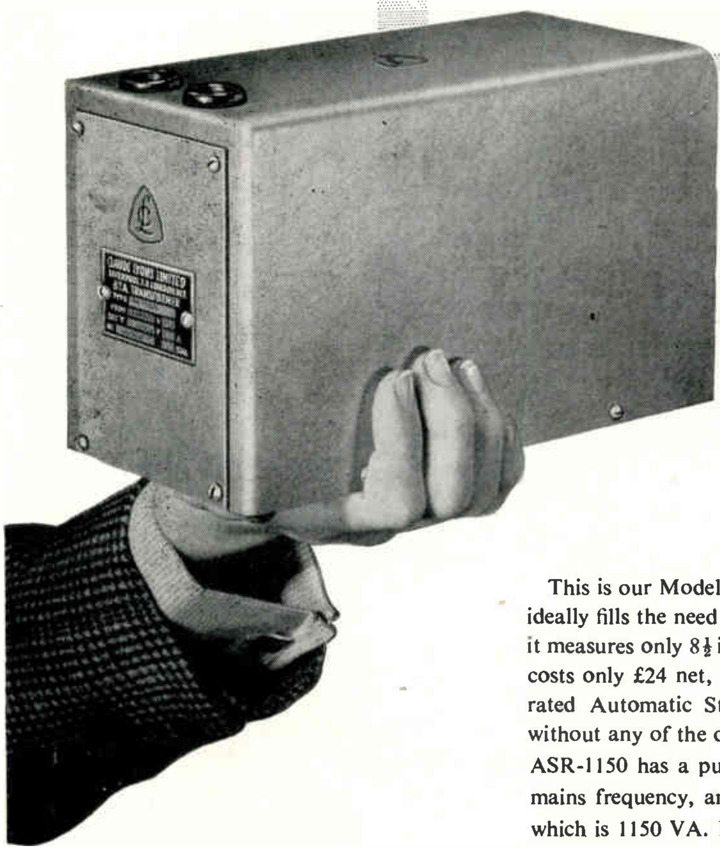
# WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

## JANUARY 1956

VOL. 33 No. 1 · THREE SHILLINGS AND SIXPENCE

# A NEW-PRINCIPLE A-C AUTOMATIC VOLTAGE STABILISER



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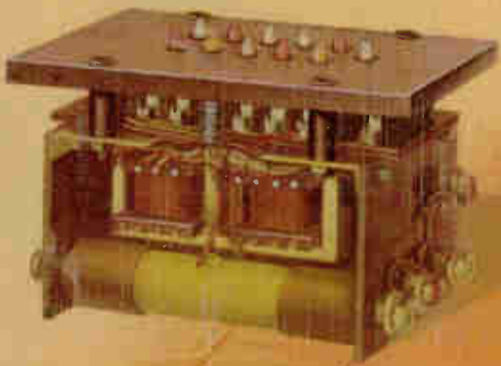

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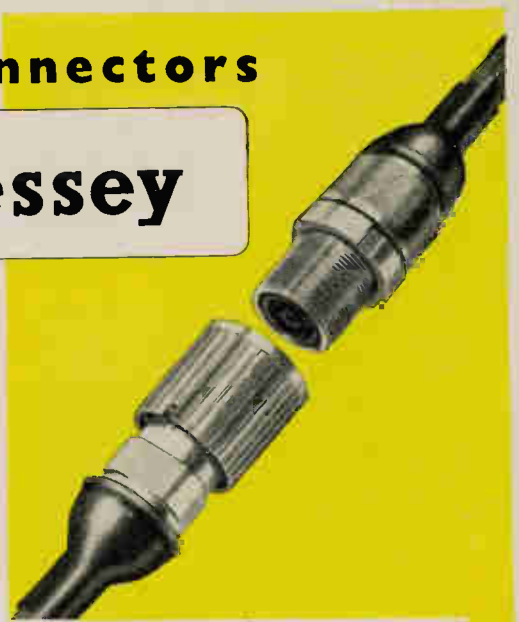

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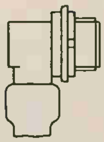

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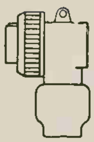

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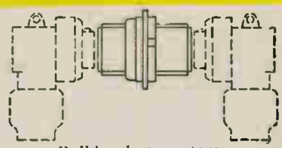

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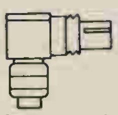
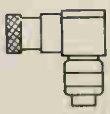
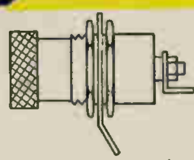


Cable plug minor

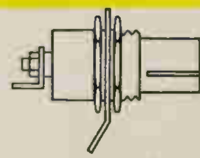


Cable socket minor

Cable plug major



Panel socket major



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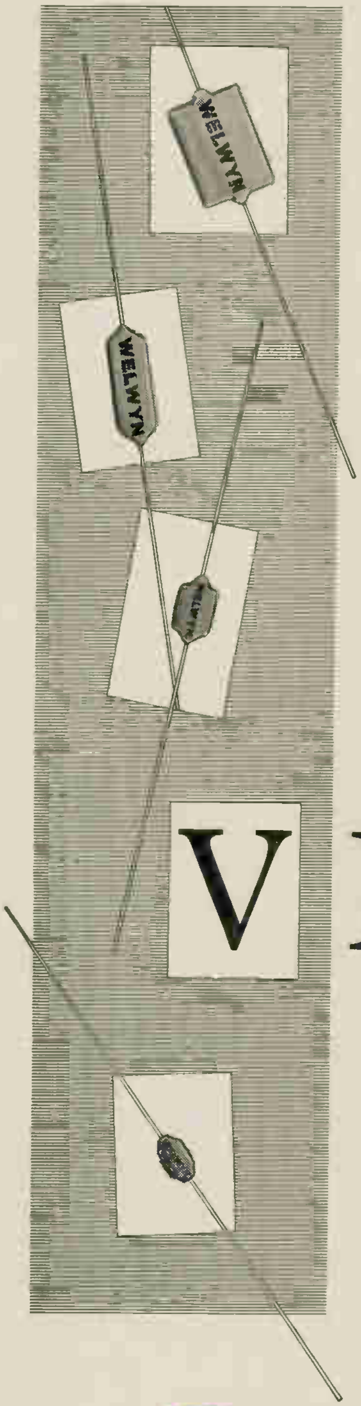
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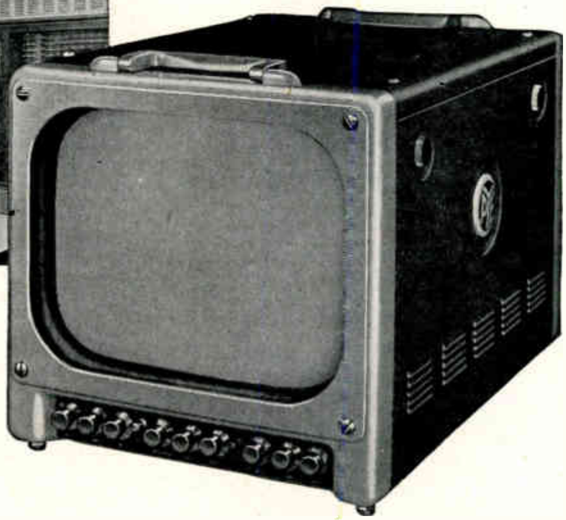
*Photograph by courtesy of the B.B.C.*



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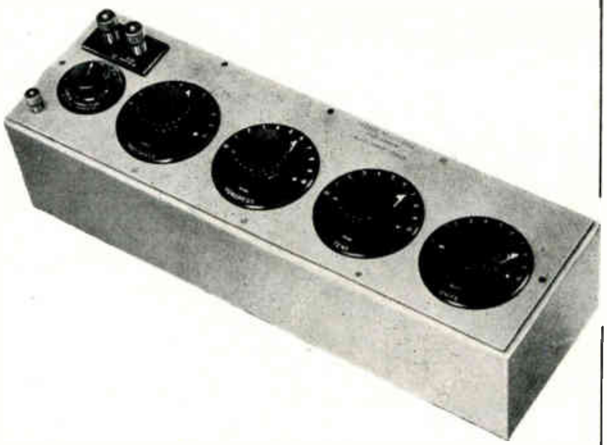

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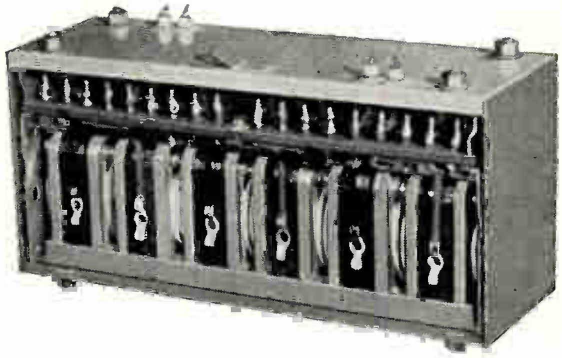

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# WIRELESS ENGINEER

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## Non-Linearity Distortion

IN television, it is a common practice to compensate for the unwanted effect of some non-linear stage by introducing an inverse non-linearity at some other point. We have seen one or two statements recently to the effect that such a process cannot be applied to audio-frequency amplifiers. We cannot see any reason why this should be so and we think the idea arises because audio engineers have the habit of thinking in terms of harmonics and combination frequencies instead of changes of waveform. A non-linear stage produces harmonics and it does seem rather incredible that a second such stage, fed with all these frequencies, should be able to produce new frequencies of precisely the right amplitudes and in the right phases for cancelling them.

We propose to examine the matter in some detail, therefore, and, to anticipate a little, it is interesting to find that in some degree this process of correcting one non-linearity by another does actually occur in every cascade amplifier. Being uncontrolled and often unsuspected its effect is rarely large, but it does result in the overall distortion of an amplifier being less than the distortion of individual stages.

Let us first of all consider an amplifier valve having the peculiar and unreal characteristic shown in Fig. 1. Over the region AB, the characteristic is quite linear and has a mutual conductance $g_{m1}$, say. Over the region BC it is also linear and has a higher mutual conductance, $g_{m2}$. The non-linearity is the abrupt change of slope at B. Let us now bias the valve to this point B and apply a signal $v \sin \omega t$ to the grid.

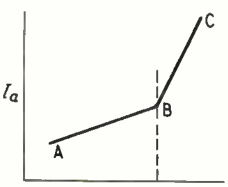


Fig. 1.

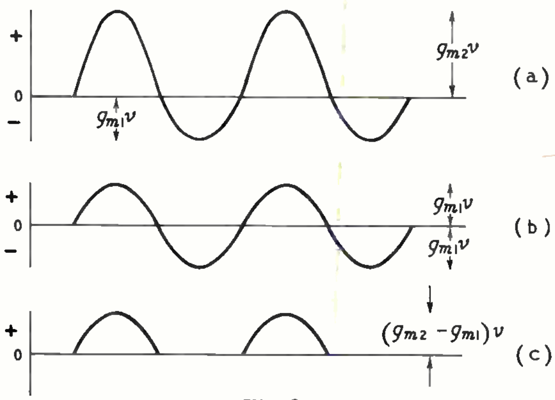


Fig. 2.

The positive excursions of the signal will all fall on BC and the negative on AB, so that the positive half-cycles of anode current will be the positive values of $g_{m2}\, v \sin \omega t$ and the negative will be the negative values of $g_{m1}\, v \sin \omega t$. The shapes of all half-cycles will be perfectly sinusoidal, but the positive ones will have a greater amplitude than the negative. The waveform is shown in Fig. 2(a) and is equivalent to the sum of (b) a sine wave $g_{m1}\, v \sin \omega t$ and (c) a rectified (half-wave) sine wave of amplitude $(g_{m2} - g_{m1})\, v$.

The Fourier series for this last is well-known to be

$$(g_{m2} - g_{m1})\, v \left[\frac{1}{\pi} + \frac{1}{2} \sin \omega t - \frac{2}{\pi}\left(\frac{1}{3} \cos 2\omega t + \frac{1}{15} \cos 4\omega t \ldots\right)\right]$$



B

and so the series for (a) is this series plus $g_{m1} v \sin \omega t$.

To correct the distortion of Fig. 2(a) we clearly need a further stage with a similar type of characteristic, but with the lower value of mutual conductance acting on the positive half-cycles and the higher value on the negative. Fig. 2(a) represents an anode-current waveform, however, and if we use resistance coupling to another valve, the anode-voltage waveform will be inverted. The signal fed to the grid of a succeeding valve will thus have its positive half-cycles smaller. It follows that the characteristic required for this second stage is identical with that of the valve which has caused the distortion!

If the coupling resistance has the value $R_a$, the positive half-cycles on the grid of the second valve are the positive values of $g_{m1} R_a v \sin \omega t$ and fall on BC (Fig. 1) where the mutual conductance is $g_{m2}$. The anode current is then the positive values of $g_{m1} g_{m2} R_a v \sin \omega t$. The negative half-cycles on the grid are the negative values of $g_{m2} R_a v \sin \omega t$ and fall in the region AB to produce an anode current which is the negative values of $g_{m1} g_{m2} R_a v \sin \omega t$. The distortion is completely cancelled.

In order that this may occur, it is essential that the bias on the second valve be correct, so that the dividing line between the positive and negative half-cycles falls at B.

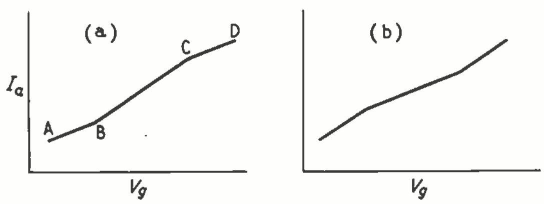



Fig. 3.

Now, if we had a valve characteristic like that of Fig. 3(a) and biased to midway between B and C, the middle section of the input sine wave would be amplified normally, but both the positive and negative peaks would be amplified less. A similar succeeding stage could now do nothing but increase the distortion further; in this case a correcting characteristic like that of Fig. 3(b) is wanted.

The form of characteristic shown in Fig. 1 is of a type which produces even-order harmonic distortion while that shown in Fig. 3 develops odd-order harmonics. Practical valve characteristics follow these forms but are, of course, curved throughout and they usually exhibit a combination of the two types. Generally, however, the form of Fig. 1 predominates, especially with triodes, but also with pentodes as long as the load impedance is not too high.

We may expect to find, therefore, that in a cascade of valve stages, there is a natural tendency for the even-order harmonic distortion to be reduced below that given by any individual stage. However, since the overall characteristic of two square-law stages in cascade tends to be S-shaped because of the reversal of phase in each stage, the odd-order distortion may be greater than that of any individual stage.

We shall now consider a more practical case, that of two valves in cascade with a.c. coupling between them and having characteristics of the general form

$$i_a = a + bV + cV^2$$

For valve 1, let $a = 0{\cdot}01$, $b = 0{\cdot}0016$ and $c = 0{\cdot}00006$, so that $i_a$ is 0·01 A for zero grid volts and cuts off at $-10$ V on the grid. Let us bias it to $-5$ V and apply a sine-wave signal of 4-V peak so that

$$V = -5 + 4 \sin \omega t$$

We then have

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

The distortion is second harmonic only and is 12%.

If the coupling resistance is $R_a$ the signal applied to the second valve is the a.c. component of $-i_a R_a$. In practice, $R_a$ might be 10 kΩ which would make the fundamental component of the input to the second valve 40 V and we should need a valve of 10 times the grid base. The coefficients $a$, $b$ and $c$ would have to be changed so that the characteristic remained of the same shape. For simplicity of computation, we shall make $R_a$ only 1 kΩ so that the fundamental input to the valve is only 4 V and we can use the same coefficients for it. With the same bias voltage, for this second valve

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

and we find

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

The second harmonic has vanished completely and in its place we have 2·8% third harmonic and 0·168% fourth. This is a considerable improvement. It is true that in our example the first stage gives no gain, but this is incidental.

In practice, to achieve this sort of result, the two valves must have characteristics of the same shape and the signals must occupy the same proportion of their grid bases. Generally speaking, it is in the output stage that most distortion occurs. The penultimate stage is relatively more linear. However, it would seem that, by suitably choosing the valve and its operating conditions to *increase* its distortion, a more linear combination for the two stages could be achieved.

This conclusion is by no means new. A good many years ago* van der Ven gave some very

* "Output Stage Distortion", by A. J. Heins van der Ven, *Wireless Engineer*, August 1939, p. 383.

relevant figures. He quoted the case of a triode output valve which introduced 5% second harmonic and 1·3% third-harmonic distortion by itself. When a penultimate stage was introduced, however, the distortion changed to 1·4% second harmonic and 2·2% third.

The practical benefits of this kind of distortion reduction are not likely to be very great in the case of a.f. amplifiers. The level of distortion aimed at is often so low that linearity correction in this way would seem to demand an impracticable degree of precision. The television case is quite different, for relatively large amounts of non-linearity are unimportant and only an approximate cancellation of non-linearities suffices.

However, it does seem as though it would be practical to reduce even-order harmonic distortion by a factor of 2 or 3 without any difficulty and this may well be beneficial in reducing the amount of negative feedback required. There is one thing of considerable importance to watch however; there must be negligible frequency or phase distortion between the two non-linearities. These introduce a change of waveform which upsets the operation.

Through thinking in terms of frequencies rather than composite waveforms, the audio engineer does not always realize how frequency, and especially phase, distortion can affect waveshape. It is, in fact, quite possible to correct for a non-linearity of amplitude response by an appropriate change of the frequency and phase characteristics. It is, however, only possible when the signal is of fixed amplitude and non-sinusoidal waveform. The method is widely used in time-bases, but is useless for audio work. If one attempts non-linear distortion cancellation in a.f. circuits, therefore, one must be very careful to avoid frequency and phase distortion as well.

There is one further aspect of distortion cancellation that we should point out. In principle, its application is not confined to amplifiers. In principle, there seems no reason why it should not be used to correct in a receiver for distortion occurring in a transmitter or en route between the two. One might thus remove interference caused by cross-modulation at some point! The essential things are two: first, to equalize the frequency and phase characteristics so that they are substantially perfect between the place of introduction of the non-linearity distortion and the place of introduction of the correcting non-linearity and, secondly, to provide a correcting non-linearity which is the inverse of the original one.

Practically, of course, it would hardly be possible to meet these requirements with sufficient accuracy and we do not suggest that there is any present likelihood of such a scheme being useful.

W. T. C.

# COMPATIBLE COLOUR-TELEVISION

## Part 1—*Two Sub-Carrier System*

**By J. Haantjes and K. Teer**

(*Philips Research Laboratories, N.V. Philips' Gloeilampenfabrieken, Eindhoven, Holland*)

**SUMMARY.**—A discussion of the requirements and possibilities for a compatible colour-television transmission system is given. A description of a transmission system using two sub-carriers for the additional colour information follows.

### Introduction

IT is well known that colour television is based on the principle that any coloured light impression can be considered to be a proper addition of light of three suitably chosen primary colours—green, red and blue[1]. Every colour-television system is, therefore, arranged in such a way that the amount of green, red and blue light in each picture element of the scene to be televised is measured. Corresponding signals are produced and transmitted and the picture at the receiving end is obtained by the addition of three television pictures each in one of the three primary colours.

It will be clear that the transmission of a coloured picture of the same quality as a standard black-and-white picture will, therefore, need three times the information necessary to transmit a black-and-white picture.

In the past, several colour-television systems have been proposed[2, 3]. Most of these systems were, however, not 'compatible', a requirement which has to be fulfilled if an existing black-and-white television service is to be developed into a colour-television service, without making the black-and-white receivers in the hands of the public obsolete.

The requirement of compatibility in a colour-television system means that the information of a colour picture must be transmitted, within a normal television channel, in such a way that

MS accepted by the Editor, July 1955

existing black-and-white receivers can reproduce the transmitted picture in monochrome, without any modification in the receiver.

The compatibility requirement restricts considerably the choice for possible colour-television systems. The N.T.S.C. system already officially adopted in the U.S.A. is an example of such a compatible system*.

## Consequences of Compatibility

One of the first consequences of the compatibility requirement is that the colour-television signal has to resemble the standard black-and-white signal as much as possible. This means that the scanning rate of the colour system has to be the same or practically the same. In our opinion, a sequential transmission of the three colour signals requires an increase in the line or frame scanning rates in order to obtain a satisfying colour image. It follows, therefore, that the three signals must be transmitted simultaneously.

The second consequence of the compatibility requirement is that it will be desirable to make one of the three signals the same as the signal which would be transmitted in black-and-white television. It will be a linear combination of the primary signals. To this signal, which is usually called the 'luminance signal', other signals called 'colour signals' have to be added in such a way that they are practically invisible in normal black-and-white reception but can be separated in a colour receiver.

The inclusion of these two signals and the ordinary black-and-white signal within a normal television channel is not a very easy matter. It can only be achieved by taking advantage of certain limitations in the human visual perception.

In the first place, there exists a certain insensitivity of the eye to errors in the colour of picture detail. When the right luminance is present in the detail, such errors are hardly noticeable. Thanks to this insensitivity of the eye, the transmitted data about picture detail can be restricted to the luminance signal only. This luminance signal has, therefore, to be transmitted with the normal bandwidth, whereas the bandwidth of the other signals can be reduced considerably[4, 5, 6, 7].

Notwithstanding this decrease in bandwidth, which simplifies the problem, the sum of the bandwidths of the individual signals will still exceed the channel bandwidth. In the transmission path, therefore, the spectra of the three signals have to overlap and, at the receiver, the signals cannot be separated without mutual cross-talk. The visual effect, however, of the cross-talk can be eliminated to a large extent by making use of another limitation of the eye, viz., the well-known integration effect. As a result of this effect it is possible to make unwanted signal components of relatively low visibility. This can be done by establishing a special frequency and phase relation between the unwanted signal frequencies and the scanning frequencies so that the interfering brightness modulation in the picture has a fine structure and is of opposite sign in successive scannings[8, 9, 10, 11].

For instance, an odd multiple of half the line frequency in the higher region of the video band will be barely visible, because such a frequency shows opposite phase in successive pictures and the successive patterns in successive pictures tend to cancel each other. This holds, also, when such a frequency is amplitude modulated by a video signal.

Experiments have shown that this principle makes it possible to decrease the visual effect of the mutual cross-talk to such a level that the black-and-white reception, as well as the colour reception, of the colour-television transmission is satisfying.

Taking into account all these things, the conclusion is that a colour-television transmission can be established as follows:—

One signal is transmitted in the normal way with standard bandwidth. This signal is a linear combination of the primary signals, and is as nearly as possible the same as the signal which would be transmitted in a monochrome system. This signal is called the luminance signal.

The other signals have relatively small bandwidths—about a quarter of the channel bandwidth. They modulate sub-carriers and these modulated sub-carriers are added to the luminance signal. These signals are called the colour-signals and are, in general, also linear combinations of the primary signals. The sub-carriers are situated within the band of the luminance signal and have special frequency and phase relations with the scanning frequencies, making the presence of these modulated sub-carriers almost unnoticeable in the picture reproduced by the luminance signal.

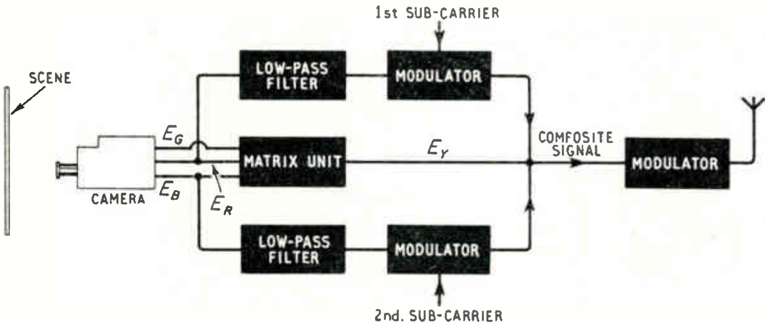
At the colour receiver, the primary signals can be obtained by demodulating the sub-carriers and by combining the demodulation results with the luminance signal. The mutual cross-talk consists of signal components which cancel out in successive lines, frames or pictures.

The colour television system accepted in the

*In the sense in which it is defined here, compatibility is not a generally-accepted requirement of a colour-television system. It is in the U.S.A., but elsewhere it has been suggested that compatibility requires only that a monochrome receiver shall, without alteration, be able to reproduce a monochrome picture from a colour transmission. In this case, all the information content of a colour transmission need not lie within the bandwidth of the monochrome receiver, but only that part of the total information which is necessary for the reproduction of a monochrome picture. Such a system might be called one of reduced compatibility and the principles of the N.T.S.C. system, as well as those of the one described in this article, can be used in it as well as in a fully compatible system.

U.S.A.—the so-called N.T.S.C. system—is based on these principles. In this system, the two sub-carriers have the same frequency and differ only in phase; in other words, only one sub-carrier is employed and it is modulated in amplitude as well as in phase[12, 13, 14].

In the rest of this article, however, we shall describe in detail a transmission system with sub-carriers of different frequencies.



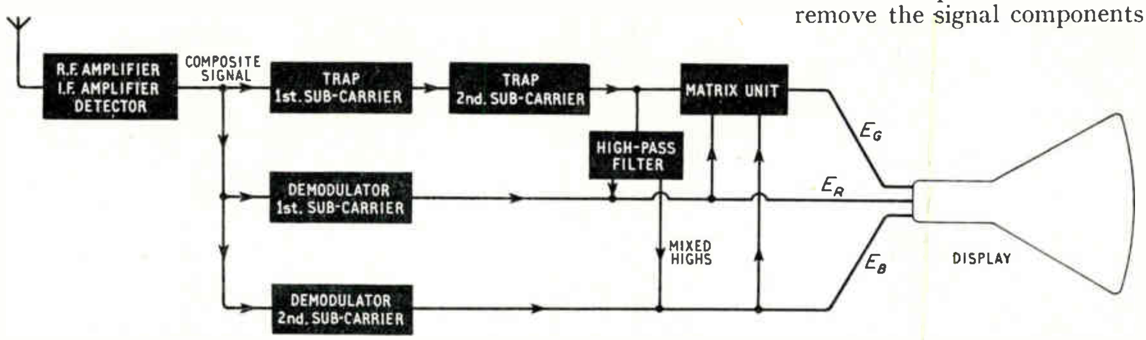




Fig. 1. *Block diagram of a colour-television transmission system, in which two sub-carriers are used.*

### Formation of the Transmission Signal

In Fig. 1 is shown a block diagram of the colour-television system to be discussed. Starting at the scene to be transmitted in colour, the first component of the system is a camera, suitable for generating three primary signals corresponding to the green, red and blue light components of the scene. These signals are proportional to the amounts of green, red and blue light present in the element of the image which is scanned. These signals are indicated by $E_G$, $E_R$ and $E_B$. They are supposed to vary between 0 and 1.

These primary signals are fed to a device (so-called matrix unit) which delivers at its output a certain linear combination of $E_G$, $E_R$ and $E_B$. The individual contributions of the three signals to this output signal correspond to the sensitivity of the human eye to green, red and blue light respectively. The signal is called the luminance signal and will be indicated by $E_y$.

It can be written as:

$$E_y = \alpha_0 E_G + \beta_0 E_R + \gamma_0 E_B \quad .. \quad (1)$$

If $\alpha_0$, $\beta_0$ and $\gamma_0$ are chosen in such a way, that

$$\alpha_0 + \beta_0 + \gamma_0 = 1,$$

the luminance signal will also vary between 0 and 1.

The luminance signal is transmitted in the normal way with normal bandwidth.

Apart from this signal two other signals are transmitted with the aid of two sub-carriers, each modulated by one of these signals. The signals are chosen to be the red and blue signals.

As already mentioned in the introduction, it is possible to reduce the bandwidths of these signals, based on the physiological effect that the eye is less sensitive to colour contrast in picture details. Hence, before the sub-carriers are modulated the red and blue signals are fed to low-pass filters which remove the signal components above a certain frequency. In our experiments on a colour system matched to the existing black-and-white television system for 625 lines and 50 frames, the bandwidth was chosen to be about 2 Mc/s for the red signal an 1 Mc/s for the blue signal.

After filtering, the red and blue signals modulate sub-carriers. These sub-carriers fall within the normal video band (0–5·5 Mc/s) and have a special frequency and phase relation with the scanning frequencies (to be discussed later). Their frequencies are chosen as high as possible; however, a frequency difference of about 1 Mc/s must exist between the higher sub-carrier frequency and the sound carrier, in order to make the separation of the sub-carriers possible with obtainable receiver filter characteristics.

After modulation the sub-carriers are added to the luminance signal. In principle, this modulation may be a double-, a vestigial-, or single-sideband modulation. The signal-to-noise conditions of the colour signals at the receiver, however, are optimum for single-sideband modulation, provided the signal amplitude is chosen to be the same in the three cases.

The combined signals form the composite video signal which is fed to a transmitter. It contains enough information to establish a satisfying colour picture at the receiver.

### Receiving Process

Up to the first video amplifier, the receiver has the same circuitry as existing black-and-white receivers. The second detector delivers a signal, which can be supposed to be the same signal as the video signal present at the transmitter.

In the colour receiver, this signal is fed to two demodulators, which are intended to demodulate the sub-carriers, and to a device where the sub-carriers in the luminance signal are suppressed to a certain extent.

Both demodulators consist of a band-pass filter tuned to the appropriate sub-carrier and a detector circuit to reproduce the modulation envelope of the applied signal. Thus, at the outputs of the demodulators the red and blue signals with small bandwidth are recovered. In order to increase the sharpness of the images reproduced by these signals, the high frequencies of the luminance signal (mixed highs) are added with proper amplitude and proper time delay. These frequencies are obtained by means of simple high-pass filters.

An equivalent green signal, composed of the lower frequencies of the signal $E_G$ and the higher frequencies of $E_y$, is formed in a matrix unit, where demodulator output signals with mixed highs and the luminance signal are combined.

The three signals obtained are fed to a display, suitable for reproducing a colour picture. This display may contain three projection tubes and dichroic mirrors, or may be some type of direct-viewing colour tube[15, 16, 17, 18]. When the system is well adjusted, large areas in the picture will appear in the original colour, whereas the picture detail is reproduced in black and white.

In this system the two colour signals can be separated from the composite signal without the need for synchronized detection. So sub-carrier generating and synchronizing circuits, as well as synchronized detectors, can be omitted from the receiver.

Also a synchronizing signal for the sub-carrier, which is necessary in the transmission signal of the N.T.S.C. system (the so-called 'burst') is not needed. However, such a burst may be introduced in the underlying system for other reasons. As only the sub-carrier amplitudes carry the colour information, the right amplitude ratio between the luminance signal and the sub-carriers must be maintained and this ratio must not be affected by differences in propagation for the various parts of the frequency band, or mis-tuning of the receiver.

In order to maintain a correct amplitude ratio in the receivers, the presence of a fixed amount of the two sub-carriers at certain predetermined moments (like the burst) may be a suitable means because it makes 'automatic ratio control' possible.

In brief, the system has now been explained. The basic idea of the system (viz., the use of two sub-carriers within the band of a normal video signal and modulated by primary signals) is not new[19, 20]. But so far as the authors know, the inherent problems of such a system have never been studied in a practical manner and solutions for these problems have never been published.

One of the major problems to be solved in the utilization of two sub-carriers is the choice of the frequency and phase of the sub-carriers in such a way that the presence of these carriers can be tolerated in black-and-white reception.

### Choice of the Sub-Carriers

It has often been shown that a frequency equal to an odd multiple of half the line frequency of the television signal is much less visible in a television picture than a frequency which has an arbitrary relation to the scanning frequency. Therefore it seems obviously desirable to use sub-carriers whose frequencies are synchronized to the line frequency and are equal to odd multiples of half the line frequency. However, when two such sine waves are added to a video signal, the reproduced picture is much more disturbed than by each sine wave individually and the picture quality is seriously affected.

This effect is caused by the difference frequency of the two sub-carriers which becomes very visible because it is a multiple of the line frequency. This frequency will always appear in some degree, due to the non-linear characteristic of the cathode-ray tube.

The visual effect of this frequency turns out to be at least as important as the perceptibility of the sub-carriers themselves. This was found during experiments, carried out to determine the most satisfying frequency and phase properties of the sub-carriers. Several possibilities were investigated. Sub-carrier frequencies equal to $n$, $n + \frac{1}{2}$, or $n + \frac{1}{4}$ times the line frequency were applied and phase resetting at the beginning of each frame or picture period (involving phase shifts of $\pm 90°$ or $\pm 180°$) was introduced. Each pair of sub-carriers was judged by three disturbing effects: the perceptibility of a dot-structure, the stroboscopic effect of moving dot-structures and the moiré effects caused by interference of dot-structure and picture detail.

At last one combination of two sub-carriers was found which presented a satisfactory solution of the problem and this was used in all further experiments with the transmission system. The lower sub-carrier frequency is made equal to an odd multiple of half the line frequency. In addition, the sub-carrier is given a sudden phase shift at the beginning of each frame of + and − 90° alternately. The higher sub-carrier frequency is made equal to a multiple of the line-frequency and this sub-carrier is given a phase shift of + or − 180° at the beginning of each frame.

The first sub-carrier can be written as:

$$A_1 \cos\left\{(n+\tfrac{1}{2})\,2\pi f_1 t + \phi_1 \pm \frac{\pi}{4}\right\} \quad .. \qquad (3)$$

where the + and − sign holds alternately in succeeding frames, $f_1$ is the line frequency and $n$ is an integer.

The second sub-carrier can be written as:

$$A_2 \cos\left\{m\,2\pi f_1 t + \phi_2 \pm \frac{\pi}{2}\right\} \quad .. \qquad .. \qquad (4)$$

in which $m$ is an integer.

In consequence, the frequency difference can be represented by:

$$A_3 \cos\left\{m\,2\pi f_1 t + \phi_2 \pm \frac{\pi}{2} - (n+\tfrac{1}{2})\,2\pi f_1 t - \phi_1 \pm \frac{\pi}{4}\right\}$$

$$= A_3 \cos\left\{(q+\tfrac{1}{2})\,2\pi f_1 t + \phi_3 \pm \frac{\pi}{4}\right\}$$

in which $q$ is an integer.

This expression is of the same nature as the expression for the first sub-carrier. Hence the frequency difference between the two sub-carriers shows the same frequency and phase properties, as the first sub-carrier.

## Dot Patterns of Sub-Carriers

The visual effect of the presence of the two sub-carriers can be described more or less by means of the dot patterns of the sub-carrier frequencies and their frequency difference.

The pattern of the first sub-carrier is represented in Fig. 2(a) [11]. The maxima in the brightness modulation along a line (appearing in the picture as bright dots) are indicated by:

A for the odd lines of a certain picture,
B for the even lines of that picture,
C for the odd lines of the succeeding picture,
D for the even lines of the succeeding picture.

Looking for possible stroboscopic effects, it will be noticed, that there is no direction in which the dots A, B, C, and D appear successively on a straight line. Therefore, in this pattern the effects can only be of minor importance. Moreover, the existing stroboscopic effects caused by the succession of A and C, and B and D in both vertical and horizontal directions are all equivalent, so that there exists no direction of preference for the perceptibility of stroboscopic effects.

Moving-dot structures are therefore less perceptible than those occurring in the pattern of an odd multiple of half the line frequency without sudden phase-shifts [Fig. 2(b)], where especially the succession of A, B, C, and D upwards gives rise to a rather striking stroboscopic effect.

With respect to the moiré effect caused by interference between the dot pattern and picture detail, no appreciable difference was found between the patterns of Figs. 2(a) and (b).

