. r.m.s. noise to the f.m. r.m.s. noise is

$$\sqrt{(W_{AM}/W_{FM})} = \sqrt{3}\,\Delta f/f_a \quad \ldots \quad (3)$$

This is the well-known f.m./a.m. improvement.[4] If the a.f. band is taken as $7\frac{1}{2}$ kc/s for both a.m. and f.m. and the r.f. band is taken greater than 15 kc/s for a.m. and $\pm$ 75 kc/s for f.m. equation (3) gives an 'improvement' of 25 dB.

It must be remembered, however, that f.m. is not '25 dB better' than double-sideband a.m. The 25-dB improvement is obtained at the expense of bandwidth. Shannon's formula for the information-carrying capacity of a communication channel shows that bandwidth and decibels of signal-to-noise ratio are equivalid ; so that f.m. is not necessarily more efficient than a.m. It has simply exchanged transmitter power for transmission bandwidth.

Pre- and De-Emphasis

The idea of pre-emphasising the higher frequency components of an audio signal at the transmitter to account for a de-emphasis (top-cut), whether intentional or not, at the receiver is not new. It does, however, appear to be most profitable when used in a transmission system which, for one reason or another, has a triangular noise spectrum at the receiver. It cannot be used, however, unless the wanted signal has a spectrum

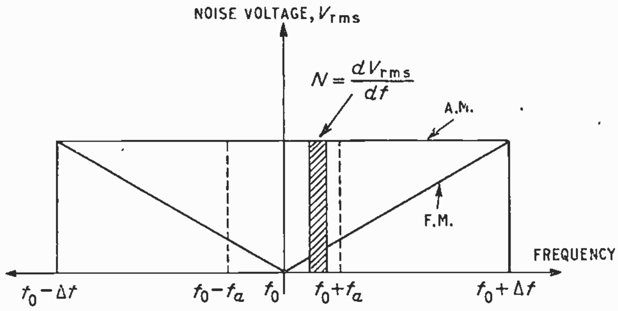


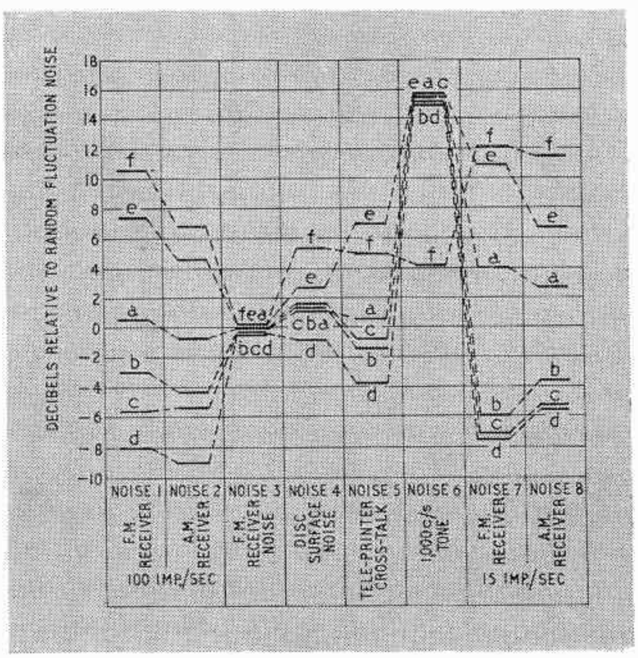
Fig. 2. *Noise spectra*

which falls off at the higher frequencies. Were this not so, then pre-emphasis would cause over-modulation at the transmitter.

Experiments conducted by the B.B.C.[5] showed beyond doubt that 50 μs was the best compromise for the pre- and de-emphasis time constants to be used for an f.m. transmission. Shorter time constants than this gave no improvement to f.m. or a.m., while longer time constants caused such over-modulation that transmitter 'line-up level' had to be so reduced as to nullify the advantage obtained in noise reduction. If account be taken of the necessary reduction in transmitter line-up level it is found that 50 μs gives a 0-dB improvement on a.m. and 4 dB on f.m.

The effect of de-emphasis in the receiver is to reduce the higher audio-frequency components of the noise but to restore the pre-emphasised components of the audio signal to their correct proportions. Its effect can

Fig. 3. *Response of meters to different noises; all meters aligned on random fluctuation noise. a, mean square meter (B.B.C. Research); b, G.P.O. speech voltmeter; c, Portable noise meter (B.B.C. Research),* 150 *ms charge time and* 1 500 *ms discharge time; d, V.U. meter; e, C.I.S.P.R.* 1 *ms charge time and* 500 *ms discharge time; f, Peak programme meter (B.B.C.)* 1·5 *ms charge time and* 1 100 *ms discharge time*



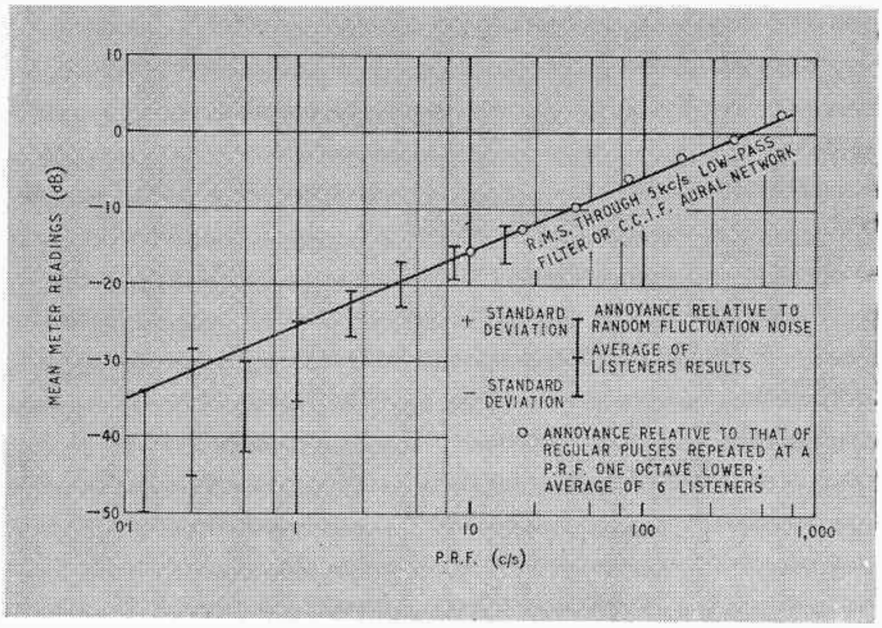


Fig. 4. *Annoyance as a function of noise p.r.f.*

thus be accounted for by a mere reduction in audio noise spectrum. Equ. (3) can be modified to account for pre- and de-emphasis by a suitable change in f_a.

Impulsive Interference

Subjective Tests Relating to Annoyance

So far we have dealt objectively with random fluctuation noise and subjectively with pre- and de-emphasis. If we wish to deal objectively with impulsive noise such as ignition interference and the noises produced by many domestic appliances, we must find or choose a suitable electrical parameter which will adequately represent the annoyance caused to listeners by this sort of interference.

A great many experiments[8] were undertaken by the B.B.C. in this connection. A proportion of the experiments consisted of asking listeners to adjust the level of an impulse noise emerging from a loudspeaker in the presense of programme to equality of annoyance with random fluctuation noise which could be interchanged instantaneously with the noise under test. The listener-adjusted levels were then measured on a number of different types of meter including power or root-mean-square meters, quasi-peak meters and V.U. meters, six meters in all. The meters were aligned on random fluctuation noise having a uniform audio spectrum. Fig. 3 shows the average change in meter readings between the reference fluctuation noise and each of eight different types of noise. It is easy to see that the high-crest-factor r.m.s. meter (a) gave readings which were in closer agreement with the listeners' results than any other meter.

This experiment was most encouraging because the power in a signal is a simpler quantity to measure or to calculate than such parameters as quasi-peak.

A further experiment tended to confirm that r.m.s. was the best criterion of annoyance of noise in the presence of a wanted programme. Listeners were asked to adjust the level of impulsive noise at a certain pulse-repetition frequency until it coincided in annoyance in one case with a previous sample of the same impulsive noise having a known but different p.r.f. and in another case with random fluctuation noise. The listener-adjusted attenuator readings could then be plotted against p.r.f. The result is Fig. 4. It can be seen that in the range 2 c/s to 700 c/s annoyance increases with p.r.f. at the rate of 3 dB per octave. This is, of course, the rate of increase of power and again favours the r.m.s. criterion.

Several other experiments confirmed the foregoing conclusions so that finally the ratio of signal power to

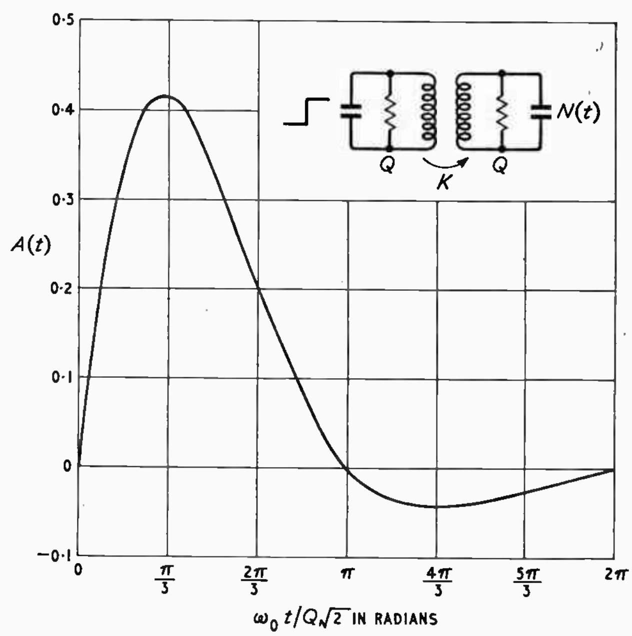


Fig. 5. *Amplitude of the i.f. impulse vector.*
$A(t) \approx e^{-\omega_0 t/Q\sqrt{2}} \sin \omega_0 t/Q\sqrt{2}$, *if* $Q \gg 1$ *and* $KQ = \sqrt{2}$

noise power, or its square root, was taken as the appropriate form of signal-to-noise ratio.

The Effect of F.M. Receiver I.F. Circuits on an Impulse

Let us consider the case of a wide-band v.h.f. f.m. superheterodyne receiver having an i.f. half-bandwidth greater than the audio bandwidth. The advantage of assuming an i.f. width greater than twice the audio band is merely one of simplicity in that noise pulses emerging from the discriminator may be treated as unit impulses in the Dirac or Heaviside sense from the point of view of the audio circuits ; thus, any small changes of shape which may result from differences in intermediate-frequency response-curve shape need not be taken into account.

As regards the audio r.m.s. noise level resulting from repeated impulses applied to the receiver aerial terminal, it will be evident that this noise level will be independent of i.f. bandwidth since the energy content within an audio-frequency range of the centre or intermediate frequency is fixed. If we double the i.f. bandwidth, the noise impulse doubles its height and halves its duration, thus extending its spectrum twice as far but the spectral density in volts per cycle per second or watts per cycle per second is unchanged. Only if the i.f. half-bandwidth were less than the a.f. bandwidth would a change in i.f. bandwidth have an aural effect. In view of this, it might be thought profitable to extend the receiver r.f. and i.f. bandwidths indefinitely whilst increasing the transmitter deviation ratio to match. Thus, it might be thought, remembering Equ. (3), the f.m./a.m. improvement could be made as great as desired. The a.m. signal-to-noise ratio remaining constant with increasing r.f. and i.f. bandwidths, it would follow that the f.m. signal-to-noise ratio could be increased without limit. In theory, this could be done, but not without increasing the transmitted power indefinitely on account of a phenomenon known as the noise barrier or improvement threshold which we must now examine.

The Roles of the Limiter and Discriminator During Reception of Impulsive Interference in the Presence of an Unmodulated (Undeviated) Carrier

In order to interfere with an f.m. broadcast receiver tuned to a v.h.f. carrier at, say 90 Mc/s, an impulse must have a spectrum containing components within the receiver r.f. bandwidth, for example (90 ± 0·075) Mc/s. If we assume for simplicity, however, that portion of the spectrum of a unit step of interference which lies within the range ($\omega_0/2\pi \pm 0{\cdot}075$) Mc/s, where $\omega_0/2\pi$ is the i.f. centre frequency, and a normal type of band-pass coupled transformer arrangement of the type used in many receiver i.f. circuits, we find that the i.f. output voltage[7] before amplitude limiting has taken place is,

$$N(t) = A(t) \sin [\omega_0 t - \psi(t)] \qquad (4)$$

whilst the unmodulated carrier has the form

$$C(t) = \sin \omega_0 t \qquad (5)$$

$A(t)$ is thus the impulse-to-carrier amplitude ratio whilst N/C is its instantaneous value. Equ. (4) is worth examining closely. Fig. 5 shows a particular example of the function $A(t)$ while Fig. 6 shows the corresponding function $\psi(t)$. We can represent the entire function $N(t)$ by an isometric drawing showing $A(t)$ along one axis, $\psi(t)$ along another and t or a quantity proportional to it along the third, Fig. 7.

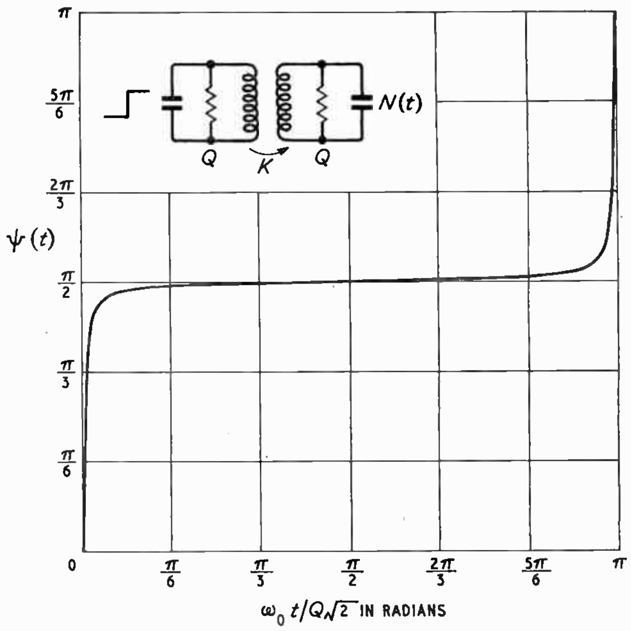


Fig. 6. *Phase of the i.f. impulse vector.* $\psi(t) \approx \tan^{-1} (Q\sqrt{2} \tan \omega_0 t/Q\sqrt{2})$ *if* $Q = 50$ *and* $KQ = \sqrt{2}$

Operation Above Improvement Threshold

We can now represent the state of affairs occurring at the discriminator input after the resultant of the carrier and impulse has passed through the amplitude limiter. Fig. 8 (a) shows a vector diagram in which the

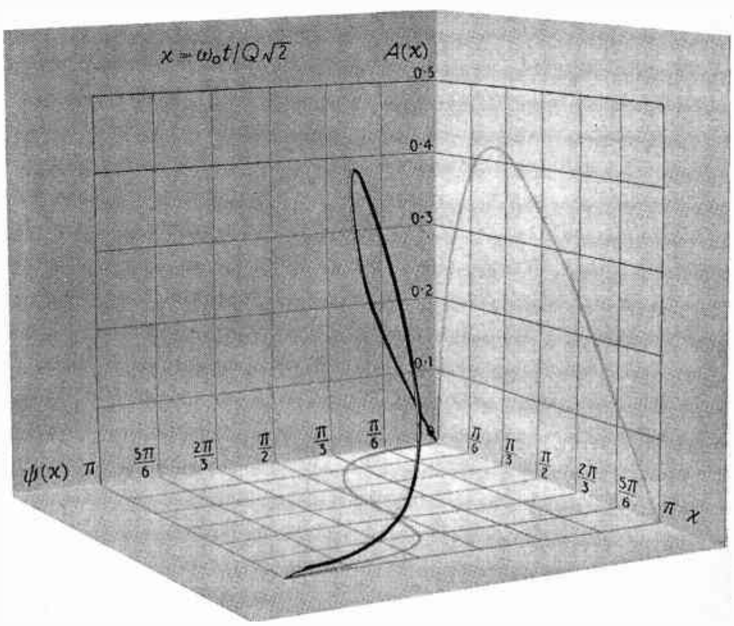


Fig. 7. *Isometric representation of* $N(t)$

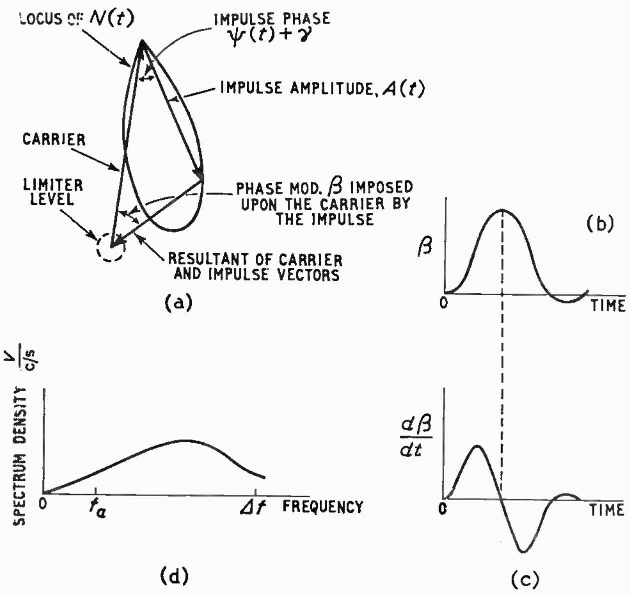


Fig. 8. *Carrier greater than that needed for the improvement threshold. (a) vector diagram of impulse and carrier; (b) phase of resultant of impulse and carrier; (c) frequency of resultant of impulse and carrier; (d) spectrum of discriminator response*

carrier vector, of unit amplitude, is made stationary, while the impulsive interference vector grows from its head, and having the same frequency but a variable phase angle, describes the locus shown. The actual position of the locus of $N(t)$ depends upon the arbitrary time at which an impulse arrives at the receiver-input terminals. With repeated impulses, therefore, all starting angles between impulse and carrier vectors are equally likely and so all locus positions will occur. Fig. 8 (a) shows a case of interference in which the peak impulse is less than the carrier amplitude. For such cases the phase angle β, Fig. 8 (b), always returns to zero thus giving rise to frequency changes, Fig. 8 (c), which are always of at least bipolar form. Such frequency changes always give rise to spectra, Fig. 8 (d), which are at least proportional to the first power of the frequency (if not to higher powers) over the audio range. Thus we see that for impulsive interference not exceeding the carrier in peak value the spectrum of the interference is triangular just as it was in the case of random fluctuation noise. Here we considered the frequency of the resultant in a direct manner and obtained the discriminator output without regarding it as a differentiating circuit because that operation (which, in fact, the discriminator performs) was done in Fig. 8 (c).

Now, depending upon the exact carrier-to-impulse ratios and the values of the arbitrary or random starting phase angles in Fig. 8 (a), we may obtain a great many different shapes in Fig. 8 (b) to (d) but the salient features will not change. The audio spectrum of the impulsive noise will be triangular or of higher power. Such noise, which is lacking in bass components, will sound like a succession of clicks or if the p.r.f. is much above 10 c/s it will resemble a continuous frying sound.

In the conditions already described where the carrier is big enough to maintain conditions above the improvement threshold the objective f.m./a.m. improvement for impulsive noise is the same as that given in Equ. (3) for random fluctuation noise.

The subjective improvement must take account of the listener's ear characteristic as well as the way in which bandwidth restriction is obtained. In what follows, the frequency-response characteristics of the listener's ear are taken conceptually as rectangular and are derived from equating noise energy output from an ideal low-pass filter to that resulting from the product of the square of the ordinates of a C.C.I.F. psophometric or weighting curve (Ref. 6 and Fig. 3, meter 'a') and the energy density of the noise spectrum under consideration. If, for instance, the f.m. receiver audio response is limited only by a de-emphasis circuit of time constant $RC = 1/(2\pi f_a)$ while the a.m. receiver audio response is wider than that of the ear, the aural or subjective improvement is approximately

$$\frac{\text{f.m. signal-to-noise ratio}}{\text{a.m. signal-to-noise ratio}} = \frac{\sqrt{5}\Delta_f}{\sqrt{f^3_a\,[8/f_a - \tan^{-1}8/f_a]}} \quad \ldots \quad \ldots \quad (6)$$

where $2\,\Delta f$ is the i.f. bandwidth in kc/s and also double the maximum deviation, and $f_a = 1/(2\pi RC)$ in kc/s, while the number 8 represents the equivalent energy bandwidth of the ear to triangular noise and the number 5 refers to the equivalent energy bandwidth of the ear to uniform noise. If $\Delta f = 75$ kc/s and $RC = 50\ \mu$s, we obtain an improvement of 28 dB. This has not accounted for the 2-dB reduction in transmitter 'line-up' level required owing to over-modulation caused by 50-μs pre-emphasis at the transmitter. In fact, an improvement of 26 dB is obtained.

The f.m./a.m. improvement factor given in Equ. (6) does not vary significantly with carrier-to-impulse ratio once the improvement threshold is surpassed. It is therefore easy to calculate an f.m. signal-to-impulse noise ratio above improvement threshold by first calculating the a.m. signal-to-impulse noise ratio and then adding the improvement figure in decibels. For an a.m. receiver having an a.f. bandwidth at least 5 kc/s wide receiving a 100% modulated carrier of η r.m.s. mV/m in the presence of impulses of area U peak mV/m × ms at a p.r.f. of f_r kc/s the a.f. aural signal-to-noise ratio is

$$\text{a.m. signal-to-noise ratio} = \frac{\eta\pi/2}{2\,U\sqrt{5f_r}} \quad \ldots \quad (7)$$

The quantity $2U\sqrt{5f_r}$ is the r.m.s. noise produced by short impulses U repeated at f_r kc/s in a receiver i.f. bandwidth of twice 5 kc/s. The $\pi/2$ in the numerator is the factor which accounts for the fact that the random phase angle γ between the oscillatory impulse and the carrier signal in the i.f. circuits makes the mean value of the r.m.s. noise pulses less than in the cases of phase coincidence and phase opposition.

If $\eta = 1$ r.m.s. mV/m, $f_r = 0{\cdot}05$ kc/s and $U = 0{\cdot}016$ mV(pk)/m × ms, Equ. (7) gives an a.m. signal-to-noise ratio of 40 dB.

Operation Below Improvement Threshold

We must now re-examine the phenomena shown in Fig. 8 (a) when the receiver is operating below the improvement threshold; that is, when the peak impulse in the i.f. output exceeds the carrier amplitude.

Two basic conditions[7,8] arise and they are shown in Fig. 9 at (a) and (a′). In Fig. 9 (a′) the carrier vector is not entirely included within the locus of $N(t)$ and it can be seen that such a condition is similar to that shown in Fig. 8 (a). A triangular spectrum will result and the impulsive interference will sound like a succession of clicks—this in spite of the fact that the impulse peak exceeds the carrier amplitude.

Fig. 9 (a) shows a different state of affairs. The carrier vector—and for simplicity, the limiter circle as well—are included inside the locus of $N(t)$. The resultant returns to its original coincidence with the carrier vector when the impulse has died away but this coincidence is achieved by a complete rotation of 360° as shown in Fig. 9 (b). The exact manner in which the 'unit step' of phase β is achieved is of secondary importance, the important feature of this phenomenon is that we now have a step of phase of duration appreciably less than $1/(\sqrt{2}\Delta f)$, say 9 μs. This duration is much shorter than any a.f. time constants and thus may be regarded from the point of view of audio frequencies as a true unit step of phase. From this results a (somewhat deformed) unit impulse of frequency, Fig. 9 (c), whence, finally, the uniform spectrum shown in Fig. 9 (d). Such a spectrum, containing the bass components which were lacking in the spectra of clicks shown in Fig. 8 (d), gives rise to a succession of pops if the impulses are repetitive.

Smith and Bradley, at that time of the Philco Corporation, were the first to point out the reason for the difference between clicks and pops although their analysis relied on the carrier being frequency-modulated by programme and the impulse being immobile with a constant phase angle γ. Their explanation did not account for pops of interference appearing during unmodulated periods of a programme. In order to have a rotation of the impulse vector it is necessary to assume some asymmetry, however slight, in either the spectrum of the interference or the transfer function of the receiver circuits or both. Such asymmetry invariably exists so that pops are always with us if the peak carrier-to-impulse ratio is less than one.

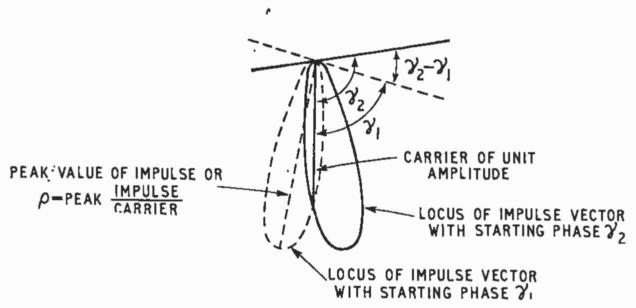


Fig. 10. *Proportion of pops in the total number of impulses*

We can obtain an approximation to the value of the improvement factor below threshold by first determining empirically the relative a.f. aural energy contents of a nominal pop and a nominal click assumed each to have the same peak-frequency deviations and durations, Figs. 8 (c) and 9 (c). Then, assuming all clicks to have the same energy content and likewise for all pops, we calculate the proportion of pops in the total of clicks and pops, Fig. 10, in terms of the peak impulse-to-carrier ratio ρ. This proportion is the ratio of the range of random starting angles $\gamma_2 - \gamma_1$ giving rise to pops, to the remainder of the 2π radians left over—the remainder which gives rise to clicks. This proportion is shown in Fig. 11 for an i.f. width of 150 kc/s at a centre frequency of 4 Mc/s, that is, a Q of 50.

The audio power ratio of a nominal pop to a nominal click may be written

$$b = \frac{23\,(\Delta f)^2 \tan^{-1}(5/f_a)}{[f^2{}_a\,(8/f_a - \tan^{-1} 8/f_a)]} \quad .. \quad (8)$$

and for $\Delta f = 75$ kc/s

$f_a = 3{\cdot}18$ kc/s ($RC = 50\ \mu$s)

we have $b = 10{,}000$ or 40 dB.

Thus if a is the proportion of pops shown in Fig. 11

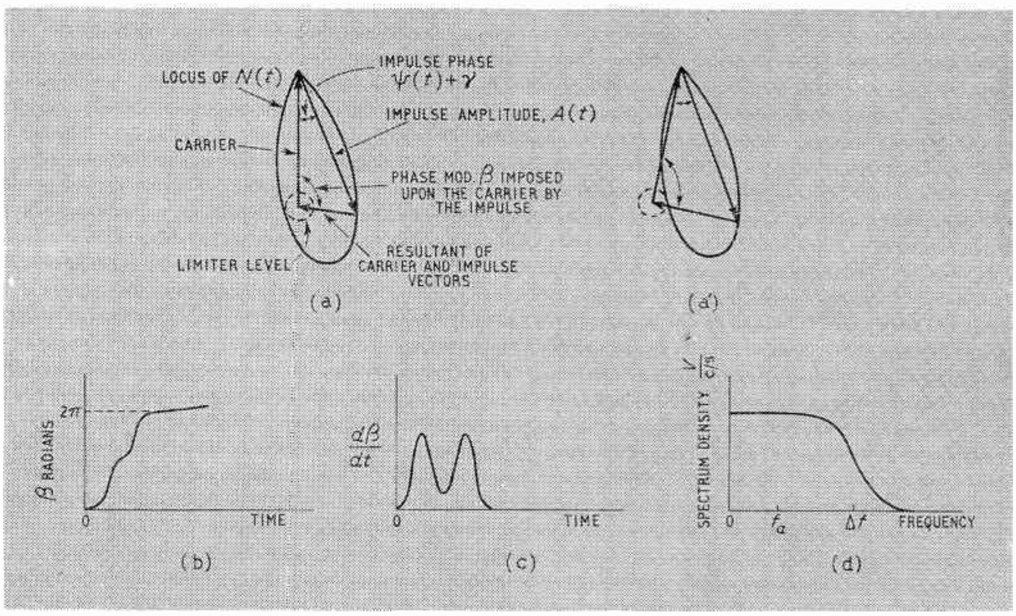


Fig. 9. *Carrier smaller than that needed for the improvement threshold. (a and a′) vector diagrams of impulse and carrier; (b) phase of resultant of impulse and carrier; (c) frequency of resultant of impulse and carrier; (d) spectrum of discriminator response*

for a given i.f. peak impulse-to-carrier ratio, ρ, we have

$$\frac{\text{aural noise power below improvement threshold}}{\text{aural noise power at improvement threshold}} = ab + 1 - a \qquad \ldots \qquad \ldots \quad (9)$$

This expression[9] is shown in Fig. 12 as a function of the carrier-to-peak impulse ratio $1/\rho$.

Of course, in spite of the various i.f. bandwidths chosen, Fig. 12 shows curves which are all coincident at the improvement threshold where $\rho = 1$ because they are each normalized to the noise power at the threshold. For $1/\rho \ll 1$ the term ab in equation (9) predominates; thus the separations between the curves in Fig. 12 for conditions well below improvement threshold tend to become equal to the ratios of the i.f. bandwidths in decibels since Equ. (8) shows the power ratio b to be proportional to the square of the bandwidth.

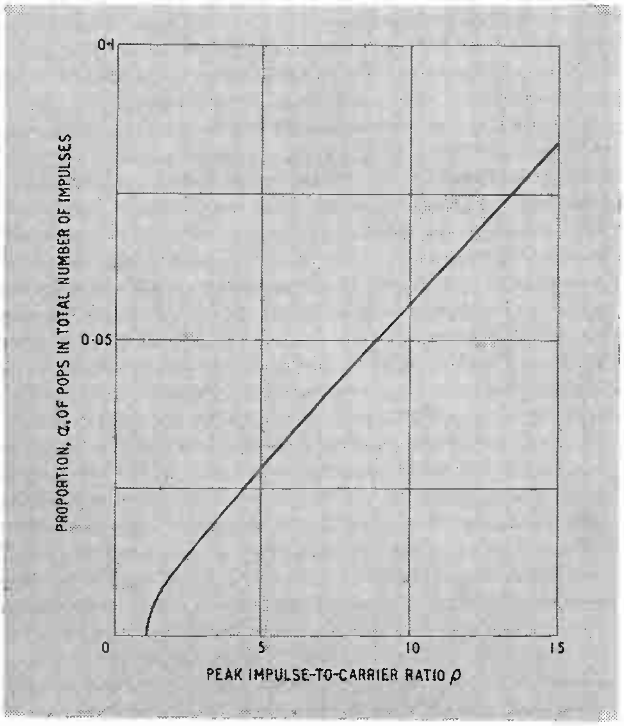


Fig. 11. *Proportion of pops as a function of impulse-to-carrier ratio.* $a = \frac{\gamma_2 - \gamma_1}{2\pi}$ (Fig. 10)

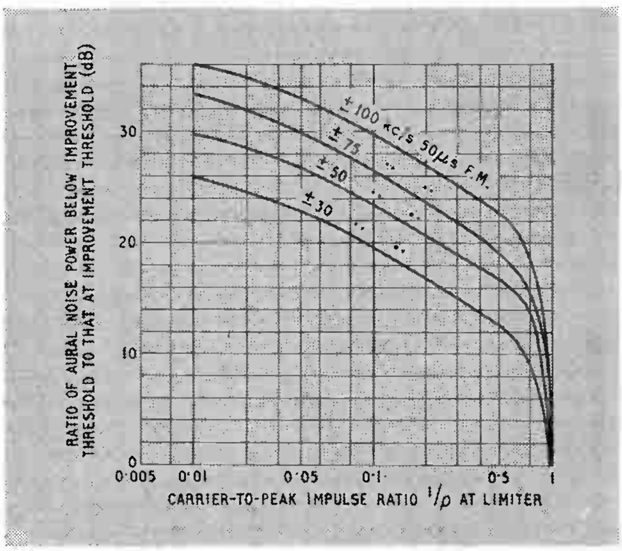


Fig. 12. *Variation in noise power in f.m. below the improvement threshold*

Variation of Signal-to-Impulsive Noise Ratio with Deviation Ratio and with I.F. Peak Impulse-to-Carrier Ratio

Fig. 12 can be used to obtain a set of curves showing the relative signal-to-impulse ratios of f.m. systems employing various maximum deviations. Equ. (6) can be used as well, in order to place an a.m. signal-to-impulse noise ratio curve upon the same graph. Thus, if we take the a.m. signal-to-noise ratio, whatever value it may have, as reference, we know from Equ. (6) that the f.m. signal-to-noise ratio is 28 dB greater at and above improvement threshold for a $\pm$75-kc/s 50 μs system, $30\frac{1}{2}$ dB greater for a $\pm$100-kc/s 50 μs system, $21\frac{1}{2}$ dB greater for a $\pm$30-kc/s 50 μs system, and so on. The shapes of the f.m. signal-to-noise ratio curves below improvement threshold can be obtained directly from Fig. 12; but the spacings between the curves on Fig. 12 rapidly become, as already stated, in proportion to the bandwidth ratios or the ratios of the maximum deviations and since the f.m./a.m. improvement factors are likewise in the same proportions, the portions of the signal-to-noise ratio curves below improvement threshold rapidly coalesce and we have Fig. 13.