The pattern of the second sub-carrier is similarly represented in Fig. 2(c). In this pattern, the dots of the odd frames (A and C) coincide and so do the dots of the even frames (B and D). Stroboscopic effects are present in horizontal and vertical directions. Just as in (b) there is no direction of preference. In this case, however, the perceptibility of stroboscopic effects has not been decreased by the introduction of the phase resetting. The pattern of the multiple of the line frequency, without the phase shifts at the beginning of each frame (d), shows stroboscopic effects of the same magnitude. Yet the pattern of (c) is much less disturbing, because it gives a finer impression and less interference with picture detail, thanks to the interlacing of the dots of the two frames (A and B).

As already shown, the difference frequency shows the same pattern as the first sub-carrier. The patterns differ only in coarseness.

| | | | | | | | | |
|---|---|---|---|---|---|---|---|---|
| C | | A | | C | | A | | C |
| | D | | B | | D | | B | |
| A | | C | | A | | C | | A |
| | B | | D | | B | | D | |
| C | | A | | C | | A | | C |
| | D | | B | | D | | B | |

(a)

| | | | | |
|---|---|---|---|---|
| C | A | C | A | C |
| B | D | B | D | B |
| A | C | A | C | A |
| D | B | D | B | D |
| C | A | C | A | C |
| B | D | B | D | B |

(b)

| | | | | |
|---|---|---|---|---|
| A | | A | | A |
| | B | | B | |
| A | | A | | A |
| | B | | B | |

(c)

| | | |
|---|---|---|
| A | A | A |
| B | B | B |
| A | A | A |
| B | B | B |

(d)

Fig. 2. *The dot pattern of the first sub-carrier with a frequency which is a multiple of half the line frequency is shown in* (a) *when there is a phase shift of* ± 90° *in alternate frames and in* (b) *with no phase shift. The dot pattern of the second sub-carrier with a frequency which is a multiple of the line frequency has the form shown in* (c) *when there is a phase shift of* ± 180° *in alternate frames in* (d) *when there is no phase shift.*

## Generation of the Sub-Carriers

As already stated, the particular frequency and phase relations for the sub-carriers were used in all the experimental work. Fig. 3 illustrates the way in which they were produced.

In both generators there is an oscillator locked to the line frequency. The oscillator frequencies are $(n + \frac{1}{2})f_1$ and $mf_1$. The locking to the line frequency can be established in several ways, for instance, by means of a divider chain and a phase discriminator. The output of the phase discriminator depends upon the phase relation between the two frequencies and controls the oscillator frequency.

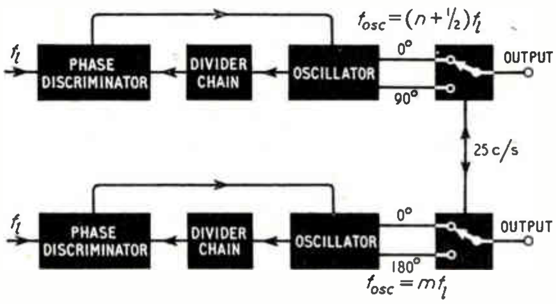
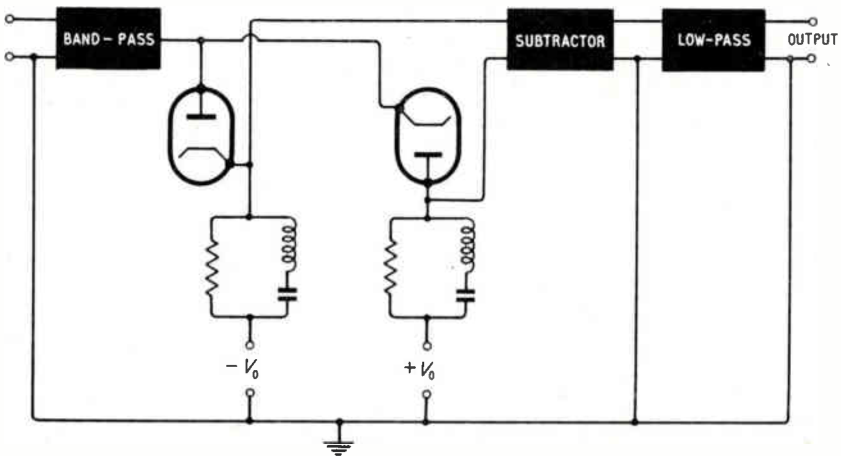



Fig. 3 (above). *Block diagram showing the method of generating the two sub-carriers.*

Fig. 4 (right). *Special detection circuit for the sub-carrier demodulation.*




Each oscillator delivers two sine waves, only differing in phase. This phase difference is 90° for the lower and 180° for the higher sub-carrier. The generator output is connected to the oscillator output by an electronic switch, which is switched at the beginning of each frame period. In this way the desired frequency and phase relations of the sub-carriers can be obtained.

## Mutual Cross-talk

In any multiplex television system, where signals with overlapping frequency spectra are present, difficulties of cross-talk between the signals involved may arise.

The modulation of the two sub-carriers by the red and blue signals, the transmission of the composite signal modulated on a picture carrier, and the receiving and detection of this picture carrier involve no serious problems, certainly none differing from those occurring in normal television technique. There are no special problems before the stage where the received signal has to be split into three colour signals. These three signals must be obtained in such a way that the mutual cross-talk, which is caused by the overlapping of the three signal spectra, does not affect the picture quality.

First, consider the signal components of the luminance signal present in the band of a modulated sub-carrier, and suppose these signal components can be described on a certain line of a certain frame as:

$$\sum_{n=1}^{\infty} a_n \cos(\omega_n t + \phi_n) \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

where $t = 0$ indicates the beginning of the line.

If the sub-carrier is detected and the disturbing signal components are interpreted by the detector as sideband frequencies of the sub-carrier $\cos(\omega_s t + \phi_s)$, the detector output will contain a disturbing signal proportional to

$$\sum_{1}^{\infty} a_n \cos\{(\omega_s - \omega_n)t + (\phi_s - \phi_n)\} \quad \ldots \quad (7)$$

As, in general, there is only a very small variation in the picture content from line to line and from one picture to the following picture, the luminance signal on adjacent lines and on successive scans on the same line will almost be the same.

Therefore, the disturbing signal in the detector output will only differ on adjacent lines and successive scans due to the differences in phase $\phi_s$ of the sub-carrier frequency.

It follows from the expression (7) that the signal components will vary in phase in the same way as the sub-carrier phase.

So cross-talk of the luminance signal in the demodulated sub-carrier consists of signal components, showing within a small region of time and space, the same phase alternation as the sub-carrier does in the luminance signal and will therefore be of reduced visibility and annoyance.

Similarly, it can also be proved, that the disturbance of the detected sub-carrier, caused by the presence of the other sub-carrier, shows phase alternation on adjacent lines and successive scans. In this case the phase alternations are equal to those which the frequency difference of the two sub-carriers shows in the luminance signal.

Thanks to the phase alternation of the cross-talk in the demodulated sub-carrier, the disturbance will be of low visibility in the corresponding picture and can be tolerated without harm to the picture quality.

However, in this reasoning it has been assumed that the interfering signal components in the band of the modulated sub-carrier are interpreted by the detector as sidebands of the sub-carrier. It will be plain that this condition may not be satisfied in some part of the picture. For instance, if in some region the sub-carrier amplitude is very small, or zero, the cross-talk in the output of the detector circuit will be the same on successive pictures and adjacent lines. In general, the undesired signal components will appear in the picture partly as a phase-alternating cross-talk and partly as a cross-talk which is independent of sub-carrier phase.

The presence of the latter type of cross-talk depends on the degree of modulation and the relative peak value of the sub-carrier. To assure its complete absence, the sub-carrier amplitude should never fall below a certain level, fixed by the peak amplitude of the undesired signal before detection.

However, for the sake of compatibility the peak value of the sub-carrier has to be kept as small as possible and it is very desirable that the sub-carrier amplitude should be zero in dark areas (100% modulation depth). We have, therefore, tried to avoid the need for a low modulation depth by applying another method of making cross-talk of low visibility. Experiments have shown this to be largely possible by the use of a special detection circuit [11]. See Fig. 4.

In this circuit, the received video signal is passed through a suitable bandpass filter and is then detected by two detectors, one following the upper envelope of this signal, the other following the lower envelope. The output signals of the two detectors are combined and the result shows a considerable reduction of the 'steady cross-talk'.

In principle, the steady cross-talk cannot be fully eliminated, because the elimination depends on the shape of the undesired signal. We can state, however, that with this detection circuit it is possible to obtain a satisfying picture from the sub-carrier information under usable conditions of peak value and bandwidth of the sub-carriers. These conditions will be specified in the second part of this article. In that part we shall give a comparison of the two sub-carrier system with the American N.T.S.C. system in order to give an impression of the advantages and disadvantages of using two sub-carriers instead of only one.

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*(To be continued)*

# ELECTRON TRAJECTORIES IN COAXIAL DIODES

## *With Combined R.F. and Steady Fields*

**By R. Dehn, B.Sc., A.R.C.S., A.Inst.P.**

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

### 1. Introduction

THE equation of motion of an electron between coaxial cylinders is given by:—

$$\frac{d^2r}{dt^2} = -\frac{e}{m}\frac{V}{r \log a/b}$$

considering only the radial component.

Here $V$ is the potential difference between the cylinders, the cathode potential being taken as zero throughout, and $r$ is the distance from the centre, $a$ and $b$ are the outer and inner radii respectively, and $e/m$ is the charge/mass ratio for the electron, a negative quantity.

If $V$ is constant this equation can be solved analytically. In most practical cases the electron transit time is short compared with the period of variation of any r.f. field applied to the diode, and $V$ can therefore be regarded as substantially constant.

The writer is using a cylindrical diode in the form of a short coaxial transmission line, mounted in a waveguide system, as a waveform and power monitor for microwaves at frequencies of about 3,000 Mc/s. This type of diode was originally developed at the Services Electronics Research Laboratory, Baldock[1].

At the frequencies mentioned the electron transit time is of the order of several r.f. cycles, and can therefore no longer be neglected. The equation of motion has to be modified accordingly, introducing $V$ as a time-dependent variable.

Experiment and theory show that a certain number of electrons will reach the anode and others return to the cathode under the influence of the r.f. field. Since no external p.d. is applied to the valve, and since the output impedance is of the same order as the dynamic impedance of the diode (i.e., about 100 ohms), the anode assumes a negative potential with respect to the cathode. The electron is therefore subject to a steady retarding field and to an r.f. field both acting simultaneously.

### 2. Theory

Very little theoretical work seems to have been published on electron trajectories in diodes involving large transit angles. A valve of different structure but designed for similar applications to the one considered in this paper has recently been described by Bronwell et al.[2] Their theoretical analysis includes the effect of space charge but considers the plane parallel case only. A brief treatment for fairly small transit angles (i.e., not greater than $2\pi$ radians) and applicable to plane parallel triodes has been given by Gavin[3].

It appeared desirable, therefore, to compute the electron trajectories for some specific cases of practical importance. Apart from the general interest of the problem such computation would allow an estimate of what proportion of electrons would reach the anode under given conditions.

In order to simplify analysis the following assumptions have been made:—

(1) The effect of space charge has been neglected.

(2) The initial or thermal velocities of the electrons have been neglected. This seems reasonable in view of the order of magnitude of these velocities compared with the anode–cathode potentials here considered.

(3) Since zero initial electron velocity has been assumed, and both r.f. and d.c. fields act radially, only radial components of motion need be considered in the analysis.

The equation of motion now becomes:—

$$\frac{d^2r}{dt^2} = -\frac{e}{m}\frac{V_0 \sin(\omega t + \alpha) + V_{dc}}{r \log r_a/r_c}$$

where $V_0$ = peak r.f. potential applied between anode and cathode
$V_{dc}$ = steady potential between anode and cathode
[cathode potential is again taken as zero]
$r$ = distance from centre
$r_a$ = anode radius
$r_c$ = cathode radius
$\alpha$ = phase of entry of electron
$\omega$ = $2\pi \times$ frequency

The equation can no longer be solved in general terms so that numerical methods have to be employed for dealing with each particular set of conditions. The programme of computation was carried out by Messrs. Elliott Bros., on one of their digital computors.

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## 3. Results

Electron trajectories have been obtained for a few parameters which correspond closely to the conditions under which the diode is used in practice. As mentioned earlier, at a frequency of 3,000 Mc/s ($\omega = 6\pi \times 10^9$) the electron transit time is of the order of several r.f. cycles. If the electron leaves the cathode at zero velocity it must receive sufficient initial acceleration to reach the anode against both the negative d.c. field and the subsequent retarding half-periods of the r.f. field. It must be remembered that the field (for a given anode–cathode potential) is non-linear, varying inversely with $r$. Hence the electron experiences maximum acceleration near the cathode.

A preliminary consideration suggested that an electron emitted early enough during the positive half cycle would reach the anode. As the phase of entry is increased a point would be reached when the initial acceleration proved insufficient, and the electron would return to the cathode during the negative half cycle. It was consequently assumed that, unless the electron thus returned to the cathode during the first cycle it would invariably reach the anode. Detailed investigation showed, however, that this conclusion was incorrect. The electron may move towards the anode for several r.f. cycles and then gradually reverse and finally return to the cathode.

Systematic variation of the phase of entry showed the existence of several critical angles. Fig. 1 gives the actual trajectories obtained for a particular set of conditions:—

$$\begin{aligned} V_0 &= 4{,}000 \text{ volts} \\ V_{dc} &= -300 \text{ volts} \\ r_a &= 0{\cdot}675 \text{ cm} \\ r_c &= 0{\cdot}045 \text{ cm} \\ \omega &= 2\pi \times 3 \times 10^9 \text{ radians/sec} \\ (\text{i.e.,}\ f &= 3{,}000 \text{ Mc/s}) \end{aligned}$$

Three points were plotted for every r.f. cycle.

For small phase angles, $\alpha = 60°, 90°, 105°$, the electron possesses positive velocity throughout its motion and reaches the anode after between one and three r.f. cycles. Between 105° and 110° there is a critical angle such that the electron just fails to reach the anode and gradually turns back towards the cathode. The most surprising fact, which was then discovered, was that even after such a reversal of velocity the electron may still reach the anode undergoing a second reversal near the cathode. Trajectories for 110° and 113° show this quite clearly. There is a further critical angle, or small range of angles, for which this second reversal does not take place, and the electron actually returns to the cathode. This

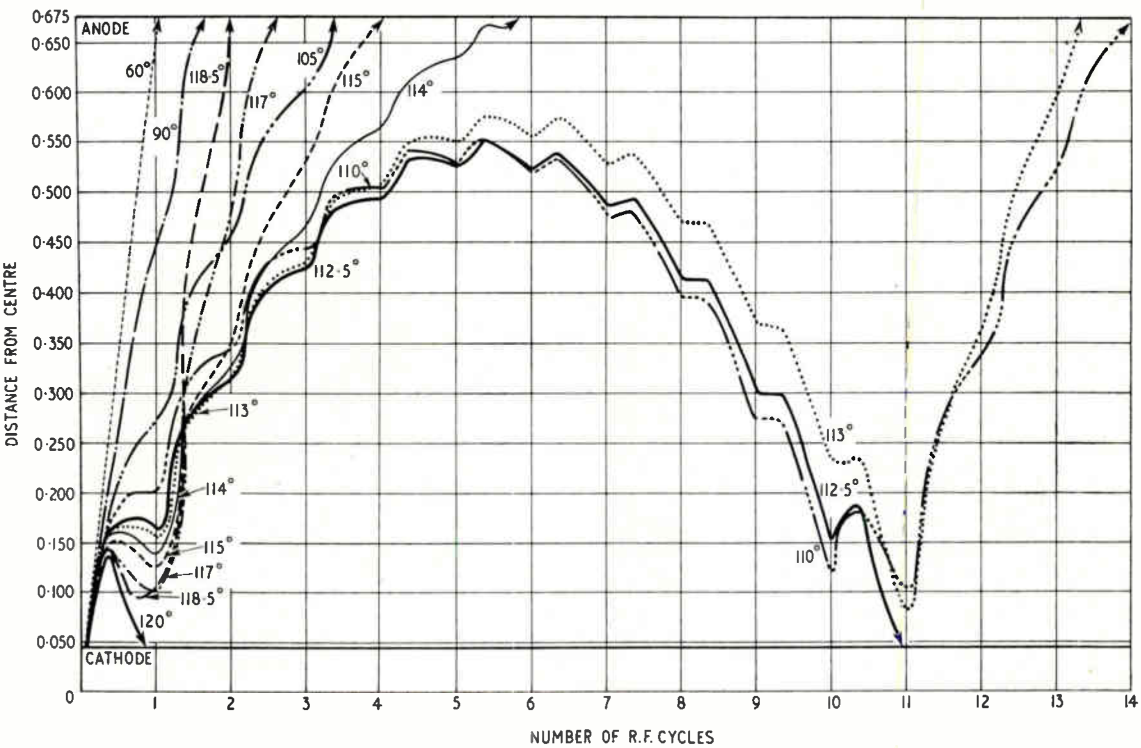


Fig. 1. *Electron trajectories for various phase angles $\alpha$.*

is here shown for an angle of 112·5°. The next critical angle is reached when the electron begins to experience a reversal during the first retarding half cycle, so that the next accelerating half cycle finds the electron nearer to the cathode and therefore in a stronger field. Thus it experiences considerable acceleration and generally reaches the anode fairly quickly. This first shows at 114° for the case here considered. As the phase angle is further increased the electron approaches the cathode more and more closely, thus becoming subject to rapidly increasing accelerating fields at the commencement of the second cycle. Angles from 114° to 118·5° clearly show this in the figure. It is evident that for some of these angles (e.g., 117°, 118·5°) the electron reaches the anode more quickly than for some angles which are less than the first critical angle (e.g., 105°). Finally, near 120° a phase angle is reached such that the electron returns to the cathode during the first cycle.

These results present a remarkable picture, much more complex than the first superficial analysis had suggested. Electrons may return to the cathode either during the first negative half cycle or after a considerable number of cycles. Likewise they may reach the anode either directly after a few cycles or after a much larger number of cycles having first suffered two reversals, one near the anode and another one near the cathode.

In the case shown in Fig. 1 electrons approach the cathode following a reversal near the anode after eleven cycles. This may lead to an increase in space-charge density due to these electrons; it also suggests the possible generation of weak transit-time oscillations of approximate frequency 3,000/11 Mc/s, or, in general, some submultiple of the applied r.f.

A few additional trajectories were obtained for different conditions, first for a larger cathode ($r_c = 0{\cdot}1125$ cm) and then for $V_0 = 2{,}000$ V (small cathode). These are shown in Fig. 2. It appears that the general picture is little changed, but that the respective critical angles are reduced.

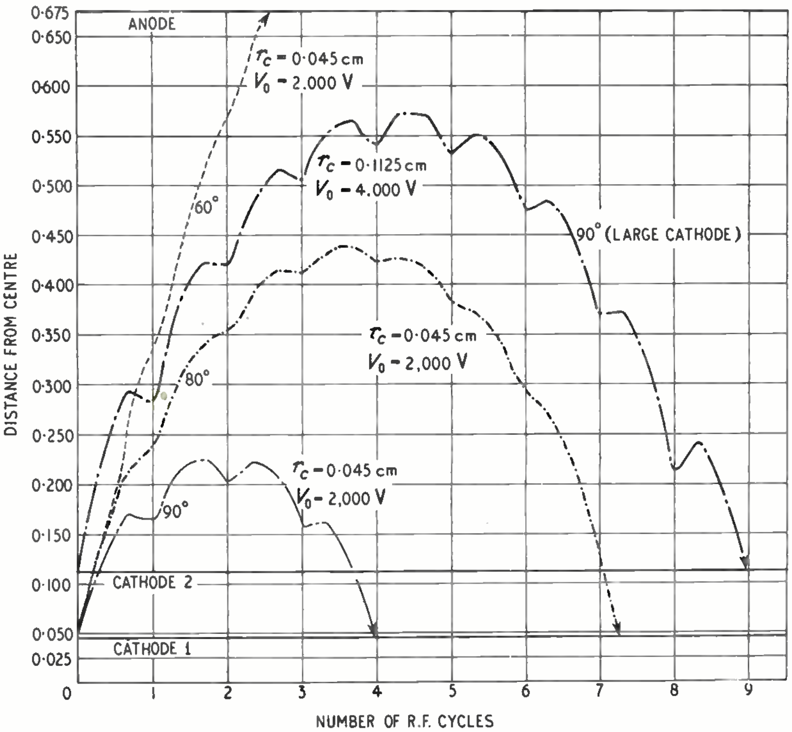


Fig. 2. *Electron trajectories for different cathode sizes and voltages.*

## 4. Conclusion

Experimental results suggest that, in general, about half the r.f. power going through the diode, regarded as a coaxial transmission line, is used up in cathode heating by electron bombardment. Obviously all electrons, whatever their path, must derive their energy from the r.f. field. If it is assumed that the total energies possessed respectively by the electrons which reach the anode and by those which return to the cathode are approximately equal, then it would follow that about half the total number of emitted electrons will reach the anode, the remainder being returned to the cathode. This seems in approximate agreement with the results obtained for the critical angles suggesting that electrons emitted during the first quarter cycle reach the anode and those emitted during the second quarter are returned to the cathode by the r.f. field. Obviously no electrons are emitted during the negative half cycle.

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# CONSTANT-FREQUENCY OSCILLATORS

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**SUMMARY.**—The conditions for the frequency of a regenerative oscillator to be independent of changes in the input and output resistances of the maintaining amplifier are derived. It is shown that the input, output, and transfer impedances of the feedback network must be resistive at the oscillation frequency. This can be achieved by inserting reactances of definite specified values in the input and output leads. Expressions for the value of these reactances are derived for the usual type of $LC$ oscillator with mutual-inductance coupling. The values depend greatly on whether the losses in the coils can be represented mainly by a series or by a parallel resistance. Expressions for the stability of an imperfectly-stabilized oscillator are also derived. The results are compared with those obtained by Llewellyn. It is shown that the values of stabilizing reactances specified by Llewellyn are not in general satisfactory, though in some instances they may produce a marked improvement in stability. Full experimental confirmation is given.

## 1. Introduction

One of the most important requirements of an oscillator is that its frequency should be substantially independent of moderate changes in the supply voltages. In the early 1930's constant-frequency oscillators were the subject of many papers and much controversy. Conflicting theories and experimental results were advanced by various writers, but despite prolonged discussion no satisfactory conclusions appear to have been reached. Interest waned, and although the situation has since become gradually clarified some problems remain unsolved. Of all this effort little survives in modern text-books except for references to the work of Llewellyn[1] and Groszkowski[2].

The two most important results of varying the supply voltages are:—

(a) Changes in inter-electrode capacitances and electron transit times.

(b) Variations in the functions relating changes of electrode currents to changes of voltages.

The relative importance of the different effects depends largely on the frequency of operation. At high frequencies it is possible, as Heising[3] has demonstrated, to secure overall stability by adjusting the oscillator in such a way that frequency changes due to (a) and (b) cancel one another. At low frequencies the effects of (a) are usually negligible. The effects of (b) are changes in the oscillation amplitude and harmonic content, both of which will, in general, change the oscillation frequency. Only (b) is considered further.

Variation of frequency with harmonic content has been adequately treated by Groszkowski and previous writers[4], and will not be considered here. It is therefore assumed that the electrode harmonic voltages are negligibly small. All frequency changes may then be ascribed to the variation of the effective input and output resistances of the valve caused by the change of oscillation amplitude.

This is the situation which was examined by Llewellyn, with the additional assumption that the oscillatory circuit is devoid of resistance, the only sources of dissipation being the grid leak and the grid-input resistance of the valve. Since in most practical oscillators this condition is not even approximately satisfied, a new and more general analysis of the problem seems desirable.

## 2. Oscillation Frequency

Fig. 1 shows the arrangement to be analysed. The feedback network is specified by the open-circuit input impedance $Z_i$ between 1 and 3, the open-circuit output impedance $Z_o$ between 2 and 3, and the open-circuit transfer impedance $Z_t$ between 1,3 and 2,3. $Z_o$ will include the grid leak and capacitor, and $Z_i$ the anode-circuit de-coupling network, if any. A triode is depicted but the results apply equally to pentodes and tetrodes.

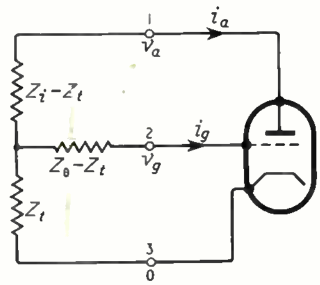


Fig. 1. *Oscillator circuit.*

It is assumed temporarily that the grid current depends only on the grid voltage $v_g$, and that the cathode current $i_k$ is a function of $v_g + v_a/\mu$. In a triode the anode current $i_a = i_k - i_g$, and in a multigrid valve it is a fraction $b$ of this. Thus in all cases

$$i_g = e(v_g),\ i_a = f(v_g + v_a/\mu) - bi_g \quad \ldots \quad (1)$$

where $e$ and $f$ are single-valued functions and $\mu$ and $b$ are assumed to be constants. Later it will be shown that when the stability requirements are satisfied these restrictions can be removed,

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and both $i_a$ and $i_g$ can be any single-valued functions of $v_g$ and $v_a$.

The object of the analysis is to find the conditions under which the oscillation frequency is independent of $\mu$, $b$, and the forms of the functions $e$ and $f$.

The currents may be non-sinusoidal, but since the alternating parts of $v_g$ and $v_a$ have been assumed sinusoidal, only the fundamental-frequency components in $i_a$ and $i_g$ need be considered. Denoting such components by the suffix 1, let

$$v_{g1} = V_{g1} \cos \omega_0 t,\ i_{g1} = I_{g1} \cos \omega_0 t$$

Then $v_{g1}/i_{g1} = V_{g1}/I_{g1} = r_g$ defines the grid-input resistance of the valve.

Let $v_{g1} + v_{a1}/\mu = V_e \cos(\omega_0 t + \theta)$ where $\theta$ is the phase angle of the cathode current with respect to the grid voltage.

Then $\quad i_{a1} = I_{a1} \cos(\omega_0 t + \theta) - b i_{g1}$

$$= g(v_{g1} + v_{a1}/\mu) - b v_{g1}/r_g \quad .. \quad (2)$$

where $g = I_{a1}/V_e$ is a real positive constant.

The anode and grid currents and voltages are also related through the impedances of the feedback network. Straightforward algebra gives these relations as

$$\left.\begin{aligned} i_{a1} &= -v_{g1}/Z_t - i_{g1}Z_o/Z_t \\ v_{a1} &= v_{g1}Z_i/Z_t + (Z_iZ_o/Z_t - Z_t)i_{g1} \end{aligned}\right\}(3)$$

The currents and voltages are now regarded as complex quantities, and the impedances are measured at the oscillation frequency $\omega_0$.

Suppose now that the feedback network is modified by connecting a resistance $r_g$ between terminals 2 and 3. The new network could be replaced by a three-element network similar to the original but with modified values for the impedances. Denoting the new elements by $Z_I$, $Z_\theta$, and $Z_T$, it is easily shown that

$$\left.\begin{aligned} Z_I &= Z_i - Z_t^2/(r_g + Z_o) \\ Z_\theta &= Z_o/(1 + Z_o/r_g) \quad .. \\ Z_T &= Z_t/(1 + Z_o/r_g) \quad .. \end{aligned}\right\} \quad (4)$$

Equations (3) can then be written as

$$i_{a1} = -v_{g1}/Z_T,\ v_{a1} = v_{g1}Z_I/Z_T$$

and substituting these into (2) gives

$$1 = -g(Z_T + Z_I/\mu) + bZ_T/r_g$$

Let $Z_T = R_T + jX_T$, $Z_I = R_I + jX_I$. Equating separately the real and imaginary parts of this equation,

$$g(R_T + R_I/\mu) - bR_T/r_g = -1 \quad .. \quad (5)$$

and $g(X_T + X_I/\mu) - bX_T/r_g = 0$, from which

$$(X_T + X_I/\mu)/(R_T + R_I/\mu) = X_T/(R_T - r_g/b) \quad (6)$$

This equation determines the oscillation frequency. Equation (5) defines the oscillation amplitude, and since $g$ is positive $R_T$ must be negative in order to maintain the oscillation.

### 3. Conditions for Frequency Stability

In the frequency equation (6) the quantities pertaining to the amplifier are $\mu$, $b$ and $r_g$, and changes in the electrode potentials or the substitution of one valve by another will usually alter all of these. Hence, if the frequency is to be independent of changes in the amplifier, the solution of (6) must be independent of $\mu$, $b$ and $r_g$.

The left-hand side of (6) is independent of $\mu$ if $X_I/X_T = R_I/R_T$, and the equation then becomes

$$X_I = X_T = 0 \quad .. \quad .. \quad .. \quad (7)$$

i.e., $Z_I$ and $Z_T$ are resistive at $\omega_0$. This makes the frequency independent also of $b$. Condition (7), together with the fact that $R_T$ must be negative, means that the alternating anode and grid voltages are in phase opposition. Conversely if these voltages are in phase opposition the frequency is independent of $\mu$ and $b$.

Equation (7) can be written in terms of the primary quantities $Z_i$, etc., as follows. From (4), $Z_T = r_g(R_t + jX_t)/(r_g + R_o + jX_o)$. Since $Z_T$ is to be real the right-hand side of this expression is real. Hence

$$X_t/R_t = X_o/(r_g + R_o) \quad .. \quad .. \quad (8)$$

Also, from (4),

$$Z_I = R_i + jX_i - (R_t + jX_t)^2/(r_g + R_o + jX_o)$$

On substituting for $r_g + R_o$ from (8) the condition for $Z_I$ to be real is

$$X_iX_o = X_t^2 \quad .. \quad .. \quad .. \quad (9)$$

(8) and (9) together are the conditions for the frequency to be independent of $\mu$ and $b$, and these can be satisfied in an infinite number of different ways. For example, $X_o$ could be given any arbitrary value by the addition of a suitable reactance to the network, and (8) could be solved to find the corresponding value of $\omega_0$. $X_i$ could then be adjusted to satisfy (9) at this frequency.

In order that $\omega_0$ should be independent also of $r_g$ further restrictions must be imposed. The solution of equation (8) must be independent of $r_g$ and this is possible only if both sides are 0. (8) and (9) then reduce to

$$X_i = X_o = X_t = 0 \quad .. \quad .. \quad (10)$$

This is the stability criterion. If $\mu = \infty$ the value of $X_i$ can have no influence, and the criterion reduces to $X_o = X_t = 0$. If $r_g = \infty$ (i.e., no grid current) the value of $X_o$ is immaterial and then $X_i = X_t = 0$ is a sufficient criterion.

When the stability requirements are satisfied the restrictions of equation (1) on the forms of the anode and grid current functions can be removed. Since the frequency is independent of $\mu$ and $b$, these can be functions of $v_g$ and $v_a$, and so $i_a$ can be any single-valued function of $v_g$ and $v_a$. Similarly, since the frequency is independent of $r_g$, $r_g$ and so also $i_g$ can be any single-valued function of $v_g$ and $v_a$. The fact that $v_g$ and $v_a$

are in phase opposition ensures that the grid-input impedance is resistive.

Any given feedback network can be made to satisfy condition (10) by the insertion of suitable reactances $X_a$ and $X_g$ in the anode and grid leads. Since these reactances cannot affect the value of $X_t$, the oscillation frequency is given by $X_t = 0$, and $X_a$ and $X_g$ are chosen to make $X_i = X_o = 0$ at this frequency. It is clear that $X_a$ and $X_g$ have definite fixed values. This conclusion is totally different from that of Llewellyn's theory, according to which stabilization can be achieved by a single reactance inserted in either anode or grid leads, and if two reactances are used one can have an arbitrary value.

Llewellyn assumed that the elements in the feedback network are pure reactances, the sole sources of dissipation being the grid leak and grid-input resistance of the valve. It might, therefore, appear that Llewellyn's results would be obtained as the limiting case of the general theory when $R_i$, $R_o$, and $R_t$ tend to 0. This expectation, however, is not fulfilled, for the criterion (10) is independent of resistance values. The reason for this failure will appear later, meanwhile, to obtain Llewellyn's results, a return is made to equation (7).

If the feedback network is purely reactive then from (4)

$$Z_T = jX_t r_g/(r_g + jX_o)$$

It is obvious that (7) cannot now be satisfied for any finite values of $X_o$ and $X_t$, but if both are very large compared with $r_g$, then

$$Z_T \approx r_g X_t/X_o + jX_t(r_g/X_o)^2 = R_T + jX_T$$

$X_T$ will approach 0 and $R_T$ will be finite if both $X_o$ and $X_t$ tend to $\infty$ while their ratio remains finite. Similarly,

$$\begin{aligned} Z_I &= jX_i - (jX_t)^2/(r_g + jX_o) \\ &\approx r_g(X_t/X_o)^2 + j(X_i - X_t^2/X_o) \\ &= R_I + jX_I \end{aligned}$$

So $X_I \to 0$ if $X_i X_o = X_t^2$, which is the same as condition (9). Hence the stability criterion is that (9) shall be satisfied and at the same time all three reactances shall be infinite. The stabilizing reactances $X_a$ and $X_g$ are not now fixed but may take any values consistent with (9).

Fig. 2 shows the arrangement analysed by Llewellyn, using the conventional equivalent circuit for the valve. It may be noted in passing that this representation is not accurate when grid current flows, but the final result is the same in this instance. Applying the criterion $X_i X_o = X_t^2$ gives

$$\begin{aligned} 2X_m(X_1 + X_m)(X_2 + X_m) \\ = (X_1 + X_a)(X_2 + X_m)^2 \\ + (X_2 + X_g)(X_1 + X_m)^2 \end{aligned}$$

which is Llewellyn's formula. The oscillation frequency is that for which $X_1 + X_2 + X_3 + 2X_m = 0$, and this makes $X_i$, $X_o$, and $X_t$ infinite as required.

The reason why Llewellyn's results cannot be obtained by taking limits in the general theory lies in the nature of the reactance function. In the neighbourhood of an antiresonance the reactive component of any physical driving-point impedance function passes from a large positive to a large negative value, passing through zero at the antiresonant frequency. The smaller the network dissipation the larger are the values of the maximum and minimum reactances but, however small the dissipation may be, the reactance still passes through zero at the antiresonance. However, if the impedance network is assumed to be completely loss-free, the reactance rises to $\infty$ on one side of the antiresonance, to $-\infty$ on the other side and is indeterminate at the antiresonant point. This explains the appearance of infinities in Llewellyn's method and zeros in the more general theory.

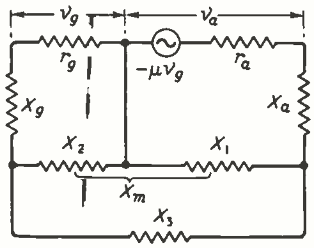


Fig. 2. *Llewellyn's circuit.*

Condition (9), which is equivalent to Llewellyn's criterion, can be regarded as the limit of the general formula provided the network dissipation is not actually zero. If this dissipation is very small compared with that in the grid-input resistance, the value of $R_o$ is very large compared with $r_g$ at the oscillation frequency. Then equation (8) will be satisfied for almost the same value of $\omega_0$ whatever the value of $r_g$, leaving (9) as the sole criterion for stability.

If $\mu = \infty$ Llewellyn's theory indicates that no stabilizing reactances are required, for the value of $X_i$ is then immaterial and the sole criterion is that $X_t$ and $X_o$ shall be infinite. The addition of any finite reactance $X_g$ to $X_o$ cannot therefore have any influence on the frequency.

## 4. Application of the Criterion

The general theory is valid for any type of harmonic oscillator, but all the important points are brought out by a consideration of the particular circuit shown in Fig. 3. $X_a$ and $X_g$ are the stabilizing reactances, and $R_b$ and $C_b$ are the grid leak and bypass capacitor. Losses in the coils $L$ and $L_1$ (coupled by the mutual inductance $M$) are represented by two series resistances $r$ and $r_1$ and the parallel resistance $R$. The reason for this choice will be discussed later.

Let $\omega_r^2 LC = 1$, $M = -k\sqrt{LL_1}$

$$Q_p = R/\omega_r L,\ Q_s = 1/\omega_r Cr,\ Q = Q_p Q_s/(Q_p + Q_s) \quad \ldots \quad (11)$$

Elementary circuit analysis gives the input, output, and transfer impedances of the network (corresponding to those indicated in Fig. 1) as

$$Z_i = jX_a + (L/M + r/j\omega M)Z_t$$
$$Z_o = jX_g + (\omega M/\omega_r L)(\omega/\omega_r - j/Q_p)Z_t + R_b/(1 + j\omega C_b R_b) + j\omega L_1 + r_1$$
$$Z_t = j\omega_r M/\{j/Q + (1 + r/R)\omega_r/\omega - \omega/\omega_r\} \quad \ldots \quad (12)$$

Let $\omega_0$ be the angular frequency of oscillation at which the three impedances are to be resistive. $Z_t$ is resistive when

$$\omega = \omega_0 = \omega_r(1 + r/R)^{\frac{1}{2}} = \omega_r(1 + 1/Q_p Q_s)^{\frac{1}{2}} \quad \ldots \quad (13)$$

and then $Z_t = R_t = \omega_r MQ$. Using these relations, $Z_i$ is resistive at $\omega_0$ if

$$X_a = Qr/(1 + r/R)^{\frac{1}{2}}$$
$$= \omega_0 L/(1 + Q_s/Q_p)(1 + 1/Q_p Q_s)$$

Since the product $Q_p Q_s$ will usually be very large compared with 1, $X_a$ is the reactance of an inductance of value

$$L_a = L/(1 + Q_s/Q_p) \quad \ldots \quad (14)$$

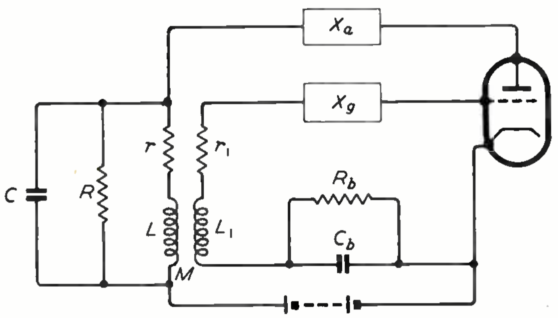


Fig. 3. *Coupled-coil oscillator.*

Similarly from (11), (12) and (13), $Z_o$ is resistive at $\omega_0$ when

$$X_g = \omega_0 C_b R_b^2/\{1 + (\omega_0 C_b R_b)^2\} - \omega_0 L_1(1 - k^2 Q/Q_p)$$

In all practical oscillators $\omega_0 C_b R_b \gg 1$. Assuming also that $Q_p Q_s \gg 1$, $X_g$ is the reactance of a capacitance $C_g$ given by

$$1/C_g = \{1 - k^2/(1 + Q_p/Q_s)\}L_1/LC - 1/C_b$$

If $C_b$ is large

$$C_g = CL/L_1\{1 - k^2/(1 + Q_p/Q_s)\} \quad \ldots \quad (15)$$

Alternatively $C_b$ may be given this value in which case $C_g = \infty$; i.e., $X_g = 0$. This arrangement would be preferred whenever possible, for a separate stabilizing capacitor would require to be shunted by a large inductor to allow the direct component of grid current to pass.

It is clear that the magnitudes of the stabilizing reactances depend greatly on the relative values of $r$ and $R$. It is often assumed that the losses in a coil can be represented either by a parallel or by a series resistance, and for many purposes it is immaterial which convention is chosen. In the present application, however, the distinction is vital.

Given an isolated coil there is no experimental method of distinguishing between a series and a parallel resistance but, when a second coil is coupled magnetically to the first, the phase of the p.d. across the second coil with respect to the current in the first coil is $\frac{1}{2}\pi$ if the losses can be represented entirely by a series resistance. If a parallel loss resistance is included the relative phase is less than $\frac{1}{2}\pi$; i.e., the mutual impedance between the coils then has a resistive as well as an inductive component.

For a single-layer air-cored coil with widely-spaced turns the losses are substantially the same as those in the same length of straight wire. There is no doubt that the losses are here correctly represented by a series resistance. However, when the turns are close together, when dielectric losses occur in the coil former, or when a ferromagnetic core of a screening can is used, the losses are greatly increased, and part of this increase must be represented by a parallel resistance.

The problem of how the various sources of loss should be represented is a complicated one and cannot be discussed here, but as an example the air-cored multi-layer coil used in the experiments had $Q_p = 32$ and $Q_s = 47$ at 50 kc/s. Thus 60% of the total loss is represented by the parallel resistance. All the loss in this coil occurs in the copper.

It might appear that a parallel resistance should also be placed across the second coil $L_1$ in Fig. 3, but this is unnecessary, for such a network could be reduced to the form of Fig. 3. The network of Fig. 3 is in fact only one of a number of possible equivalent circuits. It has the advantage of simplicity and also allows dielectric losses in the capacitor $C$ to be included in the parallel resistance $R$. Methods of measuring $R$ and $r$ are discussed in the Appendix.

This division of losses might account in part for some of the discrepancies observed by previous workers, as Clarke[5] has pointed out. It might also explain some of the errors which arise when the effective resistance of a coil is measured by methods involving mutual inductance[6].

If $r = 0$, (14) and (15) become

$$L_a = 0 \text{ and } C_g = CL/L_1(1 - k^2) \quad \ldots \quad (16)$$

independent of the value of $R$. If $R = \infty$ the results are

$$L_a = L \text{ and } C_g = CL/L_1 \quad \ldots \quad \ldots \quad (17)$$

independent of the value of $r$. When the amplifier is biased to operate without grid current the sole criterion in the last case is $L_a = L$, which is the result obtained by Mallett[7].

Formulae (14) and (15) may be compared with the results obtained by Llewellyn, which give the anode stabilizing impedance as an inductance $L_a = L\{1 + (LC/L_1C_g - 1)/k^2\}$ or a capacitance $C_a = C/\{(1 - LC/L_1C_g)/k^2 - 1\}$, the value of $C_g$ being arbitrary. In particular if $C_g = \infty$ then

$$C_a = Ck^2/(1 - k^2) \quad \ldots \quad \ldots \quad (18)$$

and if $L_a = 0$ $(C_a = \infty)$ then

$$C_g = CL/L_1(1 - k^2) \quad \ldots \quad \ldots \quad (19)$$

The values of $L_a$ and $C_g$ given by (14) and (15) are, of course, in agreement with Llewellyn's general formula, since both satisfy condition (9). If $\mu = \infty$ the value of $X_a$ and consequently also of $X_g$ can have no influence on $\omega_0$, and so stabilizing reactances should be unnecessary, as already pointed out.

## 5. Degree of Stabilization

The foregoing theory, although showing what conditions are necessary for stability, gives no indication of the magnitude of the frequency changes to be expected when these conditions are not exactly fulfilled. A general treatment of this problem is straightforward but tedious; useful information can be gained by considering two particular cases. In the first, $r_g$ is supposed very large, and the variation of $\omega_0$ with $\mu$ is calculated; in the second, $\mu$ is supposed very large and the variation of $\omega_0$ with $r_g$ is found.

If $r_g = \infty$, then from (4), $Z_I = Z_i$, etc., and the frequency equation (6) becomes $X_t + X_i/\mu = 0$. Let $X_a'$ be the correct anode stabilizing impedance, and let $\omega_0'$ be the corresponding frequency. In this state $X_i = X_t = 0$. When $X_a'$ is changed to $X_a$ let $\omega_0'$ change to $\omega_0$. Then $X_t$ changes to $(\omega_0 - \omega_0')dX_t/d\omega$, and $X_i$ to $X_a - X_a' + (\omega_0 - \omega_0')dX_i/d\omega$.

Substituting these values into the frequency equation gives

$$\omega_0 - \omega_0' = -(X_a - X_a')/(dX_i/d\omega + \mu dX_t/d\omega).$$

This assumes that $dX_a/d\omega$ is negligible compared with $dX_i/d\omega$ which is true in all practical cases. Differentiating this equation with respect to $\mu$ gives

$$\partial\omega_0/\partial\mu = \frac{(X_a - X_a')(dX_t/d\omega)}{(dX_i/d\omega + \mu dX_t/d\omega)^2}$$

If $Z_i$ and $Z_t$ have the same form, near to $\omega_0$, as the impedance of an antiresonant circuit, namely

$$Z = R/\{1 + 2jQ(\omega - \omega_a)/\omega_a\},$$

then since $\omega_a \approx \omega_0$ $dX_i/d\omega = -2R_iQ_i/\omega_0$ and $dX_t/d\omega = -2R_tQ_t/\omega_0$. Hence

$$\partial\omega_0/\partial\mu = -\tfrac{1}{2}\omega_0 R_tQ_t(X_a - X_a')/(R_iQ_i + \mu R_tQ_t)^2 \quad \ldots \quad \ldots \quad (20)$$

This expression is strictly valid only for very small oscillation amplitudes, for if the amplitude is large the value of $\mu$ will vary over the oscillation cycle.

In the second case $\mu = \infty$. Then the frequency equation reduces to $X_T = 0$. Let $X_g'$ be the correct grid stabilizing impedance and $\omega_0'$ the corresponding frequency. In this state $X_o = X_t = 0$. When $X_g'$ is changed to $X_g$ let $\omega_0'$ change to $\omega_0$. Now $Z_T = Z_t r_g/(r_g + Z_o)$. Hence if $X_t$ and $X_o$ are small

$$X_T = \{X_t - X_oR_t/(r_g + R_o)\}r_g/(r_g + R_o) = 0$$

Substituting $X_o = X_g - X_g' + (\omega_0 - \omega_0')dX_o/d\omega$, and for $X_t$ as before into this equation gives

$$\omega_0 - \omega_0' = \frac{-R_t(X_g - X_g')}{\{R_t dX_o/d\omega - (r_g + R_o)dX_t/d\omega\}}$$

Differentiating this with respect to $r_g$ gives

$$\partial\omega_0/\partial r_g = \frac{-R_t(X_g - X_g')(dX_t/d\omega)}{\{R_t dX_o/d\omega - (r_g + R_o)dX_t/d\omega\}^2}$$

If $Z_o$ and $Z_t$ have the form of antiresonant impedances near to $\omega_0$ then

$$\partial\omega_0/\partial r_g = \frac{\tfrac{1}{2}\omega_0 Q_t(X_g - X_g')}{\{R_oQ_o - (r_g + R_o)Q_t\}^2} \quad (21)$$

For the oscillator of Fig. 3 it is easily shown, provided $Q_pQ_s \gg 1$, that $Q_i = Q_o = Q_t = Q$, $Q$ being given by (11), and that $R_i = \omega_0 LQ$, $R_o = Q\omega_0 M^2/L$, $R_t = \omega_0 MQ$. If the correct anode stabilizing reactance is an inductance $L_a'$, then for $r_g = \infty$, substitution of these values into (20) gives

$$\partial\omega_0/\partial\mu = -\tfrac{1}{2}\omega_0 M(L_a - L_a')/Q^2(L + \mu M)^2 \quad (22)$$

If the correct grid stabilizing reactance is a capacitance $C_g'$ then for $\mu = \infty$ substitution into (21) gives

$$\partial\omega_0/\partial g_g = -\tfrac{1}{2}(C_g - C_g')/C_gC_g'Q \quad \ldots \quad (23)$$

where $g_g = 1/r_g$ is the grid-input conductance of the valve. The values of $L_a'$ and $C_g'$ are given by (14) and (15).

For the unstabilized oscillator $L_a = 0$ and $C_g = \infty$, and then (22) and (23) become [using (11)]

$$\partial\omega_0/\partial\mu = \tfrac{1}{2}\omega_0(M/L)/Q_sQ(1 + \mu M/L)^2 \ldots \quad (24)$$

$$\partial\omega_0/\partial g_g = -\tfrac{1}{2}\omega_0 r(L_1/L)\{1 + (1 - k^2)Q_s/Q_p\} \quad (25)$$

Variation of $r_g$ will usually be the more important source of frequency change. This can be minimized by making the coefficient of coupling as near to 1 as possible, and making $r$ small.

Equation (24) may also be expressed in terms of the equivalent anode conductance $g_a = 1/r_a$ of the valve. With this convention it is easily shown for the circuit of Fig. 3 and with $r_g = \infty$ that $-\mu\omega_0 MQ = \omega_0 LQ + 1/g_a$. Using this



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relation between $\mu$ and $g_a$, and also equations (11), equation (24) becomes

$$\partial\omega_0/\partial g_a = \tfrac{1}{2}\omega_0 r \quad \ldots \quad (26)$$

## 6. Experimental Results

Experiments were started with the arrangement shown in Fig. 3, but it was found that variations in inter-electrode capacitances and harmonic content were masking the results. These difficulties were overcome by inserting suitable resistors in the cathode lead, rearranging the connections of the grid leak and capacitor, and interposing a cathode-follower buffer stage between the network and the amplifier proper. A pentode amplifier was used and the gain was controlled by adjusting the screen voltage. Variations in $r_g$ and $\mu$ were simulated by variable resistors $r_g$ and $r_a$ connected across the network output and input terminals (2,3 and 1,3 in Fig. 1). The effective $\mu$ of the amplifier is $r_a g_m$, where $g_m$ is the mutual conductance. These resistors and the amplifier gain were adjusted simultaneously to give a very small oscillation amplitude.

Inductances $L$ and $L_1$ in Fig. 3 were air-cored multi-layer coils wound with enamelled copper wire. Their parameters, measured in the way described in the Appendix, were at 50 kc/s.

$L = 530\,\mu$H, $L_1 = 523\,\mu$H, $M = 215\,\mu$H,
$R = 5{,}340\ \Omega$, $r = 3{\cdot}55\ \Omega$, $r_1 = 8{\cdot}00\ \Omega$,

from which

$k = 0{\cdot}408$, $Q_p = 32{\cdot}1$, $Q_s = 46{\cdot}8$

By adding resistors in parallel with the coil $L$ the value of $R$ could be adjusted, and by adding resistors in series $r$ could be changed. In this way the ratio $Q_p/Q_s$ could be varied over a wide range. The main tuning capacitor $C$ was $0{\cdot}0190\,\mu$F, and the oscillation frequency 50 kc/s.

The experimental method was as follows: With $r_a = \infty$ the change in $\omega_0$ due to a definite small change in $r_g$ was measured for various values of $C_g$. With $r_g = \infty$ the effect of a small change in $r_a$ was measured for different values of $L_a$. In all cases the frequency change was a linear function of $L_a$ or $1/C_g$. Hence by drawing straight-line graphs the values of $L_a$ and $C_g$ required for exact stabilization could be found. Finally these values were checked by varying both $r_a$ and $r_g$ over a wide range. This procedure was repeated for various values of $Q_p$ and $Q_s$ with the results shown in Table 1.

The theoretical values of $L_a$ and $C_g$ are calculated from (14) and (15). In view of the various experimental uncertainties the agreement with the experimental values is considered to be satisfactory. The almost constant value of $C_g$ is due to the low value of $k^2$ ($= 0{\cdot}166$). Equation (15) shows that in such cases the value of $Q_p/Q_s$ has little effect on $C_g$ which is approximately equal to $CL/L_1$; i.e., $\omega_0 C_g = 1/\omega_0 L_1$. Thus $C_g$ is the capacitance required to tune the series circuit $L_1$, $C_g$, $r_1 + r_g$ to the oscillation frequency.

**TABLE 1**

| $Q_p$ | $Q_s$ | $L_a(\mu$H) | | $C_g(\mu$F $\times 10^{-4}$) | |
|---|---|---|---|---|---|
| | | Meas. | Calc. | Meas. | Calc. |
| 32·1 | 5·85 | 458 | 448 | 201 | 198 |
| 32·1 | 10·8 | 376 | 397 | 202 | 201 |
| 32·1 | 23·6 | 290 | 305 | 206 | 207 |
| 32·1 | 46·8 | 212 | 216 | 212 | 213 |
| 14·5 | 46·8 | 140 | 125 | 220 | 221 |
| 9·1 | 46·8 | 76 | 86 | 223 | 223 |

According to Llewellyn's theory, stabilization should be obtained with either an anode capacitance of $0{\cdot}00378\,\mu$F, given by (18), or with a grid capacitance of $0{\cdot}0231\ \mu$F, given by (19). Llewellyn's assumption that all the dissipation is in the grid-input resistance cannot be realized in practice but, by reducing $r_g$ to the lowest workable value for this particular oscillator, namely 112 $\Omega$, about 80% of the total dissipation was obtained in $r_g$. With the values of $C_a$ and $C_g$ indicated above the frequency changes produced by small variations of $r_a$ and $r_g$ were measured for different values of $r_a$. The rate of change of oscillation frequency $f_0$ ($= \omega_0/2\pi$) could then be computed to give the results of Table 2.

**TABLE 2**

| $C_a(\mu$F) | $C_g(\mu$F) | $r_a(\Omega)$ | Change in $f_0$ (kc/s per mho) | |
|---|---|---|---|---|
| | | | $g_a$ | $g_g$ |
| $\infty$ | 0·0231 | $\infty$ | 121 | −13·2 |
| | | 8820 | 121 | −13·2 |
| | | 4350 | 121 | −13·2 |
| 0·00378 | $\infty$ | $\infty$ | 2800 | −158 |
| | | 8820 | 2270 | −125 |
| | | 4350 | 1730 | −102 |
| $\infty$ | $\infty$ | $\infty$ | −4120 | −160 |
| | | 8820 | −4120 | −212 |
| | | 4350 | −4120 | −260 |

The last three rows give the results for the same oscillator without stabilizing reactances. It is seen that Llewellyn's values, though different from those required for complete stabilization produce a marked improvement when (in this particular arrangement) the reactance is in the grid circuit. Anode stabilization is much less effective. The results for $r_a = \infty$ are of particular interest, since in this case (which corresponds to $\mu = \infty$) no stabilizing reactances should be required according to Llewellyn's theory.

A value of $r_g = 112\,\Omega$ would be very unusual. Under more realistic conditions $r_g$ would be very much larger and almost all the dissipation would be in the coil resistances. Measurements of frequency stability for the limiting case of $r_a \to \infty$ and $r_g \to \infty$ were therefore carried out, see Table 3. The frequency changes were found to be linear with respect to moderate changes in $g_a$ and $g_g$ away from 0.

**TABLE 3**

| $C_a(\mu F)$ | $C_g(\mu F)$ | Change in $f_o$(kc/s/mho) | | | |
|---|---|---|---|---|---|
| | | $g_a$ | | $g_g$ | |
| | | Meas. | Calc. | Meas. | Calc. |
| $\infty$ | 0·0231 | 87 | 89 | −19 | −15·3 |
| 0·00378 | $\infty$ | 1170 | 1200 | −205 | −194 |

The theoretical values are calculated from (22), (23), (25) and (26) using the measured values of the coil parameters. The anode stabilizing capacitance of 0·00378 $\mu$F is treated as an inductance of $-2680\,\mu$H. In fact it has an anti-stabilizing effect. Stability with respect to changes in $g_g$ is about the same as for the heavily-loaded oscillator. Llewellyn's value of grid-stabilizing capacitor is very near to the correct value because of the smallness of $k$. If $k$ were near to 1, the discrepancy would be much larger unless $Q_s/Q_p$ were very large. This can be seen be comparing formulae (15) and (19).

These experiments have shown that, provided the effects of harmonics and inter-electrode capacitances are eliminated, the theoretical and actual values of the stabilizing reactances and of the frequency shifts are in good agreement. In general, the values of grid and anode stabilizing reactances specified by Llewellyn will not be satisfactory.

## APPENDIX

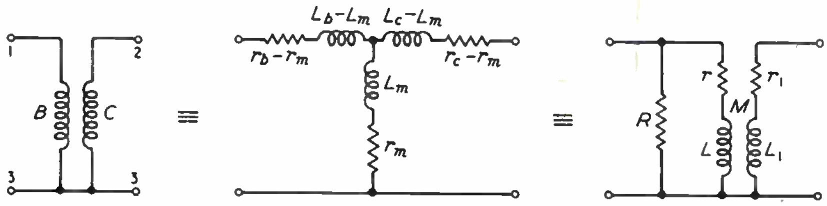
### Equivalent Circuits for Coupled Coils

The behaviour of the coupled-coil network shown in Fig. 4(a) can be specified in terms of the three impedances $j\omega L_b + r_b$, $j\omega L_c + r_c$, and $j\omega L_m + r_m$ shown at (b). To obtain the equivalent circuit (c) all that is necessary is to equate these impedances to the open-circuit input, output, and transfer impedances of (c). Thus

$$j\omega L_b + r_b = R(j\omega L + r)/(j\omega L + r + R)$$
$$j\omega L_c + r_c = j\omega L_1 + r_1 + \omega^2 M^2/(j\omega L + r + R)$$
$$j\omega L_m + r_m = j\omega MR/(j\omega L + r + R)$$

These equations may be solved to find $R$, $r$, $r_1$, $L$, $L_1$, and $M$. The general expressions are cumbersome, but if it is assumed that $Q_pQ_s \gg 1$ (which will be true in all practical cases) the relations simplify to

$$L = L_b,\ L_1 = L_c,\ M = L_m$$
$$R = \omega^2 LM/r_m,\ r = r_b - r_m L/M,\ r_1 = r_c - r_m M/L,$$ from



(a) (b) (c)

Fig. 4. *Equivalent circuits for coupled coils.*

which

$$Q_p = \omega M/r_m,\ Q_s = \omega L/(r_b - r_m L/M).\ 1 + Q_p/Q_s = r_b M/L r_m$$

$L_b$ and $r_b$ are found by measuring the impedance between terminals 1 and 3 in Fig. 4(a), and $L_c$ and $r_c$ by measuring between 2 and 3. $L_m$ and $r_m$ are obtained by measuring between 1 and 2, and using the previously found values of $L_b$, $L_c$, $r_b$, and $r_c$. The sign of $r_m$ is the same as that of $L_m$.

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# "THE ENGINEER"

Not many technical journals are yet in a position to celebrate a centenary. *The Engineer* has this distinction and, with its issue dated 4th January 1956, commemorates its first issue dated precisely 100 years earlier.

Its early days coincided with the growth of railways and it naturally devoted much space to them. Its outlook has always been broad, however, and it reported the early experiments of Marconi, although in terms which strike the modern ear as strange!

For the last 91 years the Editor-in-Chief has been a Pendred—father, son and grandson. For the present editor, therefore, the occasion is one not only of celebrating the centenary of his journal but of a long family connection with it.

# ADCOCK DIRECTION FINDER

## *Polarization Errors Due to Aerial Bending*

**By W. C. Bain, M.A., B.Sc., Ph.D.**

*(Official communication from D.S.I.R. Radio Research Station, Slough)*

**SUMMARY.**—It was found that the pull of the triatics (supporting the central sense aerial) on the main aerials of an Adcock h.f. direction finder caused the main aerials to bend inwards, so that a large polarization error was produced on frequencies near 20 Mc/s. Theoretical calculations are made for the case of aerials with an overall slope of 1°, with a distant loop transmitter at 6·3° elevation and with a loop tilt of 6° from the horizontal: the total polarization error found is 6° for a perfectly-reflecting ground and 19° for ground of conductivity 0·02 mho/m. In practical direction finding the errors are not likely to be as high as this on the average; but the figure of 3° is obtained for the root-mean-square error from a transmitter on 20 Mc/s at 20° elevation despite the small amount of aerial bending assumed.

### 1. Introduction

IN one type of U-Adcock high-frequency direction finder in common use a sense aerial of cage construction is suspended by triatics from the tops of the main aerials, which are self-supporting. The pull of the triatics varies according to the tension in the sense aerial, but can be large enough to produce appreciable inward bending in the main aerials. It is shown in this paper that this effect will lead to polarization errors, which, though negligible at the low-frequency end of the range, become large at frequencies of the order of 20 Mc/s.

### 2. Experimental Discovery of the Error

Polarization error of appreciable magnitude due to aerial bending was first discovered when investigations on a particular direction finder were made to determine the cause of an observed error which rose rapidly with frequency. The installation was of the type mentioned above, with main aerials 9·1 m high tapered towards the top and spaced 6·1 m apart; an earth mat of 31 m diameter was fitted. The polarization error was measured with the aid of a loop transmitter at 6·3° elevation, the loop of which could be rotated about a horizontal axis in the plane of propagation so that its plane was 6° from the horizontal, thus giving at the direction-finder a field in which there was a small vertically-polarized and a large horizontally-polarized component. (Details of the techniques of measurement are given by Ross[1].) On one occasion when measurements were taken the total polarization error was found to be less than 3° at 13 Mc/s but over 30° at 20 Mc/s. On removal of the triatics and straightening of the main aerials the error at 20 Mc/s fell to 10°. This residual error is about three times as great as the values commonly found at lower frequencies; but it appears to be peculiar to the installation tested and is certainly unconnected with the aerial bending.

The inward bending of the aerials due to the sense aerial triatics was subsequently measured with the aid of a theodolite. When this was done it was found that the top of the west aerial was pulled in by about 15 cm and that of the east aerial by 3 cm. The polarization errors quoted were obtained with the loop transmitter to the south of the direction finder, so the slopes of the north and south aerials were immaterial. It was confirmed that the increase in error still occurred when the bending was done by ropes in the absence of the triatics and sense aerial; this ruled out the possibility of the error being due to the presence of the conducting material in the sense aerial and its supports.

MS accepted by the Editor, March 1955

### 3. Theoretical Evaluation of the Polarization Error

The figures just cited have been taken as giving the order of magnitude of the bending usually found, and have been used in theoretical calculations of the errors likely to be produced. Expressions for the polarization error are derived in Appendix 1. Now in these calculations a number of assumptions have been made to simplify the analysis. These will be stated here and later their validity and the effects of the modification of some of them will be discussed. They are as follows:

(1) All aerials are vertical except for the top portion (one quarter of the total length), which has a constant slope inwards towards the centre of the system.

(2) The slope of every portion of the aerials from the vertical is small, say less than 6°.

(3) The ground is perfectly reflecting, the phase change on reflection being 0° for vertically-polarized waves and 180° for horizontally-polarized waves.

(4) The transmitting current distribution for any of the Adcock aerials is of the form shown in equation (9) of Appendix 1. This quantity is required for the calculation of the voltage induced in an Adcock aerial which is receiving a signal.

(5) The transmitter is at a sufficient distance

from the Adcock for the incident radiation to be considered as a plane wave and for the surface wave to be neglected.

Of these assumptions (1) is certainly not exactly true, as the bottom part of the aerial cannot remain truly vertical if the aerial is pulled inwards at the top. However, by actual observation it was discovered that the aerials were bent very little except in the top section. No doubt this happens because each aerial consists of four separate sections, of equal length but each of a different radius, which are fitted together so that the smallest is at the top of the aerial. The diameters are 4, 3, 2 and 1½ inches.

Assumption (2) is justified for the angles used in the calculation here, since the angle between the upper part of the aerial and the vertical is taken to be 4°.

The assumption (3) is probably nearly correct as regards the reflection coefficient for horizontally-polarized waves. However, the situation is more uncertain as regards the vertically-polarized waves, which will be reflected strongly by the earth mat but not so well by the ground. At 20 Mc/s, the frequency of the greatest interest here, the reflection coefficient for vertical polarization is only 0·4 and the phase change at reflection is a lag of about 160°; these figures apply to ground of conductivity 0·02 mho/m, which was the value obtained by measurement on the site. It is unlikely that the mat is responsible for giving the whole of the reflected field affecting the aerials, so the value of $E_y$ given in equation (4) is probably an overestimate at all heights below the top of the aerial. The effect of introducing a reflection coefficient different from unity will be discussed later.

The transmitting current distribution assumed is probably reasonably near the truth; a more accurate formula would almost certainly complicate the analysis to an extent which would not be justified in view of the uncertainties in other parts of the computation. No change in this distribution will therefore be considered.

Assumption (5) is incorrect in both its parts if waves from the local loop transmitter are concerned. Of course, in the consideration of errors on a signal from a distant transmitter it is fully justified.

The magnitude of the polarization error calculated on this basis is shown plotted against frequency in Fig. 1. Here the parameter $\gamma$, the angle between the vertical and the line joining aerial tip to base, has been chosen to be 1°. The error certainly rises markedly with frequency, in a way very similar to that exhibited by the error in practice. However, it reaches a value at 20 Mc/s which is less than a third of the observed figure.

The effect of errors in the initial assumptions will now be considered. First of all, a change in the assumed shape of the Adcock aerial is not likely to affect the results seriously. This has been checked by carrying out the calculations for an aerial in the shape of a uniform rod clamped at one end and with a force acting at the other, and for an aerial with a constant slope of 1° to the vertical over its whole length. Both gave similar results to the original. The cantilever shape gave somewhat higher errors at the low frequencies, and the other a lower result at 20 Mc/s.

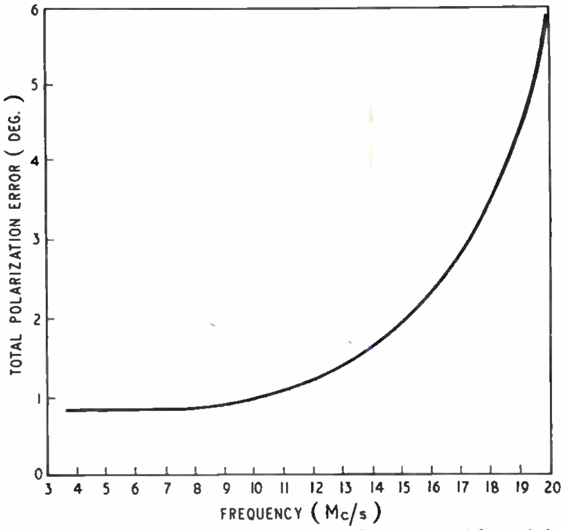


Fig. 1. *Polarization error for Adcock system with aerials bent inwards in the top portion, calculated from Equ.* (10) *with the overall slope* $\gamma = 1°$.

Because of its small size, the earth mat probably does not greatly affect the field at the aerials, particularly at their upper portions and with a low angle of elevation for the incident radiation. The results are therefore likely to be improved by assuming that the Adcock is situated on ground of finite conductivity. Now the introduction of a more realistic value of reflection coefficient for vertically-polarized waves has probably the biggest effect of all on the theoretical results. It we ignore the effect of the earth mat, this is given at 20 Mc/s by

$$R_V = 0{\cdot}4\, e^{-j160°}$$

Then the total field near the ground will be proportional to $(1 + R_V)$, approximately 0·6 in this case. The expression for the vertical field at any height $y$ is

$$E_y = E_0 \sin \alpha \cos \delta \left[e^{j\beta y \sin \delta} + R_V\, e^{-j\beta y \sin \delta}\right]$$

where $\alpha$ is the tilt of the transmitter loop from the horizontal, $\delta$ is the angle of elevation of the transmitter, and $\beta$ is the propagation constant of the waves in free space.

As $\beta y \sin \delta < 25°$ always, it turns out that quite a good approximation to $E_y$ is

$$E_y = 0{\cdot}6\, E_0 \sin \alpha \cos \delta.$$

The cosine factor in the denominator of equation (10) makes little difference, so this means that the e.m.f. induced in the north–south aerial pair will be reduced by a factor $\frac{2}{0{\cdot}6} = 3{\cdot}3$ approx. (The transmitter is assumed to be at 180° azimuth.) This would increase the polarization error at 20 Mc/s by this factor, making it 19° approximately. Such a value is of the same order as the observed error, though it should be borne in mind that on the occasion when the slope of the aerial was measured it was only the west aerial which had a 1° slope, whereas the east aerial was nearly vertical. It is interesting to note that $R_V$ will tend to + 1 as the frequency is decreased, so that the decrease of error with decrease of frequency will be more rapid than is indicated by Fig. 1.

The difference between the plane wave considered and the actual direct and reflected waves arriving at the receiver is not likely to disturb the result seriously. The use of the loop transmitter, which is a reasonable approximation to a magnetic dipole, eliminates the 'proximity' errors (see Ross[1]) in the case of a perfect Adcock, since a magnetic dipole with its axis vertical radiates everywhere a purely horizontally-polarized field; the presence of a small amount of bending in the Adcock aerials does not affect this conclusion. The presence of the surface wave will, however, upset the results to a greater extent. Calculations indicate that the field $E_y$ is altered by a factor of about 1·5 by the addition of the surface wave term. The surface wave is negligible in this case for horizontal polarization, so on this basis the error will be decreased from the value of 19° mentioned above to about 13°. The 19° error, of course, is the more appropriate to use in the case of a distant transmitter.

## 4. Effect on Practical Direction Finding

The bearing errors likely to be experienced in practice due to this aerial bending effect will not be so severe as might at first appear likely, since at 20 Mc/s radio waves will in general be incident on the direction finder at a low angle of elevation. This is because the critical frequency of the $F_2$ layer rarely approaches 20 Mc/s. Hence a wave incident on the direction finder of frequency near 20 Mc/s will be assumed to have an angle of elevation less than 20°; in such a case the horizontally- and vertically-polarized components are likely to be present in roughly equal amounts. Polarization errors have therefore been calculated for angles of elevation of 6·3°, 10° and 20°. They have been corrected for a finite ground conductivity (0·02 mho/m), and reduced by a factor of $\sqrt{2}$ to allow for averaging over all phase differences between horizontally- and vertically-polarized waves. The results appear in Table 1.

**TABLE 1**

*Estimated Root-Mean-Square Polarization Error due to 1° Overall Aerial Bending for Downcoming Waves at Different Angles of Elevation*

| Angle of Elevation (deg.) | Frequency (Mc/s) | Root-Mean-Square Polarization Error (deg.) |
|---|---|---|
| 6·3 | 16 | 0·6 |
| 10 | 16 | 0·5 |
| 20 | 16 | 0·8 |
| 6·3 | 20 | 1·8 |
| 10 | 20 | 1·3 |
| 20 | 20 | 3·0 |

An overall bend of 1° in an aerial implies that its tip is displaced inwards by 15 cm approximately. It should be noted that the errors at 20 Mc/s would be reduced if shorter aerials were used, and increased if the site were on ground of poorer conductivity.

## 5. Conclusions

It has been shown theoretically that if the aerials of an Adcock direction finder are bent inwards by the pull of triatics to give an overall slope of 1°, then polarization errors will be experienced which increase rapidly with frequency over 12 Mc/s. For a distant loop transmitter at 6·3° elevation and a loop tilt of 6° to the horizontal the total polarization error is 6° if the ground is perfectly reflecting, and 19° if the ground has conductivity 0·02 mho/m, even with this slight amount of aerial bending. The root-mean-square errors in practical direction finding with such an installation appear unlikely to exceed 3° at 20 Mc/s.


## Acknowledgments

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The author is indebted to Marconi's Wireless Telegraph Co. Ltd., for the provision of facilities to perform the experiments described in this paper. He is also indebted to Dr. E. N. Bramley for details of some early experimental work on this effect. Assistance was rendered in the observational work and in the computations by Mr. J. Bell and Mr. J. Meadows.


## APPENDIX 1

*Polarization Error due to Bending of the Aerials*

The principal symbols used are as follows.

*x, y, z.* Co-ordinate axes. The *xy* plane is a vertical plane containing the transmitter and centre of the receiver, the $x$ axis being horizontal and pointing in the direction from transmitter to receiver, the $y$ axis being directed vertically upwards. The $z$ axis is then horizontal (see Fig. 2). In the experimental arrangement considered here the transmitter is to the south of the receiver. The $x$ axis therefore points north, and the $z$ axis east.