If we wish to replace the abscissae of Fig. 13 by actual receiver r.f. input carrier-to-impulse ratios, we note immediately that the improvement thresholds for the various maximum deviations will occur for different r.f. signal-to-impulse ratios. Evidently a 100 kc/s-deviation system, for example, will have its threshold at half the r.f. impulse value of that of a system designed for a 50 kc/s maximum deviation since the bandwidth of the former will be twice that of the latter system and the i.f. impulse peak is, of course, proportional to bandwidth.

A calculation shows that improvement threshold or equality between i.f. signal amplitude and i.f. impulse peak occurs when

$$\eta / U = 4 \, \Delta f \qquad \ldots \qquad \ldots \qquad \ldots \qquad \ldots \qquad \ldots \quad (10)$$

where, as before, η = r.f. carrier field strength in r.m.s. mV/m (or r.f. voltage in r.m.s. mV)

U = field strength impulse area in peak mV/m $\times$ ms (or r.f. voltage impulse area in peak mV $\times$ ms)

Δf = i.f. half-bandwidth in kc/s or maximum deviation in kc/s.

We may now replot Fig. 13, replacing the abscissae $1/\rho$ by the r.f. input ratio η/U. From Equ. (10) the $\pm$ 100-kc/s system will have its improvement threshold at $\eta/U = 400$, the $\pm$ 75-kc/s system at $\eta/U = 300$, and so on, Fig. 14. The relative signal-to-noise ratios in Fig. 14 may be readily converted to absolute ratios by calculating the a.m. signal-to-noise ratio from Equ. (7) in any given case and then adding to this ratio expressed in decibels the difference between the appropriate f.m. curve and the a.m. line in Fig. 14. For example, consider

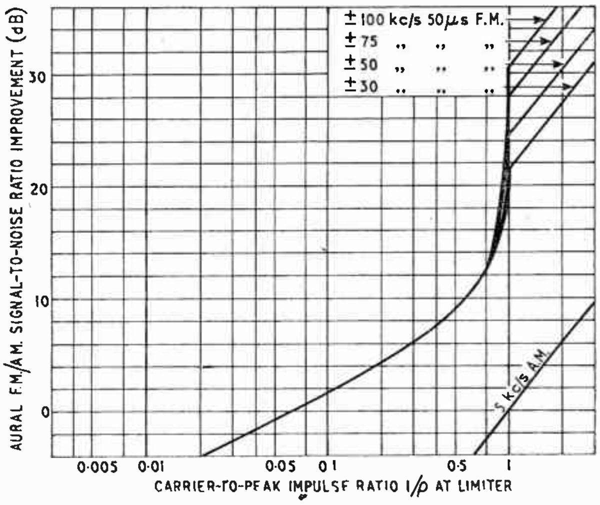


Fig. 13. *Variation of f.m. signal-to-noise ratio above and below improvement threshold*

the case of a carrier-to-impulse ratio of 250 (r.m.s. mV/m) per (peak mV/ × ms) and assume a p.r.f. of 50 c/s. The abscissa, $\eta/U = 250$, has been marked in Fig. 14 by the vertical line AB. The a.m. signal-to-noise ratio is, from Equ. (7), 52 dB: thus, if we allocate to the ordinate of the point C in Fig. 14, the number 52 dB, we can form the Table 1.

TABLE 1

Maximum Deviation of 50-μs f.m. System	Aural Signal-to-Noise Ratio
± 100 kc/s	64½ dB
± 75 kc/s	66½ dB
± 50 kc/s	76½ dB
± 30 kc/s	73 dB

which is obtained by adding 52 dB to the differences between the ordinates D–C, E–C, G–C and F–C respectively. This table shows that it is not possible to state which maximum deviation it is better to use unless the input carrier-to-impulse ratio is given. If the p.r.f. of the interference were different from the 50 c/s value chosen, the signal-to-noise ratios shown in Table 1 would each be changed by the same quantity.

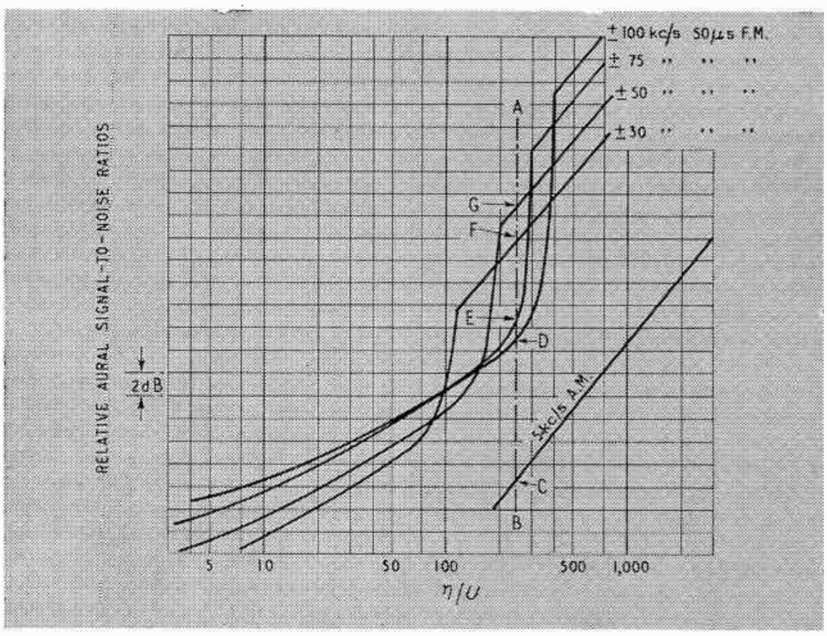
A Practical Example: Motor Car Ignition Interference

The change in the character of the sound of impulsive interference when the operating point of an f.m. system crosses the improvement threshold can be used to measure the peak value of the interference in a known f.m. receiver i.f. bandwidth. Thus, to measure the value U of impulses of field strength radiated by the ignition systems of motor cars the arrangement shown in Fig. 15 was set up at the side of a main road in Surrey[10]. The stream of traffic was practically unidirectional and less than continuous so that it was possible, as each motor car approached, to adjust the attenuator setting of the signal generator to such a value that the carrier-to-peak-impulse ratio in the receiver i.f. circuits was unity or fractionally less. Thus the record of the attenuator settings enabled the equivalent impulse U to be calculated from Equ. (10). Fig. 16 shows the results.

We now have actual values of field-strength ~~of~~ motor car ignition interference which we can use to calculate the signal-to-noise ratios which will result in any given f.m. system during reception of frequency-modulated carriers of various strengths.

Let us regard as adequate a wanted-signal field-strength which will protect a given service from 90% of all motor cars. From Fig. 16 it may be seen that only 10% of cars radiate ignition impulses having areas greater than $U = 0{\cdot}0135$ peak mV/m × ms at 67 feet. If a signal strength of 1 r.m.s. mV/m be taken, we have $\eta/U = 74$ and the accompanying a.m. signal-to-noise

Fig. 14. *Variation of f.m. signal-to-noise ratio in terms of carrier-to-impulse ratio. Ratio: η(r.m.s. mV/m or mV)/U(peak mV/m or mV × ms) of r.f. carrier to r.f. impulse*



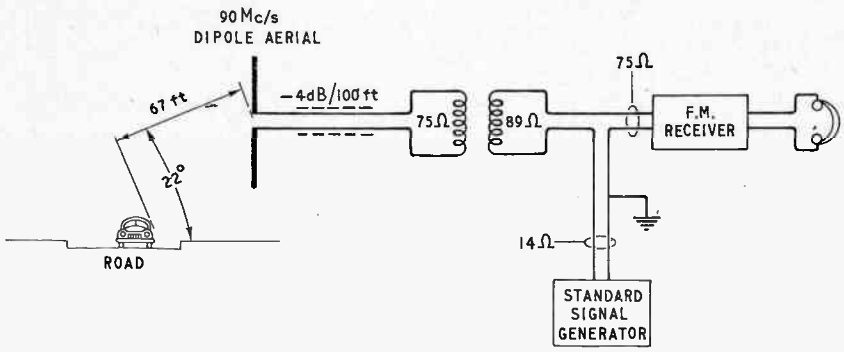


Fig. 15. *Method of measurement of car-ignition interference*

ratio for, say, 50 c/s p.r.f. (4-cylinder cars running at about 30 m.p.h.) is, from Equ. (7), 41·3 dB. If the receiving aerial were nearer to the road, for example 40 feet, we take $U = 0{\cdot}023$ on the assumption that field-strength is inversely proportional to distance. We now have $\eta/U = 44$ for $\eta = 1$ r.m.s. mV/m as before. We can now re-draw Fig. 14, replacing the η/U abscissa by wanted carrier field-strengths in r.m.s. mV/m and the ordinate by absolute signal-to-noise ratios. We have two scales of abscissae, one for receiving dipoles 67 feet from the cars and the other for dipoles 40 feet from them. The signal-to-noise ratios shown in Fig. 17 are 8 dB lower than the values obtained from Equ. (7) to allow for average modulation depth being approximately 8 dB below peak modulation.

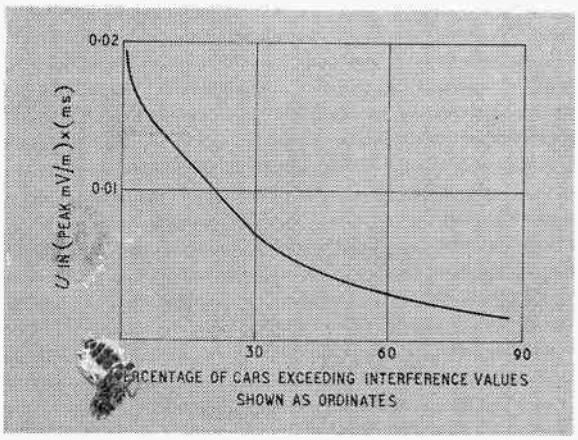


Fig. 16. *Values of ignition interference in 90-Mc/s band. The interference U was measured at a height of 25 feet and a distance of 60 feet from the centre of the carriageway and broadside on. There was no significant difference between horizontal and vertical polarization; more than 350 cars were observed*

Some of the signal-to-noise ratios shown are very great, particularly those above the improvement threshold for ± 50 kc/s-deviation systems and so all the ratios have been labelled with subjective criteria as a guide to a practical interpretation of the curves.

It is interesting to note that for low impulse repetition frequencies such as are encountered in cases of ignition interference, the signal-to-noise ratios above improvement threshold for most wide-band f.m. systems are so great that the noise is nearly if not quite inaudible in typical domestic listening conditions. Some of the misunderstandings which arose during the f.m./a.m. controversy may well have been due to the fact that when impulsive interference is actually heard it almost certainly consists of loud pops due to operation below the improvement threshold.

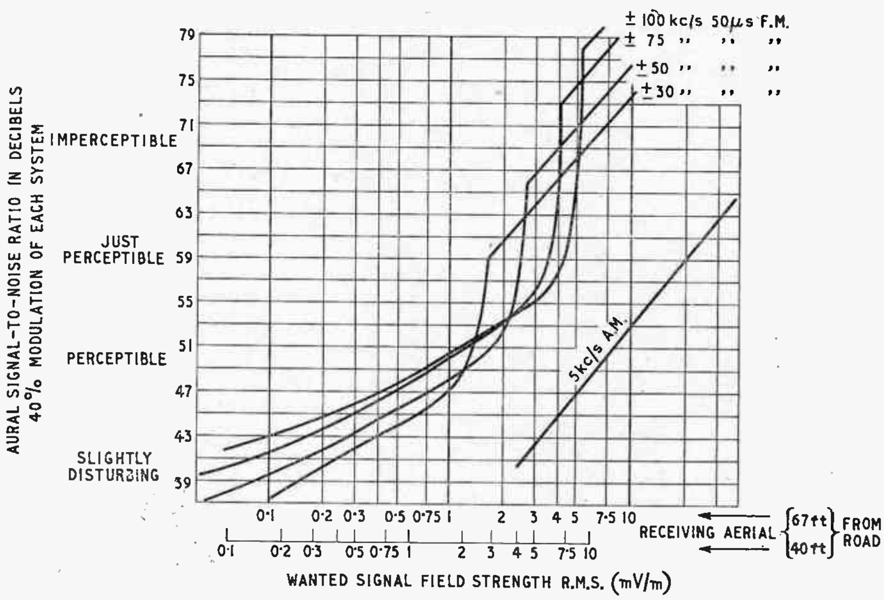


Fig. 17. *Absolute signal-to-noise ratios for ignition interference*

The Television Advisory Committee's second report, issued on the 18th January, 1954, recommended that the aim of a broadcasting authority should be to provide a median field of 0·8 mV/m in large towns and 0·25 mV/m elsewhere in the intended service area. The 0·8-mV/m figure is adequate for ignition interference if the receiving aerial is at least 40 feet from the road since it ensures signal-to-noise ratios of not less than 46 dB for $\pm$ 75 kc/s 50 μs f.m. system.

Acknowledgements

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A great deal of the information, both facts and figures, given in this work was supplied by the author's colleagues in B.B.C. Research Department. In particular, Mr. G. F. Newell has been closely connected with almost every aspect of the work and the author is greatly indebted to him.

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Transistor RC Oscillators

EFFECT OF TRANSISTOR CAPACITANCE

By M. K. Achuthan*

Resistance-capacitance tuned oscillators have become increasingly popular as sources of audio-frequency voltages. The commercially available variable-frequency oscillators use valves as the active elements. For applications where portability or light weight is of importance, the use of transistors is certainly much more desirable. The conventional *RC* tuning networks used with valves have, however, to be modified to suit the low input impedance and high output impedance of the transistor. Hooper and Jackets[1] have attempted a solution and have evolved a sine-wave transistor oscillator incorporating a zero phase-shift network as reproduced in block schematic form in Fig. 1. The frequency of oscillation is given by $f = 1/2\pi RC$ and the attenuation of the tuning network is 3 so that the gain required for maintaining oscillations is 3.

It will be noted that this expression for frequency holds good when there is no capacitance shunting R_2. If the collector–base junction capacitance C_c of the transistor is taken into account, it may be shown that the frequency is given by

$$f = \frac{1}{2\pi R\sqrt{C(C + 2C_c)}}$$

The value of C_c being usually greater than about 30 pF, we find that its presence will appreciably affect the frequency when C_c is not negligibly small in comparison with C. When $C \gg C_c$, as in the values used by Hooper and Jackets ($C = 2\mu$F, 0·5μF, 0·01 μF, and 0·005μF), the variation in frequency may be ignored. If an ordinary ganged tuning capacitor having the range 20 to 450 pF is used in a variable-frequency oscillator of the above

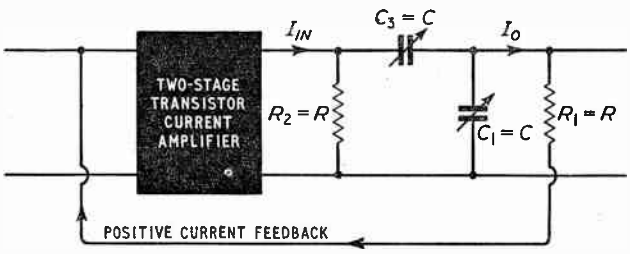

Fig. 1. *Hooper-Jackets oscillator*

type, using a typical value of R = 100 kilohms, the frequency range would be 3·5 kc/s to 80 kc/s, if C_c were neglected; whereas the presence of C_c (say equal to 40 pF) will reduce the range to 3·3 kc/s to 35·6 kc/s.

There is another and a more serious effect of the presence of C_c, which perhaps escaped the notice of Hooper and Jackets, since they used very large values of C. In the absence of C_c the attenuation of the network (for $R_1 = R_2 = R$ and $C_1 = C_2 = C$) remains constant,

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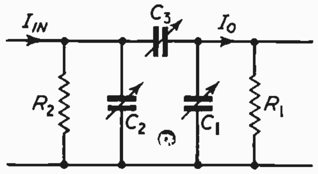
equal to 3, throughout the range of the tuning capacitor. In the presence of C_c, however, the attenuation is $3 + C_c/C$ and, in the case considered above, the attenuation will vary from 3·1 to 5 as the tuning capacitor is varied. The amplifier will have to keep track with this varying attenuation, as otherwise it will not be possible to maintain oscillations or to avoid distorted output. This means that a variable-frequency oscillator will necessarily have to involve point-to-point adjustment of



$$f = \frac{1}{2\pi\sqrt{R_1R_2\,(C_1C_2 + C_2C_3 + C_3C_1)}}$$

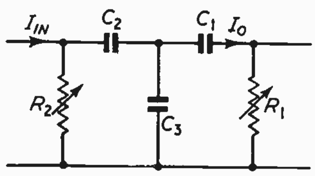
$$\textit{Attenuation} = \left[1 + \frac{R_1}{R_2} + \frac{C_2}{C_3} + \frac{C_1}{C_3}\cdot\frac{R_1}{R_2}\right]$$

Fig. 2. *Circuit and equations of π-network*

negative feedback or amplifier gain, which is highly undesirable as the oscillator will lose much of its flexibility.

We have been working on *RC* tuned transistor oscillators for some time and have developed two variable-frequency oscillators using (i) a π-network and (ii) a T-network, giving sinusoidal outputs of very good waveform. The generalized networks and the corresponding expressions for frequency and attenuation are given in Figs. 2 and 3 respectively. Both the networks are designed on the basis of positive current feedback (not voltage feedback as with valves) from the output of the transistors to the input. The part played by the transit-time effect of the current carriers inside the crystal in altering the α of the transistor and its effect on the performance of the oscillator have not been considered in this short note.

In the π-network, it will be noted that the variation



$$f = \frac{1}{2\pi}\sqrt{\frac{\dfrac{1}{C_1C_2} + \dfrac{1}{C_2C_3} + \dfrac{1}{C_3C_1}}{R_1R_2}}$$

$$\textit{Attenuation} = \left[1 + \frac{R_2}{R_1} + \frac{C_3}{C_2} + \frac{R_2}{R_1}\cdot\frac{C_3}{C_1}\right]$$

Fig. 3. *Circuit and equations of T-network*

in attenuation due to C_c can be counteracted by connecting capacitances equal to C_c across C_1 and C_3. In the symmetrical case (where $R_1 = R_2$ and the capacitances are equal to one another) the attenuation will remain constant, equal to 4, throughout the range of tuning and the expression for frequency will be

$$f = \frac{1}{2\pi R(C + C_c)\sqrt{3}}$$

The expression for frequency of the T-network shows that for the same values of C and R, in the symmetrical case, the highest frequency is obtained with the T-network ($f = \sqrt{3}/2\pi RC$). The adverse effect of C_c, as discussed above, cannot, however, be eliminated as in the π-network.

It will further be noted that the Hooper-Jackets oscillator, which is essentially of the Wien bridge type, is a special case of the π-network oscillator, in which $C_2 = 0$. Another modification of the π-type oscillator is obtained by putting $C_1 = 0$ as shown in Fig. 4. The T-type network also reduces to this type by putting $C_2 = \infty$. The expressions for frequency and network attenuation for this 'inverted-L-network oscillator' are exactly the same as in the Hooper-Jackets oscillator. Unlike the Hooper-Jackets oscillator, here the network attenuation remains constant, equal to 3, throughout the range of the tuning capacitor, even in the presence of large values of C_c, provided C_3 is shunted with a capacitance equal to C_c.

A typical oscillator of this type using low-frequency junction transistors of the type Raytheon 2N130 (with R = 125 kilohms and 15 kilohms and C varying from 0·1 μF to 20 pF, C_3 being shunted by a fixed capacitance equal to C_c) gave a frequency range continuously variable from 15 c/s to 72 kc/s. This was obtained without any point-to-point adjustment of negative feedback or amplifier gain. With point-to-point adjust-

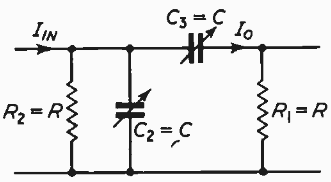


Fig. 4. *Inverted-L-network*

ment of gain an upper-frequency limit of about 150 kc/s has been attained so far. A considerably higher upper frequency limit is expected of the circuit with some of the modern high-frequency transistors.

A complete paper on this subject will be published elsewhere.

I wish to thank Dr. H. Rakshit, Head of the Department of Communication Engineering, and Dr. K. K. Bose, Assistant Professor in Communication Engineering of the Indian Institute of Technology, Kharagpur, for helpful discussions and encouragement.

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By Computer

Determination of Stability for a Four-Stage Feedback Amplifier

Last month we considered the conditions for an algebraic equation to be free from roots with positive real parts. We now apply the criteria obtained to the four-stage feedback amplifier of Fig. 1.

We shall consider only performance at frequencies so high that interstage coupling capacitances, screen-feed components and bias circuits can all be ignored. Performance is then governed mainly by the values of the shunt valve and wiring capacitances in relation to the coupling resistances.

The effective circuit of an individual stage at such frequencies can be reduced to the well-known form shown in Fig. 2. Here the input grid–cathode voltage of the valve is v_g and g_m is the mutual conductance, giving the input current $g_m v_g$ shown in Fig. 2. R is the value of the anode a.c. resistance and coupling resistance of the valve, combined in parallel with each other and with the following grid leak. C represents the total shunt capacitance of the stage and comprises the output capacitance of one valve plus the input capacitance of the next plus stray wiring capacitance. The output voltage v_0 of one stage forms the input grid–cathode voltage for the next stage.

The gain of the stage equivalent to the circuit of Fig. 2 is

$$A = \frac{v_0}{v_g} = \frac{-g_m R}{1 + pCR} \qquad \text{.. .. ..} \quad (1)$$

the minus sign being included to take account of the reversal of phase in the stage. Ignoring feedback for the moment, the gain from the grid of V_1 to the anode of V_4 is therefore

$$A_{14} = A_1 A_2 A_3 A_4 \qquad \text{.. .. .. ..} \quad (2)$$

where each of the A-terms is of the form given for Fig. 2 with the values of g_m, R and C appropriate to the stage in question.

We now have feedback to consider. C_5 (in Fig. 1) can normally be neglected at high frequencies; we shall also assume $R_6 \ll R_5$. The presence of R_5 will slightly modify the gain of the V_4 stage by shunting its coupling resistance; this can be allowed for by altering the effective value of R (in Fig. 2) for this stage.

The determination of the feedback factor is complicated by current feedback in V_1 because of the presence of R_6 in its cathode circuit. In effect, therefore, there are two kinds of feedback to take into account. The equivalent circuit of the first stage is shown in Fig. 3. If we note that

$$v_i = v_{g1} + (i_a + i_1)\, R_6 \qquad \text{.. .. ..} \quad (3)$$

and write the other equations in accordance with Kirchhoff's Laws, it is a matter of simple algebraic manipulation to find that the overall gain A_0 is given by

$$A_0 = \frac{v_{01}}{v_i}$$

$$= \frac{-\mu\, A_2\, A_3\, A_4\, z}{r_a + z + (1+\mu)\dfrac{R_5 R_6}{R_5 + R_6} + A_2 A_3 A_4 (1+\mu)\dfrac{z R_6}{R_5 + R_6}} \qquad \text{.. ..} \quad (4)$$

where $z = R'/(1 + pCR')$

Now if

$$A_1 = \frac{-\mu z}{r_a + z + \dfrac{R_5 R_6}{R_5 + R_6}(1+\mu)} \qquad \text{.. ..} \quad (5)$$

A_1 is the gain of the first stage with its own internal

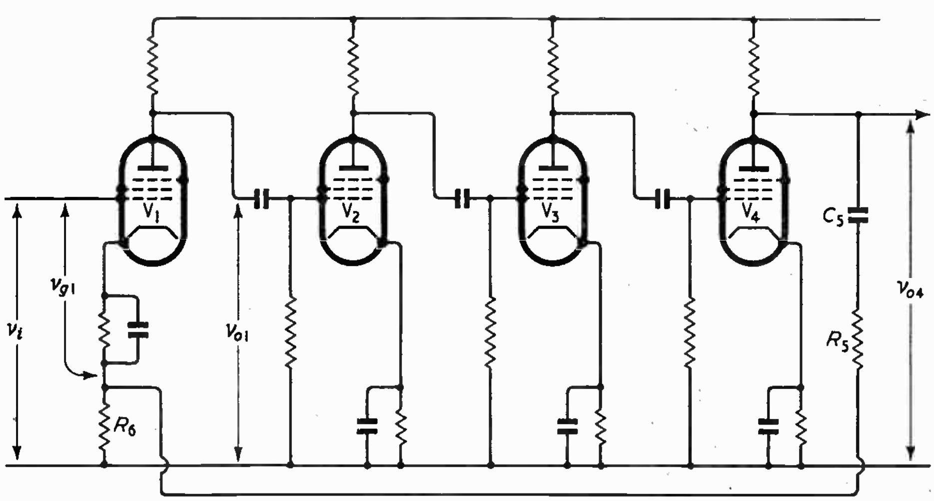


Fig. 1. *Four-stage feedback amplifier*

current feedback through $R_5R_6/(R_5 + R_6)$ acting as an unbypassed cathode resistor. This expression is conveniently expanded to

$$A_1 = \frac{-g_m r_a R'}{r_a + R' + (1+\mu)\dfrac{R_5R_6}{R_5+R_6}} \cdot K$$

where

$$\frac{1}{K} = 1 + pC\frac{R'\left[r_a + \dfrac{R_5R_6}{R_5+R_6}(1+\mu)\right]}{r_a + R' + \dfrac{R_5R_6}{R_5+R_6}(1+\mu)} \quad .. \quad (6)$$

This apparently cumbersome expression actually shows clearly that the effect of this local feedback on the transient response is merely to change the effective a.c. resistance of the valve from

r_a to $r_a + (1 + \mu)\, R_5R_6/(R_5 + R_6)$.

The value of R for this stage is thus R' in parallel with this effective resistance instead of in parallel with r_a.

If we substitute in the expression (4) for A_0 in terms of A_1 we get

$$A_0 = \frac{A_1\,A_2\,A_3\,A_4}{1 + A_1\,A_2\,A_3\,A_4\,\beta} \quad .. \quad .. \quad .. \quad (7)$$

where $\beta = R_6/(R_5 + R_6)$.

In doing this, we have introduced one approximation, which is that in the denominator $1 \ll \mu$. Because the feedback is applied to the cathode of V_1, the loop gain is dependent on $1 + \mu$ in the first stage

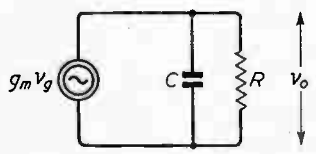


Fig. 2. *Equivalent circuit of amplifier stage*

instead of on μ. For a first-stage valve, μ will rarely be less than 40 and will be much higher if a pentode is used. The error is negligible, therefore, and considerably simplifies the equations.

Now each of the A terms is of the form

$-g_mR/(1 + pCR)$

so if $A_L = g_{m1}R_1g_{m2}R_2g_{m3}R_3g_{m4}R_4$

we have

$$A_0 = \frac{A_L}{(1+pC_1R_1)(1+pC_2R_2)(1+pC_3R_3)(1+pC_4R_4)+A_L\beta} \quad .. \quad .. \quad (8)$$

$$= \frac{A_L}{(1 + A_L\beta) + b_1p + b_2p^2 + b_3p^3 + b_4p^4} \quad .. \quad (9)$$

where the b terms are all combinations of the CR terms, and can readily be evaluated by expanding the denominator of (8) and equating the coefficients of corresponding powers of p in (8) and (9).

If β is zero, the equation has all its roots real and negative but, if β is gradually increased, the equation

$$(1 + A_L\beta) - b_2y + b_4y^2 = 0$$

whose roots we called α_1 and α_2 in last month's article is such that α_1 and α_2 move closer together but $(\alpha_1 + \alpha_2)$

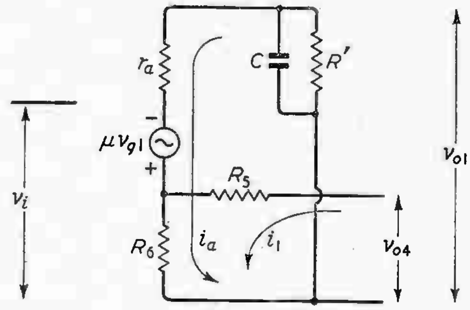


Fig. 3. *Equivalent circuit of first stage and feedback network*

remains constant and equal to b_2/b_4. The equation

$$b_1 - b_3y = 0$$

has a fixed root which we called β_1 in the earlier article. Hence, as β is gradually increased, a value must eventually be reached when the condition for stability, which is

$$\alpha_1 < \beta_1 < \alpha_2,$$

is no longer met. It is clear that

$$A_L\beta < b_2^2/(4b_4) - 1 \quad .. \quad .. \quad .. \quad (10)$$

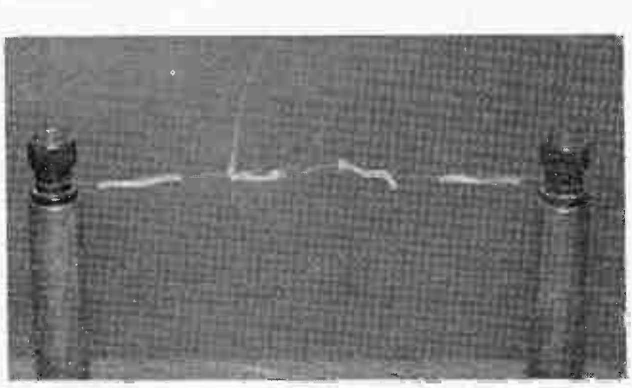
otherwise α_1 and α_2 will not be real and there cannot be stability. As the original equation is quartic, the condition for stability can be written down explicitly; it is

$$(1 + A_L\beta)b_3^2 < b_1\,b_2\,b_3 - b_1^2b_4 \quad .. \quad .. \quad (11)$$

but, in many cases, the simpler condition (10) is sufficient to give us the information required. For a quintic equation, we could still obtain a rough criterion of the form (10), but there would no longer be a simple exact one of the form (11) and, for exact information, we should be compelled to determine the quantities α_1, α_2, β_1 and β_2, as suggested in the earlier article.

IRRADIATED POLYTHENE

A form of polythene, to which desirable physical properties are imparted by bombardment with high-velocity electrons, has been introduced by Mersey Cable Works Ltd. One effect of the irradiation is to raise the maximum operating temperature of the insulant from 85°C to 100°C, with a short-term maximum of about 150°C. The photograph shows the effect of heating normal and irradiated polythene by means of a resistance wire. The normal polythene samples (in the centre) melt, while the irradiated material softens but is otherwise unaffected. Other desirable properties claimed for the new material are improved resistance to chemical attack, better tensile strength, elongation and abrasion resistance.



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.

Phase-Sensitive Discriminator

SIR,—My attention has been drawn to the comments on the "Phase-Sensitive Discriminator" by Mr. Walters in your March issue. The statement made in my letter published in the January issue of your journal that the amplitude of the sawtooth waves is proportional to the time interval between successive waves *and hence between their phase differences* was incorrect, and it is regretted that this escaped my notice at the time of communication. As has been pointed out by Mr. Walters the time interval is actually inversely proportional to the instantaneous frequency and for small frequency deviation, as is usually the case, the sawtooth amplitude bears a linear relation with the modulating waveform $f(t)$ in f.m. and $f'(t)$ in ph.m.

The point raised by Mr. Walters has, of course, been clearly explained by Mr. P. Kundu of this laboratory who originally developed the idea of utilizing the time period of an angle modulated carrier for demodulation in his recent paper[1]. He has pointed out that "An angle modulated wave (ph.m. or f.m.) is thus translated into a sawtooth amplitude-modulated wave, the amplitude, of which is proportional to $f(t)$ in case of f.m. and to $f'(t)$ for ph.m.". Mr. Kundu further adds that "The modulating signal may finally be reproduced by a low-pass filter. The wave emerging from the filter is, however, the original wave integrated in time. If the input is a phase-modulated carrier, the modulating signal is directly reproduced at the output of the filter. But in case of f.m. wave, a differentiating network is required to produce the original signal".

It was this integrating effect of the *RC* type low-pass filter that caused the output observed by me to be the same as the modulating signal.

B. CHATTERJEE

Indian Institute of Technology, Kharagpur, India.
20th May 1957.

[1] P. Kundu. "A New Method of Demodulation for Phase, Frequency and Amplitude", *Ind. J. Phys.*, 1957, Vol. 31, p. 231.

Triode Amplification Factor*

SIR,—Professor Moullin has made his paper highly confusing by insisting on theorems which are true but also irrelevant. The simple way to prove that the *effective* force inside a uniformly-charged closed container (the anode) is zero, is by taking advantage of the fact that, under inverse square-law conditions, the force due to any part of the surface is proportional to the solid angle subtended by it at the point in question. If through any point a double cone is drawn, clearly the solid angle is the same in diametrically opposite directions, and there is thus no *resultant* force. If, however, there is another body (e.g., a cathode) inside the anode, the situation is different. An engineer working on his perfectly sound and correct intuition of the situation would observe that from a point P he could see anode over a solid angle ϕ and cathode over a solid angle $(4\pi-\phi)$ and would deduce a force which he could, if he liked, ascribe to the anode as cause (e.g., anode voltage independent variable) or to the cathode shadow as cause.

Thus Möller's description of $1/\mu$ as Durchgriff "the anode sends through the grid turns lines of force into the cathode space which get hold of space charge electrons and pull them towards the anode" is not only correct and reasonable but also useful.

There is a further confusion with which every student is familiar. Every electric circuit problem is solved as a pseudo static problem by the device ultimately due to D'Alembert of introducing fictitious forces equal to mass times reversed acceleration in order to solve a dynamic problem in terms of statics. These we call 'back' voltages, equal and opposite to 'applied' voltages. Similarly there is the term 'centrifugal' force, a relatively conceptual term compared with 'centripetal' force.

If, however, we have chosen current α as independent variable, then $v=Z\alpha$ makes $Z\alpha$ a voltage source whereas with V as independent variable $Z\alpha$ is the conceptual 'back' voltage. Thus Moullin's anode voltage appears as a dependent variable 'caused' by the current. Interchanging V and α as dependent and independent variables is however merely a formal operation.

* P. Hammond. *Electronic & Radio Engineer*, April 1957, p. 135.

C. G. MAYO

British Broadcasting Corporation,
Kingswood Warren, Tadworth, Surrey.
30th May 1957.

SIR,—Mr. Mayo is in close agreement with Professor Moullin in stressing the importance of Newton's theorem about the force inside a closed container due to the gravitational attraction of that container. As I pointed out in my article, this theorem entered through Priestley and Cavendish into the very definition of electricity or as it was then called the 'electric fluid'. But Mr. Mayo's hypothetical engineer has left the company of these thinkers and is alone with his intuition, when he begins to consider the electric shadow of the cathode on the anode. It would be of interest to know what Newton would have said to the idea of a gravitational shadow. Mr. Mayo's engineer should reflect on the name of Maxwell's theory. It is known as the electro-magnetic theory of light, not the electrostatic theory of light nor the luminous theory of electricity.

I agree that there is no meaning in the question whether a voltage produces a current or a current a voltage, when we are discussing steady flow. If, however, there is a change in the flow, it is of great interest to discuss how such a change can come about. Herein lies the importance of Professor Moullin's discussion of the barrier surface. Moreover, even in the steady state, it is of interest to discuss the origin of the forces acting on the electron stream. In the somewhat analogous case of the flow of a jet of steam, it is of great interest to discuss the dynamics of steam flow, even if an energy balance calculation can be made by considering only the fuel going into the boiler. It is surely unreasonable to forbid the turbine designer to think in terms of local steam pressures. He is not likely to forget that a boiler is needed to supply the steam. Can it be that Mr. Mayo really prefers to treat a valve as the proverbial black box, whose lid must under no circumstances be opened ?

Dept. of Engineering,
University of Cambridge.
1st July 1957.

P. HAMMOND

Constant-Frequency Oscillators

SIR,—In your April issue Dr. D. A. Bell suggests that the losses in a simple resonant circuit can be represented by resistances in series with the inductance and capacitance, and that this is somehow equivalent to division of the coil losses into series and parallel components.

However, it is easy to show that whether the coil losses are represented by a series or by a parallel resistance, or by both together, the two resistances in the inductive and capacitive branches, measured by Dr. Bell's technique, are always the equivalent loss resistances of the inductor and capacitor. If he cares to write out the algebra, Dr. Bell can easily convince himself that there is no possibility of representing a parallel resistance in one branch by a series resistance in the other, nor can the series and parallel components of coil loss be distinguished in any practical way. If the inductance and the resistances are independent of frequency there exists the theoretical possibility of determining the equivalent circuit by measurements of great precision. In general, however, the coil parameters vary with frequency and even this theoretical possibility disappears. In short, the 'equivalent circuit' of a practical coil is a rather nebulous affair.

I have tried, in my paper, to explain that, at any one frequency, the division of coil losses into series and parallel components has significance only when a second coil is coupled to the first. When the coil is isolated, as in Dr. Bell's oscillator, the division is arbitrary and perhaps even meaningless. Only by making measurements over a range of frequencies could an equivalent circuit for the coil be determined, and even then the interpretation would be uncertain because of the variation of the skin and proximity effect resistances and the internal inductance with frequency.

With regard to Dr. Bell's other remarks, I am confident that no one has been mislead by the title in the way he suggests. It is surely clear from the introduction or the summary that the paper deals with only a limited aspect of the subject. Most readers would also understand that the experimental arrangement was designed, not to make the best use of the available material, but simply to separate the effects being studied from those others which, whatever their relative importance, are not relevant to the matter in hand.

A. S. GLADWIN

Faculté des Sciences, Quebec, Canada.
22nd May 1957.

Microwave Model Crystallography

SIR.—I was very interested in the article of the above title, by Messrs. J. F. Ramsay and S. C. Snook, in the May 1957 issue of *Electronic and Radio Engineer*, having had many stimulating conversations with Mr. Ramsay on the subject. In view of the closing remarks in the article on the future possibilities of the subject, perhaps it would be of some value if I elaborate somewhat my idea, to which the article refers, for simulating atoms on the microwave scale.

It is a well-known consequence of quantum mechanics that while we cannot pinpoint an electron in an atom, we can give the probability p of its being in an arbitrarily small volume. Since the X-ray frequencies used in crystallography are of the order of 10^{19}, while the orbital frequency of an electron in an atom is of the order of 10^{16}, the electron may be regarded as virtually fixed while the collision process between it and an X-ray quantum is taking place. When a large number of quanta collide with electrons in a large number of atoms, the positions of the electrons in the atoms will average out to the probability function p, so that as viewed macroscopically the X-rays are diffracted as if the atoms all consisted of a continuous charge distribution, the density at any point being given by the wave functions. Such atoms might be called 'average atoms'; it is these average atoms that we shall mean in the future in referring to atoms. Although we cannot make a model of an actual atom, we can attempt to make a model of an average atom, with some hope of imitating real effects.

The effect of the X-ray on an atom is to accelerate the extra-nuclear electrons; this motion, which is small compared with the orbital motions, appears as a reaction to the change of momentum suffered by the X-ray quanta when they collide with electrons. The problem of simulating an atom on the microwave scale is, therefore, to effect a distribution of charge, with a specified density function. A metal contains free electrons which could be accelerated by the electromagnetic fields of a microwave, but in a solid metal the electrons are uniformly distributed, so that we have no control over the density function. Also, the microwave fields would only affect a thin surface layer of electrons; they could not penetrate into the body of the 'atom'. In order for the fields to penetrate deeply into a crystal model, therefore, the metal balls used as atoms must be of a radius small compared with the inter-atomic distance, which would detract from the accuracy of representation of the crystal.

The way round the difficulty is to have small quantities of metal dispersed through the body of the atom model in such a way that the density function of the metal is the same, when averaged out statistically, as the probability function, summed over all electrons, for the electron distribution in a real atom with, of course, an appropriate change of scale. These probability functions have been calculated for many atoms. The model atom could be achieved practically by supporting the metal in some material of dielectric constant near unity, such as polystyrene foam. First, a small sphere of foam would be coated thinly with metal in a patchwork sort of way, then a layer of foam added, and another patchy film of metal, and so on until in the finished atom model there would be a large number of such films of metal. If the metal films are sufficiently thin (much thinner than the skin depth), and/or if the gaps between patches of film are sufficiently large, the incident microwave fields will have sensibly the same intensity at all points in the atom model, as is the case in the crystallographic analogue. Some provision must also be made for the generation of radial components of current; this may be done by inserting in each layer of the atom model a number of fine radial wires with the correct density distribution.

As well as having the correct density distribution, the metal films must be in the correct ratio to the radial wires so that the conductivity at any point is, statistically speaking, isotropic. We may then expect that the currents induced by the microwave fields would be similar to those induced in an atom by X-ray fields. In the case of X-ray diffraction, we may think of the X-ray scattering as being given by the reaction of the X-ray quanta when, on colliding with electrons, they impart momentum to them. In the microwave analogue, microwave quanta will be scattered in the same sort of way due to the currents they induce; i.e., as a reaction to the momentum imparted to the electrons trapped in the metal. The scattering factor of the atom model for microwaves should therefore be the same as the atomic scattering factor for X-rays.

In conclusion, I should like to point out a possible field of applica-for crystal models using atoms of the type described here which Messrs. Ramsay and Snook appear to have overlooked: it is the study of the solid state. An artificial crystal may be expected to behave at low frequencies—say tens of megacycles downwards—much as a real crystal behaves at microwave, infra-red, or optical frequencies. Single-crystal studies may thus be carried out in cases where chemico-physical techniques have not been perfected for growing single crystals. Also, when a postulate is put forward to explain an observed effect, instead of working out its consequences theoretically and comparing them with observation, an alternative approach might be to make a practical model and investigate its properties experimentally. These properties could then be compared with observations on the actual solid. Such a technique might prove useful in the study of dielectric and magnetic properties of materials at microwave or optical frequencies, and in investigating the nature of chemical bonding between the atoms of a substance. Such studies may help to solve the problem of making dielectric or magnetic materials with specified properties.

R. A. WALDRON

Marconi's Wireless Telegraph Co., Ltd., Research Department, Great Baddow, Essex.
29th June 1957.

Transistor Impedance Matching

SIR,—In your April issue Dr. H. Paul Williams pointed out that "for practical purposes" the product of the input and output matched resistances for a transistor stage is independent of the particular configuration used; i.e., common-base, common-emitter, or common-collector. This product he said was equal to the ratio h_{11}/h_{22}, this ratio itself being independent of the configuration. The implication is that this is true only provided certain approximations, appropriate for junction transistors, are made. In his derivation Dr. Williams used formulas based on such approximations.

The fact is that this relationship is true exactly, and thus holds not only for junction transistors but also for any device which can be represented as a linear three-terminal network, connected with one terminal common to input and output. This includes point-contact transistors, magnetic and dielectric amplifiers, and any combination of any of these together with coupling networks, as well as passive networks with or without energy-storage elements.

The ratio h_{11}/h_{22} is, exactly, the same not only for the three commonly-used configurations of a transistor, but also for the "inverted" configurations (which give no power gain), such as with the collector common and the emitter input. It is the same, then, for any such network in any of the six configurations.

More generally, for any terminated network of this sort, not necessarily matched, whenever the product of the source and load impedances is equal to the ratio h_{11}/h_{22}, then the product of input and output impedances is likewise equal to this ratio which, as we have seen, is independent of which of the six configurations is used. This fact, which can be proven by direct calculation, is also true exactly, and holds for any frequency whatever, provided complex impedances are used rather than real resistances.

The relationship concerning the matched resistances is thus a special case of this more general relationship. At higher frequencies, however, matching is accomplished (at any given frequency) by making the (say) output and load impedances complex conjugates, rather than equals, and the simple relationship of Dr. Williams no longer holds.

Another consequence, which may be of some value, is that the input impedance as a function of load impedance may be obtained from the output impedance as a function of source impedance, and vice versa, for any linear network of this type. The input impedance is merely the ratio h_{11}/h_{22} divided by the value of output impedance for a source equal to the ratio h_{11}/h_{22} divided by the actual load impedance. This fact simplifies the derivation of these formulas somewhat.

PAUL PENFIELD JR.

Cambridge, Mass., U.S.A.
3rd July 1957.

Uncorrelated Grid Noise

SIR,—I will endeavour to be brief in replying to Mr. Harris's letter in your July issue, though I am at a loss to identify in that letter specific questions requiring answers.

(1) As previously stated, I believe that all crucial tests fail to support the Schottky-North theory that the noise-reducing effect of space charge is due to compensating-pulses of current, and do accord with a variable-velocity analysis such as has long been established in principle by A. J. Rack and by myself. By crucial tests I mean comparison of plane and cylindrical diodes, partition-noise in multi-grid valves and noise in travelling-wave tubes.

(2) When the grid is entirely outside the potential minimum, the dense space-charge at the latter reduces the ripple on the electric field there to a negligibly small value, and therefore prevents the movements of electrons between cathode and minimum from producing a displacement current at the grid. I take it we agree on this.

(3) The chance of which electrons are reflected is random, but if an electron is elastically reflected back from anode through the grid and forward again, the pulse of anode current due to that electron is inescapably correlated with the pulse of grid current due to that same electron.

(4) It does not seem to me reasonable that terms in $(\omega T_1)^3$ should be needed to fit the theory to the results of Houlding and Glennie which were carried down to frequencies well within the working range of the valves and therefore presumably to reasonably small transit angles.

(5) Conventional theory of correlated grid noise ignores two phase factors: (a) any voltage pulse at the grid, due to induced current in the grid circuit which will be at a maximum as an electron passes through the grid, will vary emission from the potential minimum and hence generate a current-change which is behind the grid voltage by a time approximating to the cathode–grid transit-time; and (b) no circuit can predict the future, so that tuning of a circuit to give a phase-advance on a continuous sinusoid will not give a corresponding time-advance in response to a random pulse. The second factor will be small if the bandwidth of the output equipment is small compared with that of the grid–cathode circuit.

D. A. BELL

Electrical Engineering Department,
University of Birmingham.
10th July 1957.

Stop-Band Response of Filters

SIR,—It is well known that the relationship between the input and output impedances of any 2-port passive network can be represented by

$$Z_2 = Z_I \frac{Z_1 + Z_I \tanh \gamma}{Z_I + Z_1 \tanh \gamma}$$

where Z_I = image impedance
γ = propagation function
Z_1 = external load impedance at port 1
Z_2 = input impedance measured at port 2

In the case where Z_I is real and γ is pure imaginary this relationship is represented by a particularly simple transformation on the Smith Chart; viz., an angular rotation about the centre through an angle arg.γ. This corresponds to the loss-free filter operating in a pass band, and also a section of loss-free transmission line, assuming Z_I is interpreted as the characteristic impedance.

It does not appear to have been realised that the alternative situation of Z_I imaginary and γ real, corresponding to a stop-band, can also be represented very simply on the Smith Chart. It is preferable however to rearrange the equation as follows:

$$(-jZ_2) = X_I \frac{(-jZ_1) + X_I \tanh \alpha}{X_I + (-jZ_1) \tanh \alpha}$$

where $Z_I = jX_I$ and $\tanh \gamma = \tanh \alpha$ and X_I and α are both real.

By so doing we effectively treat the artificial case of a transmission line of real characteristic impedance X_I and loss α. Since α is wholly real the length of the equivalent line is zero and hence the transformation is represented by a simple radial displacement towards the centre of the chart. The extent of this displacement is given by $\alpha = \log_{10} \Gamma_1/\Gamma_2$ where Γ_1 and Γ_2 are the artificial reflection coefficients corresponding to points $-jZ_1/X_I$ and $-jZ_2/X_I$. It is of course necessary to transfer the point Z_1/X_I to $-jZ_1/X_I$ on the chart, and to reverse the process in reading off, but this is no special hardship. Alternatively the result may be read off in terms of the true reflection coefficient and the insertion loss calculated by means of the well-known formula

$$L = 10 \log (1 - |\Gamma|^2) \text{ dB.}$$

In the most usual case Z_1 is a pure resistance and hence the point $-jZ_1/X_I$ lies on the unit reflection coefficient circle of the Smith Chart. However in analysing the effect of several filters in cascade, or a single filter with a reactive load, it can occur that one or other of the points $-jZ_1/X_I$ or $-jZ_2/X_I$ has a negative real component and hence falls outside the unit circle. There is no special significance in this and the standard chart can still be used, after suitable recalibration, by application of the transformation

$$\Gamma' = \tfrac{1}{2} \{(1 - k)\Gamma + (1 + k)\}$$

which transforms the point $\Gamma = +1$ into itself, the point $\Gamma = -1$ into the point k and the centre of the Smith Chart into the point $\frac{1}{2}(1 + k)$.

Application of this graphical method is of course limited by the fact that for high values of insertion loss as are commonly encountered in filter design the point $-jZ_2/X_I$ will lie very close to the centre of the chart. The method however does have value in a certain class of problem. For example, it provides a quick and ready method of examining for the occurrence of spurious responses in the stop band. These may occur as the result of an unsuitable load or if two filter sections in cascade have dissimilar image impedance characteristics. It is likewise simple to demonstrate that the requirements for zero loss due to interaction are rather special, and hence relatively improbable.

Standard Telecommunication Laboratories J. M. C. DUKES
6th March 1957.

New Books

Electricity and Magnetism

By B. I. BLEANEY and B. BLEANEY. Pp. 676+xiv. Oxford University Press, Amen House, Warwick Square, London, E.C.4. Price 63s.

The long-established advanced-electricity books of Starling, Pidduck, and Jeans served my generation of students excellently, but their authors have passed on and, until very recently, there were no successors to interpret the subject as it stands today, in a work of comparable status. Now, however, there has appeared a book which really does promise to serve the present generation equally faithfully. The first-named author, Mrs. Bleaney, is a Fellow of St. Hugh's College, and Lecturer in Physics at Oxford; the second has just been appointed Dr. Lee's Professor of Experimental Philosophy at Oxford, where he is a Fellow of St. John's College. Their book gives an account of the principles and experimental aspects at undergraduate or *first-year graduate* level.

It goes much farther than the earlier books; and it has to, for nowadays there is so much farther to go. It is, of course, intended for the physicist, and the brighter honours student at that. But it should have a much wider field of usefulness among those people, particularly electronic and radio engineers, who want an up-to-date picture of the interpretation of fundamental physics and radio work; and it can be highly recommended as a guide and work of reference. It covers the rapidly developing practical side quite admirably, and has so much information that people like myself have never before been able to find in any single book, and have had to ferret out the hard way bit by bit from more specialized treatises. There is also a good deal of modern work in it which, in spite of fairly conscientious efforts to keep up to date, I simply had not heard about before.

The basic theory requires about the same mathematical equipment as Pidduck, and where the sums do become rather difficult (as in the general development of the dipole and quadripole field equations in terms of spherical harmonics) the authors do all they can to help. The rigour of the argument is, however, accompanied by a careful sifting of the experimental evidence for its premisses—Plimpton and Lawton's 1936 experiment which verified Coulomb's inverse-square law to one part in 10^9; Kettering and Scott's 1944 measurement of the angular momentum of a circulating current which showed (as Tolman and Stewart had found earlier) that the carriers of the current in a metal are indeed electrons; and the relation of the force between current elements to Newton's third law of motion which was sorted out by Page and Adams (1945) in terms of the rate of change of momentum of the electromagnetic field. The way in which points of this kind are emphasized shows a wholesome regard for physics as experimental philosophy in fact as well as in name. *Nature* has recently been discussing the prospects of a Science Greats school at Oxford; physics expounded with this outlook can certainly rank with the more humane intellectual disciplines.

The more practical side of the book begins at about Chapter VII, where direct-current instruments are dealt with, including the current-balance and the Lorenz ohm-determination. There is a chapter on magnetic measurements and materials, which goes fully into electromagnet design, the shaping of pole pieces, and the Bitter pattern which gives fields above the range at which iron cores are much help.

Complex numbers are used from the beginning in the alternating-current chapters, and all this work is carried out in the neat and efficient way that the electrical engineers seem to have devised for the benefit of physicists. Resonance is clearly discussed, and the interpretation of the 'quality factor' Q, in its various applications, is fully explored. Six different uses, or meanings, for Q, from

$$Q = (L/C)^{\frac{1}{2}}/R \approx \omega_0 L/R \approx (\omega_0 CR)^{-1}$$

to $$Q = \frac{\omega_0 \times \text{(stored energy)}}{\text{rate of energy dissipation}}$$

are given. There was a time when the only readily available discussion of Q in simple terms was that much-circulated article by "Cathode Ray" in *Wireless World*! It is good to find that things like this, which can cause a certain amount of ambiguity and confusion because of quite legitimate differences in meaning in different contexts, are set out in a way that will help people who will consult the book for reference.

Electromagnetic radiation, covering reflection and refraction at a dielectric boundary, and reflection at a metal surface, radiation pressure, and radiation from an oscillating dipole, are next dealt with. Then follows a chapter on filters, transmission lines and waveguides, with the cavity resonator treated as a particular kind of waveguide.

The two chapters on thermionic valves and their applications, which include details of amplifying circuits, oscillators, negative feedback and cathode followers, and the limiter-discriminator arrangement for the demodulation of frequency-modulated signals, are getting well into the electronic and radio field. The feature that I appreciated most about them is the way in which figures and orders of magnitude are given, so that one understands the reason for differences in design. Problems of transit time, and the development of the klystron, the reflex klystron, and the magnetron, occupy the next chapter.

Measurements of impedance at low and radio frequencies; the absorption, resonant cavity, and coaxial line wavemeters; Essen and Gordon-Smith's determination of the velocity of electromagnetic waves; and Froome's microwave Michelson interferometer are among the modern measuring techniques explained.

There is a good chapter on fluctuations and noise, in which the connection between resistance noise and thermal radiation is given, and the procedures for designing a receiver for minimum noise factor and for measuring receiver noise are discussed.

Modern theoretical physics, in four chapters on dielectrics, solid-state conduction, paramagnetism, and ferromagnetism and anti-ferromagnetism, provides some rather harder reading than some of the preceding work. But the results are clearly set out, and complicated matters like the splitting of energy levels in magnetic fields are explained really thoroughly here.

Magnetic resonance is very fully dealt with, including Rabi's work on molecular beams, the measurement of nuclear magnetic moments, the determination of e/M for the proton, the cyclotron resonance of the proton, and Gardner's measurement of e/m for the electron in 1951 by a cyclotron-resonance method.

While using the rationalized m.k.s. system of units, the authors have tried to ease the path of anyone previously conversant with only the c.g.s. systems by giving at the end of the book a list, equation by equation, of those cases where the two forms of expression are different. I had always heard that Oxford physicists were a little unsympathetic towards m.k.s. units, and it is pleasant to find this view corrected and brought up to date. With the assistance of the appendix, of course, the whole can, if required, be translated back into c.g.s. units—an interesting challenge to those who believe things are simpler that way for, doubtless, people will find that it is indeed a lot easier to read it in m.k.s. units than to go through a book of this size with a pencil!

To sum up, this is a quite outstanding and important book in its field. It will be particularly helpful to those of you who feel a little outstripped by the pace of things, and want some guidance in trying to keep up with developments outside (or, possibly, even within) your own special line.

G.R.N.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Because of an urgent extension of the repair work on the long-wave aerial system at Rugby, it has been necessary to discontinue the transmissions of MSF 60 kc/s and GBR 16 kc/s from 20th July for a period which may extend to several months. Some users of these transmissions may be able to employ Droitwich 200 kc/s in this interval. To assist them in obtaining the highest accuracy, the deviations of this transmission, measured at 1030 G.M.T., will be given to 1 part in 10^9 in succeeding months

Deviations from nominal frequency for June* 1957

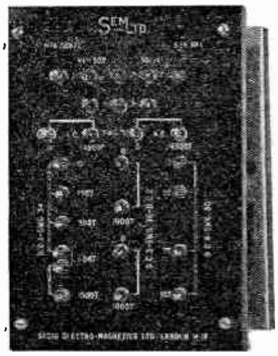
Date 1957 June	MSF 60 kc/s 2030 G.M.T. parts in 10^9	Droitwich 200 kc/s 1030 G.M.T. parts in 10^8
1	+ 1	− 1
2	+ 1	0
3	+ 1	0
4	+ 1	+ 1
5	+ 1	+ 2
6	+ 1	+ 3
7	+ 1	+ 3
8	+ 1	+ 1
9	+ 1	+ 2
10	+ 1	+ 3
11	+ 1	+ 3
12	+ 1	+ 3
13	+ 1	+ 4
14	+ 1	0
15	N.M.	N.M.
16	+ 1	N.M.
17	+ 2	+ 1
18	+ 2	+ 1
19	N.M.	+ 2
20	+ 2	+ 2
21	+ 2	+ 3
22	+ 2	− 2
23	+ 2	− 1
24	+ 1	− 1
25	+ 2	0
26	+ 2	+ 1
27	+ 2	+ 1
28	+ 2	+ 2
29	+ 2	+ 2
30	+ 2	N.M.

* *Nominal frequency is defined to be that frequency corresponding to a value of* 9 192 631 830 *c/s for the N.P.L. caesium resonator. N.M. = Not Measured.*

New Products

General-Purpose Magnetic Amplifier

The first of a range of magnetic amplifiers for 50-c/s operation has been produced by Selig Electro-Magnetics. The amplifier

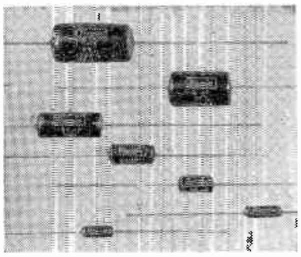


consists of a toroidal transducer with four control windings, and associated rectifiers. Numerous tappings are provided, and these and the rectifier connections are brought out to tags on an engraved front panel. The amplifier is intended for 50 V, 50-c/s operation. A typical value for the power gain is 40,000 when the device is used with auto-excitation, the output being 75 mA in 200 ohms for 1 ampere-turn input.

Selig Electro-Magnetics Ltd.,
296 *Kilburn Lane, London, W.*10.

Miniature Paper Capacitors

'Plesmin' paper capacitors for use in transistor circuits have been introduced by Plessey. The capacitance range is $0{\cdot}001\ \mu$F to $1\ \mu$F, and working voltages are normally 12, 25, 50 and 100 V. The insulation



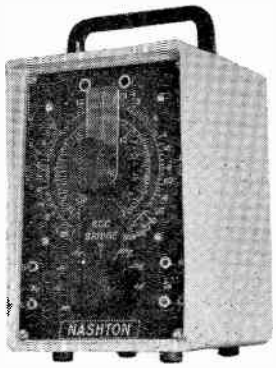
resistance at the rated working voltage is stated to be 1,000 ohms per farad.

The capacitors are tropically finished and have self-heating properties. The smallest is $\frac{3}{8}$ in. long, $\frac{1}{8}$ in. diameter, and the largest $1\frac{1}{4}$ in. by $\frac{19}{32}$ in.

The Plessey Company Ltd.,
Ilford, Essex.

R-C Comparison Bridge

A compact mains-operated bridge for measuring resistance and capacitance and comparing resistance, capacitance and inductance is available from Nash & Thompson. The resistance measurement range is 5 ohms–500 megohms, with an accuracy of $\pm 1\%$ from 100 ohms to 5 megohms. Corresponding figures for capacitance measurement are 5 pF–500 μF, $\pm 1\%$



from 100 pF to 5 μF. Balance is detected by a pair of neon lamps, which also indicate the sense of an unbalance. Measurements are made at mains frequency, the applied voltage being 18 V.

Nash & Thompson Ltd.,
Oakcroft Road, Chessington, Surrey.

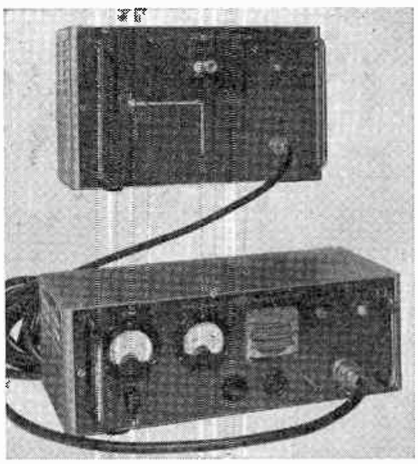
Soldering Stainless Steel

A solder paint for tinning stainless steel has been developed by Perdeck Solder Products. The paint, known as Epatam 3311 PLF, 73/11, is said to enable stainless steel to be soldered as easily as copper, and also to be suitable for tinning through iron welding scale, rust or grease, and for work on all solderable metals. It is available in 40/60 and 60/40 tin/lead combinations.

Perdeck Solder Products Ltd.,
Abbey Mills, Waltham Abbey, Essex.

20-kV Ionization Tester

The Airmec Type 209 Ionization Tester is an instrument for the non-destructive testing of insulation. A direct voltage of $2{\cdot}5$ to 20 kV is applied to the component under test. Electrical noises which occur at the onset of ionization are amplified and applied



to a loudspeaker. Direct leakage currents are displayed on a microammeter. The test voltage is displayed on another meter, so that an accurate assessment of insulation resistance can be made.