$E_x$, $E_y$, $E_z$. Components of the electric field.

$l$. Total length of each Adcock aerial.

$\gamma$. Angle to vertical of a line joining the tip of an aerial to its base. Each aerial is assumed to be bent inwards towards the central axis of the system by the same amount.

$\alpha$. The angle through which the axis of the transmitting dipole has to be rotated from the $y$ axis towards the $z$ axis in the $yz$ plane to reach its actual position; the transmitter is considered here to be a magnetic dipole. As a loop transmitter is used to simulate the magnetic dipole, the plane of the loop will then be at an angle $\alpha$ from the horizontal. If $\alpha = 90°$ the field in the $xy$ plane is vertically polarized; if $\alpha = 0°$ it is horizontally polarized.

$\delta$. The angle of elevation of the transmitter from the direction finder. In the calculations the transmitter is assumed to be at a great distance from the direction finder, but in practice $\delta$ is measured from the centre of the system.

$V_E$, $V_S$, $V_W$, $V_N$. The open-circuit e.m.fs at the base of each aerial.

$s$. Aerial spacing in each Adcock pair.

$\lambda$. Wavelength of radio waves in free space.

$\beta = 2\pi/\lambda$.

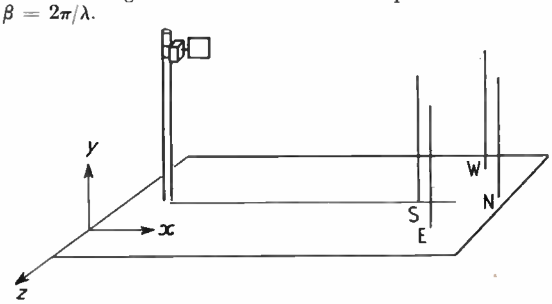


Fig. 2. *The co-ordinate system used.*

In accordance with the assumptions discussed in Section 3, the slope of the west aerial is taken to be

$$\left.\begin{aligned} \frac{dz}{dy} &= 0,\ 0 < y < \frac{3}{4}l \\ \frac{dz}{dy} &= 4\gamma,\ \frac{3}{4}l < y < l \\ \frac{dx}{dy} &= 0 \text{ everywhere} \end{aligned}\right\} \quad \ldots \quad \ldots \quad (1)$$

Now at height $y$ the field along the aerial is $E_W(y)$, where

$$\begin{aligned} E_W(y) &= E_y,\ 0 < y < \frac{3}{4}l \\ &= E_y \cos 4\gamma + E_z \sin 4\gamma,\ \frac{3}{4}l < y < l \end{aligned} \quad (2)$$

Similarly for the east aerial which is assumed to be deflected inwards precisely as in the case of the west aerial.

$$\begin{aligned} E_E(y) &= E_y,\ 0 < y < \frac{3}{4}l \\ &= E_y \cos 4\gamma - E_z \sin 4\gamma,\ \frac{3}{4}l < y < l \end{aligned} \quad (3)$$

Now suppose a plane wave is incident on the system from a distant magnetic dipole transmitter, and assume the ground is perfectly reflecting; the field components at a height $y$ above the ground in the east-west plane can then be shown to be

$$\left.\begin{aligned} E_x &= 2jE_0 \sin\alpha \sin\delta \sin(\beta y \sin\delta) \\ E_y &= 2E_0 \sin\alpha \cos\delta \cos(\beta y \sin\delta) \\ E_z &= -2jE_0 \cos\alpha \cos\delta \sin(\beta y \sin\delta) \end{aligned}\right\} \quad (4)$$

Here $E_0$ is the amplitude of the field radiated by the transmitter in free space at a point on the meridian plane of its dipole and at the same distance from it as the Adcock.

The open-circuit e.m.f. in the west aerial is given by

$$V_W = \frac{1}{i_0}\int_0^l E_W(y)\, i_t(y)\, dy,$$

where $i_t(y)$ is the transmitting current distribution on the aerial when the current at the base is $i_0$. This method of calculating the voltage at the base terminal of a receiving aerial, which is based on considerations of reciprocity, is an exact one for a linear aerial in a general incident field; it has been given by Burgess[2]. A similar expression holds for the east aerial. In this problem the e.m.f. produced in the east-west Adcock pair is required. This is

$$\begin{aligned} V_W - V_E &= \frac{1}{i_0}\int_0^l [E_W(y) - E_E(y)]\, i_t(y)\, dy \\ &= \frac{2 \sin 4\gamma}{i_0}\int_{3l/4}^l E_z(y)\, i_t(y)\, dy \quad \ldots \quad (5) \end{aligned}$$

If the north-south aerials have the same shape

$$\begin{aligned} E_S(y) &= E_y \exp\left(\frac{j\beta s \cos\delta}{2}\right),\ 0 < y < \frac{3}{4}l \\ &= (E_y \cos 4\gamma + E_x \sin 4\gamma)\exp\left(\frac{j\beta s \cos\delta}{2}\right),\ \frac{3}{4}l < y < l \quad \ldots \quad \ldots \quad (6) \end{aligned}$$

$$\begin{aligned} E_N(y) &= E_y \exp\left(-\frac{j\beta s \cos\delta}{2}\right),\ 0 < y < \frac{3}{4}l \\ &= (E_y \cos 4\gamma - E_x \sin 4\gamma)\exp\left(-\frac{j\beta s \cos\delta}{2}\right),\ \frac{3}{4}l < y < l \quad \ldots \quad \ldots \quad \ldots \quad (7) \end{aligned}$$

$$\begin{aligned} \therefore V_S - V_N &= \frac{1}{i_0}\int_0^l [E_S(y) - E_N(y)]\, i_t(y)\, dy \\ &= \frac{1}{i_0}\cdot 2j \sin\left(\frac{\beta s \cos\delta}{2}\right)\int_0^l E_y\, i_t(y)\, dy \\ &\quad + \frac{1}{i_0}\cdot 2 \sin 4\gamma \cos\left(\frac{\beta s \cos\delta}{2}\right)\int_{3l/4}^l E_x\, i_t(y)\, dy \quad \ldots \quad \ldots \quad \ldots \quad (8) \end{aligned}$$

The factor of $\cos 4\gamma$ has been neglected as it is so nearly equal to unity. $E_x$ and $E_y$ are given by equation (4) and from these it can be seen that $E_x$ is smaller than $E_y$ by a factor of $\tan\delta \tan(\beta y \sin\delta)$. The second term in equation (8) is reduced still more compared with the first by the factor $\sin 4\gamma$. The factor $\cot\left(\frac{\beta s \cos\delta}{2}\right)$ tends to increase the ratio of the second term to the first, but does not outweigh the $\tan\delta$ factor in the range of frequencies considered. Hence as $\gamma$ is a small angle the second term in equation (8) may be neglected compared with the first. We may, therefore, write

$$V_S - V_N = \frac{2j \sin\left(\frac{\beta s \cos\delta}{2}\right)}{i_0}\int_0^l E_y(y)\, i_t(y)\, dy$$

To evaluate this integral and that in equation (5) it is necessary to assume a particular transmitting current distribution $i_t(y)$. This has been selected to be

$$i_t(y) = \frac{i_0 \sin\{\mu\beta\,(l-y)\}}{\sin(\mu\beta l)} \quad \ldots \quad (9)$$

This is the common sinusoidal current distribution, but with the propagation constant $\beta$ modified empirically by the factor $\mu$ to take into account the finite thickness of the aerial. As the actual aerial varies in radius with height the value of $\mu$ was selected by examination of the resonance points in curves of aerial impedance which were obtained on the equipment tested. The value giving the best fit with the observed resonances was $\mu = 1{\cdot}14$.

$$= \frac{1}{2\beta}\left[\sin(\beta l \sin\delta)\left\{\frac{1}{\mu+\sin\delta}+\frac{1}{\mu-\sin\delta}\right\} - \frac{\sin\left\{\frac{\beta}{4}(3l\sin\delta-\mu l)\right\}}{\mu+\sin\delta} - \frac{\sin\left\{\frac{\beta}{4}(3l\sin\delta+\mu l)\right\}}{\mu-\sin\delta}\right] \quad \ldots \quad (11)$$

The total polarization error $\epsilon$ of the system in these conditions is given by

$$\tan\epsilon = \frac{V_W - V_E}{V_S - V_N},$$

since both voltages are in phase, and there will be no quadrature error.

$$= \frac{\sin 4\gamma}{j\sin\left(\frac{\beta s\cos\delta}{2}\right)} \frac{\int_{3l/4}^{l} E_z(y)\, i_t(y)\, dy}{\int_0^l E_y(y)\, i_t(y)\, dy}$$

Substituting from equations (4) and (9)

$$\tan\epsilon = \frac{-\sin 4\gamma \cot\alpha \int_{3l/4}^{l} \sin(\beta y \sin\delta)\sin\{\mu\beta\,(l-y)\}\,dy}{\sin\left(\frac{\beta s\cos\delta}{2}\right)\int_0^l \cos(\beta y\sin\delta)\sin\{\mu\beta\,(l-y)\}\,dy} \quad \ldots \quad (10)$$

The integrals have been evaluated with the following results.

$$\int_{3l/4}^{l} \sin(\beta y \sin\delta)\sin\{\mu\beta\,(l-y)\}\,dy$$

$$\int_0^l \cos(\beta y\sin\delta)\sin\{\mu\beta\,(l-y)\}\,dy$$

$$= \frac{1}{2\beta}\left\{\cos(\beta l\sin\delta) - \cos(\mu\beta l)\right\}\left(\frac{1}{\mu-\sin\delta}+\frac{1}{\mu+\sin\delta}\right) \quad \ldots \quad (12)$$

The formulae (10), (11) and (12) enable the error to be computed.

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[1] W. Ross, "The Specification and Measurement of Polarization Error in Adcock-Type Direction Finders", *Proc. Instn elect. Engrs*, Part III, 1949, Vol. 96, p. 269.

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# CIRCUITS WITH NON-LINEAR RESISTANCE

## *Calculation of Behaviour*


**By A. Liebetegger, Ph.D.**

*(University of Southampton)*


IT may be of interest to examine some simple circuits containing diodes as non-linear resistances, and to compare their behaviour with that of the corresponding linear circuits.

The particular circuits to be considered consist of a diode in series with a supply of a constant e.m.f. and an impedance $Z$, which will be taken to be either a purely ohmic resistance, an inductance, or a capacitance, or finally a series combination of all three.

The diode as a non-linear resistance obeys Child's Law

$$i = kv^{3/2} \quad \ldots \quad (1a)$$

or

$$v = (i/k)^{2/3} \quad (1b)$$

Let a constant e.m.f. $E$ be switched on at $t = 0$.

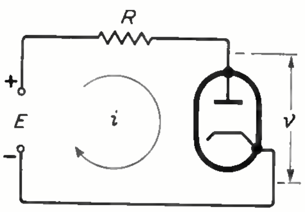


Fig. 1. *Linear and non-linear resistances in series.*


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Then we have

*Case* I

The impedance is a purely ohmic resistance $R$ (Fig. 1), and

$$Ri + v = E \quad \ldots \quad (2)$$

Substituting for $v$ from (1b) we get $i^{2/3} = k^{2/3}(E - Ri)$, or finally

$$-i^2 = k^2 R^3 (i - E/R)^3 \quad \ldots \quad (3)$$

To find $i$ we can use the well-known graphical procedure shown in Fig. 2 in which the diode characteristic, given by (1a), is plotted; and also the straight line $i = (-1/R)(v - E)$ obtained from (2) with $1/R = \tan\alpha$. The ordinate of the intersection of the straight line and the characteristic is the solution of (3).

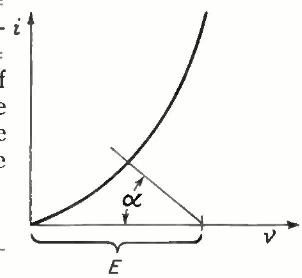



Fig. 2. *Graphical construction for Fig.* 1.

*Case 2*

The diode is in series with a pure inductance $L$ (Fig. 3). The equation of the circuit is then $L\,di/dt + v = E$, which, by (1b), gives the differential equation for $i$

$$di/dt = E/L - (1/Lk^{2/3})i^{2/3} = K(E/KL - i^{2/3}),$$

where $K = 1/Lk^{2/3}$ .. .. .. .. (4)

Putting $E/KL = a^2$ .. .. .. (5)

we have, separating variables, $di/(a^2 - i^{2/3}) = K\,dt$, and obtain

$$Kt = \text{const} + \int di/(a^2 - i^{2/3}) \quad .. \quad .. \quad (6)$$

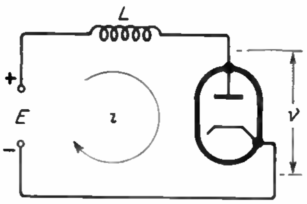



Fig. 3. *Inductance and non-linear resistance.*

If we put $i = x^3$, so that $di = 3x^2\,dx$, the integral on the right-hand side of (6) becomes

$$\int 3x^2\,dx/(a^2 - x^2) = 3\int dx\left\{a^2/(a^2 - x^2) - 1\right\}$$

$$= \frac{3a}{2}\log\frac{a + x}{a - x} - 3x$$

Thus (6) becomes

$$Kt = \text{const} + \frac{3a}{2}\log\frac{a + i^{1/3}}{a - i^{1/3}} - 3i^{1/3}.$$

Since, when $t = 0$, $i = 0$, the constant of integration is zero and

$$t = \frac{3a}{2K}\log\frac{a + i^{1/3}}{a - i^{1/3}} - \frac{3}{K}i^{1/3} \quad .. \quad .. \quad (7)$$

Equation (7) clearly satisfies the initial condition, and we see that $t \to \infty$ as $i^{1/3} \to a$, or as $i \to a^3$. By (5) and (4) we have $a^3 = kE^{3/2}$, so that $i \to kE^{3/2}$.

The limiting value of the current is the value which the current would have if $E$ were applied directly to the diode; i.e., if the inductance $L$ were absent. The same is true of the limiting value of the current in the case of a linear (ohmic) resistance in the place of the diode (although, of course, in that case the limit is determined by Ohm's not Child's Law). But the manner in which the limit is approached in the linear case is somewhat simpler according to the exponential function

$$(E/R)(1 - e^{-Rt/L}) \quad .. \quad .. \quad .. \quad (7')$$

Curve (a) in Fig. 4 shows the growth of the current $i$ with time as computed from equation (7), while curve (b) shows the current in the linear case increasing according to the exponential function (7′). The curves show that in the non-linear case the current grows more slowly than in the comparable linear case. As can be seen by differentiating (7) and (7′) the two curves (a) and (b) in Fig. 4 have a common tangent at the origin.

It may be added that in the linear case the ohmic resistance which was substituted for the diode was of such magnitude as to make the limiting values of the currents equal; i.e.,

$$E/R = kE^{3/2}, \text{ or } R = 1/kE^{1/2}.$$

*Case 2a*

If, when a current $i_0$ flows in the circuit of Case 2, the e.m.f. is short-circuited, we have $L\,di/dt + v = 0$, or by (1b): $di/dt = -i^{2/3}/Lk^{2/3}$. Separating variables, we get $di/i^{2/3} = -dt/Lk^{2/3}$, and, integrating: $i^{1/3} = \text{const} - t/3Lk^{2/3}$, where $t = 0$ at the moment of short circuit. Using the initial condition in this case, we find that the constant of integration is $i_0^{1/3}$ so that we get, finally: $i = (i_0^{1/3} - t/3Lk^{2/3})^3$ .. .. (7a)

The current is reduced to zero in the finite time $t = t_1 = 3Lk^{2/3}\,i_0^{1/3}$. This result is entirely different from that obtained with a linear resistance.

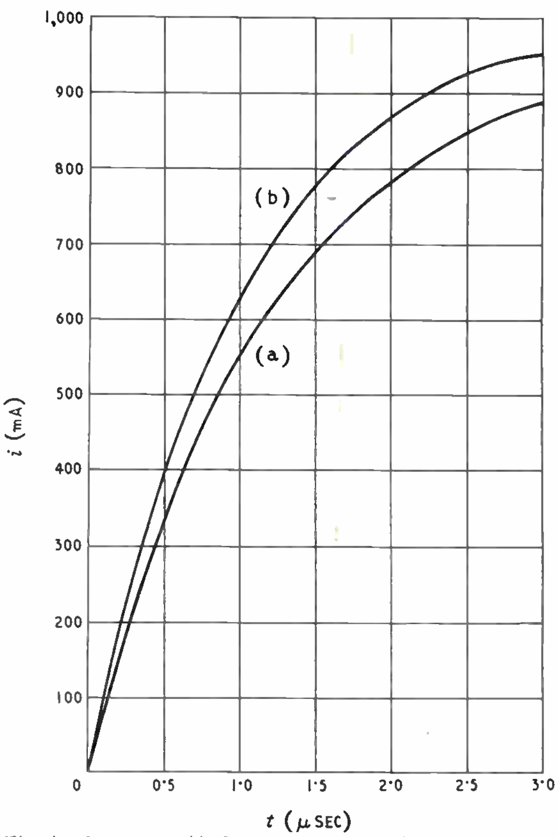


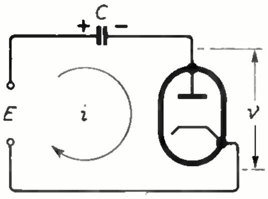

Fig. 4. *Response of inductance in series with non-linear (a) and linear (b) resistances.*

*Case 3*

The diode is in series with a capacitance $C$ (Fig. 5). In this case we have

$$v + q/C = E \quad .. \quad .. \quad .. \quad .. \quad (8)$$

where $q$ is the charge on the capacitor.




Eliminating $v$ by means of (1b) we obtain the differential equation for $q$

Fig. 5. *Capacitance and non-linear resistance.*

$$i = dq/dt = k(E - q/C)^{3/2} = kC^{-3/2}(CE - q)^{3/2}.$$

Separating variables we have

$$dq(CE - q)^{-3/2} = kC^{-3/2}\, dt,$$

and hence

$$(CE - q)^{-1/2} = \tfrac{1}{2}kC^{-3/2}\, t + \text{const.}$$

Since, when $t = 0$, $q = 0$, the constant of integration is $(CE)^{-1/2}$, so that, putting

$$2C^{3/2}/k = \alpha \quad .. \quad .. \quad .. \quad (9)$$

we have $(CE - q)^{-1/2} = (1/\alpha)(t + \alpha/\sqrt{CE})$ and, solving for $q$,

$$q = CE - \alpha^2/(t + \alpha/\sqrt{CE})^2 \quad .. \quad (10)$$

By differentiation we obtain the current

$$i = 2\alpha^2/(t + \alpha/\sqrt{CE})^3 \quad .. \quad .. \quad (11)$$

When $t = 0$, (10) gives $q = 0$, as required by the initial condition. But the initial value of the current is, by (11) and (9), $kE^{3/2}$, in agreement with (1a), since, when $q = 0$, we have, by (8), $v = E$.

As $t \to \infty$, we have $q \to CE$ and $i \to 0$. The asymptotic limits of $q$ and $i$ are the same as those in the case of a linear resistance in series with a capacitance, but the manner in which the limits are approached is different: $q = CE(1 - e^{-t/CR}) \to CE$ as $t \to \infty$, and $i = (E/R)\, e^{-t/CR} \to 0$ as $t \to \infty$ .. .. .. .. .. (11′)

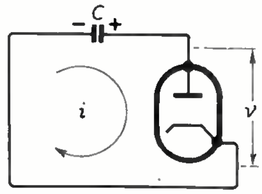



Fig. 7. *Discharge of capacitance through a non-linear resistance.*

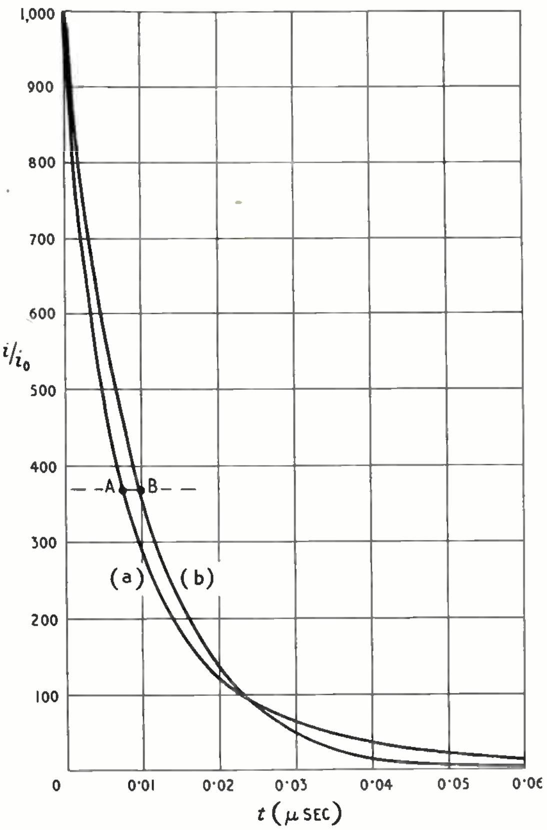



Fig. 6. *Charging current of capacitance through non-linear (a) and linear (b) resistance.*

Curve (a) in Fig. 6 shows the current according to equation (11), while curve (b) shows the current in the comparable linear case (11′). The abscissae of the points A and B are the times for the two cases when the current has been reduced to $1/e$ times the (common) initial value. It appears that the current in our non-linear case decreases more rapidly than in the linear case, at least up to a certain time (0·025 $\mu$sec in the numerical example represented in Fig. 6), after which the exponential current (11′) of the linear case is the smaller one of the two.

*Case 3a*

The difference in the manner in which the limiting values of charge and current are approached, with a linear and with our non-linear resistance in series with a capacitance, is shown more clearly when the diode is connected to a capacitor $C$ bearing an initial charge $q_0$. With the polarity shown in Fig. 7 we must put

$$i = -\,dq/dt \quad .. \quad .. \quad .. \quad .. \quad (12)$$

which gives, by (1a) and with $v = q/C$,

$$i = -\,dq/dt = k(q/C)^{3/2}.$$

Separating variables we have $-\,dq/q^{3/2} = k\,dt/C^{3/2}$, which, integrated, gives $1/q^{1/2} = \text{const} + kt/2C^{3/2}$. By the initial condition the constant of integration is $1/q_0{}^{1/2}$, and so

$$1/\sqrt{q} = kt/2C^{3/2} + 1/\sqrt{q_0} = (1/\alpha)(t + \alpha/\sqrt{q_0}),$$

where $\alpha$ is given by (9). Then we have finally the charge on the capacitor

$$q = \alpha^2/(t + \alpha/\sqrt{q_0})^2 \quad .. \quad .. \quad (13)$$

By equation (12) we obtain the current

$$i = 2\alpha^2/(t + \alpha/\sqrt{q_0})^3 \quad \ldots \quad \ldots \quad (14)$$

We see from (13) and (14) that both $q$ and $i$ tend to zero as $t \to \infty$, as charge and current also do when a linear resistance $R$ takes the place of the diode. But whereas in the linear case both $q$ and $i$ tend to zero according to the same exponential function $e^{-t/CR}$, in our non-linear

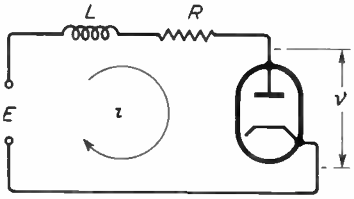



Fig. 8. *Inductance with both linear and non-linear resistance.*

case the charge and the current tend to zero as the square and the cube, respectively, of $1/(t + \alpha/\sqrt{q_0})$. In contrast to the linear case, charge and current are not in a constant ratio, which reflects the non-linearity of the diode characteristic; i.e., the fact that as $q$ tends to zero (and with it the voltage $q/C$ across the diode) the resistance of the diode increases.

*Case 4*

The e.m.f. $E$ is switched on, at $t = 0$, in series with an inductive resistance $L$, $R$ and the diode (Fig. 8). Kirchhoff's Law and (1b) give a differential equation for $i$, which can be separated and gives

$$t = L \int di/\{E - Ri - (i/k)^{2/3}\} \quad \ldots \quad (15)$$

Putting $i = x^3$, so that $di = 3x^2dx$, the integral in (15) takes the form $\int x^2\, dx/(a - bx^2 - x^3)$. By the choice of particular numerical values for $R$, $L$, $k$, $E$, it is possible to factorize the denominator and so express $t$ as a sum of logarithms, one of which tends to infinity as $x$ approaches a certain number. Thus $t \to \infty$ as $x$ tends to a certain (numerical) limit.

Moreover, putting the chosen numerical values of the constants into equation (3), the numerical limit of the current of Case 4 satisfies (3). (For a constant current $L$ has no effect and Cases 1 and 4 are identical.)

The equation of this circuit is

$L\, di/dt + Ri + v = 0$ which is, by (1b),

$di/dt = -Ri/L - i^{2/3}/k^{2/3}L =$

$-(R/L)(i + i^{2/3}/Rk^{2/3})$

Separating variables and putting

$$1/Rk^{2/3} = a$$
$$i = x^3 \quad \ldots \quad \ldots \quad (16)$$

we get $dx/(x + a) = -Rdt/3L$ and hence the integral $\log(x + a) = \text{const} - Rt/3L$. Since when $t = 0$, $i = i_0$ (and hence $x = x_0$) the constant of integration is $\log(x_0 + a)$ and so

$$(x + a) = (x_0 + a)e^{-Rt/3L} \quad \ldots \quad \ldots \quad (17)$$

As in Case 2a, we see in this case, too, that the current $i$ becomes zero ($x = 0$) in the finite time

$$t = t_1 = \frac{3L}{R} \log \frac{x_0 + a}{a} \quad \ldots \quad \ldots \quad (18)$$

where $x_0{}^3 = i_0$ and $a$ is as defined in (16).

Fig. 10 illustrates (17) and (18).

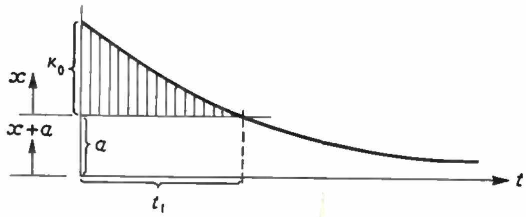



Fig. 10. *Graph of current decay for the circuit of Fig. 9.*

*Case 5*

The diode is connected in series with an inductance $L$, a resistance $R$, and a capacitance $C$. At $t = 0$ the constant e.m.f. $E$ is switched on, the initial charge and current being zero (Fig. 11).

The equation of the circuit is

$$L\, di/dt + Ri + q/C + v = E \quad \ldots \quad (19)$$

Substituting for $v$ by (1b), putting $R/L = 2\mu$; $1/Lk^{2/3} = K$; and $1/CL = \alpha^2$ and moving the constant term to the left-hand side, (19) becomes

$$d^2q/dt^2 + 2\mu\, dq/dt + K(dq/dt)^{2/3} + \alpha^2(q - CE) = 0 \quad \ldots \quad \ldots \quad \ldots \quad (20)$$

We now introduce new variables $x$, $\theta$ by putting

$$(q - CE) = (1/\alpha)(K/2\mu)^3x \quad \ldots \quad \ldots \quad \ldots \quad (21)$$

and $t = \theta/\alpha$ so that $d/dt = \alpha d/d\theta$. Denoting

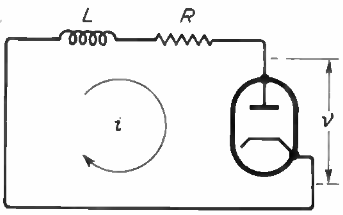
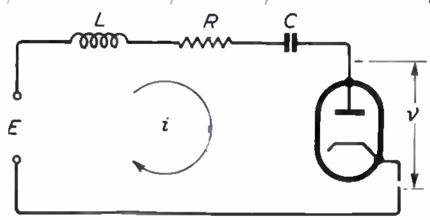



Fig. 9 (left). *Decay of current in inductive circuit.*

Fig. 11 (right). *Damped resonant circuit including a non-linear resistance.*




*Case 4a*

More interesting conditions exist when an initial current $i_0$ flows in a circuit consisting of the diode in series with an inductive resistance $L$, $R$ (Fig. 9).

differentiation with respect to $\theta$ by primes we obtain (20) in the form

$$x'' + \epsilon(x' + x'^{2/3}) + x = 0 \quad \ldots \quad \ldots \quad (22)$$

where $\epsilon = 2\mu/\alpha$.

Equation (22) is of the type $x'' + \phi(x') + x = 0$, of which van der Pol's equation is the special case $\phi(x') = - x' + x'^3$. We are here considering the special case $\phi(x') = \epsilon(x' + x'^{2/3})$.

From (21) it is clear that, whereas $x'$ is proportional to the current $i$, $x$ itself is proportional, not to the charge $q$, but to $(q - CE)$, so that the initial conditions, $i = 0$ and $q = 0$ when $t = 0$, are now $x' = 0$ and

$$x = - CE\alpha(2\mu/K)^3 \text{ when } \theta = 0 \quad .. \quad (23)$$

If the coefficient $\epsilon$ is sufficiently small, the equation is that of a damped oscillation with the damping term $\phi(x') = \epsilon(x' + x'^{2/3})$.

If $\phi(x')$ were linear in $x'$, the solution would be the familiar one of an exponentially decreasing sine or cosine.

But owing to the term $x'^{2/3}$ in $\phi(x')$, equation (22) is non-linear, and its solution in general terms becomes impracticable. We seek, therefore, a numerical solution, assuming particular values of $L$, $R$, $C$, and the constant $k$ of equation (1a).

If we take, as special values, $R = 10$ ohms, $L = 10^{-4}$ henry, $C = 10^{-10}$ farad, and $k = 10^{-3}$, we have: $1/CL = \alpha^2 = 10^{14}$ (hence $\alpha = 10^7$); $R/L = 2\mu = 10^5$; $1/Lk^{2/3} = K = 10^6$; and $2\mu/\alpha = \epsilon = 10^{-2}$. Thus the equation (22) becomes

$$x'' + 10^{-2}(x' + x'^{2/3}) + x = 0 \quad .. \quad .. \quad (24)$$

In any one of the standard methods of numerical integration it is necessary first to choose an interval $h$ of the independent variable (i.e., in our case, of $\theta$), then to calculate, with the help of Taylor's series, a few values of $x$ and $x'$, beginning with the initial values, and from the differential equation the corresponding values of $x''$. Thereafter the formulae given in the particular method chosen will yield further values of $x$ and $x'$, and again the differential equation gives the corresponding value of $x''$, and so on.

With our equation (24), however, a difficulty arises in the calculation of the first few values, owing to the fact that the non-linear power of $x'$ is fractional. For in Taylor's series for $x$ and $x'$:

$$x(\theta + h) = x(\theta) + hx'(\theta) \tfrac{1}{2}h^2 x''(\theta) + \ldots$$
$$x'(\theta + h) = x'(\theta) + hx''(\theta) + \tfrac{1}{2}h^2 x'''(\theta) + \ldots$$

we have, by differentiating (22): $x''' = - x' - x''(1 + \frac{2}{3}x'^{-1/3})$ which, initially, is not finite, since there occurs a negative power of $x'$, and $x' = 0$ when $\theta = 0$.

This difficulty can be overcome if, instead of beginning the computation at $\theta = 0$, when $x' = 0$, we start at the maximum value of $x'$ (i.e., at the current maximum) when $\theta = \theta^*$, and carry out the integration first 'inwards' (i.e., taking $h$ negative) and then 'outwards' (i.e., with $h$ positive).

In either direction the integration need only be continued until a zero of $x'$ is reached, since the current through the diode cannot reverse its sign.

The value of $x$ corresponding to the zero of $x'$ reached by integrating 'inwards' is then the initial value $x_0$ of $x$.

But as the choice of the starting point of the computation (maximum current) is arbitrary, the freedom to choose the applied e.m.f. $E$ [which determines $x_0$ by (21)] is lost. If, in any concrete practical problem, $E$ is given, the only course open is to carry out the 'inward' computation several times, beginning with different maximum values of $x'$ until we arrive at values $x_0$ corresponding to values of $E$ which are near enough to the given e.m.f.

The current, and so $x'$, is a maximum when $x'' = 0$. Then, by (24), $x = - 10^{-2}(x' + x'^{2/3})$. So, for a maximum value of $300 \times 10^{-6}$ for $x'$ which is equivalent to a current of 300 mA, the corresponding value of $x$ is found to be $-47{\cdot}81 \times 10^{-6}$. These 'starting' values for $x$ and $x'$ lead to the 'initial' value of $x$ (when $x' = 0$) of between $- 340{\cdot}45 \times 10^{-6}$ and $- 339{\cdot}01 \times 10^{-6}$, or about $-340 \times 10^{-6}$ which, by (21), implies an e.m.f. $E = 340$ volts.

Instead of using one of the more elaborate formulae for computation (e.g., Milne-Simpson's, Karman's, etc.), which yield sufficiently accurate results with comparatively larger, and therefore fewer, intervals $h$, but require four values of $x$ and $x'$ to be calculated before the step-by-step application of the formulae, the simple formulae

$$x_n = x_{n-2} + 2hx'_{n-1} \text{ and } x'_n = x'_{n-2} + 2hx''_{n-1} \quad .. \quad .. \quad .. \quad (25)$$

have been used, which require only two values of $x$ and $x'$ to be calculated beforehand, and which

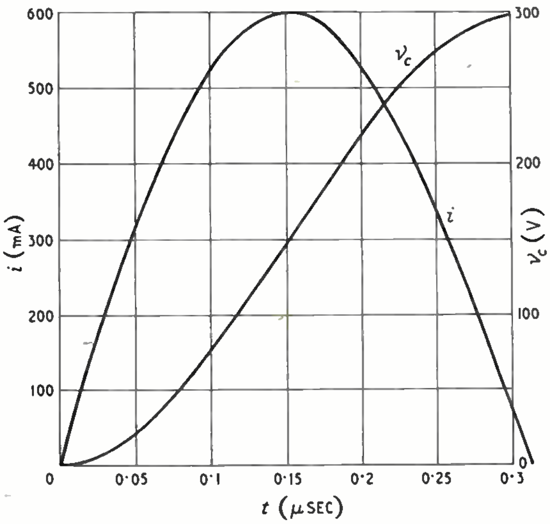



Fig. 12. *Current and voltage for the circuit of Fig. 11.*

are quite adequate provided the interval $h$ is chosen sufficiently small.

To choose $h$, let us omit, for a moment, the term $\phi(x')$, which reduces the differential equation (24) to the type $x'' + \omega^2 x = 0$, with $\omega = 2\pi/\tau = 1$, so that the period (in units of the $\theta$-time scale) is $\tau = 2\pi \approx 6$. Taking $h = 10^{-1}$, we have about 60 intervals in one period, or, since we carry out the computation over one half-period only (between two zeros of the current), there are about 30 points to be calculated.

Taking $h = 5 \times 10^{-2}$ doubles the number of points to be calculated, but makes the practical computation very convenient (since then $2h = 10^{-1}$) and increases the accuracy.

The graph (Fig. 12) shows the current $i$, obtained from $x'$ by (21), in mA, and $q/C = v_c$ the voltage across the capacitor, $q$ having been obtained from $x$ by (21). The time-scale has also been changed from $\theta$ to $t$, so that $h = 0{\cdot}01$ $\mu$sec.

If we compare this result with that of the linear case leading to the same maximum capacitor voltage, we find that the current maximum is a little greater and occurs slightly earlier. The capacitor voltage, too, is a little higher in the linear case, except at the ends of the time-range, where the voltages in the two cases coincide.

It may be added that, if our computation is carried on beyond the zero of the current, the curve obtained is similar to that of an oscillation with linear damping. It would represent the non-linear analogue to a damped oscillation, which could be realized by replacing our one diode by an anti-parallel combination of two.