The equipment consists of an e.h.t. unit and a control unit, interconnected by a cable which carries low voltages only. The voltage-control potentiometer is spring loaded so that it returns to the minimum voltage on release.

Airmec Ltd.,
High Wycombe, Bucks.

Frequency Standard

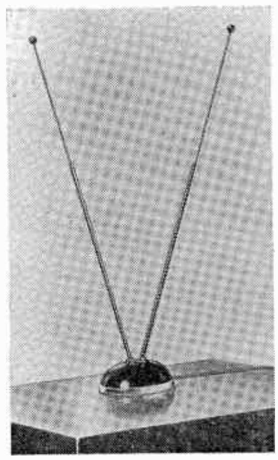
The Frequency Standard Type UE.6 consists of a 100-kc/s crystal-controlled oscillator and a chain of frequency dividers.

The crystal is housed in a thermostatically-controlled oven, and the frequency stability is quoted as 1 part in 10^6. A sine-wave output of $0{\cdot}5$ V at 100 kc/s is provided, together with rectangular-pulse outputs at 100, 10, 1 and $0{\cdot}1$ kc/s. The rise time of the pulses is $0{\cdot}1\ \mu$s (100 kc/s) or $0{\cdot}3\ \mu$s, and the output is 2 or 3 volts in 50 ohms.

G. & E. Bradley Ltd.,
Beresford House, Mount Pleasant,
Alperton, Wembley, Middx.

Multi-Band V.H.F. Aerial

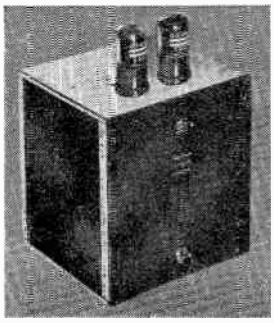
The 'Golden V' aerial illustrated is a small indoor aerial for reception of signals



in Bands I, II and III in areas of relatively high field-strength. The manufacturers state that it is suitable for both vertically-polarized and horizontally-polarized signals, and that no diplexer is required.
Belling & Lee Ltd.,
Great Cambridge Road, Enfield, Middx.

Electronic Counter

A two-decade counting unit for counting speeds up to 30 kc/s has been produced by



Millett, Levens. Vacuum-type decade counting valves (type E1T) are employed, together with monostable pulse-shaping stages.
Millett, Levens (Instruments & Engineering) Ltd.,
Stirling Corner, Barnet By-Pass, Borehamwood, Herts.

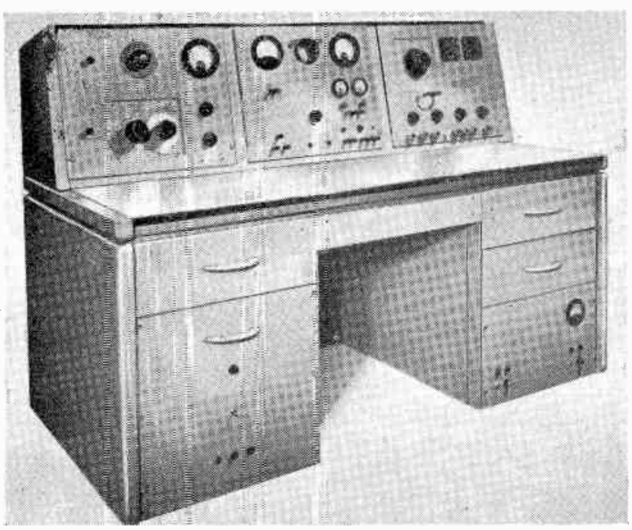
Close-Tolerance Capacitors

Paper and mica-dielectric capacitors selected to tolerances of $\pm 5\%$, $\pm 2\%$ and $\pm 1\%$, in the capacitance range $0{\cdot}001\ \mu F$ upwards, are available from Claude Lyons. The capacitors are said to have been 'aged' for a long period, so that subsequent drifts in value are small.

Decade capacitor boxes are also available, with total capacitances up to $180\ \mu F$.
Claude Lyons Ltd.,
Valley Works, Ware Road, Hoddesdon, Herts.

Fuel Gauge Test Console

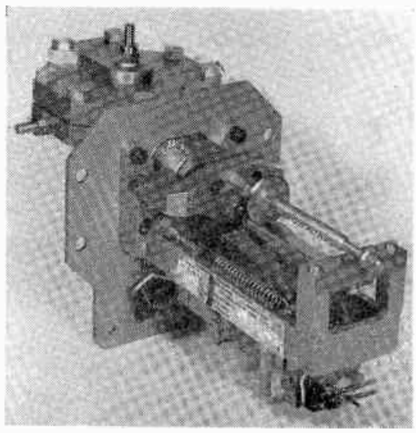
This equipment is designed for testing capacitance-operated aircraft fuel gauges. The instrument is claimed to measure capacitances of 0–1,100 pF with an accuracy of $0{\cdot}5$ pF, insulation resistance up to 200 MΩ, and to incorporate frequency standards of $33\frac{1}{3}$ kc/s ± 400 c/s and 100 kc/s.
Allied Electronics Ltd.,
28 *Upper Richmond Road, Surrey.*



Differential Pressure Transmitter

A transducer for the remote indication of differential pressures is now manufactured by Evershed and Vignoles. It is available in two versions with maximum working pressure of 500 and 1,000 p.s.i. and ranges of 0–50 and 0–20 water gauge.

The instrument employs a novel arrangement for obtaining an electrical output proportional to pressure. The movement of a pressure-sensing diaphragm is transmitted through levers to a pivoted beam. The beam carries a magnet coil which operates in the field of a pot magnet. Mechanical displacement of the beam causes a valve to deliver a restoring current to the coil. This current is therefore proportional to the pressure change.
Evershed & Vignoles Ltd.,
Acton Lane, Chiswick, London, W.4.

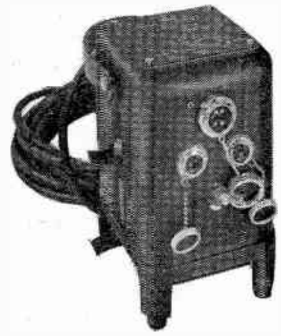


Low-Microphony Pentode for A.C./D.C. Audio Equipment

A new Mullard valve, the UF86, is a low-hum, low-microphony pentode with a heater rating of $12{\cdot}6$ volts at 100 mA. Its characteristics are said to be virtually identical with those of the audio pentode type EF86. The new valve is designed to make possible compact economical a.c./d.c. equipments having signal-to-noise ratios approaching those of the best a.c. equipments.
Mullard Ltd.,
*Mullard House, Torrington Place, London, W.C.*1.

Portable Power Transformers

Ferranti have recently introduced a range of C-core portable power transformers for 50–750 VA. A feature of the transformers is the saving in weight made possible by the



construction employed; this is stated to amount to about 50% in certain cases.
Ferranti Ltd.,
Ferry Road, Edinburgh, 5.

Seismic Instruments

The BER-62 25-trace recording oscillograph and EVS geophones (seismic transducers), as illustrated, are now being manufactured under licence by W. G. Pye & Co. Ltd., for Seismic Instruments Ltd. The

oscillograph is designed for field use, in conjunction with 25 vibration galvanometers. It is claimed that accurate timing is obtained by driving a synchronous motor from a valve-maintained tuning-fork source.
Seismic Instruments Ltd.,
Granta Works, Cambridge.

Constant-Voltage LT Transformers

A range of constant-voltage transformers with an output voltage of $6{\cdot}3$ V r.m.s. $\pm 1\%$ is announced by Advance Components Ltd. These are for 190–260 V, 50-c/s mains output. Output powers range from 8 to 50 watts.
Advance Components Ltd.,
Roebuck Road, Hainault, Ilford, Middx.

Abstracts and References

COMPILED BY THE RADIO RESEARCH ORGANIZATION OF THE DEPARTMENT OF SCIENTIFIC AND INDUSTRIAL RESEARCH AND PUBLISHED BY ARRANGEMENT WITH THAT DEPARTMENT

The abstracts are classified in accordance with the Universal Decimal Classification. They are arranged within broad subject sections in the order of the U.D.C. numbers, except that notices of book reviews are placed at the ends of the sections. U.D.C. numbers marked with a dagger (†) must be regarded as provisional. The abbreviations of journal titles conform generally with the style of the World List of Scientific Periodicals. An Author and Subject Index to the abstracts is published annually; it includes a selected list of journals abstracted, the abbreviations of their titles and their publishers' addresses. Copies of articles or journals referred to are not available from Electronic & Radio Engineer. *Application must be made to the individual publisher concerned.*

ACOUSTICS AND AUDIO FREQUENCIES

534.2 2312

Disk-Loaded Torsional Wave Delay Line: Part 1—Construction and Test.—P. Andreatch, Jr, & R. N. Thurston. (*J. acoust. Soc. Amer.*, Jan. 1957, Vol. 29, No. 1, pp. 16–19.) Two delay lines constructed of brass gave delays of 43 μs/cm-length and 114 μs/cm-length compared with 4·5 μs/cm-length for a uniform-diameter line. The former was loaded with 0·02-in.-thick disks spaced at 0·02 in. having a diameter four times that of the uniform line; the latter with 0·015-in.-thick disks spaced at 0·015 in., with a 5 : 1 diameter ratio.

534.2 2313

Disk-Loaded Torsional Wave Delay Line: Part 2—Theoretical Interpretation of Tests and Design Information.—R. N. Thurston. (*J. acoust. Soc. Amer.*, Jan. 1957, Vol. 29, No. 1, pp. 20–25.)

534.25-16-8 2314

Ultrasonic Double Refraction in Single Crystals.—P. C. Waterman & L. J. Teutonico. (*J. appl. Phys.*, Feb. 1957, Vol. 28, No. 2, pp. 266–270.) The phenomenon, methods of observation using pulse reflection, and two causes are described. Double refraction effects were observed in three Ge single crystals.

534.26 2315

Scattering in an Inhomogeneous Medium.—E. Skudrzyk. (*J. acoust. Soc. Amer.*, Jan. 1957, Vol. 29, No. 1, pp. 50–60.) "The standard mathematical procedure formally describes scattering by the superposition of a scattered pressure on the unscattered sound field. At low frequencies, because of the irregular distribution of the inhomogeneities, the phases of the scattered waves are at random and scattering is an interference phenomenon. As the frequency increases, scattering becomes highly collimated in the forward direction and the phase differences decrease to zero. At this point ray theory starts to apply. The scattered pressure, then, essentially describes only a phase change caused by the different sound velocities and the focusing and defocusing by the lens action of the patches. The medium in the neighbourhood of the receiver can be shown to contribute only by focusing, the medium farther away only by interference fluctuations. Focusing leads to normally distributed amplitude fluctuations. The distribution of the interference fluctuations, however, passes from normal to Rayleigh with increasing value of range."

534.612 2316

Acoustic Radiation Pressure.—P. J. Westervelt. (*J. acoust. Soc. Amer.*, Jan. 1957, Vol. 29, No. 1, pp. 26–29.)

534.613 2317

The Torque of an Infinite Strip Exposed to Plane Sound Waves.—H. Levine. (*Proc. Camb. phil. Soc.*, Jan. 1957, Vol. 53, Part 1, pp. 234–247.)

534.7 2318

Inapplicability of the Threshold Concept to Detection of Signals in Noise.—R. R. McPherson. (*J. acoust. Soc. Amer.*, Jan. 1957, Vol. 29, No. 1, p. 151.) Brief discussion of the concept which considers that a response occurs only if the variable is larger than a certain magnitude.

534.7 2319

Some Results of Research on Speech Perception.—A. M. Liberman. (*J. acoust. Soc. Amer.*, Jan. 1957, Vol. 29, No. 1, pp. 117–123.) "Recent experiments with synthetic speech have succeeded in isolating some of the acoustic cues which underlie the perception of speech. This paper describes, and attempts to interpret, some of the research in that area."

534.78 2320

Information Conveyed by Vowels.—P. Ladefoged & D. E. Broadbent. (*J. acoust. Soc. Amer.*, Jan. 1957, Vol. 29, No. 1, pp. 98–104.)

534.78 2321

Acoustic Properties of Stop Consonants.—M. Halle, G. W. Hughes & J. P. A. Radley. (*J. acoust. Soc. Amer.*, Jan. 1957, Vol. 29, No. 1, pp. 107–116.)

534.844.1 2322

A Reverberation-Time Meter.—G. Odin. (*Hochfrequenztech. u. Elektroakust.*, Nov. 1956, Vol. 65, No. 3, pp. 86–91.) The instrument described automatically measures and records the decay time of sound in rooms; tests can be made repetitively.

534.846 2323

Acoustics at the Rochester (New York) War Memorial Auditorium.—B. Olney & R. S. Anderson. (*J. acoust. Soc. Amer.*, Jan. 1957, Vol. 29, No. 1, pp. 94–98.) This arena-type auditorium has a volume of $3 \cdot 24 \times 10^6$ cu. ft. and a reverberation time of 2·5 s at 500 c/s when unoccupied, and 1·9 s with two-thirds of the audience. Design features are discussed.

534.846.6 **2324**

Investigation of the Acoustic Shock Waveform Used for Pulse Measurement of Room Acoustics.—H. Niese. (*Hochfrequenztech. u. Elektroakust.*, Nov. 1956, Vol. 65, No. 3, pp. 98–108.) The conditions are examined which must be fulfilled by the sound source and the pulse shape so that consistent results are achieved for the measurement of audibility at any test point, on a subjective basis. The directivity of a sound source reproducing speech was determined to find a means of measuring intelligibility. A loudspeaker specially designed to meet the requirements is shown (see also 3267 of 1956).

534.846.6.001.57 **2325**

Electroacoustic Measurements on Room Models.—W. Kraak. (*Hochfrequenztech. u. Elektroakust.*, Nov. 1956, Vol. 65, No. 3, pp. 91–98.) The assumptions necessary for measurements on a model are investigated with reference to the various conditions which have to be simulated, such as wall absorption and reflection, the presence of an audience, etc. Methods using sound pulses with a frequency spectrum around 1 kc/s extending over about 3 octaves appear most suitable. See also 2324 above.

621.395.612.45:1 **2326**

Bigradient Uniaxial Microphone.—H. F. Olson, J. Preston & J. C. Bleazey. (*RCA Rev.*, Dec. 1956, Vol. 17, No. 4, pp. 522–533.) Two uniaxial microphones (see 2199 of 1953) are connected in series opposition to form a second-order unidirectional microphone. The improved directional efficiency of this assembly makes it usable for a pickup distance three times that of a nondirectional microphone.

621.395.616 : 534.612.4 **2327**

Apparatus for the Calibration of Condenser Microphones.—A. Bressi. (*Alta Frequenza*, Dec. 1956, Vol. 25, No. 6, pp. 505–519.) Description of equipment for the absolute calibration of microphones by the pistonphone and reciprocity methods.

621.395.625.3 : 655.3 **2328**

Printed Magnetic Recording Tapes.—H. Schiesser. (*Elektrotech. Z.*, *Edn B*, 21st Dec. 1956, Vol. 8, No. 12, pp. 473–475.) A method of copying tapes is outlined in which variable-area recording based on the boundary-displacement process [1963 of 1952 (Daniels)] is used. Prints are made in colour containing suspended magnetic particles and the recording can be reproduced photoelectrically. After being magnetized to saturation the tape can be used for magnetic reproduction. Advantages and difficulties are briefly discussed; economic mass production of magnetic tape recordings may thus be feasible.

621.395.625.3 : 778.5 **2329**

Magnetic 16-mm Single-System Sound-on-Film Recording Camera Equipment.—W. Bach, E. M. Berndt, A. N. Brown & R. L. George. (*J. Soc. Mot. Pict. Telev. Engrs*, Nov. 1956, Vol. 65, No. 11. pp. 603–605. Discussion, p. 605.)

681.84 : 534.851 **2330**

Limiting Factors in Gramophone Reproduction.—D. A. Barlow. (*Wireless World*, May & June 1957, Vol. 63, Nos. 5 & 6, pp. 228–230 & 290–294.) Considers the deformation and wear of groove walls, the mechanical design of the pickup, types of styli and tracing distortion with particular reference to playing speeds and recording characteristics.

681.892 : 61 **2331**

Acoustic Mapping within the Heart.—J. D. Wallace, J. R. Brown, Jr, D. H. Lewis & G. W. Deitz. (*J. acoust. Soc. Amer.*, Jan. 1957, Vol. 29, No. 1, pp. 9–15.) A miniature barium-titanate transducer in conjunction with amplifier and recording equipment was used for intracardiac acoustic mapping. The equipment and some of the results obtained are described.

AERIALS AND TRANSMISSION LINES

621.3.091 **2332**

The Definition of Cymomotive Force.—G. Barzilai. (*Piccole Note Ist. super. Poste e Telecomunicazioni*, Nov./Dec. 1956, Vol. 5, No. 6, pp. 796–798.) The use of the term cymomotive force (c.m.f.) for defining the properties of radiating systems in free space has been adopted by the C.C.I.R. The definition given is that contained in Document 386-E of 24th May 1956 presented at the 8th Plenary Assembly of the C.C.I.R. in Warsaw, 1956, to point out that the c.m.f. is independent of frequency. For an application of this concept, see 29 of January (Micheletta) where it is termed 'radiative force'.

621.315.2.013.78 : 621.317.332 **2333**

The Measurement of the Coupling [surface-transfer] **Impedance of Cable Screens at High Frequencies.**—H. Jungfer. (*Nachrichtentech. Z.*, Dec. 1956, Vol. 9, No. 12, pp. 553–560.) General equations are derived and a suitable test arrangement is described. A comparison of results obtained by alternative methods in the frequency range 10–1 000 Mc/s shows satisfactory agreement and an approximately linear increase of transfer impedance with frequency.

621.315.212 **2334**

Screening Effect of the Outer Conductors of Flexible Coaxial Cables.—L. Krügel. (*Telefunken Ztg*, Dec. 1956, Vol. 29, No. 114, pp. 256–266. English summary, p. 293.) Tests were made on cables with braided, single and double layer-wound screens to assess the effect of the number of wires or strands, their thickness and lay. Results are given in graphical form and discussed in detail.

621.372.2 : 621.317.34 **2335**

Investigations on Helical Lines.—K. Hübener. (*Nachrichtentech. Z.*, Dec. 1956, Vol. 9, No. 12, pp. 581–584.) Inhomogeneities, such as kinks, bends and variations in pitch, are treated as lossless or lossy quadripoles. The reflection coefficients are calculated from measurements made by means of the apparatus described.

621.372.2.029.6 **2336**

Theory of Twin-Helix Coaxial Line.—V. S. Mikhalevski. (*Radiotekhnika i Elektronika*, Oct. 1956, Vol. 1, No. 10, pp. 1309–1316.) The dispersion in twin-helix lines is calculated, assuming perfect conductors and the possibility of substituting for the helices equivalent cylindrical surfaces conducting only along a helical path. Helices wound in the same sense and in opposite senses are considered.

621.372.8 **2337**

Propagation Characteristics of Low-Loss Tubular Waveguides.—H. E. M. Barlow & H. G. Effemey. (*Proc. Instn elect. Engrs*, Part B, May 1957, Vol. 104, No. 15, pp. 254–260.) The propagation of the H_{01} mode in straight lengths of circular waveguide was studied at a frequency of 35 kMc/s. Lengths of copper and aluminium tube manufactured to commercial tolerances were found to give attenuations about 30% above the theoretical values. Launching arrangements are discussed, and possible applications of the method to trunk communication and high-power transmission are pointed out.

621.372.8 **2338**

A New Type of Waveguide with Diaphragms.—R. G. Mirimanov & G. I. Zhileiko. (*Radiotekhnika i Elektronika*, Oct. 1956, Vol. 1, No. 10, pp. 1374–1377.) Coaxial waveguides with diaphragms are discussed theoretically.

621.372.8 **2339**

A New Form of Hybrid Junction for Microwave Frequencies.—P. D. Lomer & J. W. Crompton. (*Proc. Instn elect. Engrs*, Part B, May 1957, Vol. 104, No. 15, pp. 261–264.) Performance details are given for a branch-waveguide directional coupler in which the voltage coupling coefficients of the branch waveguides are proportional to the coefficients in a binomial expansion.

621.372.8 : 621.318.134 **2340**

Ferrite Slabs in Transverse-Electric-Mode Waveguide.—H. Seidel. (*J. appl. Phys.*, Feb. 1957, Vol. 28, No. 2, pp. 218–226.) A qualitative description of the nonreciprocal properties of ferrite-loaded waveguides is developed showing that a suitably loaded coaxial line may have nonreciprocal properties. The treatment is extended to gyromagnetic resonance.

621.396.67 **2341**

Calculation of the Current along a Cylindrical Aerial.—R. Dematte. (*Ann. Télécommun.*, Dec. 1956, Vol. 11, No. 12, pp. 280–287.) The author details the calculation of the numerical results used to illustrate the method developed by Poincelot (980 of 1956) who introduces the paper.

621.396.677.029.6 **2342**

The Gain of Highly Directive Aerials Used for Scatter Propagation.—A. Chinni & G. C. Corazza. (*Piccole Note Ist. super. Poste e Telecomunicazioni*, Nov./Dec. 1956, Vol. 5, No. 6, pp. 804–809.) A formula is derived on the basis of Booker & Gordon's theory of tropospheric scattering (1757 of 1950) for the change in gain as a function of aperture angle and effective scattering area. The importance of the law assumed for the distribution with height of local fluctuations of the dielectric constant is discussed.

621.396.677.029.64 **2343**

Microwave Helical Aerials.—T. G. Hame. (*Electronic Engng*, April 1957, Vol. 29, No. 350, pp. 181–183.) Dielectric-enclosed helical aerials are effective up to 10 000 Mc/s if the design allows for the detuning effect of the dielectric.

621.396.677.32 **2344**

Reduction of Side Lobes in Directional Aerials.—S. Manczarski. (*Archiwum Elektrotech.*, 1956, Vol. 5, No. 2, pp. 325–337. English summary, pp. 340–341.) Discussion of rhombic, broadside and end-fire arrays for s.w. communication followed by a description of the Type-j/8DD/1 end-fire array designed for the Polish Radio. The array comprised eight double-dipole elements spaced at λ/4 with a 90° phase difference between adjacent elements. A radiation polar diagram is given.

AUTOMATIC COMPUTERS

681.142 **2345**

A High-Speed Data Processing System.—M. L. Klein, R. B. Rush & H. C. Morgan. (*Electronic Engng*, April 1957, Vol. 29, No. 350, pp. 158–163.) Data from 20–100 independent channels, at a full-scale level of 100 mV, are taken at a rate of 100 kc/s and recorded digitally on a magnetic tape together with the source information and time.

681.142 **2346**

The 'Bizmac' Digital Data Processing System.—J. C. Hammerton. (*Electronic Engng*, April 1957, Vol. 29, No. 350, pp. 174–180.) As used at the central agency for American military supply depots.

681.142 **2347**

500 000 000-Bit Random-Access Memory.—G. E. Comstock. (*Instrum. & Automation*, Nov. 1956, Vol. 29, No. 11, pp. 2208–2211.) The machine described combines mechanical and electrical techniques to store and have accessible within less than a second 64 × 10^6 8-bit characters. Magnetic recording tape about 60 miles long is used as storage medium.

681.142 **2348**

Possibilities and Developments of the Method of Rheoelectric Analogues.—L. Malavard. (*Onde élect.*, Oct. & Dec. 1956, Vol. 36, Nos. 355 & 357, pp. 829–837 & 1046–1052.) Continuation and conclusion of a paper published in the computer issue ibid., Aug./Sept. 1956 (see 997 of April). A detailed review of the principles and experimental techniques with 200 references.

681.142 **2349**

Application of Computers in Automatic Systems.—A. A. Fel'dbaum. (*Avtomatika i Telemekhanika*, Nov. 1956, Vol. 17, No. 11, pp. 1046–1056.) A survey. 26 references including several to Russian literature.

681.142 : 517.948.34 **2350**

On the Continuous Solution of Integral Equations by an Electronic Analogue: Part 1.—M. E. Fisher. (*Proc. Camb. phil. Soc.*, Jan. 1957, Vol. 53, Part 1, pp. 162–174.) "A scheme is proposed for solving a class of integral equations by electronic analogue computing techniques in times as short as one-tenth of a second. The scheme utilizes a recently developed high-speed analogue function store for carrying out a special iterative procedure which is shown to be more efficient than the classical Neumann process. The problem of the kernel generation at high repetition rates is considered and a novel method based on pivotal function generators is described. Likely errors are analysed and an overall accuracy of the order of 1% is shown to be attainable with known techniques."

681.142 : 621.318.57 **2351**

Counters Select Magnetic Drum Sectors.—A. J. Strassman & R. E. King. (*Electronics*, 1st April 1957, Vol. 30, No. 4, pp. 161–163.) A test instrument for 'writing' a predetermined binary pattern on any selected sector of a magnetic drum memory system.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.049.75 : 621.396.6.002.2 **2352**

The 'Asmodulor' Process, a New Assembly Technique for Electronic Equipment.—M. L. Paternault. (*Onde élect.*, Dec. 1956, Vol. 36, No. 357, pp. 1031–1039.) Description of the modular design technique [see 693 of 1954 (Henry & Rayburn)] developed in the U.S.A., with particular reference to its adoption by the French electronics industry for mechanized mass production.

621.372 : 681.142 **2353**

Theory on Nonlinear Operational Elements using a Piecewise Linear Approximation.—B. Ya. Kogan. (*Avtomatika i Telemekhanika*, Dec. 1956, Vol. 17, No. 12, pp. 1081–1091.) A function generator is considered to be an operational amplifier with nonlinear conductances in the feedback circuit. Fundamental relations are derived and examples are given of the synthesis of a function generator using diode-circuit elements.

621.372.4 **2354**

Novel Method of Realizing Two-Pole Functions by Canonical Circuits or Circuits without Coupling Elements.—R. Unbehauen. (*Nachrichtentech. Z.*, Dec. 1956, Vol. 9, No. 12, pp. 565–572.) An outline is given of a generalized form of Brune's method of network synthesis by canonical circuits, which permits a reduction in transformer elements. By means of another method transformerless networks can be realized which have fewer reactive elements than those given by Bott & Duffin (1108 of 1951). See also 47 of 1955 (Reza).

621.372.41 : 621.372.8 **2355**

Resonator in the Form of a Cut-Off Waveguide.—I. V. Lebedev & E. M. Guttsait. (*Radiotekhnika i Elektronika*, Oct. 1956, Vol. 1, No. 10, pp. 1303–1308.) The input impedance of a uniform waveguide operating below cut-off frequency is calculated. The feasibility of synthesizing a resonator, with characteristics similar to a parallel tuned circuit, from the waveguide and a reactive diaphragm is shown. The resonator Q is low.

621.372.413 **2356**

The Use of Lossy Material to Suppress Unwanted Modes in Cavity Resonators.—M. Y. El-Ibiary & J. Brown. (*Proc. Instn elect. Engrs*, Part C, March 1957, Vol. 104, No. 5, pp. 25–34.) A ring of lossy material in a groove at the junction between an end wall and the cylindrical wall does not affect the H_{01n} mode but may suppress others. Analysis is given and confirmed by measurements.

621.372.413 : 621.318.134 **2357**

An Exact Solution for a Cylindrical Cavity containing a Gyromagnetic Material.—H. E. Bussey & L. A. Steinert. (*Proc. Inst. Radio Engrs*, Part 1, May 1957, Vol. 45, No. 5, pp. 693–694.)

621.372.5 **2358**

The Equivalent Transmission Line of a Linear Four-Terminal Network. Calculations with Cascade-Connected Four-Terminal Networks.—C. G. Aurell. (*Ericsson Tech.*, 1956, Vol. 12, No. 2, pp. 107–145.) The transmission-line analogy is applied to nonsymmetrical and nonreciprocal quadripoles (see also 2861 of 1955). Different formulae are derived without the use of exponential or hyperbolic functions. Steady-state conditions are assumed.

621.372.5 **2359**

Synthesis of Transmission Systems in Terms of Tandem-Connected Quadripoles.—P. W. Seymour & S. Døssing. (*Proc. Instn elect. Engrs*, Part C, March 1957, Vol. 104, No. 5, pp. 62–80.) Theoretically derived formulae aid the rapid solution of some practical transmission problems.

621.372.5 : 621.376.3 : 621.3.018.78 **2360**

Frequency-Modulation Distortion in Linear Networks.—A. S. Gladwin: R. F. Brown. (*Proc. Instn elect. Engrs*, Part B, May 1957, Vol. 104, No. 15, p. 264.) Comment on 1581 of May and author's reply.

621.372.512 **2361**

A Method for the Approximate Determination of the Impulse Response of a Number of Identical Circuits in Cascade.—K. F. Sander. (*Proc. Instn elect. Engrs*, Part C, March 1957, Vol. 104, No. 5, pp. 13–24.) The method of steepest descents is used to obtain approximate analytical expressions for various values of time. These are tested against two networks for which exact solution is possible.

621.372.54 **2362**

Outline of a Generalized Filter Theory.—E. Henze. (*Arch. elekt. Übertragung*, Dec. 1956, Vol. 10, No. 12, pp. 541–551.) This treatment of ideal linear filters by the use of function-space techniques is based on the work of Zadeh & Miller (2147 of 1952). Examples of practical applications are briefly discussed.

621.372.54 **2363**

New Types of Sections for Zig-Zag Filters.—T. Laurent. (*Ericsson Tech.*, 1956, Vol. 12, No. 2, pp. 147–164.) Continuation of an earlier paper (83 of 1954). Zig-zig and zag-zag sections, in which both attenuation peaks lie above or below the pass-band, respectively, instead of having a peak on either side of the band, are derived and their application is discussed.

621.372.54 **2364**

Practical Methods of Formulating the Hurwitz Polynomial in Filter Synthesis.—F. Bauhuber. (*Nachrichtentech. Z.*, Dec. 1956, Vol. 9, No. 12, pp. 573–580.) The application of several methods [see e.g. 3679 of 1955 (Bauer)] is discussed and a new direct iteration method is described with numerical examples. 21 references.

621.372.54 **2365**

Quartz Crystal Filters with Sharp Cut-Off and Large Bandwidth in Branch Networks.—W. Poschenrieder. (*Nachrichtentech. Z.*, Dec. 1956, Vol. 9, No. 12, pp. 561–565.) Ladder-type networks of improved efficiency and an example of their application at about 100 kc/s are briefly described.

621.372.54 **2366**

Tables of Frequency Transformation and Band-Pass Filters.—H. Weber & J. Martony. (*Tech. Mitt. schweiz. Telegr.-Teleph. Verw.*, 1st Dec. 1956, Vol. 34, No. 12, pp. 499–502. In French and German.) The tables are based on Laurent's theory of combined impedance and frequency transformation (see 723 of March). The characteristics and design parameters of band-pass filters half-sections are tabulated.

621.372.543.2 **2367**

Figure of Merit of Band-Pass Filters with a Tchebycheff Characteristic.—J. Lenkowski. (*Archiwum Electrotech.*, 1956, Vol. 5, No. 2, pp. 365–373. English summary, pp. 375–377.)

621.372.543.2 **2368**

Filter Circuits with Two Coupled Resonators for Wide Relative Pass Bands.—F. Carassa. (*Alta Frequenza*, Dec 1956, Vol. 25, No. 6, pp. 451–481.) The amplitude and phase characteristics of double-tuned filters are examined for the case where the pass-band width is great relative to the mid-frequency.

621.372.543.2 **2369**

A Survey of the Design of Lossy Filters using the Insertion-Loss Method with Special Reference to 'Zig-Zag' Band-Pass Filters.—C. Kurth. (*Frequenz*, Dec. 1956, Vol. 10, No. 12, pp. 391–396, Jan. & Feb. 1957, Vol. 11, Nos. 1 & 2, pp. 12–19 & 43–53.)

621.372.543.2 **2370**

Design of Three-Resonator Dissipative Band-Pass Filters having Minimum Insertion Loss.—J. J. Taub & B. F. Bogner. (*Proc. Inst. Radio Engrs*, Part 1, May 1957, Vol. 45, No. 5, pp. 681–687.) Universal design curves are given.

621.373 : 621.316.729 **2371**

Theory of Synchronization of Nonsinusoidal Oscillations.—I. I. Minakova & K. F. Teodorchik. (*Radiotekhnika i Elektronika*, Oct. 1956, Vol. 1, No. 10, pp. 1317–1324.)

621.373.029.42 **2372**

A Two-Phase Low-Frequency Oscillator: Part 1.—E. F. Good. (*Electronic Engng*, April 1957, Vol. 29, No. 350, pp. 164–169.) Two series-connected integrating circuits with overall feedback give oscillations of stable amplitude and low harmonic content. Two outputs are available with 90° phase difference.

621.373.029.62/.63 **2373**

Generation of Electromagnetic Oscillations by means of a Travelling-Wave Valve with a Twin-Helix Coaxial Line.—V. S. Mikhalevski, A. G. Dolganov & V. D. Ivanova. (*Radiotekhnika i Elektronika*, Nov. 1956, Vol. 1, No. 11, pp. 1383–1393.) An experimental verification of theoretical results (2336 above) is reported. Results indicate that the theory may be used for approximate calculations.

621.373.029.64 : 538.569.4 **2374**

Maser Oscillators.—Helmer. (See 2431.)

621.373.421 : 621.376.3 **2375**

Frequency-Modulated Quartz Oscillators for Broadcasting Equipment.—W. S. Mortley. (*Proc. Instn elect. Engrs*, Part B, May 1957, Vol. 104, No. 15, 239–249. Discussion, pp. 249–253.) "An f.m. system is described which employs a directly-frequency-modulated quartz-crystal oscillator, the design of the circuit and of the crystal plate being treated in some detail."

621.373.43 **2376**

Nonsinusoidal Oscillations. Investigation of Continuous Solutions.—L. Sideriades. (*C. R. Acad. Sci., Paris*, 4th March 1957, Vol. 244, No. 10, pp. 1330–1333.) Experimental verification of analysis outlined in 1369 of May.

621.373.43 **2377**

Controllable Relaxation Oscillator using a Glow Discharge Tube.—S. V. Svechnikov. (*Avtomatika i Telemekhanika*, Nov. 1956, Vol. 17, No. 11, pp. 1029–1034.) A simple neon-lamp relaxation oscillator with a linear relation between the output frequency and the input voltage is described. In the particular case considered a CdS photoresistor control element was used.

621.373.43 **2378**

A Square-Wave Converter with Feedback Control of Mark-to-Space Ratio.—J. B. Earnshaw. (*Electronic Engng*, April 1957, Vol. 29, No. 350, pp. 170–173.) Operates with sinusoidal, sawtooth or triangular input waveforms.

621.373.431.2 **2379**

The Blocking Oscillator.—(*Wireless World*, June 1957, Vol. 63, No. 6, pp. 285–289.) An analysis of the operation of this type of circuit.

621.373.431.2 **2380**

Millimicrosecond Blocking Oscillators.—J. M. Smith. (*Electronic Engng*, April 1957, Vol. 29, No. 350, pp. 184–186.) A circuit designed to deliver pulses up to 200 V amplitude to a low-impedance load such as a coaxial cable.

621.373.444 : 621.314.7 **2381**

Investigation of Transient Processes in a Point-Contact Semiconductor-Triode Trigger Circuit and the Forming of Pulses from a Sinusoidal Voltage.—V. A. Kuz'min. (*Radiotekhnika i Elektronika*, Nov. 1956, Vol. 1, No. 11, pp. 1406–1412.) The transistor switching circuit previously considered by Lebov & Baker (2592 of 1954) is further analysed. A method is presented of calculating the leading edge of the output pulses and part of the trailing edge corresponding to the transition of the circuit through the active region under the action of a negative voltage jump. The forming of pulses, using a sinusoidal input voltage, is considered and results of an experimental verification of the theory are briefly reported.

621.373.444.1 : 621.397.621 **2382**

C.R.T. Deflection Circuit has High Efficiency.—W. B. Guggi. (*Electronics*, 1st April 1957, Vol. 30, No. 4, pp. 172–175.) A transistor timebase generator for magnetic deflection. A technique of switching the paths followed by circulating currents reduces power consumption to one third.

621.373.52 **2383**

Junction-Transistor Bootstrap Linear-Sweep Circuits.—K. P. P. Nambiar & A. R. Boothroyd. (*Proc. Instn elect. Engrs*, Part B, May 1957, Vol. 104, No. 15, pp. 293–306. Discussion, 333–336.) The principles are discussed and analysed. A basic circuit employing two *p-n-p* junction transistors is given, and improvement of linearity by feedback is considered. Monostable and astable forms, and a blocking-oscillator comparator for precision timing are given. Sweep amplitudes are small, but larger values should be possible using existing transistors, with sweep durations $< 1\ \mu s$.

621.373.52 : 621.376.3 2384

Semiconductor-Triode High-Frequency *RC* Oscillator.—L. N. Kaptsov. (*Radiotekhnika i Elektronika*, Nov. 1956, Vol. 1, No. 11, pp. 1413–1418.) Analysis is presented of a circuit generating almost harmonic oscillations at frequencies of the order of the critical frequency. Experimental results indicate the feasibility of frequency modulating the oscillations.

621.374.32 2385

A Decade Pulse Counting System with Circulating Storage in a Delay Line.—D. Maeder. (*Helv. phys. Acta*, 15th Dec. 1956, Vol. 29, Nos. 5/6, pp. 459–462. In German.) Brief description of equipment which uses a Ni-wire delay line; the display is in the form of a raster on a c.r. tube screen.

621.374.33 : 531.76 2386

Gate Tube Generates Interleaved Pulse Chain.—D. Kushner. (*Electronics*, 1st April 1957, Vol. 30, No. 4, pp. 186–187.) A Type-6BN6 gated beam valve is employed to generate spaced pulses by means of ringing circuits at the anode and second grid.

621.375.012 2387

An Extension of the Noise-Figure Definition.—H. A. Haus & R. B. Adler. (*Proc. Inst. Radio Engrs*, Part 1, May 1957, Vol. 45, No. 5, pp. 690–691.) An extension of previous analysis (see *Convention Record Inst. Radio Engrs*, 1956, Vol. 4, Part 2, pp. 53–67). A new concept, the exchangeable power of a source, is suggested to overcome difficulties which arise in the use of the notion of available power when negative resistance is involved.

621.375.121.2 : 621.396.621.22 2388

Distributed Amplifiers as Antenna Multicouplers.—E. T. Pfund, Jr. (*Electronics*, 1st April 1957, Vol. 30, No. 4, pp. 176–179.) Describes tests of a Telefunken equipment employing push-pull distributed amplifiers to feed up to six receivers from one aerial. This principle gives less intermodulation than conventional systems.

621.375.132 : 621.396.822 2389

Noise in Negative Feedback Amplifiers.—C. N. W. Litting. (*Electronic Radio Engr*, June 1957, Vol. 34, No. 6, pp. 219–223.) The improvement in signal/noise ratio obtained by the use of feedback is discussed and the conditions under which it is most effective are deduced.

621.375.2 : 621.3.018.78 2390

Grid-Circuit Distortion.—E. Watkinson. (*Electronic Radio Engr*, June 1957, Vol. 34, No. 6, pp. 207–214.) The effects of control-grid currents between 10^{-6} and 10^{-9} A are analysed. It is shown that recommended operating conditions may lead to grid-circuit distortion comparable with that produced in the anode circuit.

621.375.2 : 621.317.733 2391

Intermodulation in Bridge Detector Amplifiers.—G. J. Johnson & A. M. Thompson. (*Proc. Instn elect. Engrs*, Part C, March 1957, Vol. 104, No. 5, pp. 217–221.) A discussion of the effects of intermodulation and methods for its reduction, with special reference to a practical amplifier for use at 1 and 1·6 kc/s.

621.375.4.018.7 2392

Nonlinear Distortion in Transistor Amplifiers at Low Signal Levels and Low Frequencies.—N. I. Meyer. (*Proc. Instn elect. Engrs*, Part C, March 1957, Vol. 104, No. 5, pp. 208–216.) An analytical method is presented for calculating the nonlinear distortion of sinusoidal signals in the input and output circuits. A considerable reduction of the output distortion at low frequencies in the simple common-emitter amplifier can be obtained by arranging that the input and output distortions cancel each other. The expressions derived have been verified experimentally.

621.375.4.029.5 2393

Voltage Amplification of Point-Contact Semiconductor Triodes in Tuned Amplifiers.—E. F. Vorob'eva. (*Radiotekhnika i Elektronika*, Nov. 1956, Vol. 1, No. 11, pp. 1394–1405.) The amplification factor of the grounded-base tuned amplifier at frequencies up to $1·5f_\alpha$ is expressed by $K = SR_n$, where R_n is the collector load impedance at resonance and S is a parameter calculated in terms of the l.f. parameters of the transistor and f_α. Curves are given for calculating the amplification at frequencies up to $1·5f_\alpha$ with various feedback and collector load factors. The experimental verification on four transistors shows fair agreement with calculations.

GENERAL PHYSICS

53 2394

Replacement of a Nonstationary Random Process by a Stationary Process.—V. I. Tikhonov. (*Zh. tekh. Fiz.*, Sept. 1956, Vol. 26, No. 9, pp. 2057–2059.) The nonstationary random process $\eta(t) = A\xi(t) \sin(\omega t + \phi)$, where $\xi(t)$ is a stationary random process is considered. Expressions are given for the correlation function for $\eta(t)$, using (*a*) statistical means and (*b*) a mean over a period ξ/ω. The conditions in which (*b*) can be used and their physical meaning are discussed. Application to fluctuations in a valve oscillator is considered.

530.1 2395

On a Solution of Field Equations in Einstein's Unified Field Theory: Part 1.—N. N. Ghosh. (*Progr. theor. Phys.*, Nov. 1956, Vol. 16, No. 5, pp. 421–428.)

530.145.6 : 538.566 2396

Multiple Scattering by Quantum-Mechanical Systems.—K. M. Watson. (*Phys. Rev.*, 15th Feb. 1957, Vol. 105, No. 4, pp. 1388–1398.) In addition to a general treatment of the problem, specific calculations are made of the refractive index of a medium which is 'polarized' by the scattered particle and also of a medium which has correlated structure.

531.3 : 621.396.822 2397

Influence of Non-normal Fluctuations on Linear Systems.—V. I. Tikhonov & A. A. Tolkachev. (*Bull. Acad. Sci. U.R.S.S., tech. Sci.*, Dec. 1956, No. 12, pp. 48-56. In Russian.) The influence of fluctuations is considered, the characteristics of which do not appreciably differ from a normal (e.g. Rayleigh-type) characteristic.

533.7 : 537.56 2398

A Relativistic Form of Boltzmann's Transport Equation in the Absence of Collisions.—P. C. Clemmow & A. J. Willson. (*Proc. Camb. phil. Soc.*, Jan. 1957, Vol. 53, Part 1, pp. 222-225.)

534.01 2399

On an Error in the Theoretical Analysis of Ferroresonance Phenomena at a Frequency Equal to One Third of the Frequency of the Applied Force.—A. N. Vakhrameev. (*Zh. tekh. Fiz.*, Aug. 1956, Vol. 26, No. 8, p. 1862.) Comment on paper by Hayashi (2953 of 1953).

534.01 + 538.56] : 517.9 2400

On the Short-Wave Asymptotic Theory of the Wave Equation $(\nabla^2 + k^2)\phi = 0$.—F. Ursell. (*Proc. Camb. phil. Soc.*, Jan. 1957, Vol. 53, Part 1, pp. 115–133.) "The present work appears to be the first practical and rigorous solution of a short-wave problem in optics or acoustics when a solution in closed form is not available. It is suggested that the technique (suitably combined with formal expansions) may be applicable to a wider class of radiation and diffraction problems."

535.215 + 537.533 2401

On the Additivity of Photo-emission and Secondary-Electron Emission of Metals.—L. I. Ekertova. (*Zh. tekh. Fiz.*, Aug. 1956, Vol. 26, No. 8, pp. 1665–1668.) Contrary to the assertion of a number of authors that the two effects are non-additive, it is shown theoretically, as well as experimentally, that this is not the case.

535.222 2402

A New Effort to Measure the Velocity of Light.—R. Gerharz. (*J. Electronics*, March 1957, Vol. 2, No. 5, pp. 416–424.) A preliminary report is given of apparatus being developed to measure the group velocity of light pulses originating from the dynodes of an electron multiplier as visible fluorescent radiation.

535.223 : 621.372.413 2403

Accuracy of a Microwave Resonant-Cavity Measurement of the Velocity of Light.—D. H. Janney. (*Phys. Rev.*, 15th Feb. 1957, Vol. 105, No. 4, pp. 1138–1140.) Unless the surface reactance and resistance of the cavity are known, the uncertainty of the measurement is limited to a value greater than 2 parts in 10^6.

535.34 : 535.37 2404

Universal Relation between the Absorption and Luminescence Spectra of Complex Molecules.—B. I. Stepanov. (*C. R. Acad. Sci. U.R.S.S.*, 11th Feb. 1957, Vol. 112, No. 5, pp. 839–841. In Russian.)

535.376 **2405**

Build-Up of Electroluminescent Brightness.—C. H. Haake. (*J. appl. Phys.*, Feb. 1957, Vol. 28, No. 2, pp. 245–250.) This build-up is compared with that of photoluminescence under various conditions affecting the electron population in traps. A qualitative explanation is proposed for the mechanism.

537.2/.3 : 621.3.013 **2406**

The Approximate Solution of Electric-Field Problems with the Aid of Curvilinear Nets.—L. Tasny-Tschiassny. (*Proc. Instn elect. Engrs*, Part C, March 1957, Vol. 104, No. 5, pp. 116–129.)

537.31 **2407**

The Length of the Free Path of an Electron in Liquid and Amorphous Conductors [and semiconductors].—A. I. Gubanov. (*Zh. tekh. Fiz.*, Aug. 1956, Vol. 26, No. 8, pp. 1651–1656.) The free path of an electron is determined taking into account the scattering of electrons at defects of the quasi-crystalline structure and the effect of thermal oscillations.

537.311 **2408**

Oscillographic Determination of the Energy of the Electric Explosion of Wires.—I. F. Kvartskhava, V. V. Bondarenko, A. A. Plyutto & A. A. Chernov. (*Zh. eksp. teor. Fiz.*, Nov. 1956, Vol. 31, No. 5(11), pp. 745–751.) An oscillographic method is described. Results indicate that at high voltages the resistance of the wire is not determined uniquely by the magnitude of the energy introduced.

537.311.1 **2409**

Transport Processes in Conductors taking into account Nonlinear Effects.—V. P. Shabanski. (*Zh. eksp. teor. Fiz.*, Oct. 1956, Vol. 31, No. 4(10), pp. 657–671.) Relaxation processes in the electron-lattice system are analysed and equations are obtained for the heating of the electron gas in strong electric fields. The kinetic equations are solved for particular cases and galvanomagnetic and thermoelectric phenomena are discussed. The theory is applicable to semiconductors in the region of conduction electron degeneracy.

537.311.1 : 536.3 : 621.396.822 **2410**

Thermal Fluctuations in Conductors.—N. L. Balazs. (*Phys. Rev.*, 1st Feb. 1957, Vol. 105, No. 3, pp. 896–899.) The spectrum of current fluctuations is derived for a conductor in thermal equilibrium with the surrounding radiation field, in terms of the conductor's absorptive power. For low frequencies the result reduces to Nyquist's relation, but for high frequencies the fluctuations are proportional to the skin resistance and depend on the shape of the conductor.

537.311.31 **2411**

On the Connection between Electrical Resistivity and Potential Energy in Pure Metals.—G. Borelius. (*Ark. Fys.*, 1956, Vol. 11, Part 3, pp. 291–294.) Results suggest that the zero-point energy, which at zero temperature makes no contribution to the resistivity, at higher temperatures cooperates with the thermal potential energy to maintain the lattice disturbances to which the resistivity is due.

537.312.8 : 538.632 **2412**

The Interdependence and Independence of Galvanomagnetism and of Magnetothermoelectricity.—A. Perrier. (*Helv. phys. Acta*, 15th Dec. 1956, Vol. 29, Nos. 5/6, pp. 419–423. In French.) The relation between Hall and Nernst-Ettingshausen effects is discussed.

537.52 **2413**

The Demonstration of Single Electron Avalanches and their Secondary Processes in Gases.—J. K. Vogel & H. Raether. (*Z. Phys.*, 15th Dec. 1956, Vol. 147, No. 2, pp. 141–147.) A method of obtaining voltage oscillograms of electron avalanches in pure gases is described. The need for adding vapour is eliminated by using high pressure and an electrode gap of 2 cm. Typical results are illustrated and discussed.

537.533.7 : 621.384.612 **2414**

On the Relativistic Motion of Electrons in Magnetic Fields when Quantum Effects are Taken into Account.—A. Sokolov. (*Nuovo Cim.*, 1956, Vol. 3, Supplement No. 4, pp. 743–759. In English.) Summary of theoretical results with 35 references, mainly to Russian authors.

537.533.71 **2415**

Reflection of Plane-Polarized, Electromagnetic Radiation from an Echelette Diffraction Grating.—W. C. Meecham & C. W. Peters. (*J. appl. Phys.*, Feb. 1957, Vol. 28, No. 2, pp. 216–217.)

537.534.8 **2416**

Theory of the Reflection of Positive Ions at Metal Surfaces.—O. v. Roos. (*Z. Phys.*, 15th Dec. 1956, Vol. 147, No. 2, pp. 184–209.) The theory developed yields results which are in satisfactory agreement with experiments on Mo surfaces described by Brunnée (2517 below).

537.534.8 **2417**

Theory of the Kinetic Emission of Secondary Electrons Released by Positive Ions.—O. v. Roos. (*Z. Phys.*, 15th Dec. 1956, Vol. 147, No. 2, pp. 210–227.) See also 2416 above, and for experimental verification 2517 below.

537.56 : 538.56 **2418**

Oscillations of Electron Plasma in a Magnetic Field.—A. G. Sitenko & K. N. Stepanov. (*Zh. eksp. teor. Fiz.*, Oct. 1956, Vol. 31, No. 4 (10), pp. 642–651.) Plasma oscillations at frequencies which are multiples of the gyrofrequency are considered on the basis of kinetic theory. The refractive indices for the ordinary, extraordinary and plasma waves propagated at an angle θ to the magnetic field are calculated. At the frequencies considered the plasma wave is strongly attenuated at $\theta < \pi/2$; at $\theta \approx \pi/2$ these plasma waves cannot be propagated.

537.56 : 538.566 **2419**

Note on Waves in a Homogeneous Magnetically Active Plasma.—B. N. Gershman. (*Zh. eksp. teor. Fiz.*, Oct. 1956, Vol. 31, No. 4 (10), pp. 707–709). Note on papers by Piddington (e.g. 91 and 733 of 1956).

537.56 : 538.63 **2420**

Development of a General Solution for Boltzmann's Transport Equation in the Presence of an Electric and Magnetic Field.—R. Jancel & T. Kahan. (*C. R. Acad. Sci., Paris*, 4th March 1957, Vol. 244, No. 10, pp. 1333–1336.) The general equation derived covers results previously obtained by these and other authors (see e.g. 1318 of 1955).

538.221 **2421**

On the Magnetic Interaction in a System of Doublets.—G. Karpman. (*C. R. Acad. Sci., Paris.*, 4th March 1957, Vol. 244, No. 10, pp. 1336–1339.) The interaction energy is equivalent to that postulated by Herring & Kittel in their spin-wave theory (*Phys. Rev.*, March 1951, Vol. 81, No. 5, pp. 869–880) for a ferromagnetic medium.

538.3 **2422**

Nonlinear Field Theory.—K. Bechert. (*Z. Naturf.*, March 1956, Vol. 11a, No. 3, pp. 177–182.) New formulation of the electrodynamic theory previously developed (see 84 of 1956 and back references).

538.3 : 521 **2423**

Theory of Magnetohydrodynamic Waves.—E. Richter. (*Z. Naturf.*, March 1956, Vol. 11a, No. 3, p. 251.) Brief note on applications to astrophysical problems.

538.566 + 534.2 **2424**

General Theorems on the Equivalence of Group Velocity and Energy Transport.—M. A. Biot. (*Phys. Rev.*, 15th Feb. 1957, Vol. 105, No. 4, pp. 1129–1137.) It is shown that under very general conditions there is a rigorous identity between the group velocity and the velocity of energy transport in nonhomogeneous media, with or without anomalous dispersion.

538.566 : 535.42 **2425**

The Diffraction of an Electromagnetic Wave by a Circular Aperture.—R. F. Millar. (*Proc. Instn elect. Engrs*, Part C, March 1957, Vol. 104, No. 5, pp. 87–95.) The diffraction theory previously developed (2366 of 1956) is used to calculate interaction effects in a plane, perfectly conducting screen, by asymptotic evaluation of the first-order aperture field.

538.566.029.6 **2426**

Methods of Microwave Optics.—K. Bochenek & J. Plebański. (*Archiwum Elektrotech.*, 1956, Vol. 5, No. 2, pp. 293–322. English summary, pp. 322–323.) Two approximation methods developed on the basis of geometrical optics are proposed for the treatment of e.m. fields. In the first method a solution of the wave equation is sought in the form $u = A \exp(ik_0 L)$. The expression $[1 + (1/k_0^2)(\Delta A/A)]^{\frac{1}{2}}$ stands for an effective coefficient of refraction and can be determined to a high accuracy. This interpretation is analogous to the quantum mechanical interpretation of Bohm (*Phys. Rev.*, 15th Jan. 1952, Vol. 85, No. 2, pp. 166–193). The boundary-value problem for the wave equation is transformed into a Cauchy problem for the set of equations for the successive terms of the expansion of A and L in k_0^{-1}. The second method takes the Wentzel-Kramer-Brillouin approximation as a starting point. Both methods are illustrated by examples.

538.569.4 : 538.221 **2427**

Dependence of the Ferromagnetic Resonance Line Width on the Shape of the Specimen.—A. D. Berk. (*J. appl. Phys.*, Feb. 1957, Vol. 28, No. 2, pp. 190–192.) A discussion of the relation between the line width and the damping term in the equation of motion of the magnetization.

538.569.4 : 538.221 : 539.23 **2428**

Ferromagnetic Resonance Absorption by Thin Conducting Films in Cavities.—J. O. Artman. (*J. appl. Phys.*, Feb. 1957, Vol. 28, No. 2, p. 277.) Two cases are briefly considered corresponding to film deposited on a metal and a dielectric substratum respectively.

538.569.4 : 538.222 **2429**

Some Problems of Paramagnetic Resonance.—S. A. Al'tsuler (Al'tshuler) & B. M. Kozyrev. (*Nuovo Cim.*, 1956, Vol. 3, Supplement No. 4, pp. 614–628. In English.) Brief survey and discussion of acoustic paramagnetic resonance and paramagnetic resonance in iron-group salt solutions. 52 references, mainly to Russian authors.

538.569.4 : 539.152.1 : 621.317.4 **2430**

Proton Resonance and the Measurement of Magnetic Fields.—(*Electronic Radio Engr*, June 1957, Vol. 34, No. 6, pp. 215–218.)

538.569.4 : 621.373.029.64 **2431**

Maser Oscillators.—J. C. Helmer. (*J. appl. Phys.*, Feb. 1957, Vol. 28, No. 2, pp. 212–215.) The experimental behaviour of one maser under various operating conditions has been observed using a second maser as a reference standard. Results are compared with theory obtained from a new analysis including the velocity distribution in the beam. See also 100 of 1955 (Gordon et al.).

539.15 : 538.56 **2432**

Generalized Theory of Relaxation.—F. Bloch. (*Phys. Rev.*, 15th Feb. 1957, Vol. 105, No. 4, pp. 1206–1222.) A further generalization of an earlier theory [*ibid.*, 15th Feb. 1953, Vol. 89, No. 4, pp. 728–739 (Wangsness & Bloch)], the energy of the spin system now being freed from restrictions.

539.15 : 548.0 **2433**

Interaction of an Electron Hole with Lattice Oscillations in a Homopolar Crystal.—K. B. Tolpygo & A. M. Fedorchenko. (*Zh. eksp. teor. Fiz.*, Nov. 1956, Vol. 31, No. 5(11), pp. 845–853.) The motion of a hole in a diamond-type homopolar crystal is considered on the basis of the multi-electron Schrödinger equation.

621.3.081.5 **2434**

Differences of Opinion about Dimensions.—R. O. Kapp. (*Proc. Instn elect. Engrs*, Part B, May 1957, Vol. 104, No. 15, pp. 198–204. Discussion, pp. 205–209.) An exposition of the views of different schools.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16 : 621.396.822 **2435**

Discrete Sources of Cosmic Radio Noise at 18·3 and 10·5 Mc/s.—G. R. Ellis & G. Newstead. (*J. atmos. terr. Phys.*, 1957, Vol. 10, No. 4, pp. 185–193.) Six discrete sources were observed at 18·3 Mc/s and 10·05 Mc/s using similar interferometers with a 7-λ base-line. Four of these sources have been observed at 100 Mc/s by other workers, enabling noise flux densities in the range 10–100 Mc/s to be compared. It appears that two of the sources show noise peaks at 20 Mc/s, while the remaining two peak near 10 Mc/s.

523.746 : 538.56.029.63 **2436**

Observation of Polarization of Radio Wave Emission from Sunspots at a Wavelength of 3·2 cm.—N. L. Kaidanovski, D. V. Korol'kov, N. S. Soboleva & S. E. Klaikin. (*C. R. Acad. Sci. U.R.S.S.*, 21st Feb. 1957, Vol. 112, No. 6, pp. 1012–1015. In Russian.)

523.752 **2437**

The Nature of a Type of Radio Emission associated with certain Eruptions from the Chromosphere.—A. Boischot. (*C. R. Acad. Sci., Paris*, 4th March 1957, Vol. 244, No. 10, pp. 1326–1329.) Report on observations at a frequency of 169 Mc/s by means of the interferometer at Nançay [see also 3644 of 1956 (Blum, Boischot & Ginat)].

550.384 **2438**

Linear Secular Oscillation of the Northern Magnetic Pole.—E. R. Hope. (*J. geophys. Res.*, March 1957, Vol. 62, No. 1, pp. 19–27.) "Modern data seem to support the thesis (van Bemmelen, 1899) that the secular motion of the northern magnetic pole is a nearly linear oscillation. This oscillation is along the axis of a great magnetic anomaly in the arctic. Except for the constraint of the anomaly, the motion would probably be circular or quasi-circular, as suggested by the historical declination-dip curves."

550.384 **2439**

Rotation, Pulse-Disturbance, and Drift in the Geomagnetic Secular Variation.—E. R. Hope. (*J. geophys. Res.*, March 1957, Vol. 62, No. 1, pp. 29–42.) A strong pulse disturbance which occurred between 1880 and 1920 in the relative rotation of the terrestrial core, and therefore in the westward drift of surface geomagnetic patterns, is shown to affect the 480-year rotation in a manner which helps to clarify the relations.

550.385 **2440**

On Sudden Commencements of Magnetic Storms at Higher Latitudes.—S. Matsushita. (*J. geophys. Res.*, March 1957, Vol. 62, No. 1, pp. 162–166.) 44 sudden commencements are described; in 21 a small negative impulse preceded the main positive impulse in the horizontal field, in 14 a positive impulse preceded a negative impulse, while in nine the normal low-latitude single-impulse type was observed.

550.385 **2441**

Comparison between the Ionized-Cloud Theory of Chapman and Ferraro and the Recording of the Start of a Magnetic Storm.—J. L. Bureau. (*C. R. Acad. Sci., Paris*, 4th March 1957, Vol. 244, No. 10, pp. 1396–1398.) Brief analysis of the sudden commencement of a magnetic storm on the 21st October 1952 recorded in French West Africa.

551.51 **2442**

Upper Air Pressure and Density Measurements from 90 to 220 Kilometres with the Viking 7 Rocket.—R. Horowitz & H. E. LaGow. (*J. geophys. Res.*, March 1957, Vol. 62, No. 1, pp. 57–78.) The approximate ratios of 'Viking 7' measurements to corresponding 'Rocket Panel' values (*Phys. Rev.*, 1st Dec. 1952, Vol. 88, No. 5, pp. 1027–1032) were: pressure at 90 to 105 km, one quarter; pressure at 220 km, two; densities at 120 to 185 km, one quarter to one half; density at 220 km, one; derived scale height above 140 km, two.

551.510.535 **2443**

The Ionospheric F_2 Layer over Ahmedabad, Delhi, and Tiruchirapalli during the Sunspot Minimum Period (1953–54).—K. M. Kotadia. (*J. sci. industr. Res.*, Dec. 1956, Vol. 15A, No. 12, pp. 543–550.)

551.510.535 : 523.74 **2444**

The Electron Density in the F_2 Layer and its Correlation with the Solar Activity.—F. Mariani. (*J. atmos. terr. Phys.*, 1957, Vol. 10, No. 4, pp. 239–242.) The minimum electron density is shown to be a function not only of sunspot number R but also of the areas of hydrogen filaments and flocculi, A_f and A_ϕ. The relative contribution of these quantities is different for the F_2 layer in the northern and southern geomagnetic hemispheres.

551.510.535 : 523.78 **2445**

Ionospheric Changes at Singapore during the Solar Eclipse of 20 June 1955.—C. M. Minnis. (*J. atmos. terr. Phys.*, 1957, Vol. 10, No. 4, pp. 229–236.) From vertical-incidence measurements on the eclipse and control days, it is deduced that there was a bright source of ionizing radiation in the sun's southern hemisphere, probably associated with an observed group of sunspots. An $F_{1\cdot5}$ layer appeared during the eclipse, indicating a rearrangement of electrons in the lower part of the F_2 layer.

551.510.535 : 550.385 : 523.16 **2446**

The Correlation of Radio-Star-Scintillation Phenomena with Geomagnetic Disturbances and the

Mechanism of Motion of the Ionospheric Irregularities in the F Region.—M. Dagg. (*J. atmos. terr. Phys.*, 1957, Vol. 10, No. 4, pp. 194–203.) F-region drift velocities and the magnitude of variations in the earth's magnetic field are closely correlated, while occasional correlation also exists between scintillation amplitude and magnetic variations. These results are shown to agree with Martyn's theory, ascribing F-region phenomena to the interaction of the earth's magnetic field with an electric field communicated from the dynamo region.

551.510.535 : 550.385 : 523.16 **2447**

Diurnal Variations of Radio-Star Scintillations, Spread F, and Geomagnetic Activity.—M. Dagg. (*J. atmos. terr. Phys.*, 1957, Vol. 10, No. 4, pp. 204–214.) Scintillation measurements of the radio-star in Cassiopeia for the period August 1954–July 1955 are compared month by month with the occurrence of spread F at Inverness and with K-indices at Lerwick or Eskdalemuir.