# 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.*

## X-Ray Production by Magnetron

SIR,—While investigating the power-handling capacity of waveguide rotating joints (W.G.16) it was discovered that results varied considerably depending on whether the rotating-joint assembly was connected to the magnetron (type 4J50) by means of a straight or curved length of waveguide. When a straight piece of waveguide was used, breakdown occurred at a level of 150 kW peak whereas with a curved length of connecting waveguide breakdown was in excess of 240 kW peak. The curve in the waveguide need only be sufficient to impede a direct line of sight between the magnetron window and the waveguide assembly under test. There was no difference in match for these two conditions.

By interposing an H-plane bend containing a long tube soldered into the narrow face of the bend it was possible to observe the magnetron window without upsetting the full power X-band conditions (the tube is being operated below cut off). A radiation monitor applied to the end of the tube showed the presence of X-rays. These are 'copper' X-rays having an energy potential of about 8·5 kV. The exposure of X-ray film at the end of the inspection tube confirmed the existence of X-rays.

It is concluded therefore that any equipment using a 4J50 type of magnetron and a waveguide system in W.G.16 waveguide should include a bend in the waveguide (effectively an X-ray filter) as close to the magnetron as possible.

The X-rays are also emitted through the cathode stem of the magnetron in a cone-shaped beam whose axis is that of the cathode stem. Roughly, the limits of the cone-shaped beam are the same as those which define the visibility of the magnetron heater between the cathode shield and the sole plate. The intensity of the X-rays from this source is less than obtained in the waveguide output and is very difficult to measure. Difficulty arises because of the high voltages applied to the magnetron cathode and because the radiation intensity is too high to be measured conveniently by a scintillation counter but too low to be measured by the type of ionization chamber readily available. The X-rays are so 'soft', however, that they will not penetrate to the bone marrow and any harmful effects will be confined to unprotected (e.g., by normal clothing) skin. Any such effects are cumulative (over a period, say, of a year) and it is recommended that steps are taken to prevent the exposure of operating personnel to this radiation.

It should be noted also that, in a typical equipment, the cathode-stem radiation may illuminate other components, such as pulse transformers and delay networks, thus facilitating corona or other breakdown in these components.

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Radar Department, Ferranti Ltd.,
Edinburgh.
28th November 1955.

## Transient Response Calculation

SIR,—I was interested to see Mr. D. G. Sarma's article "Transient Response Calculation" in *Wireless Engineer* (July 1955) which extended the method of linear segments for estimating the transient response associated with a given gain-frequency characteristic of a minimum-phase network. A comparison between this method and a method I suggested[1] using standard gain-frequency curves[2] might be of interest.

The main disadvantage of Mr. Sarma's method is that the number of poles and zeros in the approximating function is much greater than the number in the original function. Thus, in the given example a total of 13 poles and zeros are used to approximate a function having only 5. This means that a considerable amount of labour is necessary both in calculating the residues at the poles and in obtaining numerical values from a complicated time function. On the other hand, by using the standard gain-frequency curves a very reasonable approximation can often be obtained using a smaller number of poles and zeros than are present in the original function.

A method of approximation using straight-line segments or semi-infinite slopes, in which the straight lines can be drawn arbitrarily, will doubtless more quickly provide some sort of approximation to a given gain characteristic than a method using more complicated approximating functions. However, the full saving in time on this particular point will not be realized in Mr. Sarma's method since the straight-line segments cannot be put in arbitrarily. Further, some numerical calculation is necessary to obtain the poles and zeros corresponding to the chosen linear segments.

The method using the standard gain–frequency curves has the advantage of being easily extended to the non-minimum-phase case, the procedure being as follows:—

(a) Use the standard gain–frequency curves, $10 \log_{10} [x^2 + 2x \cos 2\delta + 1]$, to obtain an approximation to the given gain–frequency characteristic.

(b) Use the standard phase–frequency curves[1], arc $\tan[2x \cos \delta/(1 - x^2)]$, to obtain the corresponding minimum-phase characteristic, all poles and zeros being taken to lie in left-half plane. As the given system function is non-minimum-phase the phase–frequency characteristic so obtained will not be identical with the given phase–frequency characteristic.

(c) By transferring appropriate singularities (zeros in the case of a gain-type system function) to the right-half plane or by adding the phase characteristics corresponding to all-pass networks (*twice* the standard phase–frequency curves) pole-zero positions are finally obtained that correspond to good approximations to both the given gain–frequency and phase–frequency characteristics A good approximation is hence obtained for the response to a given transient. The total number of zeros to be transferred, or the total number of all-pass networks to add, is found by noting the difference between the value of the phase at frequency $+\infty$ minus the phase at frequency $+0$ for the given phase characteristic and the minimum-phase characteristic.

O. P. D. Cutteridge

Electrical Engineering Dept.,
Faculty of Technology,
University of Manchester.
19th November 1955.

## REFERENCES

[1] O. P. D. Cutteridge, "A Graphical Contribution to the Analysis and Synthesis of Electrical Networks", *Proc. Instn elect. Engrs*, Oct. 1953, Vol. 100, Part IV, pp. 83–90.

[2] J. M. Linke, "A Graphical Approach to the Synthesis of General Insertion Attenuation Functions", *Proc. Instn elect. Engrs*, 1950, Vol. 97, Part III, pp. 179–187.

## MEETINGS

### I.E.E.

9th January. "The Efficient Use of Technical Personnel", discussion to be opened by the President, J. Eccles, C.B.E., B.Sc.

11th January. "Pulse-Time Modulation Terminals for Music Transmission over Radio Links", by R. F. Rous, B.Sc.

16th January. "Advanced Courses for Engineers in Industry", by Willis Jackson, D.Sc., D.Phil., F.R.S., H. D. Morgan, M.Sc.(Eng.) and Professor G. F. Mucklow, D.Sc. Joint meeting with the Institutions of Civil and Mechanical Engineers at the Institution of Mechanical Engineers, 1 Birdcage Walk, Westminster, London, S.W.1.

23rd January. "Particle Accelerators", by E. L. Wiblin, M.A.

These meetings will commence at 5.30 and will be held at the Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2, except where otherwise stated.

### Brit.I.R.E.

5th January. "Domestic Tape Recording Applications with Special Reference to Stereophonic Reproduction", by M. B. Martin and D. L. A. Smith, B.Sc.(Eng.).

25th January. Symposium on Electronic Methods of Pictorial Reproduction:—"Facsimile Communication", by H. F. Woodman and P. H. J. Taylor; "Facsimile Transmission of Weather Charts", by J. A. B. Davidson; "Tone Reproduction with Electronic Stencils", by R. Lant, Ph.D. and "Electronic Engraving", by G. S. Allen.

These meetings will commence at 6.30 and will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.

### The Television Society

19th January. "Non-Entertainment Aspects of Television", The Fleming Memorial Lecture by Professor J. D. McGee, at 7 o'clock at the Royal Institution of Great Britain, 21 Albemarle Street, London, W.1. Admission by ticket only.

### Radar Association

13th January. "Radio Astronomy", by Professor A. C. B. Lovell, at 7.30 at the Anatomy Theatre, University College, Gower Street, London, W.C.1.

## STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

*Values for November 1955*

| Date 1955 November | Frequency deviation from nominal: parts in $10^8$ | |
|---|---|---|
| | MSF 60 kc/s 1429-1530 G.M.T. | Droitwich 200 kc/s 1030 G.M.T. |
| 1 | N.T. | +5 |
| 2 | N.T. | +1 |
| 3 | N.T. | +1 |
| 4 | N.T. | +1 |
| 5 | N.T. | +1 |
| 6 | N.T. | +1 |
| 7 | +0·1 | +1 |
| 8 | +0·1 | +1 |
| 9 | +0·1 | +1 |
| 10 | +0·1 | +1 |
| 11 | +0·1 | +1 |
| 12 | +0·1 | +1 |
| 13 | +0·1 | +1 |
| 14 | +0·1 | +1 |
| 15 | 0·0 | +1 |
| 16 | +0·1 | +2 |
| 17 | +0·1 | +1 |
| 18 | +0·1 | +1 |
| 19 | +0·2 | +1 |
| 20 | +0·2 | +1 |
| 21 | +0·1 | +1 |
| 22 | +0·2 | +1 |
| 23 | +0·1 | +1 |
| 24 | +0·1 | +2 |
| 25 | +0·1 | +2 |
| 26 | +0·1 | +2 |
| 27 | +0·1 | +2 |
| 28 | +0·1 | +2 |
| 29 | +0·1 | +2 |
| 30 | +0·1 | +3 |

The values are based on astronomical data available on 1st December 1955.

N.T. = No Transmission.

The MSF 60-kc/s aerial system was undergoing maintenance during the first six days of the month.

# ABSTRACTS and REFERENCES

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

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



## ACOUSTICS AND AUDIO FREQUENCIES

534.13 : 534.845 1

**Wedge-Shaped Cavity Resonators.**—E. Kohlsdorf. (*Hochfrequenztech. u. Elektroakust.*, July 1955, Vol. 64, No. 1, pp. 33–34.) A calculation is made of the characteristics of acoustic resonators formed in a hall by the steps and the sloping floor supporting them. The height of the resonator considered is 0·4 m, the depth 0·8 m and the width 1 to 2 m; the front wall is slotted. Maximum absorption occurs near 158 and 365 c/s; this is shown graphically.

534.213.4 2

**On Webster's Horn Equation.**—E. S. Weibel. (*J. acoust. Soc. Amer.*, July 1955, Vol. 27, No. 4, pp. 726–727.) "A wave equation for the sound propagation through tubes is derived by means of Hamilton's variational principle. It is assumed that the wave fronts can be approximated by surfaces of constant stream potential; this is the only assumption made. The variational principle ensures that the best equation that is compatible with this assumption will be obtained. The equation has the form of Webster's horn equation, however its coefficients are defined differently."

534.232 3

**On the Radiation Impedance of an Array of Finite Cylinders.**—D. H. Robey. (*J. acoust. Soc. Amer.*, July 1955, Vol. 27, No. 4, pp. 706–710.) Analysis is presented for an array of sound sources in the form of collinear cylinders.

534.232 : 546.431.824-31 4

**On the Radiation Impedance of the Liquid-Filled Squirting Cylinder.**—D. H. Robey. (*J. acoust. Soc. Amer.*, July 1955, Vol. 27, No. 4, pp. 711–714.) Analysis is presented for the underwater transducer discussed previously (1847 of 1955).

534.24 5

**Nonspecular Reflection of Sound from a Sinusoidal Surface.**—H. S. Heaps. (*J. acoust. Soc. Amer.*, July 1955, Vol. 27, No. 4, pp. 698–705.) The reflected field is obtained as the sum of an infinite number of terms whose coefficients may be found from a recurrence relation. The three central terms in the series correspond to the first-order approximation. The results are used to determine the average intensity and fluctuations of the sound reflected from a travelling sinusoidal boundary. The degree of roughness required to destroy the effect of Lloyd's mirror fringes is discussed in terms of the size of the surface corrugations relative to λ.

534.614-8 6

**Variation of Acoustic Velocity with Temperature in Some Low-Velocity Liquids and Solutions.**—G. W. Marks. (*J. acoust. Soc. Amer.*, July 1955, Vol. 27, No. 4, pp. 680–688.) Measurements on various organic liquids are reported, using an ultrasonic interferometer and a frequency of 500 kc/s.

534.75 7

**Some Parameters influencing the Pitch of Amplitude-Modulated Signals.**—A. M. Small, Jr. (*J. acoust. Soc. Amer.*, July 1955, Vol. 27, No. 4, pp. 751–760.)

534.78 8

**Effect of Duration on the Perception of Voicing.**—P. Denes. (*J. acoust. Soc. Amer.*, July 1955, Vol. 27, No. 4, pp. 761–764.) An investigation of factors affecting the recognition of phonemes is described based on the observed differences in the relative durations of the vowels and consonants according as the latter are voiced or unvoiced.

534.78 9

**Acoustic Loci and Transitional Cues for Consonants.**—R. C. Delattre, A. M. Liberman & F. S. Cooper. (*J. acoust. Soc. Amer.*, July 1955, Vol. 27, No. 4, pp. 769–773.) Experiments on the synthesis of speech are described.

534.844.5 10

**Optimum Reverberation Time of Studios.**—W. Reichardt, E. Kohlsdorf & H. Mutscher. (*Hochfrequenztech. u. Elektroakust.*, July 1955, Vol. 64, No. 1, pp. 18–25.) The reported tests, involving 85 listeners, lead to results similar to those obtained by Kuhl (620 of 1955). While the mean reverberation characteristic adjudged optimum agreed with the accepted characteristic, the preferred reverberation times varied with the type of music.

534.846 11

**Comparison of Objective and Subjective Observations on Music Rooms.**—J. Blankenship, R. B. Fitzgerald & R. N. Lane. (*J. acoust. Soc. Amer.*, July 1955, Vol. 27, No. 4, pp. 774–780.) Experiments made at the University of Texas are reported; an attempt is made to relate the terminology and viewpoints of architects, physicists and musicians.

621.395.623.7 12

**The Twin-Cone Moving-Coil Loudspeaker.**—J. J. Schurink. (*Philips tech. Rev.*, March 1955, Vol. 16, No. 9, pp. 241–249.) A description is given of a loudspeaker with a small cone arranged inside and attached to a larger cone; satisfactory response is obtained at frequencies up to about 18 kc/s. Advantages of this arrangement over the use of separate loudspeakers for the upper and lower a.f. ranges are indicated.

621.395.625 : 621.317.76 13

**Measurement of "Wow" and "Flutter".**—O. E. Dzierzynski. (*Wireless World*, Nov. 1955, Vol. 61, No. 11, pp. 547–552.) The subject is discussed generally and a detailed description of a flutter meter is given.

## AERIALS AND TRANSMISSION LINES

621.315.2 : 621.396.97 14

**Transmission Properties of Low-Frequency Programme Lines used in German Carrier-Frequency Trunk Cables.**—E. A. Pavel. (*Fernmeldetech. Z.*, Aug. 1955, Vol. 8, No. 8, pp. 455–461.) Report of tests on a 2 212-km a.f. line superimposed on carrier-frequency cable. Lines of this type more than satisfy C.C.I.R. transmission-quality requirements for distances up to at least 4 000 km. See also 1578 of 1953 (Pavel & v. Schau).

621.315.212 (083.74) 15

**Present Position on Standardization of High-Frequency Coaxial Transmission Cables.**—H. Peters. (*Fernmeldetech. Z.*, Aug. 1955, Vol. 8, No. 8, pp. 443–444.) Reasons are given for the retention in Germany of 60 Ω as the standard values of characteristic impedance for h.f. coaxial cables, in spite of international standardization on 50 Ω and 75 Ω. See also 2299 of 1954 (Gutzmann).

621.372.2 16

**Propagation of Microwaves on a Single Wire: Part 1.**—S. K. Chatterjee & P. Madhavan. (*J. Indian Inst. Sci.*, Section B, July 1955, Vol. 37, No. 3, pp. 200–223.) "Some quantitative measurements on cylindrical surface waves excited on a cylindrical bare copper conductor have been made at 3·2-cm wavelength. The results on radial field measurements indicate fair agreement with results predicted by theory. The decay coefficient for the radial field is 14·0 m$^{-1}$. It is suggested that the surface conductivity of metals in the form of wire can be found at microwave frequencies by using this method. Field distributions near the launching horn have been found to be considerably distorted due probably to the presence of complementary waves. The design and constructional details of the launching system, probe, etc. used in the experiment are presented."

621.372.8 + 621.372.413 17

**A Step-by-Step Method for designing Waveguides and Oscillatory Systems.**—M. S. Neyman. (*Radiotekhnika, Moscow*, Jan. 1955, Vol. 10, No. 1, pp. 12–22.) A method is proposed which permits calculation of the parameters of distributed electromagnetic systems such as waveguides and cavity resonators without recourse to electrodynamic wave equations. Rectangular and cylindrical waveguides and resonators are considered as examples.

621.372.8 18

**Calculation of the Angular Momentum of the Electromagnetic Field in a Waveguide.**—L. Ronchi & G. Toraldo di Francia. (*Alta Frequenza*, June 1955, Vol. 24, No. 3, pp. 204–218.) General formulae for the flux of the angular momentum of the e.m. field are applied to the case of a circular waveguide in which only the $TE_{11}$ mode is propagated; an investigation is made of the types of polarization possible. For the case of a guide containing a polarization converter calculations are made of the angular momentum delivered to the converter per unit of time when the polarization is converted from elliptical to linear; two different methods are used.

621.372.8 19

**Theory of Circularly Symmetric Standing TM Waves in Terminated Iris-Loaded Guides.**—C. C. Grosjean. (*Nuovo Cim.*, 1st July 1955, Vol. 2, No. 1, pp. 11–26. In English.) An exact treatment is developed based on matching the field components at the diaphragm apertures, and on similar lines to that described previously (2832 of 1955) in relation to infinite guides.

621.372.8 : 538.22 20

**Electromagnetic Waves in a Magnetized Ferrite in the Presence of Conducting Planes.**—A. L. Mikaelyan & A. A. Pistol'kors. (*Radiotekhnika, Moscow*, March 1955, Vol. 10, No. 3, pp. 14–24.) Theory is presented for the propagation of electromagnetic waves in an infinite space between ideally conducting planes, the space being totally or partially filled with ferrite and a constant magnetic field oriented in various directions being applied.

621.372.8 : 621.315.6 21

**Propagation Coefficients of Waves on Dielectric Tubes.**—H. G. Unger. (*Fernmeldetech. Z.*, Aug. 1955, Vol. 8, No. 8, pp. 438–443.) See 3458 of 1954.

621.372.8 : 621.396.11.029.4/.51 : 551.510.535 22

**The Radial Propagation of Electromagnetic Waves between Two [parallel] Conducting Planes (Radial Waveguide).**—H. Kaden. (*Frequenz*, July 1955, Vol. 9, No. 7, pp. 241–249.) In the radial waveguide E and H modes propagate as in a linear guide, the orientation of the modes being with respect to the normal to the conducting planes. The theory is used to calculate the 'internal' and 'external' dynamic capacitance of a parallel-plate capacitor. The radiation from electric and magnetic dipoles situated between the planes is also calculated; the result represents to a close approximation the conditions for propagation of long waves ($\lambda > 3$ km) in the guide formed between the earth and the ionosphere. The number of modes propagated varies inversely as $\lambda$ and the field-strength/distance curve at the earth's surface is of the nature of a statistical function, since many incommensurable waves interfere mutually. Resonance effects occur when the height of the ionosphere is an integral multiple of $\lambda/2$, both the distant field strength and the retroaction of the radiation on the dipole becoming very great.

621.396.677 : 523.16 23

**Chord Construction for Correcting Aerial Smoothing.**—Bracewell. (See 98.)

## AUTOMATIC COMPUTERS

681.142 24

**The Design and Construction of a High-Speed Electronic Differential Analyzer.**—N. N. Biswas, V. N. Chiplunkar & V. C. Rideout. (*J. Indian Inst. Sci.*, Section B, July 1955, Vol. 37, No. 3, pp. 186–199.)

681.142 25

**Improvement of an Iteration Process suitable for Automatic Division.**—C. Böhm. (*Ricerca sci.*, July 1955, Vol. 25, No. 7, pp. 2077–2080.) Electronic computers having no organ of division are provided with a subroutine for the computation of reciprocals. An improvement of this subroutine is proposed which permits a reduction of about 30% in the number of operations without affecting the accuracy.

681.142 : 413 26

**Use of a Computing Machine as a Mechanical Dictionary.**—A. D. Booth. (*Nature, Lond.*, 17th Sept. 1955, Vol. 176, No. 4481, p. 565.) The time required to locate a stored word is reduced by applying binary-division principles in the searching process.

681.142 : 621.3.078/.079 27

**Industrial Uses of Special-Purpose Computers.**—A. H. Kuhnel. (*Instrum. & Automation*, July 1955, Vol. 28, No. 7, Part 1, pp. 1108–1113.) An account is presented of the analysis of a simple machine-control problem and the design of a suitable special-purpose computer to perform the task. Economic factors are also briefly considered.

681.142 : 621.318.5 28

**A Method for Synthesizing Sequential Circuits.**—G. H. Mealy. (*Bell Syst. tech. J.*, Sept. 1955, Vol. 34, No. 5, pp. 1045–1079.) A theory is developed from design procedures suggested by Huffman (*J. Franklin Inst.*, March & April 1954, Vol. 257, pp. 161–190 & 275–303) and Moore (to be published in Automata Studies, Princeton Univ. Press) enabling intricate relay systems to be built up from an initial diagrammatic statement of the essential requirements. By successive application of simplifying reductions a unique circuit is finally obtained.

681.142 : 621.372.45 29

**A Nonlinear Resistance-Capacitance Circuit.**—F. A. Key & W. G. P. Lamb. (*Electronic Engng*, Oct. 1955, Vol. 27, No. 332, pp. 446–448.) An amplifier in a negative-feedback loop is used to provide the nonlinear resistance in an analogue equipment representing air-pressure variations.

681.142 : 621.383 30

**A Photoelectric Analogue Computer for investigating the Dynamic Behaviour of Linear Systems.**—S. Kitsopoulos. (*Bull. schweiz. elektrotech. Ver.*, 23rd July 1955, Vol. 46, No. 15, pp. 690–693.) Apparatus is described for determining the output function of a linear system such as an amplifier or servomechanism when the input and transfer functions are known. The two known functions are represented by appropriately shaped apertures in diaphragms interposed between light source and photocell; the mechanical system used to produce the required continuous variation of the input function is described in detail.

681.142 : 621.395.625.3 31

**High-Density Tape Recording for Digital Computers.**—(*Tech. News Bull. nat. Bur. Stand.*, Sept. 1955, Vol. 39, No. 9, pp. 121–124.)

## CIRCUITS AND CIRCUIT ELEMENTS

621.3 (083.7) 32

**Standardization of Electronic Components.**—H. W. Ghijsen. (*Tijdschr. ned. Radiogenoot.*, July 1955, Vol. 20, No. 4, pp. 227–242. In English.) A survey with particular reference to the activities of the International Electrotechnical Commission.

621.316.8 + 621.318.42 + 621.319.4].011.2 33

**The Impedance of Chokes, Resistors and Capacitors at High Frequency.**—A. Weis. (*Frequenz*, July 1955, Vol. 9, No. 7, pp. 221–227.) Analysis indicates that in order to represent the impedance variation of cylindrical inductors, resistors and capacitors at frequencies above the first resonance point, equivalent circuits are needed combining a ladder network with simple parallel or series networks; nonuniform field distribution along the component can then be taken into account. The validity of the theory is confirmed by reference to measurement results, shown graphically. The effect of an earthing lead on a lead-through capacitor is demonstrated.

621.316.86 : 621.396.822 34

**Contact Noise.**—B. S. Gal'perin. (*Zh. tekh. Fiz.*, March 1955, Vol. 25, No. 3, pp. 410–413.) A composition resistor is considered as consisting of a number of parallel circuits each containing an equal number of particles and the same number of contacts of equal resistance. It is shown that the noise energy of such a homogeneous system is equal to that of a single contact and that for a given gradient of the applied direct voltage it is determined by the degree of dispersion and the specific resistance of the conductor material. Expressions are also derived for heterogeneous systems and imperfect contacts.

621.316.86.049.75 35

**Printed Resistors.**—R. S. Marty, E. M. Davies & P. J. Franklin. (*Elect. Mfg*, Jan. 1955, Vol. 55, No. 1, pp. 56–63.) Preparation of resistors by injection moulding is described. The effect of ink composition and the printing and curing processes on the resistor characteristics are described.

621.318.435 : 621.373.43 36

**Magnetic Pulse Modulators.**—K. J. Busch, A. D. Hasley & C. Neitzert. (*Bell Syst. tech. J.*, Sept. 1955, Vol. 34, No. 5, pp. 943–993.) Mathematical analyses are given of a.c.- and d.c.-charged series-type saturable reactors for generating pulses in radar systems. Automatic core resetting is achieved. The production of pulses of duration $< 0{\cdot}1\ \mu s$ and the prevention of parasitics in the output are discussed and a circuit is described enabling the thyratron cathode in a d.c.-charged modulator to be operated at ground potential.

621.318.57 : 621.374.32 : 621.314.7 37

**A Multistable Transistor Circuit.**—R. A. Henle. (*Elect. Engng, N.Y.*, July 1955, Vol. 74, No. 7, pp. 570–572.)

621.372.4 38

**Proof of a Theorem on the Equivalence of Two-Pole Networks with only Two Types of Impedance Element.**—S. Spiess. (*Hochfrequenztech. u. Elektroakust.*, July 1955, Vol. 64, No. 1, pp. 25–32.) Note on 961 of 1955 (Weber & Schlegel).

621.372.41 39

**Theorem on the Reactive Energy in an Electric Dipole in a State of Variation.**—L. Lunelli. (*Alta Frequenza*, June 1955, Vol. 24, No. 3, pp. 246–267.) Analysis is based on differentiation of the operational impedance. A simple formula is derived giving the instantaneous difference between the stored magnetic and electric energy.

621.372.413 + 621.372.8 40

**A Step-by-Step Method for designing Waveguides and Oscillatory Systems.**—Neyman. (See 17.)

621.372.5 **41**

**Some Fundamental Properties of Networks without Mutual Inductance.**—A. Talbot. (*Proc. Instn elect. Engrs*, Part C, Sept. 1955, Vol. 102, No. 2, pp. 168–175. Digest, *ibid.*, Part B, July 1955, Vol. 102, No. 4, pp. 554–555.) Elementary methods not involving determinants are used to prove theorems on the voltage and current gains obtainable from a resistance network. The results are used to determine properties of general networks without mutual inductance; *RC* networks are investigated by a method simpler than that of Fialkow & Gerst (3369 of 1952). Attention is drawn to a neglected paper by Kirchhoff (*Ann. Phys. Chem.*, 1847, Vol. 72, p. 497) giving the foundations of the topology of networks.

621.372.5 **42**

**Synthesis of Ladder Networks to give Butterworth or Chebyshev Response in the Pass Band.**—E. Green. (*Proc. Instn elect. Engrs*, Part C, Sept. 1955, Vol. 102, No. 2, p. 290.) Addendum to 370 of 1955.

621.372.5 **43**

**Ladder Networks with Similar Quadripole Sections.**—U. Ruelle. (*Alta Frequenza*, June 1955, Vol. 24, No. 3, pp. 268–283.) Formulae for the section input and output currents and voltages and input impedances are derived directly from Kirchhoff's laws, using determinants.

621.372.5 : 512.83 **44**

**On Node and Mesh Determinants.**—S. Okada. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, p. 1527.) An extension of the analysis presented by Seshu (1581 of 1955).

621.372.512.3.015.3 **45**

**Initial Transients in some Special Passive Networks with Amplitude and Frequency Modulation.**—E. Henze. (*Arch. elekt. Übertragung*, July 1955, Vol. 9, No. 7, pp. 326–338.) It is shown that for multistage selective circuits there is an optimum coupling for which the output corresponding to unit-step input is aperiodic and free from overshoot. The performance of two frequency-shift circuits is studied, and the existence of an optimum ratio of shift to keying frequency is demonstrated. A formula is derived for calculating distortion factor for two-circuit band-pass filters with f.m.

621.372.54 **46**

**New Filter Theory of Periodic Structures.**—W. K. R. Lippert. (*Wireless Engr*, Oct. & Nov. 1955, Vol. 32, Nos. 10 & 11, pp. 260–266 & 305–310.) Theory for linear passive quadripoles is developed in terms of characteristic reflection and transmission factors which can be specified by reference to the standing-wave pattern on lines connected to the input and output. The relations between these factors and the usual circuit constants, particularly the image function of symmetrical circuits, are studied. The method is illustrated by examining multisection structures combining quadripoles and loss-free lines. The use of these characteristic factors leads to major simplifications in dealing with loss-free structures, where most of the work can be done graphically.

621.372.542.2 : 621.396.61 **47**

**A Ten-Kilowatt Low-Pass Filter.**—E. R. Broad & E. J. P. May. (*P.O. elect. Engrs' J.*, July 1955, Vol. 48, Part 2, pp. 94–96.) The filter is designed for use in the aerial feed lines of transmitters working in the frequency range 4–28 Mc/s and discriminates by about 60 dB against any frequencies in the television band 40–70 Mc/s which may be present as harmonics. In measuring the effectiveness of the filter account must be taken of radiation from sources other than the aerial.

621.372.542.32 : 621.372.412 **48**

**A Piezoelectric High-Pass Filter.**—Ya. I. Velikin, Z. Ya. Gel'mont & E. V. Zelyakh. (*Radiotekhnika, Moscow*, March 1955, Vol. 10, No. 3, pp. 41–49.) A lattice filter is considered (Fig. 1) whose special feature is the use of transformers in its arms, with crystals included in the secondary circuits. The main parameters of the filter are determined, and formulae are derived for calculating the resonance and antiresonance frequencies of the arms. A method is proposed for designing the filter elements and for calculating its basic loss. The experimental curve of a filter designed in this manner (Fig. 7) shows a sharp cut-off.

621.372.543.2.029.4 **49**

**A High-*Q* *RC* Feedback Filter.**—M. J. Tucker & L. Draper. (*Electronic Engng*, Oct. 1955, Vol. 27, No. 332, pp. 451–453.) Description of the design and construction of a 175-c/s acceptor filter which will maintain a *Q* of 1 000 stable to within 2% over one hour and a *Q* of 500 to within 2% over considerable periods. The circuit uses a twin-T rejector filter in the negative-feedback path of an amplifier.

621.372.56 : 621.314.632 **50**

**A Voltage-Controlled Attenuator.**—G. M. Ettinger. (*Electronic Engng*, Oct. 1955, Vol. 27, No. 332, pp. 458–459.) "A three-stage voltage controlled germanium diode attenuator is described. Transmission may be varied over a range exceeding 35 decibels by a 12 V control signal. Push-pull operation is discussed and various applications are suggested."

621.372.56.029.6 : 621.372.8 **51**

**An S-Band Variable Attenuator For High-Power Working.**—B. E. Kingdon. (*J. Brit. Instn Radio Engrs*, Sept. 1955, Vol. 15, No. 9, pp. 471–478.) "The description is given of a waveguide insertion attenuator in which the loss in r.f. power is produced by the action of a column of water flowing in a glass tube mounted in the waveguide, continuous variation of attenuation being obtained by adjustment of the water column height. Resonance effects in the water column are exploited by choosing the cross-sectional dimensions of the glass such that an optimum attenuation per unit length of tube is obtained. Experimental results are given for a model covering the range 0 to 30 dB, the input v.s.w.r. at $\lambda_g = 10{\cdot}00$ cm being $\nless 0{\cdot}87$ at any setting and $\geqslant 0{\cdot}97$ over the greater part of the range. Satisfactory operation has been achieved in an evacuated test-system fed by a magnetron delivering 2 MW peak power (200 W mean) at a wavelength of 10 cm. Various modifications in design are suggested for use with higher mean powers."

621.372.6 **52**

**The Graphs of Active Networks.**—W. S. Percival. (*Proc. Instn elect. Engrs*, Part C, Sept. 1955, Vol. 102, No. 2, pp. 270–278. Digest, *ibid.*, Part B, Sept. 1955, Vol. 102, No. 5, pp. 727–729.) Results obtained previously (2587 of 1953 and 379 of 1955) are used to develop the properties of the graphs of linear networks including valves and transformers.

621.373.4 + 621.376.32 **53**

**Linear and Nonlinear Effects in Reactance and Oscillator Circuits.**—T. Zagajewski. (*Hochfrequenztech. u. Electroakust.*, July 1955, Vol. 64, No. 1, pp. 8–18.) The analysis presented of the behaviour of a reactance valve in various circuits is based on an expression for the anode current containing terms up to the third degree in the grid and anode voltages. Two nonlinear effects are considered: (*a*) the primary effect of the curvature in the valve characteristic, and (*b*) the secondary effect due to the presence of harmonic components in the applied signal. These were also investigated experimentally and the results are presented graphically. A circuit for reducing the effect of nonlinearity was also investigated.

621.373.4 **54**

**Stability of Oscillation in Valve Generators.**—A. S. Gladwin. (*Wireless Engr*, Aug.–Nov. 1955, Vol. 32, Nos. 8–11, pp. 206–214, 246–253, 272–279 & 297–304.) "A comprehensive theory of stability is developed which is applicable to a large class of harmonic oscillators. The paper is concerned mainly with the 4-terminal regenerative type of oscillator with grid-leak bias but the analysis applies also to circuits with fixed grid-bias and to 2-terminal oscillators. All known forms of instability appear as special cases of the general theory and some new forms are predicted. Stability is determined by the nature of the roots of a characteristic equation and stability criteria are obtained in the form of inequalities between the parameters of the amplifier and feedback network. A modified form of Nyquist locus-diagram can also be used. When the feedback network is symmetrical with respect to the oscillation frequency the characteristic equation can be factorized to give independent criteria for frequency and amplitude stability. Hysteresis effects and periodic instability are analysed in detail. In addition to the general treatment, specific forms of the various parameters which appear in the stability criteria are worked out for a three-halves-law amplifier with a semi-linear or exponential grid-current characteristic. Quantitative experimental confirmation is provided."

621.373.4 : 621.396.822 **55**

**Effect of Electrical Fluctuations on a Valve Oscillator.**—P. I. Kuznetsov, R. L. Stratonovich & V. I. Tikhonov. (*Zh. eksp. teor. Fiz.*, May 1955, Vol. 28, No. 5, pp. 509–523.) The effect of 'slow' normal fluctuations is considered. Expressions are derived for the one-dimensional probability density functions for amplitude and phase. An approximate method is shown for calculating the correlation functions for amplitude and phase variations.

621.373.421 : 621-526 **56**

**The Frequency-Response Analysis of Nonlinear Systems.**—P. E. W. Grensted. (*Proc. Instn elect. Engrs*, Part C, September 1955, Vol. 102, No. 2, pp. 244–253. Digest, *ibid.*, Part B, Sept. 1955, Vol. 102, No. 5, pp. 721–723.) If the waveform at the input to the nonlinear elements in the system is approximately sinusoidal, the analysis may be carried out on the assumption that all harmonics generated by the nonlinear elements can be ignored. Analysis of on-off control systems on this basis gives results within about 10% of the exact solutions for both transient and steady oscillations.

621.373.421.13 **57**

**Synthesizing Crystal Stability.**—W. A. Hayes. (*Radio-Electronic Engng*, May 1955, Vol. 24, No. 5, pp. 20, 30.) The output from two crystal oscillators, combined to give a 10-kc/s beat frequency, is squared and triggers a blocking oscillator, producing a 10-kc/s pulse which is used to quench a tunable oscillator. The tunable oscillator locks to harmonics of the 10-kc/s signal.

621.373.423 : 621.316.726 **58**

**An Electromechanical Stabilizer of the Frequency of a Klystron.**—A. I. Barchukov, G. A. Vasil'ev, M. E. Zhabotinski & B. D. Osipov. (*Radiotekhnika, Moscow*, March 1955, Vol. 10, No. 3, pp. 29–32.) In the method proposed a portion of the klystron output is applied to a cavity resonator, the natural frequency of which is modulated by a separately excited vibrating diaphragm. The output of the resonator is detected, amplified and, through a phase detector, applied to the klystron reflector. The theory of the method is discussed and its accuracy is determined. A comparison with the resonance frequency of the absorption line of ammonia has shown that the relative frequency error of the method is of the order of $5 \times 10^{-6}$.