551.510.535 : 621.396.11 **2448**

Geographical Distribution in High Latitudes of the Anomalous Absorption of Radio Waves in the Ionosphere.—A. P. Nikol'ski. (*C. R. Acad. Sci. U.R.S.S.*, 1st Feb. 1957, Vol. 112, No. 4, pp. 628–631. In Russian.) Discussion based on data from North American ionospheric observatories and Kiruna, Tromsö, Reykjavik, Spitzbergen and Bukhta Tikhaya.

551.510.535 : 621.396.11.029.45 **2449**

Calculations of Ionospheric Reflection Coefficients at Very Low Radio Frequencies.—J. R. Wait & L. B. Perry. (*J. geophys. Res.*, March 1957, Vol. 62, No. 1, pp. 43–56.) "A set of calculated curves are presented for the reflection coefficients at a sharply bounded homogeneous ionized medium with a superimposed magnetic field. The results are plotted parametrically to permit general comparisons with experimental data. Both steady-state and transient cases are considered."

551.594.5 : 621.396.11.029.62 **2450**

The Frequency Dependence of Radio Reflections from Aurora.—P. A. Forsyth & E. L. Vogan. (*J. atmos. terr. Phys.*, 1957, Vol. 10, No. 4, pp. 215–228.) It is concluded that within small regions of the aurora the ionization is often sufficient to cause complete reflection at four frequencies in the range 30–50 Mc/s over a distance of 860 km. Absorption may contribute significantly to observed frequency dependence.

551.594.6 **2451**

Some Statistical Properties of Atmospherics.—Ya. I. Likhter. (*Radiotekhnika i Elektronika*, Oct. 1956, Vol. 1, No. 10, pp. 1295–1302.) An experimental determination of the amplitude probability distribution is reported; a block diagram of the apparatus and the simplified circuit-diagram of a ten-channel statistical analyser are shown. Results of measurements at 50 kc/s (bandwidth 750 c/s) indicate that the probability distribution function is given approximately by the relation $P(V) = (1-c)$ exp $(-aV^2) + c$ exp $(-bV^2)$, where a, b, and c are constants.

LOCATION AND AIDS TO NAVIGATION

621.396.9 **2452**

The Expected Error of a Least-Squares Solution of Location from Direction-Finding Equipment.—B. Harkin. (*Aust. J. appl. Sci.*, Dec. 1956, Vol. 7, No. 4, pp. 263–272.) "Formulae are derived for the error variances and covariances of the coordinate components of an unweighted least-squares solution for the location of a missile simultaneously reported by a number of optical or electronic direction-finding stations. Practical applications of the formulae are suggested."

621.396.932 **2453**

An Investigation of the Sensitivity of the Direction-Finder Telegon III with and without Rectification.—K. Baur. (*Telefunken Ztg*, Dec. 1956, Vol. 29, No. 114, pp. 288–290. English summary, pp. 295–296.) In the direction-finder Telegon III [see also 1455 of May (Troost)] signal strength is indicated by a vertical trace on a 'magic-eye' c.r.tube. Three methods of applying the signal to the indicator, one direct and two using rectification, are compared to determine their merits regarding the indication of low signal levels. Results are inconclusive.

621.396.96 : 621.396.934 **2454**

Subminiature Beacon for Guided Missiles.—M. Cohen & D. Arany. (*Electronics*, 1st April 1957, Vol. 30, No. 4, pp. 144–147.) A transponder for installation in a missile to assist radar tracking. The transponder, which uses transistors, also provides an audio command channel to the missile.

621.396.969.3 : 538.569.4.029.6 **2455**

Investigation of Radar Absorption Materials.—A. Giger & F. Tank. (*Schweiz. Arch. angew. Wiss. Tech.*, Dec. 1956, Vol. 22, No. 12, pp. 414–416.) Brief discussion of theory of centimetre-wave absorption by dissipative dielectric layers on a conductive metallic base. Reflection-coefficient/depth-of-coating curves are given for an iron-dust material for 3 and 10 cm λ and theoretical curves are given for materials with typical dielectric constants and loss tangents.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 : 546.482.21 **2456**

The Properties of Cadmium Sulphide Photoresistors Irradiated with γ and β Rays.—S. V. Svechnikov. (*Zh. tekh. Fiz.*, Aug. 1956, Vol. 26, No. 8, pp. 1646–1650.) The voltage/current characteristics of the photoresistors and the build-up and decay of photocurrent characteristics were determined. Results are presented graphically.

535.215 : 546.817.231 **2457**

Photoconductivity in Lead Selenide: Theory of the Dependence of Sensitivity on Film Thickness and Absorption Coefficients.—J. N. Humphrey & R. L. Petritz. (*Phys. Rev.*, 15th Feb. 1957, Vol. 105, No. 4, pp. 1192–1197.) The wavelength dependence of the absorption coefficient of a photoconductor calculated from the photoconductive spectral response of thin films, is consistent with that predicted from the theory of indirect main-band transitions.

535.215 : 546.863.221 : 621.397.331.2 **2458**

Semiconductor Photosensitive Layers for Photoresistance Television Tubes.—Ya. A. Oksman. (*Radiotekhnika i Elektronika*, Oct. 1956, Vol. 1, No. 10, pp. 1340–1343.) The experimental determination of the characteristics of Sb_2S_3 layers for use in vidicon-type tubes is described and results are presented graphically.

535.215.2 : [546.817.221 + 546.56.231 **2459**

External Photoeffect in Lead Sulphide and Copper Selenide.—P. S. Popov. (*Radiotekhnika i Elektronika*, Oct. 1956, Vol. 1, No. 10, pp. 1334–1339.) Measurements on stable PbS and CuSe photocathodes gave the following results: cut-off wavelength (2919 ± 10)Å and (2950 ± 10)Å for two PbS specimens, and (3172 ± 10)Å for CuSe; contact potential difference to Au in high vacuum: 0·350 V and 0·470 V for PbS and CuSe respectively. PbS specimens with different energy gaps δ between the filled zone and the Fermi level had equal thermionic work functions. The width of the forbidden zone, ΔE, calculated from the results of measurements, was 0·30 eV and 0·39 eV for the two specimens of PbS respectively; these values agree with that predicted by Bell et al. (2037 of 1953). Results of measurements are presented graphically.

535.215.2 : 546.863.36 **2460**

Some Results of an Investigation of the Energy Distribution of Photoelectrons from an Antimony-Caesium Cathode.—N. M. Politova. (*Radiotekhnika i Elektronika*, Oct. 1956, Vol. 1, No. 10, pp. 1325–1333.) The photoelectric work function of Cs_3Sb, determined from measurements of the red-threshold wavelength of the photoeffect was found to be greater than ϕ_E, the work function determined from retardation potential measurements. ϕ_E is only slightly greater than the thermionic work function and its value increases with $h\nu$. These results are discussed.

535.37 **2461**

Impurity-Activated Crystalline Phosphors: their Production and Thermoluminescence Curves.—A. Wizesinska. (*Acta phys. polon.*, 1956, Vol. 15, No. 3, pp. 151–162. In English.)

535.37 **2462**

Tungstate-Silicate Mixed Phosphors.—H. Witzmann & W. Plamann. (*Naturwissenschaften*, Dec. 1956, Vol. 43, No. 24, p. 580.) Brief preliminary report of investigations on the luminescence of the following systems activated by ultraviolet light: (*a*) $CaWO_4$-$CaSiO_3$; (*b*) $CaWO_4$-$CaSiO_3$: (Mn, Pb); (*c*) $ZnWO_4 \cdot ZnO$-Zn_2SiO_4 : (Mn).

535.37 **2463**

Luminescence of Potassium Iodide.—K. J. Teegarden. (*Phys. Rev.*, 15th Feb. 1957, Vol. 105, No. 4, pp. 1222–1227.) The excitation and emission spectra are presented.

535.37 : 538.569.4 **2464**

Paramagnetic Resonance Spectrum of Manganese in Cubic MgO and CaF_2.—W. Low. (*Phys. Rev.*, 1st Feb. 1957, Vol. 105, No. 3, pp. 793–800.) Experimental results at 1·2 and 3·3 cm λ.

535.37 : 538.569.4 **2465**

Paramagnetic Resonance and Optical Absorption Spectra of Cr^{3+} in MgO.—W. Low. (*Phys. Rev.*, 1st Feb. 1957, Vol. 105, No. 3, pp. 801–805.)

535.376 **2466**

Aging Characteristics of Electroluminescent Phosphors.—S. Roberts. (*J. appl. Phys.*, Feb. 1957, Vol. 28, No. 2, pp. 262–265.) A discussion of the decay of brightness with time, and of the rate of aging with varying voltage and frequency.

535.376 **2467**

Low-Field Electroluminescence in Insulating Crystals of Cadmium Sulphide.—R. W. Smith. (*Phys. Rev.*, 1st Feb. 1957, Vol. 105, No. 3, pp. 900–904.) Report of measurements of green electroluminescence obtained with field strength about 1 kV/cm at the emitting centres. The associated abrupt increase of current through the crystal is attributed to the injection of free carriers from the electrodes.

535.376 : 537.311.33 : 546.26-1 **2468**

Electroluminescence of Semiconducting Diamonds.—R. Wolfe & J. Woods. (*Phys. Rev.*, 1st Feb. 1957, Vol. 105, No. 3, pp. 921–922.) Light, whose spectrum consists of a single broad band centred at 4 400 Å, is emitted in the vicinity of a negatively biased point-contact electrode. Voltage and frequency characteristics of the phenomenon are described.

537.226 + 537.311.33 **2469**

Dipole Moments of Dielectric and Semiconductor Particles.—D. V. Kuz'min. (*Zh. tekh. Fiz.*, Sept. 1956, Vol. 26, No. 9, pp. 1880–1883.) Dipole moments of solid particles with various permittivities and conductivities are calculated both for constant and for alternating electric fields.

537.226/.227 : 546.431.824-31 **2470**

Neutron Diffraction Study of Orthorhombic $BaTiO_3$.—G. Shirane, H. Danner & R. Pepinsky. (*Phys. Rev.*, 1st Feb. 1957, Vol. 105, No. 3, pp. 856–860.)

537.226/.227 : 546.431.824-31 **2471**

Symmetry of the Low-Temperature Phase of $BaTiO_3$.—F. Jona & R. Pepinsky. (*Phys. Rev.*, 1st Feb. 1957, Vol. 105, No. 3, pp. 861–864.)

537.226/.227 : 546.431.824-31 **2472**

Electrostatic Considerations in $BaTiO_3$ Domain Formation during Polarization Reversal.—R. Landauer. (*J. appl. Phys.*, Feb. 1957, Vol. 28, No. 2, pp. 227–234.) An analysis of Merz's concept (445 of 1955) of spike-shaped domains of reversed polarization, with special consideration of the electrostatic aspects.

537.226/.227 : 546.824-31 **2473**

Anomalous Polarization of Polycrystalline Titanium Dioxide.—N. P. Bogoroditski, I. D. Fridberg & N. M. Tsvetkov. (*Zh. tekh. Fiz.*, Sept. 1956, Vol. 26, No. 9, pp. 1890–1901.) An experimental investigation of the effect of Group II, III and V oxide impurities. Results of measurements of the dielectric constant and the loss tangent at r.f. are tabulated and the temperature characteristics are presented graphically. A phase diagram for the Ti-TiO_2 system is also given.

537.226 : 546.87.824-31 **2474**

Dielectric Properties of Bismuth Titanates.—G. I. Skanavi & A. I. Demeshina. (*Zh. eksp. teor. Fiz.*, Oct. 1956, Vol. 31, No. 4(10), pp. 565–568.) The properties of non-ferroelectric dielectrics of composition $TiO_2 : Bi_2O_3$ between 22·3 : 1 and 1 : 1 were investigated and results are tabulated and partly presented graphically. The dielectric constant at a temperature of 20°C and a frequency of 2 Mc/s lies between 68 and 121, tan δ between 0·0015 and 0·0054 and the temperature coefficient of ε between -540×10^{-6} and $+590 \times 10^{-6}$.

537.226 : 621.315.61 **2475**

The Aging of the Insulation of Ceramic Materials at High Temperatures.—I. E. Balygin & K. S. Porovski. (*Zh. tekh. Fiz.*, Aug. 1956, Vol. 26, No. 8, pp. 1714–1722.) Experiments are reported on the effects of aging in specimens of ultra-porcelain, radio-porcelain, steatite and spinel at a temperature of 380°C in steady electric fields between about 0·4 and 1·3 kV/mm. For the investigation of the electrolytic processes some of the experiments were carried out at a temperature of approximately 700°C.

537.226 : 621.315.61 **2476**

An Investigation into the Dielectric Polarization and Losses of Polytrifluormonochlorethylene.—G. P. Mikhailov & B. I. Sazhin. (*Zh. tekh. Fiz.*, Aug. 1956, Vol. 26, No. 8, pp. 1723–1729.) The permittivity and the loss angle of the material were investigated over a frequency range from 50 to 10^7 c/s and a temperature range from −100 to +230° C. Two types of relaxation polarization were observed.

537.226.3 **2477**

Dielectric Losses in Glasses: Parts 2 & 3.—N. M. Verebeichik & V. I. Odolevski. (*Zh. tekh. Fiz.*, Aug. 1956, Vol. 26, No. 8, pp. 1696–1703 & 1704–1713.) In Part 2 results are given of an experimental investigation of the electrical properties, thermal expansion, density and refractive index of aluminosilicate sodium glasses. A theoretical interpretation of these results is given. In Part 3 the various existing theories of the 'high-temperature' relaxation dielectric losses in alkali glasses are reviewed and their insufficiency is pointed out. A structural model of this type of glass is proposed and a theory is developed which explains the peculiar role which is played by aluminium atoms in these glasses. A report is also presented on an experimental investigation the results of which confirm the theory. For Part 1, see *Zh. tekh. Fiz.*, Jan. 1952, Vol. 22, No. 1, pp. 12–15 (Verebeichik et al.).

537.226.3 **2478**

Dependence of Dielectric Losses in Ceramic Materials on the Strength of the Electric Field.—I. E. Balygin & A. I. Obraztsov. (*Zh. tekh. Fiz.*, Sept. 1956, Vol. 26, No. 9, pp. 1917–1923.) Experimental investigation of tan δ of several Russian ceramic materials at a frequency of 50 c/s.

537.226.31 **2479**

Nature of the Temperature Dependence of Dielectric Losses in the Polarization of Ionic Compounds.—N. P. Bogoroditski & I. D. Fridberg. (*Zh. tekh. Fiz.*, Sept. 1956, Vol. 26, No. 9, pp. 1884–1889.) The temperature dependence of tan δ at a frequency of 1 Mc/s was experimentally investigated in borate and silicate glasses and in several ceramic materials; results are presented graphically. The dielectric losses are probably due to phenomena of (*a*) relaxation during polarization, (*b*) relaxation during electrical conduction and (*c*) ionization (usually of the free or distributed gas in the solid); (*a*) and (*b*) are connected with the thermal motion of particles.

537.227 **2480**

Behaviour of Ferroelectrics in Strong Electric Fields.—V. A. Bokov. (*Zh. tekh. Fiz.*, Sept. 1956, Vol. 26, No. 9, pp. 1902–1911.) Investigation of the coefficient of nonlinear distortion K and of ϵ_1 in Ba (Ti, Sn)O_3, Ba (Ti, Zr)O_3 and (Ba, Sr) TiO_3 in electric fields of frequency 70 c/s. Maximum nonlinearity, at room temperature, was observed in Ba ($Ti_{0\cdot9}$, $Sn_{0\cdot1}$)O_3 and Ba ($Ti_{0\cdot9}$, $Zr_{0.1}$)O_3.

537.227 **2481**

Causes of Formation of a Curie Region in some Ferroelectric Solid Solutions.—V. A. Isupov. (*Zh. tekh. Fiz.*, Sept. 1956, Vol. 26, No. 9, pp. 1912–1916.) The spread of Curie temperature in specimens of ferroelectric materials can largely be accounted for by fluctuations in the composition, assuming that the 'Kanzig regions' decrease with an increase in non-ferroelectric content.

537.3 : 621.315.6 **2482**

An Investigation of the Electrical Conductivity of Insulating Materials Prior to, During and After Irradiation.—I. M. Rozman & K. G. Tsimmer. (*Zh. tekh. Fiz.*, Aug. 1956, Vol. 26, No. 8, pp. 1681–1688.) A new and simple method based on the use of a capacitor ionization chamber is described, and results are given of measurements on pressed amber, polystyrene, polymethylmetacrylate, polythene and polymonochlortrifluorethylene.

537.311 : 536.2.08 **2483**

The Thermal and Electrical Conductivities of Metals at High Temperatures.—M. R. Hopkins. (*Z. Phys.*, 15th Dec. 1956, Vol. 147, No. 2, pp. 148–160.

In English.) In the method described simultaneous measurements of electrical potential and maximum temperature are made on a short current-carrying wire; observations can be extended into the molten range.

537.311.33 **2484**

The Properties and Structure of Ternary Semiconductor Systems: Part 3—Conductivity and Photoconductivity of Systems based on the Sulphides of Thallium, Antimony and Bismuth.—N. A. Goryunova, B. T. Kolomiets & A. A. Mal'kova. (*Zh. tekh. Fiz.*, Aug. 1956, Vol. 26, No. 8, pp. 1625–1633.) Experimental results show that the x Tl_2S. $(1-x)$ Sb_2S_3 system has a very complex structure, and that within this system there exists a new ternary compound. The materials of which this system is made up are semiconductors with a photoconductivity not exceeding that of the initial binary compounds. The x Sb_2S_3. $(1-x)$ Bi_2S_3 system forms a series of semiconductor materials based on solid solutions of the substitution type. Materials with a considerably lower conductivity than that of the initial materials, and with the maximum of spectral sensitivity displaced towards longer waves can be obtained in this system. By introducing an excess of sulphur, materials of a considerably higher absolute sensitivity can be obtained. Part 2: 3417 of 1956 (Goryunova & Kolomiets).

537.311.33 **2485**

Bipolar Diffusion in Semiconductors at Heavy Currents.—A. I. Gubanov & L. L. Makovski: K. B. Tolpygo. (*Zh. tekh. Fiz.*, Sept. 1956, Vol. 26, No. 9, pp. 2126–2128.) Comment on 1752 of 1956 (Tolpygo & Zaslavskaya) and author's reply.

537.311.33 **2486**

The Stability of Vertical Fusion Zones.—W. Heywang. (*Z. Naturf.*, March 1956, Vol. 11a, No. 3, pp. 238–243.) Mathematical treatment of the stability conditions for stationary zones [see also 1381 of 1955 (Heywang & Ziegler)]. The validity of simplifying assumptions is discussed.

537.311.33 **2487**

One-Dimensional Treatment of the Effective Mass in Semiconductors.—I. Adawi. (*Phys. Rev.*, 1st Feb. 1957, Vol. 105, No. 3, pp. 789–792.) The width of the forbidden band and the effective mass of electrons are calculated as a function of the atomic spacing and potential asymmetry, using a Kronig-Penney model of a one-dimensional semiconductor. Contrary to the result of Seraphin (1379 of 1955), the effective mass, for a specific average potential, is found to increase monotonically with potential asymmetry.

537.311.33 **2488**

Hall and Drift Mobilities; their Ratio and Temperature Dependence in Semiconductors.—F. J. Blatt. (*Phys. Rev.*, 15th Feb. 1957, Vol. 105, No. 4, pp. 1203–1205.) It is shown that the temperature-dependence of the Hall and drift mobilities in the impurity-scattering temperature range should be less rapid than $T^{3/2}$. The conclusion is in agreement with experiment.

537.311.33 : 535.215 **2489**

Theory of Atomic Semiconductors.—A. G. Samoilovich & V. M. Kondratenko. (*Zh. eksp. teor. Fiz.*, Oct. 1956, Vol. 31, No. 4 (10), pp. 596–608.) Problems in the theory of absorption of light and the theory of photoconductivity are considered on the basis of the polar crystal model of Shubin & Vonsovski (*C. R. Acad. Sci. U.R.S.S.*, 1934, Vol. 1, p. 449), taking into account excitons.

537.311.33 : 535.215 **2490**

The Determination of Volume- and Surface-Recombination of Charge Carriers in Semiconductors.—W. Heywang & M. Zerbst. (*Z. Naturf.*, March 1956, Vol. 11a, No. 3, pp. 256–257.) A photoconductivity method for determining separately the bulk lifetime and surface recombination velocity is outlined and some typical results are briefly discussed.

537.311.33 : 535.215 **2491**

Influence of Vapours and Gases on the Internal Photoeffect in Oxide Semiconductors and Sensitization by Clorophyll.—E. K. Putseiko. (*Radiotekhnika i Elektronika*, Oct. 1956, Vol. 1, No. 10, pp. 1354–1373.) The effect of oxygen, water vapour, and other gases and vapours on the photoeffect in ZnO, HgO and PbO was investigated. In the presence of O_2, thermal activation up to 100° C increases the photo-e.m.f. of HgO and increases the sensitivity of the photo-e.m.f. of PbO to long-wave light. The sensitization of ZnO and HgO by chlorophyll and its analogues is discussed.

537.311.33 : 535.215 : 538.61 **2492**

Spectral Distribution of the Photomagnetoelectric Effect in Semiconductors: Theory.—W. Gärtner. (*Phys. Rev.*, 1st Feb. 1957, Vol. 105, No. 3, pp. 823–829.) An extension of the theory of the photomagnetoelectric (p.m.e.) effect to cover its dependence on the wavelength of the incident light. The p.m.e. response is shown graphically as a function of absorption coefficient, with bulk lifetime, surface recombination velocities and slab thickness as parameters.

537.311.33 : 536.21.022 **2493**

On Thermal Conduction in Semiconductors.—A. F. Ioffe. (*Nuovo Cim.*, 1956, Vol. 3, Supplement No. 4, pp. 702–715. In English.) Brief review and discussion of the results of recent investigations mainly by Russian authors.

537.311.33 : [538.63 + 538.66 **2494**

Theory of Isothermal Galvano- and Thermo-magnetic Phenomena in Semiconductors.—F. G. Bass & I. M. Tsidil'kovski. (*Zh. eksp. teor. Fiz.*, Oct. 1956, Vol. 31, No. 4 (10), pp. 672–683.) The phenomena are considered theoretically for the case of magnetic fields at which $(uH/c)^2 \approx 1$ or $\gg 1$, where H is the magnetic field strength, u the mobility of charge carriers, and c the velocity of light.

537.311.33 : [546.23 + 546.24 **2495**

Electronic Band Structure of Selenium and Tellurium.—J. R. Reitz. (*Phys. Rev.*, 15th Feb. 1957, Vol. 105, No. 4, pp. 1233–1240.) The band structure of Se and Te has been calculated according to the tight-binding scheme in which only nearest-neighbour interactions are presumed to be important.

537.311.33 : 546.24 **2496**

The Effect of Impurities and Heat Treatment on the Electrical Properties of High-Purity Tellurium.—H. Kronmüller, J. Jaumann & K. Seiler. (*Z. Naturf.*, March 1956, Vol. 11a, No. 3, pp. 243–250.) The influence of small additions of As, Sb, Br and I on the Hall effect and conductivity was investigated in the temperature range −140 to 300° C. Results are shown graphically and discussed.

537.311.33 : 546.28 **2497**

Ionization Rates for Holes and Electrons in Silicon.—S. L. Miller. (*Phys. Rev.*, 15th Feb. 1957, Vol. 105, No. 4, pp. 1246–1249.) "The ionization rates for holes and electrons in silicon at high electric fields have been evaluated from data on the multiplication of reverse-biased junctions. In Si, electrons have a higher ionization rate than holes. The variation of ionization rate with field strength is in good agreement with theory."

537.311.33 : 546.28 **2498**

Properties of Gold-Doped Silicon.—C. B. Collins, R. O. Carlson & C. J. Gallagher. (*Phys. Rev.*, 15th Feb. 1957, Vol. 105, No. 4, pp. 1168–1173.) Measurements of the temperature dependence of resistivity and Hall coefficient show an acceptor level at 0·54 eV from the conduction band and a donor level at 0·35 eV from the valence band. Concentrations of these levels are equal within experimental accuracy.

537.311.33 : [546.28 + 546.289] : 534.13-8 **2499**

Ultrasonic Attenuation in Germanium and Silicon.—F. J. Blatt. (*Phys. Rev.*, 1st Feb. 1957, Vol. 105, No. 3, pp. 1118–1119.) It is suggested that measurements of ultrasonic attenuation would throw light on the intervalley scattering of electrons and might permit a direct determination of the relevant coupling constant.

537.311.33 : 546.289 **2500**

Determination of Germanium in some Italian Coals.—G. Leonardi & E. Mariani. (*Piccole Note Ist. super. Poste e Telecomunicazioni*, Nov./Dec. 1956, Vol. 5, No. 6, pp. 799–803.)

537.311.33 : 546.289 **2501**

Surface Electrical Conductivity of Germanium.—V. I. Lyashenko & T. N. Sytenko. (*Zh. eksp. teor. Fiz.*, Nov. 1956, Vol. 31, No. 5(11), pp. 905–907.) Experimental results indicate the existence of surface zone conductivity in the specimens investigated.

537.311.33 : 546.289 **2502**

Energy of Ionization by Electrons in Germanium Crystals.—V. S. Vavilov, L. S. Smirnov & V. M. Patskevich. (*C. R. Acad. Sci. U.R.S.S.*, 21st Feb. 1957, Vol. 112, No. 6, pp. 1020–1022. In Russian.) An experimental determination of the mean energy of ionization ϵ in Ge bombarded by 5–15-keV electron beams; results indicate that $\epsilon = 3{\cdot}7 \pm 0{\cdot}4$ eV in this range.

537.311.33 : 546.289 **2503**

Experimental Evidence of the Anisotropy of Hot Electrons in *n*-Type Germanium.—W. Sasaki & M. Shibuya. (*J. phys. Soc. Japan.*, Nov. 1956, Vol. 11, No. 11, pp. 1202–1203.) Brief note on experimental technique used and results obtained. For a statement of the problem, see 464 of 1956 (Shibuya).

537.311.33 : 546.289 **2504**

Effect of Annealing in Various Gases on the Bulk Lifetime of Germanium.—K. Weiser. (*J. appl. Phys.*, Feb. 1957, Vol. 28, No. 2, pp. 271–272.) The lifetime was increased markedly by annealing in oxygen having a trace of water vapour. A decrease, after a period of no change, was observed for dry oxygen, wet nitrogen, helium, argon and hydrogen.

537.311.33 : 546.289 **2505**

The Influence of Omnidirectional Pressure on the Drift Mobility of Holes in Germanium.—G. Landwehr. (*Z. Naturf.*, March 1956, Vol. 11a, No. 3, p. 257.) Brief account of measurements at pressures up to 10 000 kg/cm² at 22° C, which show that drift mobility is not affected and that the surface recombination velocity increases with pressure.

537.311.33 : 546.289 **2506**

Effect of Impurities on Free-Hole Infrared Absorption in *p*-Type Germanium.—R. Newman & W. W. Tyler. (*Phys. Rev.*, 1st Feb. 1957, Vol. 105, No. 3, pp. 885–886.) The spectrum structure becomes less pronounced with increasing carrier and total impurity concentration. The effects are consistent with changes in the Fermi level and with nonvertical transitions induced by charged impurity centres.

537.311.33 : 546.289 **2507**

***K* X-ray Absorption Spectrum of a Single Crystal of Germanium.**—D. G. Doran & S. T. Stephenson. (*Phys. Rev.*, 15th Feb. 1957, Vol. 105, No. 4, pp. 1156–1157.)

537.311.33 : 546.289 **2508**

The Breakdown of *p-n* Junctions in Germanium by a Voltage Impulse.—A. P. Shotov. (*Zh. tekh. Fiz.*, Aug. 1956, Vol. 26, No. 8, pp. 1634–1645.) Junctions prepared by alloying indium with *n*-germanium and diffusing of antimony in *p*-germanium were investigated. The specific resistance of the initial germanium varied between 0·15 and 50 Ω.cm. The breakdown voltage was measured with 10^{-5}–10^{-6}-s pulses. It was established that the breakdown of all junctions investigated is caused by shock ionization. The breakdown voltage increases with temperature owing to the dependence of the ionization coefficient on temperature. For junctions prepared by alloying indium with germanium, the value of the ionization coefficient is determined.

537.311.33 : 546.289 : 537.533.9 **2509**

Formation of Crystal Lattice Defects in Germanium under Bombardment by Fast Electrons.—V. S. Vavilov, L. S. Smirnov, G. N. Galkin, A. V. Spitsyn & V. M. Patskevich. (*Zh. tekh. Fiz.*, Sept. 1956, Vol. 26, No. 9, pp. 1865–1869.) The dependence of the defect-formation cross-section on irradiation electron energy was investigated experimentally by measurement of the electrical conductivity of monocrystalline *n*-type Ge films bombarded with 400–1 000-keV electrons. No effects were observed at energies below 500 ± 20 keV.

537.311.33 : 546.289 : 537.534.9 **2510**

Work-Function Studies of Germanium Crystals Cleaned by Ion Bombardment.—J. A. Dillon, Jr, & H. E. Farnsworth. (*J. appl. Phys.*, Feb. 1957, Vol. 28, No. 2, pp. 174–184.) Measurements of the effects of adsorption of gases, strong electric fields and intense illumination upon the work functions of single germanium crystals are reported and discussed.

537.311.33 : 546.289 : 538.63 **2511**

Magnetoconductivity in *p*-Type Germanium.—C. Goldberg, E. N. Adams & R. E. Davis. (*Phys. Rev.*, 1st Feb. 1957, Vol. 105, No. 3, pp. 865–876.) "Measurements of the Hall coefficient and resistivity of *p*-type germanium have been made as a function of magnetic field, temperature, and carrier concentration between 77° K and 300° K. An attempt is made to interpret the data quantitatively using a two-carrier model but no completely satisfactory quantitative interpretation is possible."

537.311.33 : 546.482.21 **2512**

Connection between Changes in Electrical Conductivity and Redistribution of Electron Density in a Cadium Sulphide Crystal.—Yu. N. Shuvalov. (*Zh. tekh. Fiz.*, Sept. 1956, Vol. 26, No. 9, pp. 1870–1879.) An X-ray crystallographic investigation is reported.

537.311.33 : 546.682.86 **2513**

Interband Magneto-optic Effects in Semiconductors.—E. Burstein & G. S. Picus. (*Phys. Rev.*, 1st Feb. 1957, Vol. 105, No. 3, pp. 1123–1125.) With magnetic fields as low as 15 000 oersted the absorption spectrum shows several peaks, which move to higher photon energies with increasing field. Tentative mechanisms for the absorption peaks are suggested, in terms of the energy bands as affected by the magnetic field.

537.311.33 : 546.682.86 **2514**

Elastoresistance Constants of *p*-Type InSb at 77° K.—A. J. Tuzzolino. (*Phys. Rev.*, 15th Feb. 1957, Vol. 105, No. 4, pp. 1411.) Report of a preliminary experiment.

537.311.33 : 546.811-17 **2515**

On Factors Influencing the Transformation of White Tin into Grey Tin.—A. I. Bykhovski. (*Zh. tekh. Fiz.*, Aug. 1956, Vol. 26, No. 8, pp. 1799–1801.) A brief review of the literature is presented and the effect of impurities and of their redistribution during heating on the transformation of white tin into grey tin ('tin plague') is discussed.

537.533 : [546.289 + 546.56 + 546.811 **2516**

Secondary-Electron Emission of Copper, Germanium and Tin in Solid and Liquid States.—V. G. Bol'shov & V. K. Seleznev. (*Zh. tekh. Fiz.*, Aug. 1956, Vol. 26, No. 8, pp. 1657–1664.) Experimental results show that when the temperature is increased from room temperature to 232° C (in the case of tin) and to 1 033° C (in the case of copper) σ_{max} varies by not more than 1%. In the case of Ge, when temperature is increased to 959° C, σ_{max} decreases by approximately 5–6%. As a result of melting σ changes abruptly in the same direction for the whole range of energies of primary electrons from 100 to 1 500 eV, viz., in the case of copper σ_{max} decreases by 5%, in the case of tin and germanium it increases by 14% and 9% respectively. The energy distribution of secondary electrons for solid and liquid tin has also been considered.

537.534.8 **2517**

The Ion Reflection and Secondary Electron Emission at the Impact of Alkali Ions on Clean Molybdenum Surfaces.—C. Brunnée. (*Z. Phys.*, 15th Dec. 1956, Vol. 147, No. 2, pp. 161–183.) Report and detailed discussion of measurements in the energy range 0·4–4 keV. Over 60 references.

537.534.9 : [546.74 + 546.883 **2518**

Disintegration of Tantalum and Nickel in the Form of Ions under Bombardment by Positive Caesium Ions.—V. I. Veksler & M. B. Ben'yamovich. (*Zh. tekh. Fiz.*, Aug. 1956, Vol. 26, No. 8, pp. 1671–1680.) In a mass-spectrometer investigation of the products of the secondary emission of Ta and Ni targets, positive ions of these two metals were observed in quantities greatly exceeding the theoretical values.

537.583 : 546.883 **2519**

Autoelectron [thermionic] **Emission of Tantalum.**—M. I. Elinson & G. F. Vasil'ev. (*Zh. tekh. Fiz.*, Aug. 1956, Vol. 26, No. 8, pp. 1669–1670.)

538 **2520**

Conference on the Physics of Magnetic Phenomena [Moscow, 23rd–31st May, 1956].—S. V. Vonsovski. (*Uspekhi fiz. Nauk*, Dec. 1956, Vol. 60, No. 4, pp. 709–722.) Review of papers presented at the conference.

538.22 **2521**

Magnetic Susceptibility of Dilute Alloys of Nickel in Copper between 2·5° K and 295° K.—E. W. Pugh, B. R. Coles, A. Arrott & J. E. Goldman. (*Phys. Rev.*, 1st Feb. 1957, Vol. 105, No. 3, pp. 814–818.)

538.221 **2522**

New Magnetic Anisotropy.—W. H. Meiklejohn & C. P. Bean. (*Phys. Rev.*, 1st Feb. 1957, Vol. 105, No. 3, pp. 904–913.) Report and discussion of measurements relating to 'exchange anisotropy' (see 3803 of 1956).

538.221 **2523**

Effect of a Cavity on a Single-Domain Magnetic Particle.—W. F. Brown, Jr, & A. H. Morrish. (*Phys. Rev.*, 15th Feb. 1957, Vol. 105, No. 4, pp. 1198–1201.) A detailed investigation of the effect of internal cavities in single-domain particles on the coercive force, shows that they may account, at least in part, for the low values sometimes encountered.

538.221 : 538.245 **2524**

On the Statistical Nature of the Remagnetization of Ferromagnetics.—F. V. Bunkin. (*Zh. tekh. Fiz.*, Aug. 1956, Vol. 26, No. 8, pp. 1782–1789.) A mathematical investigation of the phenomenon is presented, on the assumption that all domains have an equal probability of remagnetization during each cycle of the field variation. It is shown that the probability of the transitions of a domain into a new state at a given value of the magnetizing field is proportional to the rate of the increase in the magnetization of the specimen at this value of the field. Formulae are also derived for determining the scatter of transition instants.

538.221 : 538.245 **2525**

Noise due to the Cyclic Remagnetization of Ferromagnetics.—F. V. Bunkin. (*Zh. tekh. Fiz.*, Aug. 1956, Vol. 26, No. 8, pp. 1790–1798.) A general expression is derived for the spectral intensity of the induction e.m.f. during the cyclic remagnetization of a ferroelectric specimen. The signal/noise ratio in transformers is also determined.

538.221 : 546.74-3 **2526**

Crystal Structure of Ferromagnetic Nickel Oxide.—Y. Shimomura, M. Kojima & S. Saito. (*J. phys. Soc. Japan*, Nov. 1956, Vol. 11, No. 11, pp. 1136–1146.)

538.221 : 621.318.12 **2527**

Structure and Magnetic Properties of Permanent-Magnet Alloys during Isothermal Precipitation Hardening: Part 2—The Process of Segregation and the Interpretation of Magnetic Behaviour.—E. Biedermann & E. Kneller. (*Z. Metallkde*, Dec. 1956, Vol. 47, No. 12, pp. 760–774.) Further analysis of investigations made on Cu-Ni-Fe and Cu-Ni-Co alloys. Part 1: 202 of January.

538.221 : 621.318.134 **2528**

Formation of Manganese Ferrite by Solid-State Reaction.—H. H. Kedesdy & A. Tauber. (*J. Amer. ceram. Soc.*, Dec. 1956, Vol. 39, No. 12, pp. 425–431.) The effect of four different firing cycles on the formation of the ferrite is investigated, and the magnetic properties are compared for toroidal specimens subjected to three different forms of heat treatment.

538.221 : 621.318.134 **2529**

Low-Loss Magnesium Manganese Ferrites.—L. C. F. Blackman. (*J. Electronics*, March 1957, Vol. 2, No. 5, pp. 451–456.) These are prepared by thermal decomposition of mixed nitrates. The normalized energy-loss-factor for the Q-band is 0·04 ± 0·02 dB.

538.221 : 621.318.134 : 538.566 **2530**

Effects of Zero Ferrite Permeability on Circularly Polarized Waves.—B. J. Duncan & L. Swern. (*Proc. Inst. Radio Engrs*, Part 1, May 1957, Vol. 45, No. 5, pp. 647–655.) If the imaginary part of the effective permeability of a ferrite is negligible, the ferrite can be made to exhibit zero permeability to a wave with positive sense of circular polarization. The application of this to a rod placed axially in a circular waveguide propagating the TE_{11} circularly polarized mode is examined theoretically and experimentally. It is shown that the ferrite can be made to expel practically all microwave energy from its interior, and this behaviour can be used to obtain large nonreciprocal attenuations.

538.221 : 621.318.134 : 538.6 **2531**

Microwave Frequency Doubling from 9 to 18 kMc/s in Ferrites.—J. L. Melchor, W. P. Ayres & P. H. Vartanian. (*Proc. Inst. Radio Engrs*, Part 1, May 1957, Vol. 45, No. 5, pp. 643–646.) The theory of the process is given and it is found experimentally that conversion efficiency depends markedly on the geometry of the ferrite; efficiencies as high as −6 dB have been observed. See also 2138 of 1956 (Ayres et al.).

538.632 : 539.23 : 546.87 **2532**

The Hall Effect in Thin Bismuth Films.—A. Colombani & P. Huet. (*C. R. Acad. Sci., Paris*, 4th March 1957, Vol. 244, No. 10, pp. 1344–1347.) Experimental results are given in graphical form for fields up to 34 300 oersted and thicknesses between 60 and 6 000 Å. See also 2225 and 2226 of July.

538.652 : 546.74 **2533**

Theory of Magnetostriction of Single Crystals of Nickel.—W. F. Brown, Jr.: N. S. Akulov. (*C. R. Acad. Sci. U.R.S.S.*, 11th Feb. 1957, Vol. 112, No. 5, pp. 827–830. In Russian.) Comment on 2143 of 1956 comparing the theory with Heisenberg's theory of magnetostriction, and author's reply.

548.5 : 537.228.1 **2534**

Properties of Synthetic Quartz Oscillator Crystals.—C. S. Brown & L. A. Thomas. (*Proc. Instn elect. Engrs*, Part C, March 1957, Vol. 104, No. 5, pp. 174–184.) A process for the growing of well-formed crystals weighing about 135 g is described. The crystals possess electrical and mechanical properties which closely resemble those of the natural Brazilian quartz. Comparative measurements on oscillator crystals cut from natural and synthetic quartz are also described.

621.315.6 **2535**

The Electrical Properties and Structure of Silicone Polymers.—K. A. Andrianov & G. E. Golubkov. (*Zh. tekh. Fiz.*, Aug. 1956, Vol. 26, No. 8, pp. 1689–1695.) The electrical properties of polydimethylsiloxanes and polydiethylsiloxanes were studied at various temperatures and frequencies, as well as the variation of their thermal properties in the process of cooling and heating.

621.318.134 : 621.318.424 **2536**

Ferrite Thermomagnets.—L. I. Rabkin & B. Sh. Epshtein, (*Radiotekhnika i Elektronika*, Oct. 1956, Vol. 1, No. 10, pp. 1357–1363.) Temperature-sensitive ferrite core materials suitable for temperature compensation of inductors are described.

621.79 : 621.3.049.001.4 **2537**

Leakage Testing of Sealed Electronic Enclosures.—D. C. Bedwell & E. A. Meyer. (*Elect. Mfg*, Dec. 1955, Vol. 56, No. 6, pp. 127–133.) Methods of testing electronic components suitable for airborne applications are reviewed, and an evaluation of the qualities of various types of seals based on a series of tests is given in tabular form.

MATHEMATICS

512.3 : 621.372 **2538**

An Elimination Technique for Certaiu Impedance Equations.—C. D. Allen. (*Electronic Engng*, April 1957, Vol. 29, No. 350, pp. 187–188.) "A set of linear simultaneous equations of a type which frequently arises in the analysis of linear networks is considered. A short method of performing the elimination necessary to obtain the impedance equations of the network is proved."

512.831 : 621.372 **2539**

Some Matrix Theorems.—W. P. Wilson. (*Electronic Radio Engr*, June 1957, Vol. 34, No. 6, pp. 229–231.) A theorem relating to the nth power of a matrix and another having particular relevance in the study of tapered networks are included.

517 **2540**

Limiting Properties of Mathieu Functions.—M. A. Jaswon. (*Proc. Camb. phil. Soc.*, Jan. 1957, Vol. 53, Part 1, pp. 111–114.) A new analysis is made of the coefficients appearing in periodic Mathieu functions and this is applied to establish the limiting properties of certain solutions of the equation

$$r\frac{d}{dr}\left(r\frac{d}{dr}\right)y-\left(a+\alpha^2r^2+\beta^2r^{-2}\right)y=0$$

where a, α and β are parameters.

517.512.2 2541

The Closed-Form Summation of some Common Fourier Series.—C. C. Chao. (*Quart. J. Mech. appl. Math.*, Dec. 1956, Vol. 9, Part 4, pp. 508-512.) "A method is presented for the summation in closed form of Fourier series whose coefficients are ratios of polynomials of certain types frequently encountered in practice. The result is obtained as the solution of a linear differential equation with constant coefficients, which can be solved by elementary methods."

MEASUREMENTS AND TEST GEAR

531.76 : 621.372.632 2542

Millimicrosecond Time Analyser.—C. Cottini & E. Gatti. (*Nuovo Cim.*, 1st Dec. 1956, Vol. 4, No. 6, pp. 1550–1557. In English.) A fuller account is given of the arrangement for measurement of time intervals discussed by Cottini et al. (219 of January).

621.3.087 2543

A Method of Measuring Integrated Values with Very High Time Constants.—G. Bartels & R. Steinert. (*NachrTech.*, Nov. 1956, Vol. 6, No. 11, pp. 500–502.) The application is briefly discussed of a high-resistance valve-voltmeter circuit using electrometer or 'inverted' valves for recording average values of variables with time constants up to several hours. A circuit for u.s.w. field strength measurement with a time constant of 900 s is shown together with a typical record made by it.

621.317.3 : 621.314.63 2544

Dynamic Methods of Testing Semiconductor Rectifier Elements and Power Diodes: Part 1.—A. H. B. Walker & R. G. Martin. (*Electronic Engng*, April 1957, Vol. 29, No. 350, pp. 150–157.) In full dynamic tests the forward and reverse characteristics of the diode are measured simultaneously. The forward voltage is measured by a low-range voltmeter unaffected by the reverse half cycle, and the reverse current by an ammeter of substantially zero impedance, unaffected by the forward current. The design and performance of negative-feedback amplifiers which provide these characteristics are fully described.

621.317.3.029.6 2545

Single-Oscillator Microwave Measuring System.—D. H. Ring. (*Bell Lab. Rec.*, Dec. 1956, Vol. 34, No. 12, pp. 465–468.) A simplified version of the continuous waveguide phase-shifter described by Fox (1255 of 1948) is used to obtain a 2° phase shift for each degree of rotation of the motor-driven middle section. By passing a microwave signal through the waveguide and rotating the section at e.g. 4 500 r.p.m. a frequency change of ±150 c/s is obtained so that measurements can be made at the i.f. of 150 c/s. The advantages of the conventional two-oscillator circuit are thus combined with the stability of the single-oscillator system.

621.317.3/.41].029.64 : 621.318.134 2546

New Method of Measuring the Parameters of Magnetized Ferrites at Centimetre Wavelengths.—V. N. Vasil'ev. (*Radiotekhnika i Elektronika*, Nov. 1956, Vol. 1, No. 11, pp. 1444–1460.) The ferrite specimen in the form of a lamina is placed along the axis of a rectangular resonator; using H_{210} and H_{310} modes and varying the direction of the steady magnetic field in a specified way, the resulting changes of the resonance frequency and the Q of the resonator will be functions of only one or two parameters of the ferrite. The required formulae are given. The method is suitable for the 0·8–20-cm-λ range. Results of measurements at 3 cm λ are presented graphically.

621.317.328 : 621.396.822 2547

Logarithmic Amplifier measures Noise.—J. D. Wells. (*Electronics*, 1st April 1957, Vol. 30, No. 4, pp. 169–171.) An equipment for recording atmospheric noise level is described in which the logarithm of the voltage equalled or exceeded for 50% of the time is determined.

621.317.33+[621.317.39 : 531.71 2548

Regenerative Measuring Pickups.—L. L. Dekabrun. (*Avtomatika i Telemekhanika*, Dec. 1956, Vol. 17, No. 12, pp. 1114–1122.) Analysis is presented of devices for measuring r.f. conductivity or linear displacement by means of the damping effect on a tuned circuit.

621.317.334 : 621.362 2549

Inductance-Type Thermocouple Tester.—P. R. Morris. (*Instrum. & Automation*, Nov. 1956, Vol. 29, No. 11, pp. 2217–2219.) A battery-operated inductance-bridge and 1·5-Mc/s oscillator are used to detect short-circuits, four or more inches from the hot junction of a 5-ft thermocouple.

621.317.335.3.029.64 : 537.226.2 2550

Waveguide Method of Measuring the Dielectric Properties of Materials at Elevated Temperatures.—V. I. Aksenov & M. Ya. Borodin. (*Radiotekhnika i Elektronika*, Nov. 1956, Vol. 1, No. 11, pp. 1435–1443.) Application of the short-circuit and open-circuit waveguide methods of determining the dielectric properties of materials at a wavelength of 3·2 cm and temperatures of 20–200° C is discussed. Formulae are given for calculating ϵ' and ϵ'' and the results of determinations of ϵ' and tan δ of some materials of high molecular weight are tabulated and presented graphically.

621.317.36 : 621.385.029.6 2551

The Measurement of Magnetron Frequency Pulling.—Twisleton. (See 2645.)

621.317.38.029.6 2552

Hall Effect and its Counterpart, Radiation Pressure, in Microwave Power Measurement.—H. E. M. Barlow. (*Proc. Instn elect. Engrs*, Part C, March 1957, Vol. 104, No. 5, pp. 35–42.) Theory shows the close link between high-frequency Hall effect and radiation pressure. It predicts an equivalent Hall e.m.f. in a dielectric and offers a possible new approach in exploration of properties of dielectrics. Aspects of applications for microwave power measurement are discussed.

621.317.729.1 2553

A New Form of Electrolytic Tank.—K. F. Sander & J. G. Yates. (*Proc. Instn elect. Engrs*, Part C, March 1957, Vol. 104, No. 5, pp. 81–86.) Measurements with capillary probes in a plane defined by an insulating surface eliminate meniscus effects with probes in a free surface.

621.317.733 : 621.375.2 2554

Intermodulation in Bridge Detector Amplifiers.—Johnson & Thompson. (See 2391.)

621.317.755 : 621.397.001.4 : 535.623 2555

The Vectorscope.—Parker Smith & Matley. (See 2602.)

621.317.784.029.64 2556

An Absolute Microwave Wattmeter.—A. Macpherson. (*Proc. Inst. Radio Engrs*, Part 1, May 1957, Vol. 45, No. 5, pp. 688–689.) The wattmeter is of the calorimetric type and versions for X-band and K-band use are described. The estimated least detectable power is about 0·1 W at K-band and 0·01 W at X-band frequencies.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621.384.6 2557

Theory of Betatron Oscillations of Particles in Magnetic-Field Systems.—A. A. Kolomenski. (*Zh. tekh. Fiz.*, Sept. 1956, Vol. 26, No. 9, pp. 1969–1990.)

621.384.612 2558

Synchrotron Oscillations in Strong-Focusing Accelerators: Part 1—Linear Theory.—L. L. Gol'din & D. G. Koshkarev. (*Zh. eksp. teor. Fiz.*, Nov. 1956, Vol. 31, No. 5(11), pp. 803–814.)

621.384.613 2559

Stereo-betatron.—V. A. Moskalev. (*Zh. tekh. Fiz.*, Sept. 1956, Vol. 26, No. 9, pp. 2060–2061.) Brief note with a section drawing of a 10-MeV stereo-betatron for flaw detection in materials.

621.384.622.2 2560

A Theoretical and Experimental Investigation of Anisotropic-Dielectric-Loaded Linear Electron Accelerators.—R. B. R. Shersby-Harvie, L. B. Mullett, W. Walkinshaw, J. S. Bell & B. G. Loach. (*Proc. Instn elect. Engrs*, Part B, May 1957, Vol. 104, No. 15, pp. 273–290. Discussion, pp. 290–292.) The theory of waveguides loaded with closely-spaced dielectric disks is given, and the shunt loss is derived. Advantages over the conventional corrugated waveguides are discussed. Details of practical construction and measurements using titanium-oxide disks are given; excess attenuation at high r.f. power probably being due to deposits of carbon from the oil-diffusion pumps, the use of mercury pumps is suggested. 40 references.

621.385.833 **2561**

The Treatment of the Transition Regions by the Parabola Method and Similar Computation Techniques.—P. Schiske. (*Optik, Stuttgart*, Dec. 1956, Vol. 13, No. 12, pp. 529–536.) Methods of approximating the third-order electron trajectories in electrostatic systems are compared and discussed.

621.385.833 **2562**

The Electron-Microscope Transmission Factor of Thin Films.—W. Lippert. (*Optik, Stuttgart*, Nov. 1956, Vol. 13, No. 11, pp. 506–515.)

621.385.833 **2563**

The Lower Limit of the Aperture Error in Magnetic Electron Lenses.—W. Tretner. (*Optik, Stuttgart*, Nov. 1956, Vol. 13, No. 11, pp. 516–519.) Addition to earlier work (236 of 1955 and 539 of 1956).

621.385.833 : 535.317.3 **2564**

The Lower Limit of the Chromatic Aberration of Magnetic Lenses for a Given Maximum Field Strength.—P. Schiske. (*Optik, Stuttgart*, Nov. 1956, Vol. 13, No. 11, pp. 502–505.)

621.385.833 : 537.533.73 **2565**

The Treatment of Electron-Optical Aberrations by Diffraction Theory.—J. Picht. (*Optik, Stuttgart*, Nov. 1956, Vol. 13, No. 11, pp. 494–501.)

PROPAGATION OF WAVES

538.566 **2566**

Reflection of Waves by an Isotropic Inhomogeneous Layer.—A. G. Zharkovski & O. M. Todes. (*Zh. eksp. teor. Fiz.*, Nov. 1956, Vol. 31, No. 5(11), pp. 815–818.) An approximate method of calculating the reflection coefficient of an isotropic inhomogeneous layer for a plane e.m. wave is developed for media whose permittivity and permeability are functions of a single space coordinate.

538.566 : 537.56 **2567**

On the Interaction of Electromagnetic Waves with Charged Particles and on the Oscillations of the Electronic Plasma.—A. Ahiezer (Akhiezer). (*Nuovo Cim.*, 1956, Vol. 3, Supplement No. 4, pp. 591–613. In English.) Survey of theoretical work in this field with 34 references, mainly to Russian authors.

538.566 : 537.56 **2568**

A Property of the Field of an Electromagnetic Wave Propagated in an Inhomogeneous Plasma.—N. G. Denisov. (*Zh. eksp. teor. Fiz.*, Oct. 1956, Vol. 31, No. 4 (10), pp. 609–619.) The growth of the field in the region where the plasma permittivity tends to zero is investigated theoretically. The influence of absorption is clarified. The connection between field growth and plasma resonance is established.

621.396.11 **2569**

Radio Propagation over a Discontinuity in the Earth's Electrical Properties: Part 1.—T. B. A. Senior. (*Proc. Instn elect. Engrs*, Part C, March 1957, Vol. 104, No. 5, pp. 43–53.) A method avoiding the analytical complications of rigorous theory is developed using known results for diffraction at a straight edge. Solutions reveal all the important features of mixed-path propagation and are suitable for numerical application.

621.396.11 **2570**

Radio Propagation over a Discontinuity in the Earth's Electrical Properties: Part 2—Coastal Refraction.—T. B. A. Senior. (*Proc. Instn elect. Engrs*, Part C, March 1957, Vol. 104, No. 5, pp. 139–147.) By adopting a model for the coastal region, analysis of Part 1 (2569 above) provides expressions for the angle of refraction appropriate to various positions of a transmitter and receiver relative to the coast.

621.396.11 : 550.372 **2571**

Influence of a Ridge on the Low-Frequency Ground Wave.—J. R. Wait & A. Murphy. (*J. Res. nat. Bur. Stand.*, Jan. 1957, Vol. 58, No. 1, pp. 1–5.) "The problem of a plane wave incident on a semielliptical boss on an otherwise perfectly conducting flat ground plane is considered. A solution in terms of elliptic wave functions is obtained. Numerical values of the field on the near and far side of this idealized ridge are given for a base width of about two-thirds of a wavelength and various ellipticity ratios."

621.396.11 : 551.510.52 **2572**

Scatter-Field Strengths and Large-Ion Concentration.—S. C. Coroniti & N. C. Gerson. (*J. atmos. terr. Phys.*, 1957, Vol. 10, No. 4, pp. 237–239.) The authors suggest that the concentration of large ions may be more sensitive than refractive index as an indicator of the more basic atmospheric parameters which influence forward-scatter signal strength.

621.396.11 : 551.510.535 **2573**

Doppler Effect in Ionospheric Propagation.—S. Borowski, S. Jasiński & S. Manczarski. (*Archiwum Elektrotech.*, 1956, Vol. 5, No. 2, pp. 343–351. English summary, pp. 352–353.) A mathematical analysis of the general problem and an analysis of experimental evidence is presented. A formula expressing the frequency variation due to motion of the ionosphere is derived on the basis of geometrical optics; calculations made using this formula predict a small frequency change only. The formula applies at frequencies not too near the m.u.f. where the ionosphere behaves like a selective low-pass filter, for waves returning to the earth, with constantly changing filter parameters. The returning signal is analysed. The treatment presented provides an explanation of the correlation between the Doppler effect and the variability of the angle of arrival of ionospheric waves.

621.396.11.029.4 : 551.510.535 **2574**

Multiple Reflections between the Earth and the Ionosphere in V.L.F. Propagation.—J. R. Wait & A. Murphy. (*Geofis. pura appl.*, Sept.–Dec. 1956, Vol. 35, No. 3, pp. 61–72. In English.) The ionosphere is treated as a sharply bounded ionized medium and, using geometrical-optical methods, reflection coefficients are calculated for very low radio frequencies. The amplitudes and phase delays of the sky waves are derived, and the computed field-strength/distance curves for 16 kc/s are compared with experimental values for an all-sea path.

621.396.11.029.62 **2575**

Long-Distance Propagation at 94·35 Mc/s over the North Sea.—R. A. Rowden & J. W. Stark. (*Proc. Instn elect. Engrs*, Part B, May 1957, Vol. 194, No. 15, pp. 210–212.) Measurements over long sea paths suggest that higher field strengths are reached for a given percentage of the overall time than for overland paths.

621.396.11.029.62 : 551.510.535 **2576**

Radio-Frequency and Scattering-Angle Dependence of Ionospheric Scatter Propagation at V.H.F.—A. D. Wheelon. (*J. geophys. Res.*, March 1957, Vol. 62, No. 1, pp. 93–112.) "The weak and fluctuating radio signals observed at distances of 1 500 km on v.h.f. are attributed to scattering from E-region turbulence. It is noted that propagation constants $k=4\pi/\lambda \sin(\theta/2)$, corresponding to the experimental frequencies (28 to 108 Mc/s), just straddle the viscosity cutoff wave-number $ks = (2 \text{ metres})^{-1}$ of the region, thereby giving a qualitative explanation for the curious dichotomy found in the experimental data. The two competitive turbulence theories are then developed in detail near the viscosity transition range." It is concluded that both theories explain recent experimental results quite well, and that therefore more precise, simultaneous measurements will be required.

621.396.11.029.63 : 523.5 **2577**

Meteor Echoes at Ultra High Frequencies.—W. A. Flood. (*J. geophys. Res.*, March 1957, Vol. 62, No. 1, pp. 79–91.) "It is proposed that, at ultra-high frequencies, underdense meteor echoes have an effective scattering length L, which is much less than a Fresnel zone. Consequently, u.h.f. meteoric echoes may be analyzed in terms of Fraunhofer diffraction theory, resulting in a relaxation of the requirement that a meteor trail be perpendicular to the radar line-of-sight before an echo can be received. Formulas for the back-scattered power, time duration, and echo rate are deduced."

RECEPTION

621.396.621 : 621.3.018.41 **2578**

The Effect on a Radio Receiver of an Input Voltage with an Instantaneous Frequency with Sawtooth Variation.—P. Poincelot. (*Ann. Télécommun.*, Dec. 1956, Vol. 11, No. 12, pp. 262–266.) The input voltage is expressed as a Fourier integral and a numerical example is given to show that erroneous results will arise from the incorrect application of the concept of instantaneous frequency. See also 2908 of 1953.

621.396.621 : 621.314.7 **2579**

Tetrajunction Transistor Simplifies Receiver Design.—R. J. Farber, A. Proudfit, K. M. St. John & C. R. Wilhelmsen. (*Electronics*, 1st April 1957, Vol. 30, No. 4, pp. 148–151.) "Dual-triode transistor, with emitter of one unit and collector of the other section of same germanium region, provides performance of two triode units with considerable reduction of circuit components when compared to two individual units. Superheterodyne receiver in which first four stages are replaced with two tetrajunction units is described."

621.396.621 : 621.376.33 **2580**

Limiters and Discriminators for F.M. Receivers: Part 4.—G. G. Johnstone. (*Wireless World*, June 1957, Vol. 63, No. 6, pp. 275–280.) The operating principles and practical design of the gated beam discriminator, the synchrotector and the counter circuit are described. Part 3: 2277 of July.

621.396.621 : 621.376.33 **2581**

Methods of Compensating Quasi-static and Dynamic Distortions in the I.F. Section of F.M. Receivers.—E. G. Woschni. (*NachrTech.*, Nov. 1956, Vol. 6, No. 11, pp. 488–491.) Brief outline of possibilities of eliminating nonlinear distortion by appropriate filter design.

621.396.621.22 : 621.375.121.2 **2582**

Distributed Amplifiers as Antenna Multicouplers.—Pfund, Jr. (See 2388.)

621.396.621.59 **2583**

Reception of Pulse Trains of Arbitrary Shape by a Superregenerator.—L. R. Yavich. (*Radiotekhnika i Elektronika*, Nov. 1956, Vol. 1, No. 11, pp. 1419–1427.) Theoretical treatment.

621.396.621.59 **2584**

Analysis of Superregenerator Circuit with Quenching causing Frequency Modulation.—F. Wiśniewski. (*Archiwum Elektrotech.*, 1956, Vol. 5, No. 2, pp. 263–289. English summary, pp. 290–292.)

621.396.8 **2585**

The Effect of Fading on Communication Circuits Subject to Interference.—F. E. Bond & H. F. Meyer. (*Proc. Inst. Radio Engrs*, Part 1, May 1957, Vol. 45, No. 5, pp. 636–642.) The statistical performance of such circuits is analysed for the cases where either the desired or undesired signal or both are subject to Rayleigh fading over the propagation path and the improvement which dual diversity reception offers is discussed.

621.396.812.029.62 **2586**

Observations of Long-Distance Reception in the 3-m Broadcast Band.—L. Klinker. (*Hochfrequenztech. u. Elektroakust.*, Nov. 1956, Vol. 65, No. 3, pp. 77–86.) Analysis of field-strength measurements for eight transmission paths of 100 to 500 km length made at Kühlungsborn since 1951 [see also 248 and 538 of 1955 (Lauter & Klinker)] confirms the validity of the C.C.I.R. curves for the Central European area. Diurnal field-strength variations have a mean amplitude of 10 dB in summer, and seasonal propagation changes are only noticeable for a sea path, with slight improvement during early summer. Fading range, day-by-day changes and the mean amplitude of diurnal variations reach a maximum for about 200 km path length; day-by-day changes are greatest in winter.

621.396.812.3 **2587**

The Effect of the Correlation between the Received Field Strengths in Diversity Reception.—K. H. Schmelovsky. (*Hochfrequenztech. u. Elektroakust.*, Nov. 1956, Vol. 65, No. 3, pp. 74–76.) A formula is derived for calculating approximately the probability that the field strengths at two spaced aerials will drop simultaneously below a given value in cases of low to medium correlation between the two signals.

621.396.822 **2588**

The Influence of Threshold Action on the R.M.S. Value of Input Gaussian Noise.—R. D. Teasdale & A. H. Benner. (*Proc. Inst. Radio Engrs*, Part 1, May 1957, Vol. 45, No. 5, p. 697.)

621.396.822 : 621.317.328 **2589**

Logarithmic Amplifier measures Noise.—Wells. (See 2547.)

STATIONS AND COMMUNICATION SYSTEMS

621.376.3 : 621.3.018.78 : 621.372.5 **2590**

Frequency-Modulation Distortion in Linear Networks.—A. S. Gladwin: R. F. Brown. (*Proc. Instn elect. Engrs*, Part B, May 1957, Vol. 104, No. 15, pp. 264.) Comment on 1581 of May and author's reply.

621.396.3 **2591**

Binary Data Transmission Techniques for Linear Systems.—M. L. Doelz, E. T. Heald & D. L. Martin. (*Proc. Inst. Radio Engrs*, Part 1, May 1957, Vol. 45, No. 5, pp. 656–661.) Problems associated with utilizing fully the binary data transmission potential of an s.s.b. voice channel, including stability of the propagation media and multipath reception, are discussed. Equipment designed to transmit 3 000 bits/sec is described.

621.396.3 **2592**

The Frequency [of occurrence] **of Errors in A1- and F1-Telegraphy Transmission Systems, particularly in the Presence of White Noise.**—H. Beger. (*Telefunken Ztg*, Dec. 1956, Vol. 29, No. 114, pp. 245–255. English summary, p. 294.) Report of investigations carried out to compare the performance of on-off and frequency-shift keying systems. Tests were made on teleprinter systems to determine the influence of transmission bandwidth, modulation index and limiting level. For transmissions at constant level with white-noise interference both methods of modulation are almost equivalent. The importance of maintaining the optimum limiting level to minimize errors is shown; during fading this is only possible for F1 operation. Stability and accurate alignment of transmitter and receiver are necessary to secure the maximum advantage of the F1 system.