621.373.424.029.42 **59**

**Subaudio Oscillator Tunes 0 to 50 cycles.**—L. Fleming & W. W. Follin. (*Electronics*, Oct. 1955, Vol. 28, No. 10, pp. 144–145.) A beat-frequency oscillator using carrier frequencies of about 2 kc/s provides an output of up to 3 V constant within 2% over the frequency range, with $< 0{\cdot}2$ c/s per hour drift and $< 1\%$ distortion. The need for low-frequency amplification is avoided by the use of a high-level mixer circuit consisting of two Se rectifiers with balancing adjustment to eliminate the d.c. component.

621.374.3 : 621.315.212 **60**

**Some Transmission Line Devices for Use with Millimicrosecond Pulses.**—I. A. D. Lewis. (*Electronic Engng*, Oct. 1955, Vol. 27, No. 332, pp. 448–450.) A phase inverter, impedance transformers and a valve-heater isolating transformer are described, all comprising coaxial-cable sections.

621.374.4 : 621.385.832 **61**

**Generation of Harmonics of a Given Fundamental.**—H. Katz & K. L. Rau. (*Frequenz*, July 1955, Vol. 9, No. 7, pp. 234–240.) Theory and experimental results are given for a frequency multiplier based on use of a c.r. tube in which a ribbon beam is deflected across a metal target with an appropriate profile. Suitable beam-forming structures are described.

621.375.1.029.63 : 621.396.822 **62**

**Noise in U.H.F. Amplifiers.**—E. P. Dement'ev. (*Radiotekhnika, Moscow*, Jan. 1955, Vol. 10, No. 1, pp. 45–52.) General noise effects in u.h.f. amplifiers of any type are considered. Relations between the noise parameters are established using the method of the transfer of the current generators proposed previously (*ibid.*, 1954, No. 4.) General expressions for the noise factors are derived, and the minimum number of noise parameters required to characterize a circuit is determined. A concept of 'partial' noise parameters is introduced.

621.375.232 : 623.011.21 **63**

**A Negative Impedance Converter for Use as a Two-Terminal Amplifier.**—J. R. Tillman. (*P.O. elect. Engrs' J.*, July 1955, Vol. 48, Part 2, pp. 97–101.) A specified stable negative impedance may be generated by connecting a positive impedance between the output and input of an amplifier possessing a high degree of negative feedback. An amplifier suitable for use in transmission-line systems is described.

621.375.3.049.75 **64**

**Etched Wiring simplifies Magnetic Amplifier Design.**—K. H. Sueker. (*Elect. Mfg*, Jan. 1955, Vol. 55, No. 1, pp. 106–107.)

621.373/.376 65

**Active Networks.** [Book Review]—V. C. Rideout. Publishers: Prentice-Hall, New York, and Constable, London, 1954, 485 pp., 42s. (*Sci. Progr.*, July 1955, Vol. 43, No. 171, pp. 534–535.) Deals with circuits including valves, transistors and other devices. There are sections on the basic principles of analogue computers, servo systems and wave-shaping circuits as well as on general aspects of amplifiers and oscillators.

## GENERAL PHYSICS

53.087/.088 66

**The Correction of Observational Data for Instrumental Bandwidth.**—F. D. Kahn. (*Proc. Camb. phil. Soc.*, July 1955, Vol. 51, Part 3, pp. 519–525.) The formula relating the readings, $\phi(x)$, of an instrument having the apparatus function $g(x)$ to the true values $f(x)$ of an observed function is stated. A three-point interpolation operator can be chosen to simulate the spreading action of a class of functions $g(x)$. Under certain conditions the interpolation operator has an exact inverse which, when applied to $\phi(x)$, gives a function which is a better approximation to $f(x)$. The accuracy obtained is discussed, and the effect of the method on the standard errors in the measurement is found.

531 + 537].001.362 67

**Electromechanical Analogies.**—G. Meyer. (*Frequenz*, July 1955, Vol. 9, No. 7, pp. 227–232.) A discussion of the physical significance of various analogies and of circuit concepts and terminal representation in mechanics.

535.215 68

**State of Research on Photoelectricity.**—W. Kluge. (*Z. angew. Phys.*, June 1955, Vol. 7, No. 6, pp. 302–311.) A survey is presented of research since 1949. 122 references.

535.22 69

**Velocity of Light from the Molecular Constants of Carbon Monoxide.**—E. K. Plyler, L. R. Blaine & W. S. Connor. (*J. opt. Soc. Amer.*, Feb. 1955, Vol. 45, No. 2, pp. 102–106.) From infrared spectroscopy measurements the value of 299 792 km/s is deduced for the velocity of light, with a standard deviation of 6 km/s. A short account of the work is given in *Tech. News Bull. nat. Bur. Stand.*, Jan. 1955, Vol. 39, No. 1, pp. 1–3.

535.642.08 70

**Simple Color Computer Gives Tristimulus Values.**—R. P. Burr & J. R. White. (*Electronics*, Oct. 1955, Vol. 28, No. 10, pp. 166–169.) Trichromatic coefficients for the CIE colour-specification system are automatically computed on a simple analogue machine.

537.223/.224 71

**An Interesting Phenomenon Observed on some Dielectrics.**—I. Sumoto. (*J. phys. Soc. Japan*, June 1955, Vol. 10, No. 6, p. 494.) A glass tube, either empty or containing a polar liquid or a metal rod, rotates continuously when placed in an inhomogeneous electric field of sufficient strength if the electrodes and part of the tube are immersed in carnauba wax or a higher alcohol, stearic acid, palmitic acid or xylol.

537.29 : 621.315.61 72

**Field-Dependent Conductivity in Nonuniform Fields and its Relation to Electrical Breakdown.**—P. T. G. Flynn. (*Proc. Instn elect. Engrs*, Part C, Sept. 1955, Vol. 102, No. 2, pp. 264–269. Digest, *ibid.*, Part B, Vol. 102, No. 5, pp. 723–725.) "A method is given for calculating the resulting field distribution when the conductivity is a function of the field strength. This is applied to a point-plane electrode system in an amorphous solid, with the assumption that the conductivity at low field-strengths is electronic and that the conductivity at high field-strengths is described by a relation due to Fröhlich. The results show that for polymethylmethacrylate with average fields of the order of $10^6$ volt/cm, the reduction in maximum field-strength from that calculated for constant conductivity can be a factor of five or more, the magnitude of the reduction depending on the degree of non-uniformity of the field." Various other insulating materials are also discussed.

537.5 : 538.561 73

**Study of Gyromagnetic Resonance in the Penning Gauge.**—G. Dumas. (*Rev. gén. Élect.*, July 1955, Vol. 64, No. 7, pp. 331–350.) Conditions in the Penning gauge are particularly favourable for studying r.f. radiation resulting from gyromagnetic resonance of electrons [see 1139 of 1951 (Laffineur & Pecker)]. Both longitudinal oscillations and thermal agitation of the electrons are involved, resulting in an accumulation in the anode plane. The presence of plasma has an important effect on the potential distribution, introducing a radial electric field. The equations of electron motion are established; some particular solutions indicate the possibility of a circular path in the anode plane, with an angular frequency about half the gyrofrequency. With values of magnetic induction between 100 and 1 000 G radiation at decimetre wavelengths was produced. Measurements of the radiation are described. Investigation of the complete spectrum emitted indicates that coherent oscillations may occur. The results are compared with those of other workers.

537.52 : 535.215.9 74

**Influence of Light on the Mechanism of Gas Discharges (Joshi Effect).**—S. R. Mohanty. (*Z. Phys.*, 20th April 1955, Vol. 140, No. 4, pp. 370–375.) Results of previous experimental work are reviewed. The variation of the discharge current with the intensity and frequency of the incident light and in particular the reversal of this current variation in a certain voltage range are interpreted in terms of space-charge effects.

537.533 75

**The Mechanism of High-Field Electron Emission from Tarnished Metal Surfaces.**—T. J. Lewis. (*Proc. phys. Soc.*, 1st Aug. 1955, Vol. 68, No. 428B, pp. 504–512.) A theory of field-dependent thermal emission is proposed which takes into account the existence of patch fields due to variation in work function over the surface.

537.533 76

**Numerical Ray Tracing in Electron Lenses.**—J. C. E. Jennings & R. G. Pratt. (*Proc. phys. Soc.*, 1st Aug. 1955, Vol. 68, No. 428B, pp. 526–536.) Corrections are suggested to Liebmann's formulae for tracing paraxial rays (966 of 1950); a more accurate method is developed, applicable also to marginal rays, and is compared with methods suggested by other workers. The use of the constancy of the Wronskian determinant as a criterion of accuracy of paraxial ray tracing is discussed.

537.533 77

**Effect of Space Charge on the Propagation of Intense Beams of Charged Particles.**—M. D. Gabovich. (*Uspekhi fiz. Nauk*, June 1955, Vol. 56, No. 2, pp. 215–256.) A review. About a third of the 81 references given are to Russian work.

537.533 : 538.691 **78**

**A Method of finding a Large Class of Electrostatic and Magnetic Fields for which the Solutions of the Fundamental Equation of Electron Optics may be expressed in Terms of Known Functions.**—B. E. Bonshtedt. (*Zh. tekh. Fiz.*, March 1955, Vol. 25, No. 3, pp. 541–543.) Exact solutions of equation (1) are of great interest for various types of e.s. and magnetic fields. Until recently such solutions were known only for a very limited number of cases. Grinberg in particular indicated a class of fields for which the trajectories of electrons are expressed in terms of cylindrical functions (*Zh. tekh. Fiz.*, 1953, Vol. 23, p. 1904). Solutions are now found for a much larger class of fields.

537.533.1 **79**

**Velocity Distribution of Electrons in Presence of an Alternating Electric Field and Constant Magnetic Field.**—V. M. Fain. (*Zh. eksp. teor. Fiz.*, April 1955, Vol. 28, No. 4, pp. 422–430.) Two cases are considered theoretically: (*a*) the electric field varies sinusoidally, and (*b*) the variation is amplitude modulated. Case (*b*) is of interest in the investigation of ionospheric cross-modulation. The effect of collisions between electrons is taken into account.

537.562 : 621.317.335.3 : 621.372.413 **80**

**Resonant-Cavity Measurements of the Relative Permittivity of a D.C. Discharge.**—K. W. H. Foulds. (*Proc. Instn elect. Engrs*, Part C, Sept. 1955, Vol. 102, No. 2, pp. 203–216. Digest, *ibid.*, Part B, July 1955, Vol. 102, No. 4, pp. 564–567.) Measurements at 2·1, 2·5 and 3 kMc/s on low-pressure d.c. discharges in (*a*) a mixture of Hg vapour and Ar, and (*b*) Hg vapour alone. Decrease of permittivity with increase of discharge current was noted in all cases. In case (*b*) the cavity $\underset{\sim}{Q}$-factor exhibited a minimum at a value of discharge current depending on the resonance frequency; this is attributed to electron resonance. Theory of the method is given.

538.1 **81**

**General Theory of Magnetic Masses at Rest and in Movement.**—É. Durand. (*Rev. gén. Élect.*, July 1955, Vol. 64, No. 7, pp. 350–356.) Theory is presented based on the analogy with electric charge.

538.1 **82**

**A New Vectorial Transformation of a Curved-Line Integral into a Surface Integral, and its Application to Magnetostatics.**—É. Durand. (*C. R. Acad. Sci., Paris*, 17th Aug. 1955, Vol. 241, No. 7, pp. 594–596.)

538.1 : 538.221 **83**

**Magnetostatics with Ferromagnetic Materials.**—V. I. Skobelkin & R. N. Solomko. (*Zh. eksp. teor. Fiz.*, April 1955, Vol. 28, No. 4, pp. 385–393. Correction. *ibid.*, June 1955, No. 6, p. 766.) A variational principle is formulated and proved in the theory of ferromagnetic materials in the field of electric currents. A direct method is developed for investigating magnetic fields in the general case of a nonlinear dependence of permeability on magnetic field strength.

538.3 **84**

**Comments on the Nonlinear Electrodynamics.**—K. Bechert. (*Ann. Phys., Lpz.*, 15th July 1955, Vol. 16, Nos. 3/4, pp. 97–110.) Some solutions of the equations presented earlier (1969 of 1953) are examined.

538.3 : 512.831 **85**

**Description of the Electromagnetic Field by means of Matrices.**—G. A. Zaitsev. (*Zh. eksp. teor. Fiz.*, May 1955, Vol. 28, No. 5, pp. 524–529.) The properties of 'matrix-tensors' are briefly considered and the quantities characterizing an e.m. field are then expressed in matrix form.

538.56 : 537.5 **86**

**Note on Plasma-Electron Oscillations.**—K. G. Emeleus & T. K. Allen. (*Aust. J. Phys.*, June 1955, Vol. 8, No. 2, pp. 305–306.) Brief report of the results of experiments on hot-cathode discharges in Hg vapour or Ar at pressures of the order of $10^{-3}$ mm Hg, and discussion of their relation to the work of Wild et al. (704 of 1955).

538.566 **87**

**The Perturbation of an Electromagnetic Field by Small Deformations of a Metal Surface.**—B. Z. Katsenelenbaum. (*Zh. tekh. Fiz.*, March 1955, Vol. 25, No. 3, pp. 546–557.) A general expression is derived for the perturbation due to the deformation. If the deformation is slight this perturbation can be found by the usual method of quadratures. The method is applied to the problem of the diffraction of a plane wave at a step, the height of which is small compared with $\lambda$. From this the well-known formulae of Rayleigh for reflection from small and smooth irregularities are easily derived.

538.566 **88**

**Study of the Local Electromagnetic Field at the Junction of Three Dielectric Media.**—G. Weill. (*C. R. Acad. Sci., Paris*, 8th Aug. 1955, Vol. 241, No. 6, pp. 554–556.) Analysis is given for the case of three media with a common edge.

538.566 : 535.42 **89**

**A Diffraction Problem.**—P. Poincelot. (*C. R. Acad. Sci., Paris*, 22nd Aug. 1955, Vol. 241, No. 8, pp. 625–627.) Analysis is presented for the case of a plane e.m. wave incident normally on an infinitely long perfectly conducting strip.

538.566 : 537.31 **90**

**Scattering of Electromagnetic Waves from a 'Lossy' Strip on a Conducting Plane.**—J. R. Wait. (*Canad. J. Phys.*, July 1955, Vol. 33, No. 7, pp. 383–390.) An examination is made of the error introduced in the analysis of radiation and propagation problems by assuming perfect conductivity in media, such as the ground, which are in fact imperfectly conducting. The result of the approximate calculation is compared with that of an exact calculation involving elliptic wave functions. For the approximation to be permissible, the surface impedance of the lossy regions should not exceed 0·1 times the impedance of free space.

538.566 : 537.56 **91**

**Hydromagnetic Waves in Ionized Gas.**—J. H. Piddington. (*Nature, Lond.*, 10th Sept. 1955, Vol. 176, No. 4480, p. 508.) The equation representing the e.m. field in a moving isotropically conducting medium is discussed and expressions are derived for the velocities of the ordinary, extraordinary and $S$ (sound-type) waves, for gas pressures (*a*) much less than and (*b*) much greater than the 'magnetic pressure'. See also 3579 of 1955.

538.569.4 : 535.33 **92**

**New Technique for High-Resolution Microwave Spectroscopy.**—R. H. Romer & R. H. Dicke. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, pp. 532–536.) Spectral lines of width substantially less than the Doppler width are produced by containing the gas in a pillbox-shaped cavity of height about $\lambda/2$ and diameter large compared to $\lambda$.

538.615 : 538.569.4.029.6 : 546.171.1 **93**

**The Zeeman Effect in Ammonia Microwave Spectra.**—Y. Torizuka, Y. Kojima, T. Okamura & K. Kamiryo. (*J. phys. Soc. Japan*, June 1955, Vol. 10, No. 6, pp. 417–420; *Sci. Rep. Res. Inst. Tohoku Univ., Ser. A*, June 1955, Vol. 7, No. 3, pp. 316–320.) The Zeeman effect was investigated experimentally using magnetic fields up to 15 000 oersted. Results indicate that the $J = 1, K = 1$ line shows a change from the Zeeman effect to the Paschen-Back effect at field strengths near 10 000 oersted.

538.63 : 538.221 **94**

**Theory of the Nernst Effect in Ferromagnetic Metals.**—K. Meyer. (*Z. Naturf.*, Feb. 1955, Vol. 10a, No. 2, p. 166.) On the basis of a two-band model with one band nearly empty and the other nearly fully occupied, theory for the Nernst effect can be derived simply from that for the Hall effect.

538.632 **95**

**The Hall Effect and Related Phenomena.**—F. A. Vick. (*Sci. Progr.*, July 1955, Vol. 43, No. 171, pp. 454–461.) A brief review is presented of recent work on the theory of the Hall effect in ferromagnetic, in strongly paramagnetic, and in semiconducting materials. Practical applications of the effect, such as in the InSb power amplifier [2576 of September (Ross & Thompson)] are also mentioned.

539.155 **96**

**Reliability of Atomic Masses in the Chromium-Germanium Region.**—J. T. Kerr, J. G. V. Taylor & H. E. Duckworth. (*Nature Lond.*, 3rd Sept. 1955, Vol. 176, No. 4479, pp. 458–459.) Discrepancies between atomic-mass values obtained from transmission data and those obtained by mass-spectrometer methods are discussed. The mass-spectrometer values for Ni may be in error, but those for the other elements from Cr to Ge are considered to be accurate.

## GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16 : 523.3 **97**

**Lunar Occultation of a Radio Star and the Derivation of an Upper Limit for the Density of the Lunar Atmosphere.**—B. Elsmore & G. R. Whitfield. (*Nature, Lond.*, 3rd Sept. 1955, Vol. 176, No. 4479, pp. 457–458.) Brief report of observations made at Cambridge, England, of the passage of the moon across the large-diameter radio source in Gemini. Wavelengths of 7·9 and 3·7 m were used.

523.16 : 621.396.677 **98**

**Chord Construction for Correcting Aerial Smoothing.**—R. N. Bracewell. (*Aust. J. Phys.*, June 1955, Vol. 8, No. 2, pp. 200–205.) A graphical method of correcting for aerial smoothing in radio astronomy, simpler and no less accurate than methods described previously.

523.5 : 621.396.96 **99**

**The Influence of Noise on Radar Meteor Observations.**—C. D. Ellyett & G. J. Fraser. (*Aust. J. Phys.*, June 1955, Vol. 8, No. 2, pp. 273–278.) An experimental investigation has been made of the minimum detectable echo power from meteor trails using 69-Mc/s radar equipment with incoherent detection. For optimum signal/noise ratio the pulse width must exceed the width of the c.r.-tube spot. Extraterrestrial noise is predominant, receiver noise figure being of second-order importance. Variations in the effective aerial temperature introduce corresponding variations in the observed meteor rate. Using artificial echoes, a minimum signal/noise ratio of 8 dB is found to be necessary for detection; the most important parameter is the total received signal energy. It is not likely that meteors of smaller size can be made observable by improving receivers.

523.5 : 621.396.96 **100**

**Diffusion Coefficients from the Rate of Decay of Meteor Trails.**—A. A. Weiss. (*Aust. J. Phys.*, June 1955, Vol. 8, No. 2, pp. 279–288.) An estimate of the diffusion coefficient in the meteor zone is obtained by an extension of Huxley's theory of ambipolar diffusion (*Aust. J. sci. Res., Ser. A*, 1952, Vol. 5, p. 10). The absolute value of the diffusion coefficient thus calculated, and its gradient with height, are confirmed by comparison with results of measurements made at Adelaide, using c.w. radar technique at a wavelength of 11·2 m [1424 of 1954 (Robertson et al.)]. The theory appears adequate to account for the observed variation or diffusion coefficient with height.

523.7 **101**

**The Identification of Sunspots by means of Chromospheric Faculae.**—M. d'Azambuja. (*C. R. Acad. Sci., Paris*, 17th Aug. 1955, Vol. 241, No. 7, pp. 592–594.)

550.372 : 621.396.11 **102**

**Influence of Temperature and Humidity on the Electrical Constants of the Ground.**—M. Argirovic. (*Ann. Télécommun.*, May 1955, Vol. 10, No. 5, pp. 113–116.)

550.38 : 551.510.535 **103**

**Comparison of Records of the Horizontal Component of the Terrestrial Magnetic Field (*K* index) at Kerguelen and at Heard Island.**—A. Luchet. (*C. R. Acad. Sci., Paris*, 8th Aug. 1955, Vol. 241, No. 6, pp. 569–571.) The records confirm that magnetic activity is particularly strong during the equinoxes. The correlation coefficient for the two stations is about 0·7. Ionograms obtained during the strong magnetic variations are characteristic of a highly disturbed ionosphere.

550.380.8 **104**

**NOL Vector Airborne Magnetometer Type 2A.**—E. O. Schonstedt & H. R. Irons. (*Trans. Amer. geophys. Union*, Feb. 1955, Vol. 36, No. 1, pp. 25–41.) The instrument previously described (719 of 1954) has been modified so as to be suitable for use in all latitudes and to simplify the reduction of results.

550.380.87 **105**

**A Recording Magnetic Variometer.**—J. H. Meek & F. S. Hector. (*Canad. J. Phys.*, July 1955, Vol. 33, No. 7, pp. 364–368.) Description of an electronic magnetometer designed for continuous measurement of changes in the earth's magnetic field at one place; variations as small as $10^{-5}$ oersted can be recorded.

550.385 **106**

**Notes on the Distribution of *SC** in High Latitudes.**—T. Nagata & S. Abe. (*Rep. Ionosphere Res. Japan*, March 1955, Vol. 9, No. 1, pp. 39–44.) An examination of experimental data shows that the equivalent overhead currents for the preliminary reverse impulse in this type of magnetic storm in the northern hemisphere are represented by current flows across the polar cap from the longitude corresponding to 10 h to that corresponding to 22 h and by the resulting two vortices extending to lower latitudes, a clockwise vortex in the afternoon hemisphere and a weaker counter-clockwise vortex in the forenoon hemisphere.

550.385 : 523.78 **107**

**The Effect on Geomagnetism of the Solar Eclipse of 30th June 1954.**—J. Egedal & N. Ambolt. (*J. atmos. terr. Phys.*, Aug. 1955, Vol. 7, Nos. 1/2, pp. 40–48.) An examination of records of magnetic declination from 11 observatories near the path of totality shows that a maximum magnetic effect occurs at the time of totality, appearing as a diminution of normal daytime departures from the night-time value of the declination. Quantitative agreement with theory is shown.

551.510.52 : 535.325 **108**

**Survey of Airborne Microwave Refractometer Measurements.**—C. M. Crain. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, pp. 1405–1411.) Results of measurements made at various places in the U.S.A., over land and off the east and west coasts, are summarized. Variations of the vertical distribution of refractive index are studied in relation to time and place. Particular attention is paid to the fine detail of the refractive-index structure. See also e.g. 2662 of 1954 (Crain et al.).

551.510.52 : 535.325 **109**

**Amplitude, Scale and Spectrum of Refractive-Index Inhomogeneities in the First 125 Meters of the Atmosphere.**—G. Birnbaum & H. E. Bussey. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, pp. 1412–1418.) Report of an extensive series of observations made using two refractometers at various levels on a 128-m tower. The correlation between the refractive-index variations and variations of meteorological conditions is studied.

551.510.534 **110**

**Seasonal and Latitudinal Temperature Changes in the Ozonosphere.**—J. Pressman. (*J. Met.*, Feb. 1955, Vol. 12, No. 1, pp. 87–89.) Computations are presented relevant to the elucidation of the temperature structure in the middle atmosphere. See also 1028 of 1955.

551.510.535 **111**

**Hydromagnetic Resonance in Ionospheric Waves.**—C. O. Hines. (*J. atmos. terr. Phys.*, Aug. 1955, Vol. 7, Nos. 1/2, pp. 14–30.) The large-scale travelling ionospheric disturbances detected by various workers in the $F_2$ layer may be partially explained as an electron resonance effect, produced when an initial atmospheric disturbance generates natural modes of electromagnetic oscillation, as a result of hydromagnetic coupling.

551.510.535 **112**

**Some Features of the $E_2$ Layer observed at the Ionosphere Field Station, Haringhata, Calcutta.**—A. K. Saha & S. Ray. (*J. atmos. terr. Phys.*, Aug. 1955, Vol. 7, Nos. 1/2, pp. 107–108.) Photographic records are reproduced showing how the $E_2$ ionization originates at $F_1$-layer level, moves downwards and forms a sporadic E layer.

551.510.535 **113**

**Recombination and Ion Production from the Total Electron Content.**—N. J. Skinner & R. W. Wright. (*J. atmos. terr. Phys.*, Aug. 1955, Vol. 7, Nos. 1/2, pp. 105–107.) Revised values are given for the figures suggested previously (1338 of 1955) for the mean recombination coefficient and the mean rate of ion production in the F layer. Results of observations at Ibadan for magnetically quiet and disturbed days are plotted and discussed.

551.510.535 **114**

**A Consideration of the Mechanism of Electron Removal in the $F_2$ Layer of the Ionosphere.**—T. Yonezawa. (*Rep. Ionosphere Res. Japan*, March 1955, Vol. 9, No. 1, pp. 17–37.) Possible mechanisms of electron removal are discussed in the light of observations. No satisfactory theory of the phenomenon exists. See also 1036 of 1955.

551.510.535 : 523.78 **115**

**Variation of $fE_s$ during Solar Eclipses.**—W. Stoffregen. (*Nature, Lond.*, 24th Sept. 1955, Vol. 176, No. 4482, p. 610.) Records obtained by Minnis at Khartoum in 1952 (2298 of 1955) and by Stoffregen in the north of Sweden in 1945 (1778 of 1947) are discussed and compared with observations from various stations in 1954. A decrease of $fE_s$ during the eclipse is found only at places where the eclipse is total or nearly so. The typical eclipse configuration of sun-moon-earth is thought to be in some way responsible for the associated preliminary increase of $fE_s$.

551.510.535 : 525.624 **116**

**The Lunar Semidiurnal Oscillation in the Ionospheric Absorption of 150-kc/s Radio Waves.**—A. P. Mitra. (*J. atmos. terr. Phys.*, Aug. 1955, Vol. 7, Nos. 1/2, pp. 99–100.) Vertical-incidence absorption measurements made at Pennsylvania State University during 1949–1953 appear to confirm earlier work by Appleton & Beynon (3133 of 1949).

551.510.535 : 550.385 **117**

**Geomagnetic Activity and Average Deviations of Daytime $F_2$-Layer Critical Frequencies in Different Geomagnetic Latitudes.**—G. Lange-Hesse. (*J. atmos. terr. Phys.*, Aug. 1955, Vol. 7, Nos. 1/2, pp. 49–60. In German.) Analysis of data from six ionospheric sounding stations and one geomagnetic observatory shows a distinct dependence of the average daytime deviations from median values on the geomagnetic latitude, the seasons and the sunspot cycle. The effect on the prediction of ionospheric disturbances is discussed and the reliability of predictions is assessed. See also 2942 of 1954.

551.510.535 : 551.55 **118**

**Sporadic-E Movements.**—N. C. Gerson. (*J. Met.*, Feb. 1955, Vol. 12, No. 1, pp. 74–80.) Reports of radio amateurs on v.h.f. reception in N. America are consistent with atmospheric movements at a height of 110 km with speeds averaging about 300 km/h. Data on winds at heights up to 120 km, determined by various methods, indicate that at all heights the winds are predominantly from the east during summer and from the west during winter.

551.510.535 : 551.594.5 **119**

**The Auroral E-Layer Ionization and the Auroral Luminosity.**—A. Omholt. (*J. atmos. terr. Phys.*, Aug. 1955, Vol. 7, Nos. 1/2, pp. 73–79.) An empirical relation is established between the photon emission within the negative nitrogen bands of aurorae in zenith and the maximum electron density in the E layer. The recombination coefficient in the layer during aurorae is $> 10^{-7}$ cm³/s. In medium to strong aurorae the mean electron density, deduced from critical frequencies, is between $2 \cdot 10^5$ and $10 \cdot 10^5$ electrons/cm³. V.h.f. echoes observed during aurorae may be explained as backscatter from the ground reflected at low incidence from the auroral E layer.

551.510.535 : 621.396.11 **120**

**The Effects of Anisometric Amplitude Patterns in the Measurement of Ionospheric Drifts.**—G. J. Phillips & M. Spencer. (*Proc. phys. Soc.*, 1st Aug. 1955, Vol. 68, No. 428B, pp. 481–492.) Measurement of ionospheric drift by determination of time shifts between fading curves of ionospherically reflected waves received at three spaced points is discussed and the errors involved in assuming that the amplitude pattern on the ground is statistically isometric are evaluated. From 15 recordings of fading curves on 2·4 Mc/s a mean error of 15° in the direction of drift was found, the greatest error being 53°. The errors are not generally systematic.

551.510.535 : 621.396.11 **121**

**The Shape of Irregularities in the Upper Ionosphere.**—M. Spencer. (*Proc. phys. Soc.*, 1st Aug. 1955, Vol. 68, No. 428B, pp. 493–503.) Observations of radio stellar scintillations made with three spaced receivers on a frequency of 38 Mc/s show that irregularities in the amplitude pattern on the ground are, regarded statistically, elliptically shaped, the ratio of major to minor axis being at least 5 : 1. The direction of the major axis of the corresponding ellipsoidal ionospheric irregularities appears to be along the lines of force of the earth's magnetic field. Estimates of the drift velocity of ionospheric irregularities based on the assumption that the ionospheric pattern has spherical symmetry are hence subject to error.

551.594.6 **122**

**Some Waveforms of Atmospherics and their Use in the Location of Thunderstorms.**—F. Horner & C. Clarke. (*J. atmos. terr. Phys.*, Aug. 1955, Vol. 7, Nos. 1/2, pp. 1–13.) The main types of atmospherics waveforms recorded in southern England are illustrated and their use in estimating range is discussed. In an analysis of recordings made during one year, 3% of the waveforms recorded by day and 5% of those by night were suitable for this purpose. Most night-time storms within a distance of 2 000 km produce some waveforms from which range may be determined with an accuracy within ±30%; by day the corresponding distance is 500 km in summer and 1 000 km in winter.

## LOCATION AND AIDS TO NAVIGATION

621.396.9 **123**

**International Conference on Lighthouses** [and other aids to navigation] **at Scheveningen** [31st May–9th June 1955].—C. B. Broersma. (*Tijdschr. ned. Radiogenoot*, July 1955, Vol. 20, No. 4, pp. 257–264.) A report is given of the papers and discussion. Navigation aids discussed included radar, loran, consol, radar beacons, radio-range beacons, radiocommunication.

621.396.93 **124**

**Microwave Harbour Beacon.**—A. L. P. Milwright. (*Wireless World*, Nov. 1955, Vol. 61, No. 11, pp. 569–570.) A 3-cm-λ pulse transmitter feeds alternately two vertical resonant-slot aerials erected at the harbour entrance and arranged on either side of a separator plate, so that the radiated beams overlap, defining the safe course for entry along a sector 1° wide within which a continuous signal is received. Reception is by a pre-tuned transistor receiver, coupled to a horn aerial having a horizontal beam width of 16°. On trial the equipment, which is intended for small craft, operated satisfactorily at ranges up to 7 miles.

621.396.933 **125**

**TACAN Navigation System shows Bearing and Distance.**—(*Electronics*, Oct. 1955, Vol. 28, No. 10, pp. 174, 176.) Information is automatically transmitted by a ground installation to an aircraft on interrogation by means of correctly spaced pulses. Bearing accuracy is within ±1° and range is within 600 ft + 0·15% of the distance measured. Operating frequency is around 1 kMc/s. An outline is given of the principles involved.

621.396.96 **126**

**Portable Precision Approach Radar.**—J. B. Levin. (*Electronics*, Oct. 1955, Vol. 28, No. 10, pp. 154–159.) The system described uses two corner reflectors to determine the touch-down point, the runway centre-line and the elevation reference. The equipment takes six hours to set up and provides touch-down accuracy within ±20 ft.

621.396.96 : 551.577 **127**

**Radar Observations of Rain at Poona.**—B. K. Gupta, A. M. Mani & S. P. Venkiteshwaran. (*Indian J. Met. Geophys.*, Jan. 1955, Vol. 6, No. 1, pp. 31–40.) Observations made with a search radar Type SCR 717 C operating at 9·1 cm λ are reported and illustrated by photographs.

621.396.96 : 621-526 **128**

**The Stability and Time Response of Fast-Operating Closed-Loop Pulsed Radar Circuits.**—McDonnell & Perkins. (See 269.)

621.396.963.3 : 621.396.822 **129**

**Detection of Pulse Signals in Noise.**—D. G. Tucker & J. W. R. Griffiths. (*Wireless Engr*, Nov. 1955, Vol. 32, No. 11, pp. 290–297.) Visual detection of pulse signals against a noise background on an intensity-modulated display is considered; an examination is made of the relation between the probability of detection at any given signal/noise ratio and the measured qualities of the c.r. tube and its phosphor, or of the chemical recorder and its iodized paper. Theory is presented in terms of just-noticeable differences (j.n.ds), defined as the increase of voltage or current which can with certainty cause a just-noticeable increase in visible intensity; the size and number of j.n.ds constitute a subjective measure of the display characteristics. Numerical examples illustrating the theory are presented.

## MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 : 546.482.21 **130**

**Influence of Infrared Rays on the Photoconductivity of the Polycrystalline CdS.**—S. Yoshimatsu, C. Kanzaki, S. Ibuki & N. Murai. (*J. phys. Soc. Japan*, June 1955, Vol. 10, No. 6, p. 493.) In the specimens investigated, the photocurrent under irradiation by visible light decreased when infrared illumination was added. The dependence of this effect on the wavelength of the infrared radiation is shown graphically for temperatures of 18°, 110° and 150°C.

535.37 + 535.215 **131**

**Infrared Stimulation and Quenching of Photoconductivity in Luminescent Powders.**—H. Kallmann, B. Kramer & A. Perlmutter. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, pp. 391–400.) Continuing previous work (2021 of 1953), measurements are reported on a range of (ZnCd)S phosphors, indicating that photoconductivity excited by ultraviolet or gamma rays may be stored for days and revealed by infrared stimulation, whereas application of the infrared radiation at the same time as the excitation produces quenching of both fluorescence and photoconductivity. An explanation of the observations is given in terms of the action of the infrared radiation on trapped and valence-band electrons.

535.37 : 537.311.33 **132**

**Present-Day Problems in Crystal Luminescence.**—M. Curie & D. Curie. (*Cah. Phys.*, March–June 1955, Nos. 55–58, pp. 1–36, 29–48 & 52–80.) A comprehensive survey. The structure of inorganic crystal phosphors, as typified by ZnS, is studied by the methods used for investigating semiconductors generally as well as by specialized luminescence methods. Numerous electrons at the bottom of the conduction band contribute to the luminescence but are not sufficiently displaced to contribute effectively to photoconduction. Radiationless transitions are discussed and a synthesis of various theories is presented. Experimental work, to be effective, should be conducted with transparent films.

535.37 + 537.226.2] : 546.472.21 **133**

**Change of the Dielectric Constant of Phosphors under the Influence of Infrared Radiation.**—E. E. Bukke. (*Zh. eksp. teor. Fiz.*, April 1955, Vol. 28, No. 4, pp. 507–508.) In all cases when luminescence was increased by illumination with radiation of wavelength 0·8 $\mu$ or 1·3 $\mu$, the dielectric constant also increased; in cases when the luminescence decreased the dielectric constant remained unchanged. Results for ZnS-Cu,Tm and ZnS.Cds-Cu,Ni are presented graphically; results for other phosphors are tabulated.

535.37 : 546.472.21 **134**

**State of Cu Activator in Zinc-Sulphide Phosphors.**—A. A. Cherepnev. (*Zh. eksp. teor. Fiz.*, April 1955, Vol. 28, No. 4, pp. 458–462.) Experimental results on ZnS-Cu phosphors prepared by several different methods and excited by ultraviolet radiation indicate that the Cu is in a dispersed state in the phosphor.

535.37 : 546.482.21 **135**

**An Experimental Contribution on the Problem of Field Breakdown of CdS Single Crystals.**—K. W. Böer & U. Kümmel. (*Ann. Phys., Lpz.*, 15th July 1955, Vol. 16, Nos. 3/4, pp. 181–191.)

535.372 **136**

**A Comparison of the Luminescence of Calcium Silicate (Mn, Pb) and Zinc-Beryllium Silicate (Mn).**—H. Dziergwa & H. Lange. (*Z. Phys.*, 20th April 1955, Vol. 140, No. 4, pp. 359–369.) Fundamental differences between the emission spectra of the two materials at temperatures between − 180° and + 400°C and between their decay characteristics are discussed.

535.376 **137**

**Brightness Waves in the Luminescence of Phosphors excited by Alternating Electric Fields.**—H. Gobrecht, D. Hahn & F. W. Seemann. (*Z. Phys.*, 20th April 1955, Vol. 140, No. 4, pp. 432–439.) C.r.o. traces are shown of the brightness variations of electroluminescent phosphors. These exhibit characteristic differences for different phosphors but are not greatly influenced by the frequency, field strength or binding dielectric. The variations are more advanced in phase with respect to the voltage waveform than those of a glow discharge. The asymmetry of the waveforms with increased layer thickness indicates an additional surface effect at the anode.

535.376 : 539.23 **138**

**Field-Enhanced Cathodoluminescence in Magnesium Oxide.**—J. Woods & D. A. Wright. (*Proc. phys. Soc.*, 1st Aug. 1955, Vol. 68, No. 428B, pp. 566–569.)

537.226/.227 **139**

**On Interactions among Ions of a $BaTiO_3$ Crystal and on its 180° and 90° Type Domain Boundaries.**—W. Kinase. (*Progr. theor. Phys.*, May 1955, Vol. 13, No. 5, pp. 529–539.)

537.226 : 621.3.029.6 **140**

**Methods for calculating the Dielectric and Magnetic Permeabilities of Artificial Media.**—A. L. Mikaelyan. (*Radiotekhnika, Moscow*, Jan 1955, Vol. 10, No. 1, pp. 23–36.) Artificial dielectrics are considered in which the size of the included particles and the distance between them are much smaller than the operating wavelength, so that the properties of the medium are independent of frequency. Several methods are proposed for calculating the dipole moments of conducting and nonconducting particles; the results are tabulated for different shapes of particle. Media can be produced having a refractive index smaller than that of free space.

537.226.2/.3 : 621.317.335.029.6 **141**

**Determination of Dielectric Properties from the Brewster Angle of Incidence and the Restored Azimuth at Wavelengths below 3 cm.**—Rabenhorst. (See 208.)

537.226.8 : 621.315.616.9 **142**

**Creep-Path Formation in Synthetic Materials.**—K. Schumacher. (*Elektrotech. Z., Edn A*, 1st June 1955, Vol. 76, No. 11, pp. 369–376.) An experimental investigation is reported. Local heating was produced by means of an electrically heated filament stretched near the specimen, across which a potential was applied. The changes in the specimens up to the time of breakdown were recorded.

537.227 : 547.476.3 **143**

**Some Peculiarities of the Polarization of Rochelle Salt exposed to Radioactive Irradiation.**—I. S. Zheludev, M. A. Proskurnin, V. A. Yurin & A. S. Boberkin. (*C. R. Acad. Sci. U.R.S.S.*, 11th July 1955, Vol. 103, No. 2, pp. 207–208. In Russian.) The effect of irradiation at an intensity of $10^{19}$ eV.h$^{-1}$.g$^{-1}$ is shown by a series of hysteresis oscillograms recorded (*a*) at the beginning of the experiment, (*b*) after 30 min irradiation, (*c*) after 1 h, (*d*) after 2 h, and (*e*) after 13 h 30 min. The polarizing field applied was 1 340 V/cm. As the period of irradiation increases, the loop becomes more and more constricted in the middle, the constriction changes into a straight line the length of which increases until the whole characteristic is linear.

537.228.1 : 534.2-8 **144**

**Ultrasonic Velocities in Polarized Barium Titanate Ceramics.**—H. B. Huntington & R. D. Southwick. (*J. acoust. Soc. Amer.*, July 1955, Vol. 27, No. 4, pp. 667–679.) "Measurements have been made of the ultrasonic velocities in polarized barium titanate ceramic blocks for both compressional and transverse waves, parallel and perpendicular to the axis of polarization. Some variation in compressional velocity with the degree of binding was evident. The largest variation between the velocities for transverse waves causing shear amounted to nearly 12 percent and proved to be accounted for by electromechanical coupling. The compressional velocity parallel to the axis of polarization ran some four percent higher than that at right angles. Here again it appeared that the effect was caused principally by differences in electrical boundary conditions and that the constant field elastic constants varied only slightly with polarization."

537.311.33 145

**Semiconductors.**—(*J. Electronics*, Sept. 1955, Vol. 1, No. 2, pp. 103–230.) This issue is devoted to a series of papers presented at a symposium noted previously [3634 of 1955 (Ross)]; the titles are as follows:—

Speculations on the Energy-Band Structure of Zinc-Blende-Type Crystals.—F. Herman. (pp. 103–114.)

Bond Relationships in Diamond-Type Semiconductors.—C. H. L. Goodman. (pp. 115–121.)

The Validity of the Hydrogen-Like Approximation for Impurity Levels.—G. Rickayzen. (pp. 122–125.)

Photovoltaic and Photoconductive Theory applied to InSb.—T. S. Moss. (pp. 126–133.)

The Photoelectromagnetic Effect in Indium Antimonide.—C. Hilsum, D. J. Oliver & G. Rickayzen. (pp. 134–137.)

Some Implications of the Small Energy Gap and Small Effective Electron Mass in InSb.—J. W. Allen & I. M. Mackintosh. (pp. 138–144.)

Measurements of Diffusion Length in Indium Antimonide.—D. G. Avery & D. P. Jenkins. (pp. 145–151.)

Accurate Measurements of Absorption in Indium Antimonide and Gallium Antimonide.—V. Roberts & J. E. Quarrington. (pp. 152–160.)

A Study of the Conduction Band of InSb.—R. Barrie & J. T. Edmond. (pp. 161–170.)

The Electrical Properties of Indium Antimonide at Low Temperatures.—B. V. Rollin & A. D. Petford. (pp. 171–174.)

The Magneto-resistance Effect in Indium Antimonide.—R. Mansfield. (pp. 175–177.)

A Simple Method to visualize Degeneracy of an Intrinsic Semiconductor.—G. Busch. (pp. 178–180.)

Optical and Electrical Properties of GaAs, InP and GaP.—H. Welker. (pp. 181–185.)

On the Degeneracy Effect in InAs.—R. Talley & F. Stern. (pp. 186–189.)

Preparation and Electrical Properties of CdTe Single Crystals.—F. A. Kröger & D. de Nobel. (pp. 190–202.)

The Preparation, Electrical and Optical Properties of $Mg_2Sn$.—W. D. Lawson, S. Nielsen, E. H. Putley & V. Roberts. (pp. 203–211.)

Zone Melting of Decomposing Solids.—J. van den Boomgaard, F. A. Kröger & H. J. Vink. (pp. 212–217.)

Thermoelectric Applications of Semiconductors.—H. J. Goldsmid. (pp. 218–222.)

Applications of Indium Antimonide.—I. M. Ross & E. W. Saker. (pp. 223–230.)

537.311.33 146

**Physical Theory of Semiconductor Surfaces.**—C. G. B. Garrett & W. H. Brattain. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, pp. 376–387.) Properties associated with the space-charge region and with surface states are discussed in the light of experimental evidence. The theory of the space-charge region is fairly precisely established, but knowledge about the exact structure of the surface states under given chemical conditions is still incomplete. Measurements of surface photoeffect are not alone sufficient to fill the gaps; field-effect experiments with controlled chemical environment would probably give useful results.

537.311.33 147

**Measurement of Minority Carrier Lifetimes in Semiconductors.**—N. F. Durrant. (*Proc. phys. Soc.*, 1st Aug. 1955, Vol. 68, No. 428B, pp. 562–563.) A modification of the method proposed by Haynes & Westphal (2233 of 1952) is used in which a short pulse of minority carriers is injected into the specimen and made to drift towards the collector by a sweeping field triggered with variable delay by the emitter pulse. The collected pulse is proportional to the number of carriers surviving. The method permits the detection and elimination of errors due to trapping effects and conductivity modulation and gives results consistent within 5%.

537.311.33 148

**On Lattice Scattering in Homopolar Semiconductors.**—J. M. Radcliffe. (*Proc. phys. Soc.*, 1st Aug. 1955, Vol. 68, No. 428A, pp. 675–680.) A particular solution of the Boltzmann equation for the stationary states of a system subject to electric and magnetic field is obtained, assuming a non-degenerate conduction band of arbitrary shape. The theory is applied in a discussion of the temperature variation of the mobility of holes in Ge and Si.

537.311.33 149

**Scattering of Electrons and Holes by Charged Donors and Acceptors in Semiconductors.**—R. B. Dingle. (*Phil. Mag.*, Aug. 1955, Vol. 46, No. 379, pp. 831–840.) The problem of determining the time of relaxation due to scattering of electrons and holes by charged impurity donors and acceptors in semiconductors is reconsidered and new formulae are derived using an approach which is basically the same as that used by Mott (*Proc. Camb. phil. Soc.*, 1936, Vol. 32, p. 281). Results obtained by these formulae are compared with those derived by other workers.

537.311.33 : 546.28 150

**Optical Properties of Indium-Doped Silicon.**—R. Newman. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, pp. 465–467.) Absorption and photoconduction measurements at low temperatures ($< 77°K$) indicate that the ionization energy of In as an impurity in Si is 0·16 eV. Sharp absorption lines observed at 20°K broaden as the temperature rises and practically disappear at 77°K.

537.311.33 : 546.28 151

**Spin Resonance of Impurity Atoms in Silicon.**—C. P. Slichter. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, pp. 479–480.) "An explanation is proposed for the weak satellite lines observed in the microwave electron spin resonance of Group V impurity atoms in silicon. The satellites occur halfway between the main hyperfine lines. It is proposed that the satellites represent the resonance of pairs of impurities close enough together for their exchange interaction to be very large compared to the hyperfine coupling. Mathematical details are worked out."

537.311.33 : 546.28 : 539.32 152

**The Determination of the Elastic Constants of Silicon by Diffuse X-Ray Reflexions.**—S. C. Prasad & W. A. Wooster. (*Acta Cryst.*, 10th June 1955, Vol. 8, Part 6, p. 361.)

537.311.33 : 546.289 153

**Electrical Properties of Germanium Semiconductors at Low Temperatures.**—H. Fritzsche. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, pp. 406–419.) The low-temperature maximum of Hall coefficient and the saturation of resistivity observed by Hung & Gliessman (1696 of 1955) have been confirmed by measurements over the temperature range 1·5°–300°K, using *n*- and *p*-type crystals with various impurity concentrations. Field effects were excluded, and the anomalous observations were shown not to be due to surface conduction or crystal anisotropy. To explain the results, conduction in at least two energy bands must be assumed, one of which is characterized by very low carrier mobility.

537.311.33 : 546.289 154

**Optical Properties of Plastically Deformed Germanium.**—H. G. Lipson, E. Burstein & P. L. Smith. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, pp. 444–445.) Samples of 1-Ω.cm *n*-type Ge were deformed to various degrees by compression at about 700°C. Resistivity and thermoelectric-power measurements indicated that the larger deformations were accompanied by conversion to *p* type. Transmission measurements over the wavelength range 1–15 $\mu$ indicated that in all cases the deformation gave rise to a shift of the intrinsic absorption edge towards longer wavelengths, together with an extended tail. A theoretical explanation of the changes is indicated briefly.

537.311.33 : 546.289 155

**Hall Theory in *n*-type Germanium.**—L. Gold. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, p. 596.) Brief interim report of a study of the galvanomagnetic properties of *n*-type Ge based on the value of the scattering time.

537.311.33 : 546.289 156

**Some Observations on Growth and Etching of Crystals with the Diamond or Zincblende Structure.**—E. Billig & P. J. Holmes. (*Acta Cryst.*, 10th June 1955, Vol. 8, Part 6, pp. 353–354.) Characteristic features are described of dendritic plates of Ge and other crystals which have been prepared in the form of long thin lamellae.

537.311.33 : 546.289 : 621.314.632 157

**Rectification Properties of Metal/Semiconductor Contacts.**—E. H. Borneman, R. F. Schwarz & J. J. Stickler. (*J. appl. Phys.*, Aug. 1955, Vol. 26, No. 8, pp. 1021–1028.) Experiments were made using *n*- and *p*-type Ge specimens with various metals plated on. No correlation was found between the reverse saturation current densities and the work function of the metal. For the Ge-metal combinations giving the lowest saturation-current densities, calculations indicate that most of the current is carried by holes, the magnitude of this hole current depending primarily on the diode geometry. This is confirmed by the observed variation of the $I/V$ characteristic with Ge resistivity in the case of Ge-In contacts, and by the temperature variation of the zero-voltage conductance. It is inferred that while it is easy to establish a high barrier to electrons flowing on Ge surfaces, no barrier to hole flow was established with any of the metals used. For a given metal there is an upper limit to the electron current, corresponding to the lowest Ge resistivity.

537.311.33 : 546.46-31 158

**Electrical Conductivity induced in MgO Crystals by 1·3-MeV Electron Bombardment.**—M. A. Pomerantz, R. A. Shatas & J. F. Marshall. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, pp. 489–490.)

537.311.33 : 546.561 159

**Effect of Temperature of Oxidation on the Electrical Conductivity of the Cuprous-Oxide/Copper System.**—M. Moldovanova. (*C. R. Acad. Sci. U.R.S.S.*, 11th July 1955, Vol. 103, No. 2, pp. 223–225. In Russian.) The effect of oxidation at temperatures between 980° and 1 040°C was investigated experimentally. Specimens oxidized at 1 000°C were found to have a minimum dissociation energy $u$, and a minimum value of constant $A$ in the expression for specific conductivity $\sigma = A \exp(-u/kT)$. The conductivity passed through a minimum for oxidation temperatures between 1 000° and 1 020°C.

537.311.33 : 546.682.86 160

**Electrical Properties of *p*-Type Indium Antimonide at Low Temperatures.**—H. Fritzsche & K. Lark-Horovitz. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, pp. 400–405.) Measurements over the temperature range 1·5°–370°K indicate anomalies similar to those observed in Ge by Hung & Gliessman (1696 of 1955) for the temperature variation of resistivity and Hall coefficient. The magnetoresistance changes sign from positive to negative as the crystals are cooled through a temperature somewhat lower than that of maximum Hall coefficient; no theory capable of explaining this negative magnetoresistance has yet been advanced.

537.311.33 : 546.682.86 161

**Effect of Pressure on the Electrical Properties of Indium Antimonide.**—D. Long. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, pp. 388–390.) Resistivity and Hall-coefficient measurements were made at pressures up to 2 000 atm and temperatures of 0°, 24·3° and 54·3°C, on a sample which was *i*-type down to about 220°K and *p*-type at lower temperatures. The results indicate that the energy gap increases by $14{\cdot}2 \times 10^{-6}$ eV/atm and the electron mobility decreases by 14% for a pressure increase of 2 000 atm at room temperature.

537.311.33 : 546.682.86 162

**Effect of Pressure on the Electrical Conductivity of InSb.**—R. W. Keyes. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, pp. 490–495.) Measurements have been made over the temperature range $-78°$ to $+300°$C at pressures up to 12 000 kg/cm². The activation energy increases with pressure at a rate $15{\cdot}5 \times 10^{-6}$ eV/(kg/cm²). The electron mobility is approximately inversely proportional to the activation energy, and the hole mobility is independent of pressure. Conclusions are drawn regarding the energy-band structure.

537.311.33 : 546.682.86 163

**Magnetic Susceptibility of Indium Antimonide.**—D. K. Stevens & J. H. Crawford, Jr. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, pp. 487–488.) Measurements of the susceptibility of *n*- and *p*-type samples over the temperature range 65°–650°K are reported. The value deduced for the energy gap at 0°K is 0·262 eV, with an effective electron mass of $0{\cdot}028\ m_0$.

537.311.33 : 549.351.12 164

**Electrical Characteristics of Chalcopyrite. The Effect of Surface Treatment on the Rectifying Properties of Crystals.**—B. I. Boltaks & N. N. Tarnovski. (*Zh. tekh. Fiz.*, March 1955, Vol. 25, No. 3, pp. 402–409.) Experiments were conducted over a wide range of temperatures to determine the electric and thermoelectric properties of natural and synthetic $CuFeS_2$. The temperature dependence of the electrical conductivity of these materials is of a more complex character than in ordinary impurity semiconductors. By etching the surface it is possible considerably to improve the rectifying properties of the crystals and in a number of cases to obtain static characteristics not inferior to those of Ge and Si detectors.

537.311.33 : 621.314.634 165

**The Effectiveness of Cadmium Selenide Layers in Selenium Rectifiers.**—H. Strœsche. (*Z. Phys.*, 20th April 1955, Vol. 140, No. 4, pp. 409–413.) To determine the lowest useful thickness for the intermediate CdSe layer formed in a Se rectifier with a counter electrode containing Cd, measurements were made of the voltage/current characteristics of samples with CdSe or Cd layers evaporated on to a Se-coated plate and with a Bi counter electrode. No deleterious effects were noticed until the layer thickness dropped below about $10^{-7}$ cm. If the coating is dense, a monomolecular thickness is sufficient to maintain the rectifying properties.

537.323 **166**

**Thermoelectric Properties of Alloys of the Bismuth-Tellurium System.**—F. I. Vasenin. (*Zh. tekh. Fiz.*, March 1955, Vol. 25, No. 3, pp. 397–401.) Application of powder-metallurgy pressing and sintering technique enabled much higher values of thermo-e.m.f. to be obtained than hitherto. It was established experimentally that the inversion of the sign of the thermo-e.m.f. in pressed and sintered alloys of the system occurs for compositions in close proximity (within 0·2%) to the compound $Bi_2Te_3$. It is shown that the theory of Telkes (*J. appl. Phys.*, 1947, Vol. 18, No. 12, pp. 1116–1127) that there exists a second compound in the system with negative thermo-e.m.f., is erroneous.

537.533 **167**

**Photoelectric Work Functions of the Borides of Lanthanum, Praseodymium, and Neodymium.**—R. W. Decker & D. W. Stebbins. (*J. appl. Phys.*, Aug. 1955, Vol. 26, No. 8, pp. 1004–1006.) Using the DuBridge method of determining the photoelectric threshold, the values 2·74 eV, 3·12 eV and 4·57 eV were found for the work function of $LaB_6$, $PrB_6$ and $NdB_6$ respectively.

538.22 : 546.3-1-71 1-621-26 **168**

**The Structure and Magnetic Properties of the Alloy $Mn_3AlC$.**—R. G. Butters & H. P. Myers. (*Phil. Mag.*, Aug. 1955, Vol. 46, No. 379, pp. 895–902.) In the as-cast state an alloy of composition close to $Mn_3AlC$ is feebly magnetic at room temperature and the magnetic property is retained after homogenization at 1 000°C. The ordered structure is similar to that of $Mn_3ZnC$ (2333 of 1955). The alloy $Mn_3AlC$ is strongly magnetic at low temperatures; the variation of the paramagnetic susceptibility over the range 65°–800°C is different from that of a normal ferromagnetic material.

538.221 **169**

**Domain Configurations and Crystallographic Orientation in Grain-Oriented Silicon Steel.**—W. S. Paxton & T. G. Nilan. (*J. appl. Phys.*, Aug. 1955, Vol. 26, No. 8, pp. 994–1000.) The domain structure was investigated by powder-pattern technique and the orientation of the individual domains was determined by the etch-pit optical-goniometer technique. The relations between the two sets of observations are discussed and some conclusions are drawn regarding the three-dimensional domain configuration.

538.221 **170**

**Dependence of the Coercive Force on the Density of Some Iron Oxide Powders.**—A. H. Morrish & S. P. Yu. (*J. appl. Phys.*, Aug. 1955, Vol. 26, No. 8, pp. 1049–1055.) $Fe_3O_4$ and $\gamma$-$Fe_2O_3$ are investigated. Calculations are made of the critical size below which the particle comprises a single domain; the value is about $3\mu$ for the major axis of a prolate spheroid of $Fe_3O_4$ with ratio of major to minor axis 10 : 1, and slightly greater for $\gamma$-$Fe_2O_3$. The observed variation of coercive force with size and shape of particles is consistent with the calculated value of critical size, but the absolute value of the coercive force of the single-domain particles is much smaller than predicted by theory.

538.221 **171**

**Magnetic Transformation in MnBi.**—R. R. Heikes. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, pp. 446–447.) "Manganese bismuth loses its spontaneous magnetization very sharply at 633°K. At the same temperature drastic changes occur in the lattice constants. Previous workers have interpreted both these phenomena as arising from a ferromagnetic-antiferromagnetic transition. The present work indicates that the phase existing above 633°K is actually a ferromagnetic one with a Curie temperature of about 470°K."

538.221 **172**

**Multiple Resonances in Cobalt Ferrite.**—P. E. Tannenwald. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, pp. 463–464.) "Double microwave resonances have been observed in cobalt ferrite when a magnetic field was applied in crystal directions near the hard axis. One resonance occurs before the magnetization vector is lined up with the external magnetic field and another occurs in the aligned state. The experimental results exhibit characteristic features expected before complete line-up and yield, at 90°C, $K_1/M = 2\,200 \pm 200$ and $g = 2{\cdot}7 \pm 0{\cdot}3$."

538.221 **173**

**Electrical Conductivity and Seebeck Effect in $Ni_{0{\cdot}80}Fe_{2{\cdot}20}O_4$.**—F. J. Morin & T. H. Geballe. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, pp. 467–468.) Measurements over a temperature range from about 100° to 350°K indicate thermal hysteresis in the conductivity but not in the Seebeck-effect variation. This suggests that the hysteresis involves the charge-carrier transfer process but not the production of carriers. The activation energy associated with the transfer process is estimated to be 0·10 eV in the high-temperature state and 0·06 eV in the low-temperature state.

538.221 **174**

**Electronic Energy Bands in Iron.**—J. Callaway. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, pp. 500–509.)

538.221 **175**

**Ferromagnetism in Noble-Metal Alloys containing Manganese.**—F. A. Otter, Jr, P. J. Flanders & E. Klokholm. (*Phys. Rev.*, 15th July 1955, Vol. 99, No. 2, pp. 599–600.)

538.221 **176**

**A Study of Domain Structures in Alnico.**—L. F. Bates & D. H. Martin. (*Proc. phys. Soc.*, 1st Aug. 1955, Vol. 68, No. 428B, pp. 537–540.)

538.221 **177**

**A Theory of the Uniaxial Ferromagnetic Anisotropy induced by Magnetic Annealing in Cubic Solid Solutions.**—S. Taniguchi. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. A*, June 1955, Vol. 7, No. 3, pp. 269–281.) A theory is developed on the basis of van Vleck's model. The results obtained are similar to those of Néel (2983 of 1954).

538.221 **178**

**Magnetic Behaviour of Fe-Ni Compacts under Sintering Process.**—H. Kojima. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. A*, June 1955, Vol. 7, No. 3, pp. 321–328.) The magnetic properties and density variations of specimens containing between 45% and 85% Ni were investigated as functions of the sintering temperature between 600° and 1 100°C. Results are presented graphically.

538.221 **179**

**Demonstration of Magnetization-Reversal Starting Centres in Iron-Nickel Wires under Tension, using the Bitter-Pattern Method.**—C. Greiner. (*Ann. Phys., Lpz.*, 15th July 1955, Vol. 16, Nos. 3/4, pp. 176–180.)

538.221 : 539.23 **180**

**Energy of the Bloch Walls in Thin** [spontaneously magnetized] **Layers.**—L. Néel. (*C. R. Acad. Sci., Paris*, 8th Aug. 1955, Vol. 241, No. 6, pp. 533–536.) It is shown that as the thickness of the layer is reduced the energy of the Bloch walls first increases, then passes through a maximum and regains its usual value when the layer becomes extremely thin.

538.221 : 539.234 **181**

**Preparation of Thin Magnetic Films and their Properties.**—M. S. Blois, Jr. (*J. appl. Phys.*, Aug. 1955, Vol. 26, No. 8, pp. 975–980.) Evaporation technique is described for producing films of ferromagnetic material, particularly Ni-Fe alloys, with thicknesses from 1 000 Å to 100 000 Å and coercivities of 0·1–40 oersted. Magnetic orientation is used to produce a rectangular hysteresis loop. Questions of suitable substrate materials, evaporable dielectrics for use in insulating multilayer films, and control of alloy composition are discussed.

538.221 : 621.3.042.14 **182**

**Compressional Stress on Magnetic Laminations.**—R. E. Fischell. (*Elect. Engng, N.Y.*, July 1955, Vol. 74, No. 7, p. 585.) Digest of paper to be published in *Trans. Amer. Inst. elect. Engrs, Communication and Electronics*, 1955. Report of an experimental investigation of the effect of stress, such as that produced by clamping, on the magnetic properties of laminations of various materials.

538.221 : 621.318.134 **183**

**Ferromagnetic Resonance of Gadolinium Ferrites as a Function of Temperature at 9 000 Mc/s.**—J. Paulevé. (*C. R. Acad. Sci., Paris*, 8th Aug. 1955, Vol. 241, No. 6, pp. 548–550.) The value of the field strength $H$ corresponding to resonance becomes very low in the neighbourhood of 306°K, the compensation temperature. The results of measurements are consistent with the two-sublattice theory of rare-earth ferrites developed by Néel (3275 of 1954); the discontinuity in the value of $H$ is due to reversal of the magnetization of one of the sublattices near the compensation point.

538.221 : 621.318.134 **184**

**Dilatometer Thermal Analysis of Samarium Ferrite and Magnetic Transitions of Rare-Earth Ferrites.**—G. Guiot-Guillain & X. Waché. (*C. R. Acad. Sci., Paris*, 8th Aug. 1955, Vol. 241, No. 6, pp. 550–552.)

538.221 : 621.318.134 **185**

**Resonance Field Strength for a System of Magnetic Sublattices (Ferrimagnetic Resonance.)**—B. Dreyfus. (*C. R. Acad. Sci., Paris*, 8th Aug. 1955, Vol. 241, No. 6, pp. 552–554.) Formulae are presented facilitating calculation of the resonance field strength for ferrites.

538.222/.224 **186**

**Some Remarks on S. G. Salikhov's Work.**—N. N. Neprimerov. (*Zh. eksp. teor. Fiz.*, June 1955, Vol. 28, No. 6, pp. 719–724.) Detailed critical discussion of recently published papers by Salikhov (e.g. 1833 of 1954) on paramagnetic microwave resonance absorption in metals.

546.3-1-681 **187**

**Composition of Binary and Multiple Compounds of** [sub-group] **B Metals: Part 1—Systems comprising Gallium with** [sub-group] **B Metals.**—H. Spengler. (*Z. Metallkde*, June 1955, Vol. 46, No. 6, pp. 464–467.)

546.3-1-682 **188**

**Indium Alloys.**—S. Valentiner. (*Z. Metallkde*, June 1955, Vol. 46, No. 6, pp. 442–449.) A comprehensive review of research on In alloys, having regard to the position of In in the periodic system. Over 86 references.

621.315.61 **189**

**Breakdown of Solid Dielectrics in Divergent Fields.**—J. H. Mason. (*Proc. Instn elect. Engrs*, Part C, Sept. 1955, Vol. 102, No. 2, pp. 254–263. Digest, *ibid.*, Part B, Sept. 1955, Vol. 102, No. 5, pp. 725–727.) Experiments with electrode systems in which one electrode is a needle point embedded in polythene or polyisobutylene are reported. Dependence of breakdown stress on the polarity of the point is discussed in relation to the stress-enhanced conductivity and space-charge accumulation. The electric strength of various insulators under industrial conditions is discussed in relation to test data.

621.315.612 **190**

**Structure in Vitreous Silicates as shown by Low-Angle Scattering of X-Rays.**—L. C. Hoffman & W. O. Statton. (*Nature, Lond.*, 17th Sept. 1955, Vol. 176, No. 4481, pp. 561–562.)

621.315.612.4 **191**

**Solid-State Reactions and Dielectric Properties in the System Magnesia Lime Tin-Oxide Titania.**—L. W. Coughanour, R. S. Roth, S. Marzullo & F. E. Sennett. (*J. Res. nat. Bur. Stand.*, March 1955, Vol. 54, No. 3, pp. 149–162.) A detailed study is reported of the solid-state reactions occurring among the oxide components in this system and of the effects of composition and temperature on the dielectric properties; the results are presented in numerous tables and graphs.

621.318.134.002.2 **192**

**Magnetic Ceramics: Part 1—General Methods of Magnetic Ferrite Preparation.**—G. Economos. (*J. Amer. ceram. Soc.*, 1st July 1955, Vol. 38, No. 7, pp. 241–244.) Various methods with commercial possibilities are discussed.

666 : 621.791.3 **193**

**Review of High-Temperature Metal-Ceramic Seals.**—H. Palmour, III. (*J. electrochem. Soc.*, July 1955, Vol. 102, No. 7, pp. 160C–164C.) Emphasis is on electronic applications. Practical designs are illustrated and properties of several ceramics and glasses are listed.

## MATHEMATICS

512.3 **194**

**Solution of the Cubic Equation.**—E. Astuni. (*Ricerca sci.*, July 1955, Vol. 25, No. 7, pp. 2048–2068.) A method derived from the analysis for coupled circuits under transient conditions (3189 of 1955). Numerical examples are given.

517.512.2 **195**

**Convergence and Summation of a Fourier Series corresponding to an Analytical Function.**—R. Cazenave. (*Ann. Télécommun.*, May 1955, Vol. 10, No. 5, pp. 102–108.)

519.2 **196**

**The Analysis of Non-Markovian Stochastic Processes by the Inclusion of Supplementary Variables.**—D. R. Cox. (*Proc. Camb. phil. Soc.*, July 1955, Vol. 51, Part 3, pp. 433–441.)

519.2 **197**

**The Probability Distribution of the Amplitude of a Constant Vector plus a Rayleigh-Distributed Vector.**—K. A. Norton, L. E. Vogler, W. V. Mansfield & P. J. Short. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, pp. 1354–1361.) It is emphasized that both amplitude distribution and phase distribution are required to describe a Rayleigh-distributed vector. A summary is presented of physical conditions which must be satisfied for a given phenomenon to exhibit statistical properties of a Rayleigh-distributed vector. The application of the theory to various aspects of radio propagation is indicated.

517 **198**

**Reihenentwicklungen in der mathematischen Physik.** [Book Review]—J. Lense. Publishers: de Gruyter, Berlin, 1953, 216 pp., D.M. 26. (*Proc. phys. Soc.*, 1st Aug. 1955, Vol. 68, No. 428A, pp. 764–765.) An account of the use of expansions in terms of orthogonal functions, and of related techniques.

517.564.3 : 518.2 **199**

**Bessel Functions of Integer Orders and Large Arguments.** [Book Review]—L. Fox. Publishers: University Press, Cambridge, 1954, 28 pp., 6s. 6d. (*Sci. Progr.*, July 1955, Vol. 43, No. 171, p. 526.) Royal Society Shorter Mathematical Tables, No. 3. Auxiliary functions are given permitting computation of $J_n(x)$, $Y_n(x)$, $I_n(x)$, $K_n(x)$ for integral values of $n$ from 0 to 20 and $x > 20$, using in addition merely tables of trigonometric or exponential functions.

## MEASUREMENTS AND TEST GEAR

531.761 **200**

**New Definition of the Unit of Time and Frequency.**—L. Essen. (*Wireless Engr*, Nov. 1955, Vol. 32, No. 11, p. 312.) At the 9th General Assembly of the International Astronomical Union, 1955, the second was defined as 1/31556925·975 of the tropical year for 1900. Some implications of this decision are briefly indicated.

621.317.3 : 537.311.33 **201**

**Apparatus for measuring Resistivity and Hall Coefficient of Semiconductors.**—T. M. Dauphinee & E. Mooser. (*Rev. sci. Instrum.*, July 1955, Vol. 26, No. 7, pp. 660–664.) Use of the isolating potential-comparator circuit [3064 of 1953 (Dauphinee)] secures the elimination of errors due to thermal e.m.f.s. while permitting the use of d.c. measuring apparatus. Contact and lead resistances do not affect the results.

621.317.3 : 621.372.412 **202**

**Spectrum Analyzer for Quartz Crystals.**—T. E. McDuffie. (*Electronics*, Oct. 1955, Vol. 28, No. 10, pp. 160–162.) The output from a frequency-sweep oscillator having a sweep range variable up to ± 300 kc/s from a 20 Mc/s mean frequency is mixed with a selected signal from a 20·5–40-Mc/s oscillator. The crystal under test is included in an amplifier circuit fed by the difference-frequency signal, and the detected output is displayed on a c.r.o.

621.317.32 : 621.314.63 **203**

**Measurement of [voltage] Ratios using Dry Rectifiers as Voltage-Dependent Resistors.**—H. Matusche. (*Frequenz*, July 1955, Vol. 9, No. 7, pp. 232–234.) Theory and experimental results are given for circuits based on a voltage divider comprising a linear resistor in series with the rectifier.

621.317.32 : 621.385.832 **204**

**Method of measuring the Screen Potential of Cathode-Ray Tubes.**—J. R. Young. (*Rev. sci. Instrum.*, July 1955, Vol. 26, No. 7, pp. 647–648.) The Kelvin vibrating-capacitor method is described. A small Al disk arranged parallel to the screen and earthed through a resistor is made to vibrate at mains frequency. When the screen and the disk are at different potentials a.c. flows in the resistor. In typical experiments the screen potential could be determined to within about 5V.

621.317.331 **205**

**Residual Effects in the Campbell Bridge Method for the Absolute Measurement of Resistance, and a Note on a New Bridge.**—N. F. Astbury. (*Proc. Instn elect. Engrs*, Part C, Sept. 1955, Vol. 102, No. 2, pp. 279–283. Digest, *ibid.*, Part B, Sept. 1955, Vol. 102, No. 5, pp. 729–730.) Systematic errors associated with coil reversals are shown to be due to uncompensated eddy-current coupling between supply and detector meshes. In the proposed modified bridge this effect can be eliminated without external compensation if three mutual impedances, two of which are nearly equal, are determined instead of two.

621.317.332 **206**

**Corrections for Line Losses in [microwave] Impedance Measurements.**—F. Mayer. (*Ann. Télécommun.*, May 1955, Vol. 10, No. 5, pp. 109–111.) A formula is derived giving the true s.w.r. as a function of the measured s.w.r. and the attenuation constant of the line. A nomogram is hence developed, and its use is illustrated by examples.

621.317.335.029.6 **207**

**Microwave Measurements on Low-Loss Dielectrics.**—D. Missio. (*Alta Frequenza*, June 1955, Vol. 24, No. 3, pp. 219–237.) Report of measurements made on various dielectrics, at a frequency of about 3 kMc/s and normal temperature, using the method of standing-wave measurements in a short-circuited coaxial line. Various corrections are included in order to improve the accuracy of determination of losses. Results are tabulated and discussed.

621.317.335.029.6 : 537.226.2/.3 **208**

**Determination of Dielectric Properties from the Brewster Angle of Incidence and the Restored Azimuth at Wavelengths below 3 cm.**—H. Rabenhorst. (*Ann. Phys., Lpz.*, 15th July 1955, Vol. 16, Nos. 3/4, pp. 163–175.) The refractive index and the absorption coefficient are determined from (*a*) the Brewster angle and (*b*) the ratio of intensities of the reflected-wave components parallel and perpendicular to the plane of incidence. The incident wave is plane polarized at 45° to the plane of incidence. From (*a*) and (*b*) the dielectric characteristics can be calculated using the Fresnel relations. Measurements were made at 2·52 cm $\lambda$ on trolitul, glass, phenol and calcium sulphate. The apparatus is described.