621.396.71.029.55 (43) **2593**

Modern High-Power Transmitting Stations for the Short-Wave Overseas Service.—A. Heilmann. (*Telefunken Ztg*, Dec. 1956, Vol. 29, No. 114, pp. 225–236. English summary, pp. 292–293.) A survey with details of the telegraphy transmitter of the German Post Office near Frankfurt (Main).

621.396.712 : 621.376.3 **2594**

The B.B.C. Sound Broadcasting Service on Very High Frequencies.—E. W. Hayes & H. Page. (*Proc. Instn elect. Engrs*, Part B, May 1957, Vol. 104, No. 15, pp. 213–224. Discussion, pp. 249–253.) This paper describes the developments leading to the inauguration of B.B.C. sound broadcasting in the band 87·5–100 Mc/s. The need for the use of v.h.f. and the choice of frequency modulation are explained. By 1958 96% of the population of Great Britain and Ireland will be covered by this service. The experience of the first year's operation is described, with special reference to the performance of commercial receivers.

621.396.712.029.55 **2595**

New Short-Wave Transmitters for Single-Sideband Telephony.—W. Burkhardtsmaier. (*Telefunken Ztg*, Dec. 1956, Vol. 29, No. 114, pp. 236–244. English summary, p. 293.) Description and some design and test data of the 20/100-kW transmitters for the 3·3–26·4-Mc/s frequency range installed at Usingen by the West-German Post Office.

621.396.73 : 621.396.6 **2596**

Superregenerative Transistor Transceiver.—W. F. Chow. (*Electronics*, 1st April

1957, Vol. 30, No. 4, pp. 180–182.) The transceiver which employs three transistors, uses a tetrode transistor as a 52-Mc/s oscillator for transmission and as a 'forced quench' superregenerative detector for reception. The range of communication is half a mile.

621.396.933 **2597**

Aircraft Radiophone speeds Communications.—B. R. Rashkow. (*Electronics*, 1st April 1957, Vol. 30, No. 4, pp. 164–168.) The increase of transoceanic air traffic has necessitated the replacement of W/T by R/T. The article discusses the operational, equipment and propagation problems involved.

SUBSIDIARY APPARATUS

621-52 **2598**

The Describing-Function Analysis of a Nonlinear Servomechanism Subjected to Stochastic Signals and Noise.—P. N. Nikiforuk & J. C. West. (*Proc. Instn elect. Engrs*, Part C, March 1957, Vol. 104, No. 5, pp. 193–203.) A method is given for evaluating the response of a specific type of nonlinear mechanism to random signals, and to sinusoidal and random signals contaminated by noise.

621.314.5 : 621.373.52 **2599**

Design Considerations of Junction-Transistor Oscillators for the Conversion of Power from Direct to Alternating Current.—F. Oakes. (*Proc. Instn elect. Engrs*, Part B, May 1957, Vol. 104, No. 15, pp. 307–317. Discussion, pp. 333–336.) Basic principles and graphical methods of design are given. The design of a Class-B oscillator is described in detail and illustrated by means of a practical numerical example. Calculated data agreed well with practical measurements.

621.316.726 : 621.376.3 **2600**

A New Control Element for Automatic Frequency Retuning.—A. Karaminkov. (*Nachr Tech.*, Nov. 1956, Vol. 6, No. 11, pp. 497–500.) The device described performs the function of reactance valve and motor in automatic tuning control equipment for u.s.w. f.m. transmitters or receivers. It consists of a suitably damped moving-coil instrument assembly, without control springs, which drives the vane of an air capacitor. The frequency stability is claimed to be within about 1 part in 10^6.

TELEVISION AND PHOTOTELEGRAPHY

621.397 : 621.39.001.11 **2601**

Bandwidth Compression of a Television Signal.—G. G. Gouriet. (*Proc. Instn elect. Engrs*, Part B, May 1957, Vol. 104, No. 15, pp. 265–272.) "Two sets of data are fundamentally required to describe a television picture, one giving the significant changes of brightness, and the other the positions of such changes. The total information content is calculated according to Shannon, and the means are discussed for reducing bandwidth by redistributing the data in time so as to achieve a constant rate of transmission."

621.397.001.4 : 535.623 : 621.317.755 **2602**

The Vectorscope.—N. N. Parker Smith & C. J. Matley. (*Electronic Radio Engr*, June 1957, Vol. 34, No. 6, pp. 198–206.) The instrument displays as a pattern of vectors the chrominance component of the N.T.S.C. type of signal in which the component is transmitted by means of a subcarrier modulated both in amplitude and phase.

621.397.331.2 : 535.215 : 546.863.221 **2603**

Semiconductor Photosensitive Layers for Photoresistance Television Tubes.—Oksman. (See 2458.)

621.397.5 **2604**

High Definition on 405 Lines.—(*Wireless World*, June 1957, Vol. 63, No. 6, pp. 254–255.) Spot wobble at a frequency of 6 Mc/s is synchronised in frequency and phase between camera and monitor c.r. tube. Primarily suggested for suppressed-frame telerecording, it could be used for a high-definition service compatible with the existing 405-line standard.

621.397.5 : 621.396.41 **2605**

Bilingual Television by Pulse-Multiplex System.—B. Pouzols. (*Électronique, Paris*, Dec. 1956, No. 121, pp. 21–25.) Outline of a single-channel multiplex system with p.a.m. for providing alternative sound transmissions. The circuits described are suitable for French standards. Tests show that crosstalk is negligible and only small modifications to normal receivers are required. See also 2300 of July (Dubec).

621.397.5.001.4 : 535.623 **2606**

Test Signal for Measuring 'On-the-Air' Colour-Television System Performance.—R. C. Kennedy. (*RCA Rev.*, Dec. 1956, Vol. 17, No. 4, pp. 553–557.) Description of a signal used by the N.B.C. which provides a reference level for the amplitude adjustment of the various video signals and permits the measurement of differential gain and phase distortion.

621.397.61 **2607**

Transistorized Television Cameras using the Miniature Vidicon.—L. E. Flory, G. W. Gray, J. M. Morgan & W. S. Pike. (*RCA Rev.*, Dec. 1956, Vol. 17, No. 4, pp. 469–502.) Description of miniature camera and portable outside-broadcasting equipment. See also 1925 of June and 2610 below.

621.397.61 **2608**

A Video Automatic-Gain-Control Amplifier.—J. O. Schroeder. (*RCA Rev.*, Dec. 1956, Vol. 17, No. 4, pp. 558–570.) Design and performance data of an amplifier successfully used by the N.B.C. are discussed.

621.397.611 : 535.623 **2609**

Camera Tubes for Colour-Television Broadcast Service.—R. G. Neuhauser. (*J. Soc. Mot. Pict. Telev. Engrs*, Dec. 1956, Vol. 65, No. 12, pp. 636–642.) The general characteristics required for colour-television tubes are discussed and compared with the characteristics of the vidicon and image orthicon.

621.397.611.2 **2610**

A Miniature Vidicon of High Sensitivity.—A. D. Cope. (*RCA Rev.*, Dec. 1956, Vol. 17, No. 4, pp. 460–468.) The experimental tube described is 3 in. long; its diameter is ½ in. A photoconductive layer of increased sensitivity is used covering a range from 400 to 800 mμ with a maximum near 600 mμ. For applications of this tube, see 2607 above.

621.397.62 **2611**

Television Receiver for Metre Wavelengths.—V. Biggi. (*Onde élect.*, Dec. 1956, Vol. 36, No. 357, pp. 1021–1030 & Jan. 1957, Vol. 37, No. 358, pp. 55–67.) The design and application are discussed of a receiver capable of driving a relay transmitter for long-range television transmission. A special a.g.c. circuit for positive modulation systems is described.

621.397.62 : 539.234 : 546.45 **2612**

The Properties of Thin Beryllium Foils, and their Application to Light Modulation and Television.—M. Auphan. (*Onde élect.*, Dec. 1956, Vol. 36, No. 357, pp. 1040–1045.) A method is described of obtaining foils of Be 200 to 1 000 Å thick tightly drawn across a supporting graticule in close proximity of a glass screen. Electrostatic charges deposited on such a screen inside a c.r.t. will result in local deflections of the foil thus providing a means of modulating light by reflection. The application to monochrome television is outlined; experiments have been made using a foil specially corrugated to suit the optical system. Colour television methods are discussed, including a compatible projection system with two tubes using foil screens: one low-definition tube reproduces colour, the other provides the high-definition luminance detail.

621.397.62 : 621.3.049.7 **2613**

Mechanized Production of TV Wiring Boards.—J. Markus. (*Electronics*, 1st April 1957, Vol. 30, No. 4, pp. 138–143.) Description of the latest semi-automatic machines and assembly lines for manufacturing printed circuits in use by the Philco Corporation. This article brings up to date an earlier report (3811 of 1955).

621.397.62 : 621.373.4 **2614**

Analytical Approaches to Local-Oscillator Stabilization.—W. Y. Pan & D. J. Carlson. (*RCA Rev.*, Dec. 1956, Vol. 17, No. 4, pp. 534–552.) The analytical treatment of frequency stability under complex thermal conditions indicated by Yau (593 of 1956) is applied to two commercial television receiver circuits. Results show that satisfactory stabilization even for frequencies beyond 1 kMc/s can be achieved with conventional temperature-sensitive devices.

621.397.621 : 621.373.444.1 **2615**

C.R.T. Deflection Circuit has High Efficiency.—Guggi. (See 2382.)

621.397.621.2 : 535.623 : 621.385.832 **2616**

Methods of Local [post-deflection] **Spot Position Control for Electron Beams.**—U. Pellegrini. (*Alta Frequenza*, Dec. 1956, Vol. 25, No. 6, pp. 482–504.) Two methods of accurately focusing the beam in a colour c.r. tube are analysed; in one a wire grid close to the phosphor screen is used, as e.g. in the post-deflection-focus colour kinescope [3894 of 1956 (Carpenter)]; in the other the screen consists of separate aluminized strips. Differential equations are derived for the electron trajectories, and the equipotential contours and lines of flux are plotted.

621.397.8 **2617**

Television Interference from Sea Reflections.—J. K. S. Jowett. (*Wireless World*, June 1957, Vol. 63, No. 6, pp. 262–266.) A possible explanation of the flutter experienced at certain localities served by the North Hessary Tor transmitter. Quasi-specular reflection at oblique incidence may occur from sea wave fronts and beat with the direct signal attenuated by local hills. The flutter frequency is consistent with the sea wave velocity.

621.397.8 **2618**

Reduction of Co-channel Television Interference by Precise Frequency Control of Television Picture Carriers.—W. L. Behrend. (*RCA Rev.*, Dec. 1956, Vol. 17, No. 4, pp. 443–459.) In offset carrier-frequency operation interference minima occur at some offset frequencies which are multiples of the frame frequency. To assess the importance of precise frequency control subjective tests were made of interference visibility and the results of field tests with experimental equipment are reported.

621.397.81 **2619**

A Method of Predicting the Coverage of a Television Station.—J. Epstein & D. W. Peterson. (*RCA Rev.*, Dec. 1956, Vol. 17, No. 4, pp. 571–582.) A method of estimating the median field strength along a radial line from the transmitter is described. The estimate is based on the plotted elevation profile and the use of theoretical and empirical curves resulting from the analysis of extensive field surveys and investigations (see e.g. 2423 of 1953). The entire v.h.f. and u.h.f. spectrum is covered and the accuracy so far achieved is adequate for surveys involving large numbers of stations.

621.397.813 **2620**

Frequency - Dependent Equalization of Gradation for Television Signals.—H. Schonfelder. (*Arch. elekt. Übertragung*, Dec. 1956, Vol. 10, No. 12, pp. 512–534.) By limiting the equalization to the lower frequencies of the video signal a picture of acceptable quality can be obtained; the signal/noise ratio in the dark portions is considerably improved in comparison with that for uniform equalization of gradation. An upper limiting frequency of about 3 Mc/s is chosen according to system conditions and the amount of distortion tolerable. The design of a suitable equalizing circuit is detailed and experimental results are illustrated by oscillograms and reproductions of test pictures.

TRANSMISSION

621.376.3 : 621.373.421 **2621**

Frequency-Modulated Quartz Oscillators for Broadcasting Equipment.—Mortley. (See 2375.)

621.396.712 : 621.376.3 **2622**

Frequency-Modulated V.H.F. Transmitter Technique.—A. C. Beck, F. T. Norbury & J. L. Storr-Best. (*Proc. Instn elect. Engrs*, Part B, May 1957, Vol. 104, No. 15, 225–238. Discussion, pp. 249–253.) The general planning and design of the complete v.h.f. transmitting equipment used by the B.B.C. on its sites is surveyed, including problems of unattended operation, automatic phasing of paralleled amplifiers and 3 - program common - aerial working; special reference is made to a current 10 kW f.m. transmitter operating in band II (87·5–100 Mc/s).

VALVES AND THERMIONICS

621.314.63 **2623**

The Apparent Contact Potential of a Pseudo-Abrupt *p-n* Junction.—H. Kroehmer. (*RCA Rev.*, Dec. 1956, Vol. 17, No. 4, pp. 515–521.) Junctions are studied which have constant impurity densities on both sides without abrupt transition. A finite region of transition within the space-charge region is assumed. The difference between the 'apparent' contact potential found from capacitance measurements and the true potential calculated from the impurity densities provides quantitative information about the internal structure of the junction.

621.314.63 **2624**

On the Tail in the Transient Behaviour of Point-Contact Diodes.—H. L. Armstrong. (*Proc. Inst. Radio Engrs*, Part 1, May 1957, Vol. 45, No. 5, pp. 696–697.) Although the main part of the transient response of small hemispherical *p-n* junctions is almost independent of lifetime, there is a 'tail' which is almost exponential and is determined by the lifetime.

621.314.63 : 537.311.33 **2625**

Minority-Carrier Storage in Semiconductor Diodes.—J. C. Henderson & J. R. Tillman. (*Proc. Instn elect. Engrs*, Part B, May 1957, Vol. 104, No. 15, pp. 318–333. Discussion, pp. 333–336.) Transient reverse-current effects in planar and hemispherical diodes of *n*-type base material are analysed under various conditions, although the extent to which the analyses apply to point-contact diodes remains debatable. Experiments to test the theory and to deduce the hole lifetimes are given.

621.314.63 : 537.311.33 **2626**

Rectification Properties of Metal-Silicon Contacts.—E. C. Wurst, Jr, & E. H. Borneman. (*J. appl. Phys.*, Feb. 1957, Vol. 28, No. 2, pp. 235–240.) The rectifying properties of metal/silicon contacts are related to the work function of the metal. The measured reverse saturation currents of metal/silicon diodes are greater than predicted values; the discrepancy is not yet explained.

621.314.632 **2627**

Strain-Energy Calculations in the Design of Cat's Whiskers for Semiconductor Devices.—S. J. Morrison. (*Proc. Instn elect. Engrs*, Part C, March 1957, Vol. 104, No. 5, pp. 148–153.) A general theory of contact pressures and deflections is derived and experiments on large-scale whisker models have shown close agreement with theory.

621.314.632 : 546.289 **2628**

A Millisecond Relaxation Process in the Reverse Current of Germanium Point-Contact Diodes.—R. E. Burgess. (*Brit. J. appl. Phys.*, Feb. 1957, Vol. 8, No. 2, pp. 62–63.) When a reverse voltage is suddenly applied to a diode initially having zero bias a decrease of current to the final steady state is usually found, the relaxation time being about 1 ms. It is suggested that redistribution of surface charge is the most likely cause.

621.314.7 **2629**

Transistor Junction Temperature as a Function of Time.—K. E. Mortenson. (*Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, pp. 504–513.) The step and impulse temperature responses are obtained for a one-dimensional heat flow model and the junction temperature determined for periodic rectangular pulse excitation. The maximum, average and minimum temperatures are plotted in terms of duty cycle, p.r.f. and thermal time constant, and the theoretical results confirmed experimentally. The maximum junction temperature can be several times the average at low duty cycles and repetition rates.

621.314.7 **2630**

Temperature Dependence of Junction Transistor Parameters.—W. F. Gärtner. (*Proc. Inst. Radio Engrs*, Part 1, May 1957, Vol. 45, No. 5, pp. 662–680.) Four representative types of transistor are considered, namely Ge *p-n-p* alloy, Ge *n-p-n* grown, Ge *n-p-n* rate-grown, and Si *n-p-n* grown types. The temperature dependence of their characteristics is calculated from existing theories of transistor performance and the known temperature behaviour of the semiconductor properties. The results are expressed in terms of the four-pole and equivalent-circuit parameters.

621.314.7 **2631**

Point-Contact Transistor Studies using Radioactive Collectors.—D. Haneman & A. J. Mortlock. (*Proc. phys. Soc.*, 1st Jan. 1957, Vol. 70, No. 445B, pp. 145–147.) Pile-activated pure antimony points containing ^{122}Sb in contact with single crystals of *n*-type Ge were formed. The activity transferred and the surface distribution were then measured.

621.314.7 **2632**

The Characteristics and the Charge-Carrier Distribution of Alloy-Junction Transistors.—W. Engbert. (*Telefunken Ztg*, Dec. 1956, Vol. 29, No. 114, pp. 277–287. English summary, p. 295.) The distribution of carrier and hole densities and potential in the *p-n-p* regions is derived from the transistor characteristics and the amount of doping.

621.314.7 : 621.396.621 **2633**

Tetrajunction Transistor Simplifies Receiver Design.—Farber, Proudfit, St. John & Wilhelmsen. (See 2579.)

621.314.7.012.8 **2634**

The Effect of Non-ideal Emitter Junctions on the Behaviour of Junction Transistors.—E. Baldinger, W. Czaja & M. Nicolet. (*Helv. phys. Acta*, 15th Dec. 1956, Vol. 29, Nos. 5/6, pp. 428–430. In German.) In the transistor equivalent circuit [see e.g. 1193 of 1955 (Giacoletto)] the emitter 'lead' resistance R_e, introduced to allow for the difference between theory and experimental behaviour, varies according to the operating point. By modifying the slope S by a factor m between 0·5 and 1, a value of R_e constant over the whole range can be assumed.

621.383.2 : 621.385.83 **2635**

Shutter Image-Converter Tubes.—B. R. Linden & P. A. Snell. (*Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, pp. 513–523.) A triode type is described which uses a mesh, near the cathode, to gate the photo-electron emission. The electron-optical theory is presented for electrostatically and magnetically focused types and illustrated by practical results.

621.383.27 **2636**

Optical Feedback in Photomultipliers.—R. Gerharz. (*J. Electronics*, March 1957, Vol. 2, No. 5, pp. 409–415.) The optical feedback to the photo cathode, arising from electro-fluorescence induced by the dark-current, has been investigated for commercial photomultipliers Type 1P21 and 931A.

621.383.27(47) **2637**

Commercial Types of Multistage Photomultipliers.—I. Ya. Breido, B. M. Glukhovskoi & L. G. Leiteizen. (*Radiotekhnika i Elektronika*, Oct. 1956, Vol. 1, No. 10, pp. 1344–1356.) Description of Russian 'commercial' types with characteristic curves and tabulated data.

621.383.42.011.4 **2638**

On the Capacitance of Selenium Photocells.—M. Bichara. (*C. R. Acad. Sci., Paris*, 4th Feb. 1957, Vol. 244, No. 6, pp. 742–744.) Results of measurements on 46 samples are tabulated.

621.383.5 : 537.311.33 **2639**

A New Semiconductor Photocell using Lateral Photoeffect.—J. T. Wallmark. (*Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, pp. 474–483.) A non-uniform illumination produces a lateral voltage parallel to a germanium-indium junction. The output voltage is a linear function of the position of the light spot and the sensitivity is approximately 200 μA/lumen. The spot position and the direction of the light beam can be measured to less than 100 Å and 0·1 seconds of arc respectively.

621.385-71 **2640**

Vapotron Technique.—C. Beurtheret. (*Rev. tech. Comp. franç. Thomson-Houston*, Dec. 1956, No. 24, pp. 53–83.) The development of this method of cooling high-power transmitting valves is outlined (see also 542 of 1952) together with the underlying theory of heat transfer. Results of tests on various forms of anode dissipating surface and for various metals are discussed. Illustrations show the evolution of the vapotron from 1950 to date, including associated equipment. 32 references.

621.385 : 621.317.331 **2641**

A New Method for the Detection of Thin Conducting Films in Thermionic Valves.—F. H. Reynolds & M. W. Rogers. (*Proc. Instn elect. Engrs*, Part B, May 1957, Vol. 104, No. 15, pp. 337–340.) A system of three electrodes is painted on the outside of the bulb over the region where the film is to be detected, and connected to an impedance bridge.

621.385 : 621.318.57.032.2 **2642**

Improved Keep-Alive Design for T.R. Tubes.—L. Gould. (*Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, pp. 530–533.) By insulating the cone wall and using a stainless-steel cathode a life in excess of 1 000 h is possible.

621.385.029.6 **2643**

International Congress on Microwave Valves.—(*Onde élect.*, Nov. 1956, Vol. 36, No. 356, pp. 866–996.) Selection of papers presented at the 1956 Congress in Paris. See 1267 of April.

621.385.029.6 **2644**

The Assembly of Microwave Valves by Brazing.—R. Paliès. (*Rev. tech. Comp. franç. Thomson-Houston*, Dec. 1956, No. 24, pp. 41–51.) The method of brazing by resistance heating is described and illustrated by two examples of its application.

621.385.029.6 : 621.317.36 **2645**

The Measurement of Magnetron Frequency Pulling.—J. R. G. Twisleton. (*Proc. Instn elect. Engrs*, Part C, March 1957, Vol. 104, No. 5, pp. 8–12.) The theoretical frequency pulling when multiple reflected waves exist in the output feeder is estimated and the performance of a typical testing waveguide system is examined.

621.385.029.6 : 621.396.822 **2646**

Experiments on Noise Reduction in Backward-Wave Amplifiers.—M. R. Currie & D. C. Forster. (*Proc. Inst. Radio Engrs*, Part 1, May 1957, Vol. 45, No. 5, p. 690.) Using a tube with a gun structure which produces an annular electron beam noise figures approaching the theoretical ones were obtained.

621.385.029.65 **2647**

Backward-Wave Oscillator Experiments at 100 to 200 Kilomegacycles.—A. Karp. (*Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, pp. 496–503.) The circuit structure consists of a ridged waveguide with transverse slots in the broad wall. For electron beam velocities between 650 and 2 700 V and current densities between 3 and 10 A/cm², probable power outputs of a few tenths of a milliwatt were obtained.

621.385.032.3 **2648**

High-Temperature Properties of Tungsten which influence Filament Temperatures, Lives and Thermionic-Emission Densities.—R. N. Bloomer. (*Proc. Instn elect. Engrs*, Part B, March 1957, Vol. 104, No. 14, pp. 153–157.) Calculations using reliable published data are compared with experimental results.

621.385.3 **2649**

On the Amplification Factor of the Triode.—E. B. Moullin. (*Proc. Instn elect. Engrs*, Part C, March 1957, Vol. 104, No. 5, pp. 222–232.) A new treatment of the characteristic of the triode, as expressed by the equation $I_a = f(V_a + \mu V_g)$. The limitations in the conventional method are discussed. See also 1985 of June (Hammond).

621.385.5 : 621.373.422 **2650**

Transitron Negative Resistance.—A. G. Bogle. (*Electronic Radio Engr*, May 1957, Vol. 34, No. 5, pp. 170–174.) A detailed investigation shows that the negative resistance is approximately inversely proportional to the cathode current within ±10% over the range 3–90 kΩ.

MISCELLANEOUS

621.3.002.2 **2651**

Electronic Subminiaturization Techniques.—(*Tech. News Bull. nat. Bur. Stand.*, Jan. 1957, Vol. 41, No. 1, pp. 3–5.) Techniques developed in connection with the program for the U.S. Navy Bureau of Aeronautics are discussed; items mentioned include a commutator-type gain control using a tape resistor on a glass tube carrying longitudinal stripes of conducting paint, a temperature-compensated permeability-tuned inductor, high-temperature litz wire, and i.f. transformers with ferrite sleeve and tuning screw.

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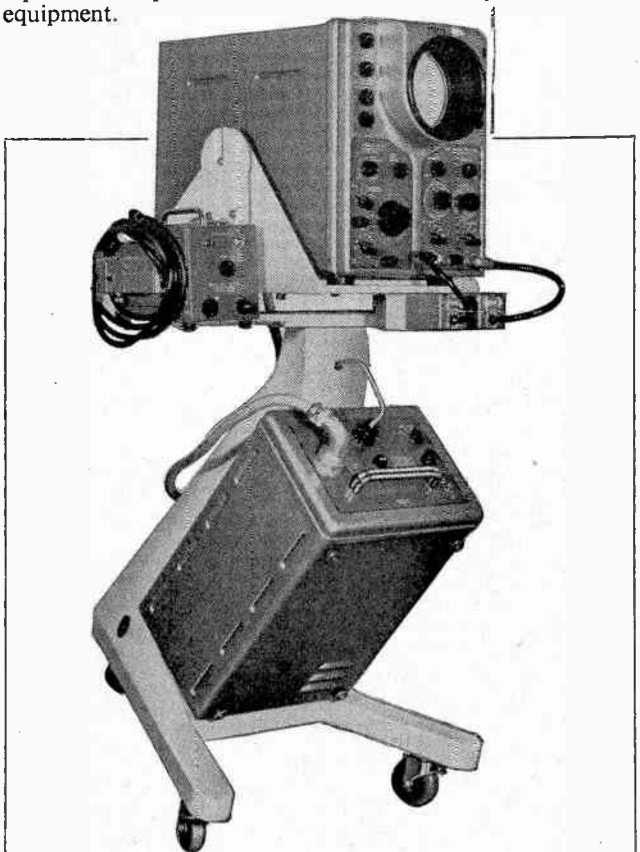

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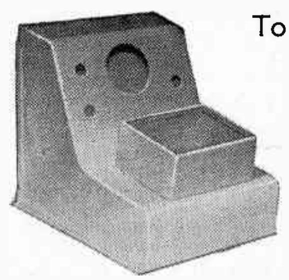

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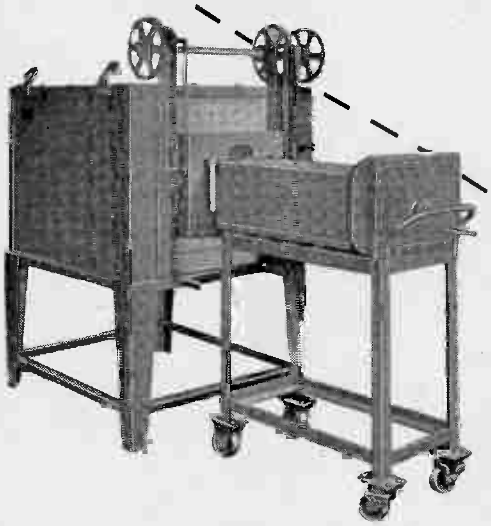

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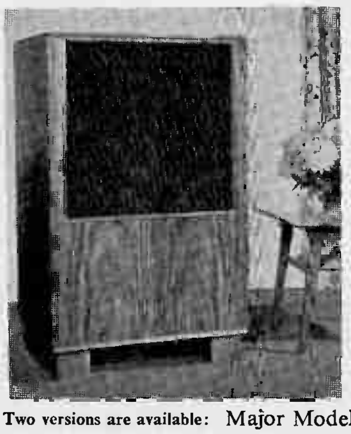

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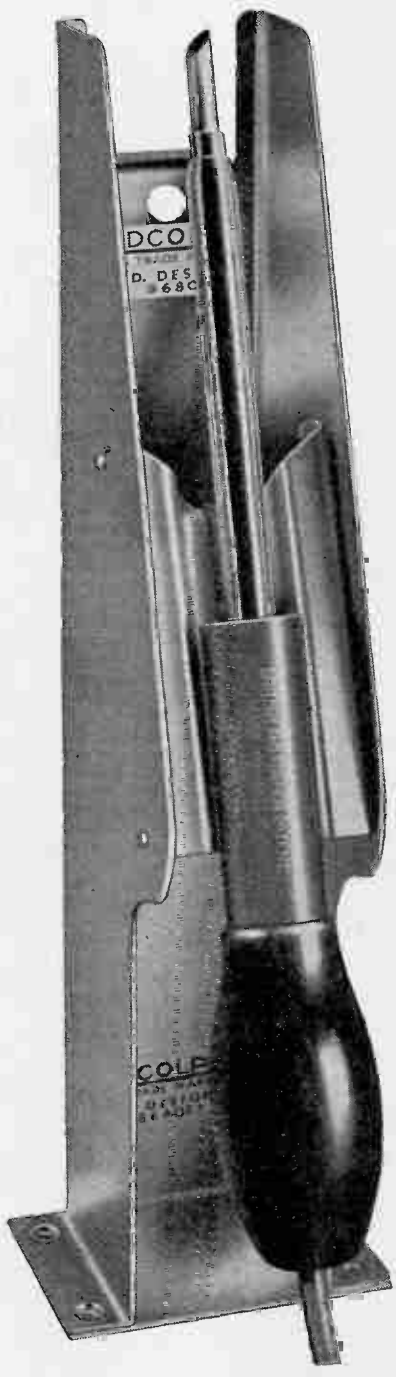
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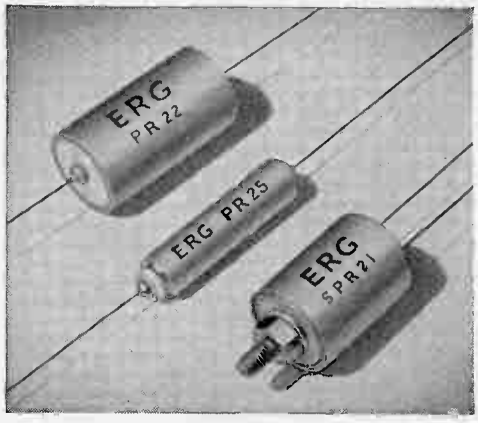

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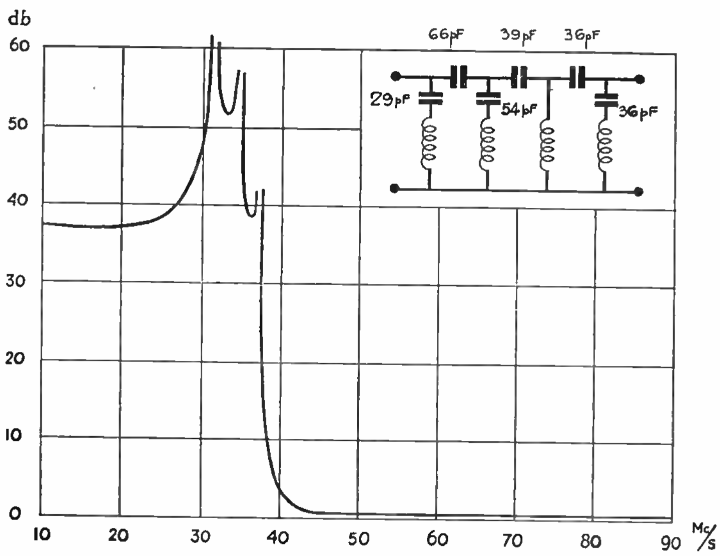

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