621.317.335.3 : 621.372.413 : 537.562 **209**

**Resonant-Cavity Measurements of the Relative Permittivity of a D.C. Discharge.**—Foulds. (See 80.)

621.317.335.3.029.6 **210**

**A Method for the Determination of the Dielectric Constant of Loss-Free Solid Bodies in Uniform Transmission Lines.**—H. Lueg & F. Gemmel. (*Arch. elekt. Übertragung*, July 1955, Vol. 9, No. 7, pp. 307–310.) Measurements at dm $\lambda$ and cm $\lambda$ are considered. The test specimen is shaped so as accurately to fill the cross-section of a coaxial line or waveguide, the system then constituting a line-type transformer. The dielectric constant of the specimen is found from the ratio of the characteristic impedances of the line with and without the specimen, and this ratio is determined from the transformer characteristic without requiring precise knowledge of the position of the line short-circuit or of the specimen.

621.317.382 : 621.373.43 **211**

**A Method for determining the Power of a Pulse Generator.**—G. F. Novozhenov. (*Radiotekhnika, Moscow*, March 1955, Vol. 10, No. 3, pp. 33–40.) The peak power of a pulse generator can be calculated from the mean power measured by a wattmeter and the 'energy coefficient' of pulses, which is the ratio of the actual pulse energy to the product of pulse period and peak power. This coefficient can be determined from observations of pulses on the screen of a c.r.o. Since these are voltage rather than power pulses, it is necessary to introduce correction factors, and these are calculated for sinusoidal, bell-shaped, trapezoidal and exponential-rise pulses. Curves of the correction factors are also given.

621.317.4 **212**

**Measurement of the D.C. Magnetization Curve and of the Hysteresis Loop.**—H. Neumann. (*Arch. tech. Messen*, May 1955, No. 232, pp. 105–108.) A brief survey of practical methods.

621.317.4 **213**

**A New Astatic Magnetometer for Measurements on Ferromagnetic Sheet in Direct and Alternating Fields.**—K. Mellentin & H. Lange. (*Z. Metallkde*, June 1955, Vol. 46, No. 6, pp. 450–456.) The main features of the instrument described are (*a*) use of only one magnetizing coil, and (*b*) the particular arrangement of the system in relation to the specimen. The shape of the latter is such that it becomes uniformly magnetized in a uniform field. Loss measurements can be made very rapidly.

621.317.733 : 621.372.542.2 **214**

**A General-Purpose 50-c/s Null Indicator, particularly Suitable for Use with High-Voltage Bridges.**—W. P. Baker. (*J. sci. Instrum.*, July 1955, Vol. 32, No. 7, pp. 268–269.) A tapered low-pass *RC* filter, in which every third passive resistance element is replaced by the anode impedance of an amplifying valve, forms the basis of an indicator having a sensitivity of 10–20 $\mu$V and a dynamic range of 80 dB, with third-harmonic attenuation 130 : 1.

621.317.742 : 621.372.8 **215**

**A Direct-Reading Waveguide Standing-Wave Detector for Use at Low Power Levels.**—H. V. Shurmer. (*Proc. Instn elect. Engrs*, Part C, Sept. 1955, Vol. 102, No. 2, pp. 176–180. Digest, *ibid.*, Part B, July 1955, Vol. 102, No. 4, p. 563.) A directional coupler is used to measure reflected power from the load, which is a function of its v.s.w.r. Errors introduced by the coupler are compensated. The main application is to measurements on crystal rectifiers at 3·2 cm $\lambda$, at power levels from 5 to 650 $\mu$W.

621.317.755 **216**

**Electronically Controlled Electromagnetic Scanning System for High-Voltage Cathode-Ray Oscillograph.**—J. Lagasse & J. Clot. (*C. R. Acad. Sci., Paris*, 17th Aug. 1955, Vol. 241, No. 7, pp. 598–600.) The scanning system for a multibeam c.r.o. [1624 of 1954 (Fert et al.)] is discussed. The scanning current is obtained by discharging a capacitance through the scanning coil via a vacuum triode controlled by pulses applied to the grid. Synchronization of blanking and scanning is performed by means of an auxiliary thyratron which operates a delaying circuit on receipt of the primary triggering pulse. Photographic records can be obtained corresponding to a trace speed of 70 mm/$\mu$s.

621.317.755 : 621.396.41.029.62 **217**

**An Instrument for the Measurement and Display of V.H.F. Network Characteristics.**—J. S. Whyte. (*P. O. elect. Engrs' J.*, July 1955, Vol. 48, Part 2, pp. 81–83.) Group-delay/frequency and gain/frequency characteristics are displayed on a c.r.o. A discrimination of $10^{-9}$ sec is obtainable over a bandwidth of 20 Mc/s centred on 60 Mc/s. Gain/frequency characteristics can be measured with a discrimination of about 0·1 dB.

621.317.76 : 621.395.625 **218**

**Measurement of "Wow" and "Flutter".**—Dzierzynski. (See 13.)

621.317.761.029.6 **219**

**High-Precision [cavity-] Resonator Frequency Meter for Decimetre Wavelengths.**—H. Groll & G. Pusch. (*Fernmeldetech. Z.*, Aug. 1955, Vol. 8, No. 8, pp. 462–466.)

621.317.791 **220**

**A Quasistatic Multipurpose Meter for Direct and Alternating Voltages.**—K. Müller-Lübeck. (*Arch. tech. Messen*, May 1955, No. 232, pp. 113–116.) The universal instrument described, which also includes a facility for capacitance measurements between 50 pF and 20 $\mu$F, is based on a modified cathode-follower stage. Some of the characteristics of the instrument are presented graphically.

621.317.794 : 621.396.822 **221**

**The Sensitivity of Radiometers.**—F. V. Bunkin & N. V. Karlov. (*Zh. tekh. Fiz.*, March & April 1955, Vol. 25, Nos. 3 & 4, pp. 430–435 & 733–741.) A precise definition of the sensitivity of a radiometer is given and the value is calculated for a compensation-type radiometer, taking into account the fluctuation of the amplification factor at high frequency. Similar calculations are carried out for a radiometer of the modulation type. In both cases particular attention is given to the effect of the width of the band received.

621.317.794 : 621.396.822 **222**

**Null Method of measuring Weak Electrical Fluctuations.**—V. S. Troitski. (*Zh. tekh. Fiz.*, March 1955, Vol. 25, No. 3, pp. 478–496.) In this method the amplifier is used only as a null indicator, hence errors due to variations of the amplification factor are completely eliminated. The theory of the method is discussed in application to long and medium waves, and conditions of balance are established for both linear and quadratic detectors. The errors due to various characteristics of the circuit used, as well as to noise in the apparatus itself are discussed in detail.

621.373.1.029.42 **223**

**A Sine-Wave Generator for Very Low Frequencies.**—W. A. Penton. (*J. sci. Instrum.*, July 1955, Vol. 32, No. 7, pp. 282–283.) Two probes rotating continuously in an electrolyte cell, across which there is a potential difference, generate a sinusoidal voltage with frequency range 0·1–2·0 c/s, and output level variable between 2·5 and 60 V peak to peak. The generator has been used in measurements on high-pass filters to cut off below 0·5 c/s.

## OTHER APPLICATIONS OF RADIO AND ELECTRONICS

413 : 681.142 **224**

**Use of a Computing Machine as a Mechanical Dictionary.**—Booth. (See 26.)

531.768 : 546.431.823-31 **225**

**Some Developments in Vibration Measurement.**—S. Edelman, E. Jones & E. R. Smith. (*J. acoust. Soc. Amer.*, July 1955, Vol. 27, No. 4, pp. 728–734.) Two types of $BaTiO_3$ accelerometer and associated equipment and technique developed at the National Bureau of Standards are described.

534.6-8 **226**

**Microacoustic Interferometer using 30-Mc/s Pulses.**—R. A. McConnell & W. F. Mruk. (*J. acoust. Soc. Amer.*, July 1955, Vol. 27, No. 4, pp. 672–676.) An instrument for determining ultrasonic velocity and absorption in liquids uses ultrasonic pulses of duration about 10 $\mu$s and frequency 30 Mc/s, and detects the reflection from a layer of the test material whose thickness is varied to pass through a resonance condition.

621.3.078/.079 : 681.142 **227**

**Industrial Uses of Special-Purpose Computers.**—Kuhnel. (See 27.)

621.317.39 : 531.71 **228**

**Single-Tube Capacitance Transducer Circuit.**—L. Fleming. (*Electronics*, Oct. 1955, Vol. 28, No. 10, pp. 182..190.) A capacitance probe with a resistor in series is connected across the tuned circuit of a r.f. oscillator to give a simple and stable transducer for measurement of mechanical displacements.

621.384.611 **229**

**A Variable-Energy Cyclotron.**—D. E. Caro, L. H. Martin & J. L. Rouse. (*Aust. J. Phys.*, June 1955, Vol. 8, No. 2, pp. 306–309.) Brief description of a cyclotron designed to accelerate protons in the range 2–12·5 MeV, deuterons in the range 4–6·3 MeV, and $\alpha$ particles in the range 8–12·5 MeV.

621.384.622.2 **230**

**The Theory of Electron Beam Loading in Linear Accelerators.**—G. Saxon. (*Metrop. Vick. Gaz.*, July 1955, Vol. 26, No. 432, pp. 219–225.) "An analysis is made of the waveguide circuit of a linear accelerator of electrons employing radio-frequency power feedback to show how the power flowing into such an accelerator varies with the beam loading."

## PROPAGATION OF WAVES

538.566 : 551.510.535 **231**

**Notes on the Propagation of Long-Wave Lightning Atmospherics around the Earth.**—H. G. Stäblein. (*Z. angew. Phys.*, June 1955, Vol. 7, No. 6, pp. 290–291.) The discussion presented by Schumann (528 of 1955) is supplemented by an analysis of the propagation of a signal represented by exponential time functions.

538.566.029.43 **232**

**The Influence of the Earth's Magnetic Field on the Propagation of Very Long Electrical Waves.**—W. O. Schumann. (*Z. angew. Phys.*, June 1955, Vol. 7, No. 6, pp. 284–290.) Analysis is presented in which the ionosphere is regarded as an anisotropic poor conductor. Ordinary and extraordinary wave components penetrate the ionosphere travelling normal to the boundary, with resulting attenuation of and interference with the components travelling in the waveguide formed by the earth and the ionosphere. To satisfy the boundary conditions, both an E mode and an H mode must be propagated in the guide; an original E mode gives rise to an auxiliary H mode, and vice versa. The ratio of the horizontal components at the earth's surface of magnetic field strength perpendicular to and parallel to the propagation direction is a measure of the relative intensities of the E and H modes. A short account of the work is given in *Naturwissenschaften*, Feb. 1955, Vol. 42, No. 4, pp. 91–92.

621.396.11 **233**

**Radio Communication by Wave Scattering.**—R. L. Smith-Rose. (*Wireless Engr*, Nov. 1955, Vol. 32, No. 11, pp. 287–290.) A brief survey of investigations carried out in America and in Europe; both tropospheric and ionospheric scattering are discussed. Experimental results indicate that over the frequency range from about 25 to 60 Mc/s a signal steady enough for communication with a bandwidth of $\not>$ 3 kc/s, over distances up to about 1 200 miles, is propagated by scattering from the E layer.

621.396.11 **234**

**Scatter Propagation.**—(*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, pp. 1173–1526.) The main part of this issue is devoted to a group of papers together providing a comprehensive survey of the state of knowledge on scatter propagation, including both extended-range ionospheric and beyond-horizon tropospheric propagation. The ionospheric mode is of interest for frequencies from about 25 to 60 Mc/s and path lengths of 600 to 1 200 miles; the tropospheric mode is of interest for frequencies from about 100 Mc/s to at least 10 kMc/s and path lengths of 100 miles upwards. Abstracts of several of the papers are given individually; titles of the others are as follows:—

Characteristics of Beyond-the-Horizon Radio Transmission.—K. Bullington. (pp. 1175–1180.)

Aerodynamical Mechanisms producing Electronic-Density Fluctuations in Turbulent Ionized Layers.—R. M. Gallet. (pp. 1240–1252.)

Some Remarks on Scattering from Eddies.—R. A. Silverman. (pp. 1253–1254.)

Factors affecting Spacing of Radio Terminals in a U.H.F. Link.—I. H. Gerks. (pp. 1290–1297.)

Demonstration of Bandwidth Capabilities of Beyond-Horizon Tropospheric Radio Propagation.—W. H. Tidd. (pp. 1297–1299.)

Results of Propagation Test at 505 Mc/s and 4 090 Mc/s on Beyond-Horizon Paths.—K. Bullington, W. J. Inkster & A. L. Durkee. (pp. 1306–1316.)

Some Tropospheric Scatter-Propagation Measurements near the Radio Horizon.—H. B. Janes & P. I. Wells. (pp. 1336–1340.)

The Probability Distribution of the Amplitude of a Constant Vector plus a Rayleigh-Distributed Vector.—K. A. Norton, L. E. Vogler, W. V. Mansfield & P. J. Short. (pp. 1354–1361.) (See 197.)

Trans-horizon Microwave Propagation over Hilly Terrain.—Y. Kurihara. (pp. 1362–1368.)

Tropospheric Overwater Measurements far beyond the Radio Horizon.—L. A. Ames, P. Newman & T. F. Rogers. (pp. 1369–1373.)

Note on Scatter Propagation with a Modified Exponential Correlation.—A. D. Wheelon. (pp. 1381–1383.)

Measurements of the Phase of Radio Waves received over Transmission Paths with Electrical Lengths varying as a result of Atmospheric Turbulence.—J. W. Herbstreit & M. C. Thompson. (pp. 1391–1401.)

Phase-Difference Variations in 9 350-Mc/s Radio Signals arriving at Spaced Antennas.—A. P. Deam & B. M. Fannin. (pp. 1402–1404.)

Some Fading Characteristics of Regular V.H.F. Ionospheric Propagation.—G. R. Sugar. (pp. 1432–1436.)

Line-of-Sight Propagation Phenomena: Part 1—Ray Treatment.—R. B. Muchmore & A. D. Wheelon. (pp. 1437–1449.)

Line-of-Sight Propagation Phenomena: Part 2—Scattered Components.—A. D. Wheelon & R. B. Muchmore. (pp. 1450–1458.)

Near-Field Corrections to Line-of-Sight Propagation.—A. D. Wheelon. (pp. 1459–1466.)

Certain Mode Solutions of Forward Scattering by Meteor Trails.—G. H. Keitel. (pp. 1481–1487.)

621.396.11 **235**

**Radio Propagation Prediction considering Scattering Wave on the Earth's Surface.**—K. Miya & S. Kanaya. (*Rep. Ionosphere Res. Japan*, March 1955, Vol. 9, No. 1, pp. 1–15.) M.u.f. predictions for multi-hop communication circuits can take into account scattered reflections at the earth's surface by use of a new 'control-line' method, in which the distance of the control points of the older method is taken as the radius of arcs round transmitter and receiver forming the control lines, the m.u.f.s being then determined in the normal way. By including propagation paths other than great circles m.u.f.s may be increased. A comparison of predictions with experience on several commercial circuits confirms the greater efficiency of the new method and shows that land scatter is of chief importance at frequencies below 20 Mc/s, while sea scatter should also be considered in the range 20–30 Mc/s.

621.396.11 : 523.5 **236**

**The Role of Meteors in Extended-Range V.H.F. Propagation.**—O. G. Villard, Jr, V. R. Eshleman, L. A. Manning & A. M. Peterson. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, pp. 1473–1481.) The main factors influencing the propagation of continuous radio signals by reflection from meteor ionization trails are reviewed. A method is indicated for calculating the required values of system parameters to maintain the received field strength above the background-noise level for 95% of the time. Measurement results indicate that in some cases forward-scatter propagation in the ionosphere is predominantly governed by meteor ionization; in these cases the performance of existing communication circuits could be improved by appropriate redesign of aerials.

621.396.11 : 551.510.52 **237**

**Characteristics of Tropospheric Scattered Fields.**—L. G. Trolese. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, pp. 1300–1305; *Trans. Inst. Radio Engrs*, July 1955, Vol. AP-3, No. 3, pp. 117–122.) Measurements made on cm-$\lambda$ signals transmitted over a 46·3-mile path are compared with values calculated from the theory of Booker & Gordon (1757 of 1950). Tests with a narrow-beam aerial indicate that the scattered field on reception is spread over an angle five to seven times as large as indicated by the theory, assuming that the scale of tropospheric turbulence is large compared to $\lambda$. Fading observations support the view that the phenomenon is due to beating between scattered components whose frequencies differ owing to motion of the scatterers.

621.396.11 : 551.510.52 **238**

**Investigations of Angular Scattering and Multipath Properties of Tropospheric Propagation of Short Radio Waves beyond the Horizon.**—J. H. Chisholm, P. A. Portmann, J. T. deBettencourt & J. F. Roche. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, pp. 1317–1335.) Report of experiments carried out in the N.E. coastal region of the U.S.A., over paths of lengths 161–188 miles, to determine the possibilities of establishing communication circuits at u.h.f. and microwave frequencies. The results indicate that reliable communication with bandwidths of the order of thousands of Mc/s can be established if power of the order of several kilowatts is used together with highly directional aerials; the tropospheric scattering mechanism is directive and multipath delays are less serious than indicated by previous investigations.

621.396.11 : 551.510.52 **239**

**Forward Scattering of Radio Waves by Anisotropic Turbulence.**—H. Staras. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, pp. 1374–1380.) Theory of tropospheric scattering is extended to derive propagation formulae for the case where the scale of turbulence is different in the horizontal and vertical directions. The frequency dependence of the scattered radiation is the same for anisotropic large-scale turbulence as for isotropic. Those radio-system parameters which depend only on the rate of decrease of scattered energy with elevation angle (e.g. vertical correlation function and height gain) are not affected by the anisotropy, while those which depend on the departure from the great-circle plane (e.g. horizontal correlation function) may be substantially affected. The effect on longitudinal correlation function, bandwidth and effective aerial gain is also discussed.

621.396.11 : 551.510.52 **240**

**Propagation of Short Radio Waves in a Normally Stratified Troposphere.**—T. J. Carroll & R. M. Ring. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, pp. 1384–1390.) An explanation of the main features of beyond-horizon propagation is presented on the basis of careful calculations assuming absence of turbulence in the troposphere. Internal reflection effects associated with coherent molecular scattering in the atmosphere must be taken into account.

621.396.11 : 551.510.52 **241**

**Some Applications of the Monthly Median Refractivity Gradient in Tropospheric Propagation.**—B. R. Bean & F. M. Meaney. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, pp. 1419–1431.) Results of an extensive program of transmission-loss measurements at 100 Mc/s in the U.S.A. are found to indicate consistent correlation with $\Delta N$, the difference between the atmospheric refractive index at the earth's surface and that at a height of 1 km at the mid-point of the propagation path. Six-year average values of $\Delta N$ are presented as an aid in predicting annual transmission-loss variations. The observed dependence of transmission loss on $\Delta N$ is much greater than indicated by standard propagation theory.

621.396.11 : 551.510.52 **242**

**The Use of Angular Distance in estimating Transmission Loss and Fading Range for Propagation through a Turbulent Atmosphere over Irregular Terrain.**—K. A. Norton, P. L. Rice & L. E. Vogler. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, pp. 1488–1526.) A comprehensive investigation of beyond-horizon tropospheric propagation; theoretical formulae are developed on the assumption that the energy may be propagated by diffraction or by forward scattering. The formulae involve the 'angular distance', defined as the angle in the great-circle plane between the radio-horizon rays from transmitting and receiving aerials. For fixed linear and angular distances the nature of the terrain has a much larger influence on the diffracted-mode intensity than on the scattered-mode intensity. Comparison of calculated and observed values of transmission loss over 122 paths at frequencies between 66 Mc/s and 1·046 kMc/s provides strong evidence in favour of the Weisskopf-Villars theory of tropospheric forward scatter. Theoretical curves of the average value of the path aerial gain to be expected in tropospheric forward-scatter propagation are presented as a function of the angular distance, the asymmetry factor and the free-space gains of the transmitting and receiving aerials.

621.396.11 : 551.510.535 **243**

**Radio Transmission at V.H.F. by Scattering and Other Processes in the Lower Ionosphere.**—D. K. Bailey, R. Bateman & R. C. Kirby. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, pp. 1181–1231.) The results of a comprehensive investigation extending over four and a half years are presented; the transmissions used were mainly c.w. Solar wave radiation, corpuscular radiation of presumably solar origin, and meteors are recognized as ionizing agents producing different effects on propagation. The dependence of the received signal strength on time of day, season, solar activity, path details, frequency and scatter angle was studied. The height at which the scattering occurred was determined by a pulse method. The results do not fit existing propagation theories entirely satisfactorily. The significance of the aerial patterns for the results is examined. The possibilities of this type of propagation for communication over distances from about 1 000 to 2 300 km are discussed. Notes are appended by two of the authors and by the editor regretting that certain material has been eliminated from the paper on government instructions.

621.396.11 : 551.510.535 **244**

**On the Scattering of Radio Waves by Turbulent Fluctuations of the Atmosphere.**—F. Villars & V. F. Weisskopf. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, pp. 1232–1239.) A theoretical analysis is presented of the mechanism by which v.h.f. signals are transmitted over distances of the order of $10^3$ km. It is found that turbulent mixing at the lower edge of the E layer produces fluctuations of electron density large enough to account for the observed signals. The basic assumptions are the existence of an electron-density gradient $> 10^3/cm^3$ per km and a reasonable level of turbulent activity. Errors in previous work (2747 of 1954) are indicated.

621.396.11 : 551.510.535 **245**

**Investigations of Scattering and Multipath Properties of Ionospheric Propagation at Radio Frequencies exceeding the M.U.F.**—W. G. Abel, J. T. deBettencourt, J. H. Chisholm & J. F. Roche. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, pp. 1255–1268.) Data obtained from measurements covering about 13 000 h are analysed. Paths used were mostly about 1 000 miles long and situated in medium latitudes. The main frequency used was 49·6 Mc/s, but some measurements were also made on 27·8, 22·9 and 21·6 Mc/s. Short-period fading on 49·6 Mc/s exhibited Rayleigh distribution, long-period fading was essentially Gaussian. The median received signal levels were of the order of 100 dB below free-space values and were higher in winter and summer than in spring and autumn; there was some indication of correlation with sunspot activity and of dependence on direction of polarization. Effective scattering heights of 84–94 km were found in December 1954, using a pulse method. Comparison of 49·6-Mc/s and 27·8-Mc/s signals indicates that received power probably varies as the inverse third or fourth power of frequency.

621.396.11.029.4/.51 : 551.510.535 : 621.372.8 **246**

**The Radial Propagation of Electromagnetic Waves between Two [parallel] Conducting Planes (Radial Waveguide).** Kaden.—(See 22.)

621.396.11.029.62 **247**

**Abnormal Radio Propagation December 3, 1954.**—R. F. Jones. (*Met. Mag., Lond.*, July 1955, Vol. 84, No. 997, pp. 225–226.) Abnormally strong signals on 180·4 Mc/s were received at Beddingham, near Lewes, from the B.B.C. station at Sutton Coldfield, distant 147 miles, on the evening of 3rd December 1954. Meteorological records for this period indicate the presence of extraordinarily dry air at levels as low as 2 200 ft, together with a temperature inversion.

621.396.11.029.63 **248**

**Transmission Tests at 530 Mc/s to Distances beyond Optical Range.**—E. Prokott. (*Fernmeldetech. Z.*, Aug. 1955, Vol. 8, No. 8, pp. 430–437.) Tests were carried out over two transmission paths of nearly 200 km in order to determine the practical requirements for satisfactory communication. A 50-W transmitter was used with an aerial system comprising dipole array and reflectors. The effect of meteorological conditions at the horizon point is noted. The value of radio-path attenuation corresponding to the value of receiver input voltage attained or exceeded during 50% of the test time is estimated as 162 dB. This value is compared with previous results based partly on theory and partly on observations. The results show that optical visibility and freedom from obstructions in the first Fresnel zone are not absolutely essential though reliable operation would be difficult at grazing incidence.

621.396.81 **249**

**Long-Period Fading in Medium-Wave Radio Signals.**—M. S. Rao & B. R. Rao. (*Nature, Lond.*, 3rd Sept. 1955, Vol. 176, No. 4479, pp. 459–460.) Fading of 1·42-Mc/s signals observed over the 600-km path between Madras and Waltair during May 1954, between 0700 and 0800 h I.S.T., is illustrated by a typical record. The average period of the variations is about 7 min; calculations give a value in good agreement with this.

621.396.812.3 : 551.510.52 **250**

**The Rate of Fading in Propagation through a Turbulent Atmosphere.**—K. A. Norton, P. L. Rice, H. B. James & A. P. Barsis. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, pp. 1341–1353.) Fading rate is defined as the number of times per minute that the envelope of the received field strength crosses its median level with a positive slope. The value thus obtained is numerically related to the relevant parameters of the propagation medium under conditions generally satisfied in ionospheric and tropospheric propagation; for beyond-horizon propagation these parameters are the location and shape of the scattering volume and the turbulent and drift velocities of the scatterers. Fading-rate data obtained with 92-Mc/s and 1·046-kMc/s signals over transmission paths of lengths 70 to 394 miles are analysed. Fading over within-sight paths is also discussed. See also 2399 of 1955 (Barsis et al.).

## RECEPTION

621.376.23 : 621.396.822 **251**

**Signal/Noise Performance of Multiplier (or Correlation) and Addition (or Integrating) Types of Detector.**—D. G. Tucker. (*Proc. Instn elect. Engrs*, Part C, Sept. 1955, Vol. 102, No. 2, pp. 181–190. Digest, *ibid.*, Part B, July 1955, Vol. 102, No. 4, pp. 558–563.) Discussion is mainly directed to narrow-band pulse systems in which post-detection integration plays little part. The analysis indicates that pulse-to-pulse multiplication is not in general more effective than adding; the slight advantage afforded by the square-law response can be obtained more simply by using square-law rectifiers. When the multiplication is performed before rectification the output contains no direct component in the absence of correlated signals; this is an important advantage when the noise background is varying. Numerical results obtained by various alternative addition and multiplication techniques are tabulated for values of input signal/noise ratio from 0·1 to 4.

621.376.333 : 621.396.62 **252**

**The Ratio Detector.**—K. R. Sturley. (*Wireless World*, Nov. 1955, Vol. 61, No. 11, pp. 532–537.) The working of the ratio detector in f.m. receivers is simply explained.

621.396.62 : 621.396.3 : 621.376.3 **253**

**Automatic Frequency Control in Frequency-Shift Radiotelegraphy.**—G. Bronzi. (*Alta Frequenza*, June 1955, Vol. 24, No. 3, pp. 284–287.) A theoretical discussion indicates that receivers for frequency-shift radiotelegraphy should be provided with a.f.c. of the second local oscillator, in order to keep the required bandwidth low.

621.396.62 : 621.396.812.3 **254**

**Diversity Reception in U.H.F. Long-Range Communications.**—C. L. Mack. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, pp. 1281–1289.) "Several diversity techniques employed in u.h.f. beyond-the-horizon systems are discussed. Field experience is evaluated in terms of equipment reliability, flexibility and performance. A nonswitching parallel combiner has become the standard military diversity circuit for u.h.f. long-range receivers. The circuit is described and analyzed."

621.396.822 : 621.317.794 **255**

**The Sensitivity of Radiometers.**—Bunkin & Karlov. (See 221.)

621.396.822 : 621.317.794 **256**

**Null Method of measuring Weak Electrical Fluctuations.**—Troitski. (See 222.)

## STATIONS AND COMMUNICATION SYSTEMS

621.376.2 + 621.376.4 **257**

**Quadrature Amplitude-Phase Modulation.**—E. M. Vereshchagin. (*Radiotekhnika, Moscow*, March 1955, Vol. 10, No. 3, pp. 72–77.) The energy of sidebands can be increased and the width of the radiated spectrum halved by using the proposed method, in which the carrier is simultaneously subjected to a.m. and ph.m. To do this it is necessary to displace the sidebands so obtained by 90°. Two circuits satisfying the required conditions are considered and the theory of the method is discussed under the following headings: (*a*) frequency spectrum of modulation; (*b*) estimated distortion; (*c*) energy relations and (*d*) effects of interference. Experiments with both circuits gave satisfactory results and confirmed the validity of the theory.

621.376.56 : 621.39 **258**

**Methods of reducing the Effects of Noise and increasing Channel Capacity with Pulse-Code Modulation.**—F. Benz. (*Arch. elekt. Übertragung*, July & Aug. 1955, Vol. 9, Nos. 7 & 8, pp. 299–306 & 381–387.) Noise entering the r.f. part of the system is considered; its effect is reduced in a pulse system because the pulses occupy only a fraction of the total time. The ratio of theoretical channel capacity $C_{th}$ to actual capacity $C$ is a measure of the noise reduction attained. This ratio can be improved by using the bandwidth-economy system previously described (3415 of 1953) in which a.m. is combined with a type of phase modulation termed 'transverse' modulation. Theory based on that of Shannon (1649 of 1949) is used to determine the probability of confusion between the partial signals and hence to determine that arrangement of individual signals which makes $C$ as nearly as possible equal to $C_{th}$. Practical circuits for this purpose are described briefly.

621.39.001.11 **259**

**On the Limiting Transmission Capacity of a Communication System.**—A. A. Kharkevich & E. L. Blokh. (*Radiotekhnika, Moscow*, Feb. 1955, Vol. 10, No. 2, pp. 14–20.) Shannon's geometrical proof of his well-known expression (1) determining the transmission capacity of a communication system (1649 of 1949) is criticized, and a more rigorous proof is developed. The results obtained indicate that expression (1) is valid only at the limit when signal/noise ratio becomes infinite.

621.39.001.11 : 061.3 **260**

**Information Theory.**—(*Wireless World*, Nov. 1955, Vol. 61, No. 11, pp. 545–546.) A review of papers presented at the conference held at the Royal Institution, London, in 1955.

621.39.001.11 : 621.376.5 **261**

**Electrical Pulse Communication Systems: Part 1—Information Theory and Pulse System.**—R. Filipowsky. (*J. Brit. Instn Radio Engrs*, Sept. 1955, Vol. 15, No. 9, pp. 451–467.) A survey including an account of the historical development of pulse systems and information theory. The four main classes of system distinguished are communication, data transmission, location and computers. The main features of a system are source, message encoding, signal formation, multiplexing (train formation), wave formation (carrier modulation), transmission over the medium, reception, decoding, destination; individual steps may be omitted in particular systems. The various steps are described.

621.395.44 + 621.397.242 + 621.315.212 **262**

**The Position of the Italian Coaxial Network at 31st December 1953.**—L. Niccolai. (*Elettrotecnica*, May 1954, Vol. 41, No. 5, pp. 242–250.) The network provides 960 telephone channels in a 4-Mc/s band, together with a television channel a little over 6 Mc/s wide. The average spacing of repeaters is 9·3 km. The cables and repeater characteristics are described.

621.396.41 : 551.510.52 **263**

**U.H.F. Long-Range Communication Systems.**—G. L. Mellen, W. E. Morrow, A. J. Poté, W. H. Radford & J. B. Wiesner. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, pp. 1269–1281.) Aspects of scatter propagation relevant to the design of multichannel systems operating over distances of 200 miles or more beyond the horizon are considered. Details of aerial systems, modulation techniques and f.m. transmitting and receiving equipment are discussed. Space-diversity reception is used to overcome fading. System reliability is defined and its value is calculated for a particular system.

621.396.41 : 621.376.3 **264**

**Multiplexing F.M. Broadcast Transmitters.**—J. H. Bose. (*Electronics*, Oct. 1955, Vol. 28, No. 10, pp. 146–150.) The problems of background noise in auxiliary f.m. subcarrier channels and cross modulation in the multiplier stages of the transmitter have been solved by narrowing the bandwidth of the early multiplication stages of the main channel and by introducing the auxiliary modulation at a late stage. Field tests have proved that the main f.m. carrier can be modulated to a depth of ± 20 kc/s with a 30-kc/s f.m. subcarrier without affecting quality. See also 849 of 1954 (Armstrong & Bose).

621.396.41 : 621.396.65 **265**

**Beyond-Horizon Signals extend Communications.**—J. R. Day. (*Electronics*, Oct. 1955, Vol. 28, No. 10, pp. 122–127.) Klystrons giving a power output of 10 kW and aerial systems giving gains of 40 dB over a single dipole, together with diversity reception, make possible u.h.f. multiplex communication circuits with repeater spacings of 175 miles; propagation is by tropospheric forward scatter.

621.396.97 : 621.315.2 : 621.396.8 **266**

**The Influence of Delay Distortion and Frequency-Band Limitation on the Quality of Broadcast Transmissions.**—E. Belger, E. A. Pavel & H. Rindfleisch. (*Fernmeldetech. Z.*, Aug. 1955, Vol. 8, No. 8, pp. 445–455.) Report of tests carried out in 1950 and 1951 on the Federal German broadcasting line network. Subjective tests indicate that the tolerable delay in the transmission lines is about 8 ms at the top of the a.f. band, 24 ms at 100 c/s and 120 ms at 50 c/s. The transmission band should go up to at least 10 kc/s, and about 12 kc/s for high-quality transmissions. See also 3709 of 1954 (Belger).

## SUBSIDIARY APPARATUS

621-526 : 621.3.015.7 **267**

**The Pulse Transfer Function and its Application to Sampling Servo Systems.**—R. H. Barker. (*Proc. Instn elect. Engrs*, Part C, Sept. 1955, Vol. 102, No. 2, pp. 291–292.) Addendum to 1131 of 1953.

621-526 : 621.373.421 **268**

**The Frequency-Response Analysis of Nonlinear Systems.**—Grensted. (See 56.)

621-526 : 621.396.96 **269**

**The Stability and Time Response of Fast-Operating Closed-Loop Pulsed Radar Circuits.**—D. McDonnell & W. R. Perkins. (*Proc. Instn elect. Engrs*, Part C, Sept. 1955, Vol. 102, No. 2, pp. 191–202. Digest, *ibid.*, Part B, July 1955, Vol. 102, No. 4, pp. 567–568.) "The paper deals with the stability and time responses of a sampling servo-system typical of a.g.c., a.f.c., range-measuring and overall-feedback circuits used in pulsed radar equipments in which the loop-response time is not many times greater than the pulse-repetition time. The case of a high-speed radar-controlled missile may be such a system. Conditions under which Nyquist plots can be made are given. Expressions for the output both at and between the sampling times are obtained in terms of the input functions of the system. The operation of the system in the presence of noise is considered in the Appendix. A short list of transformations suitable for analysis is included."

621.314.632.1 **270**

**The Temperature Dependence of the Capacity of the Copper-Oxide Rectifier.**—T. Niimi. (*J. phys. Soc. Japan*, June 1955, Vol. 10, No. 6, pp. 444–453.) The variation of the capacitance at the metal/semiconductor boundary is calculated on the assumption that it is caused by the temperature variation of the free-charge-carrier concentration in the semiconductor. The calculated results are compared with experimental results for the temperature range from −160°C to +60°C.

621.314.634 **271**

**Properties of Selenium Rectifiers with Various Impurities.**—F. Eckart & A. Schmidt. (*Ann. Phys., Lpz.*, 15th July 1955, Vol. 16, Nos. 3/4, pp. 134–152.) Se rectifiers were prepared with halogen and halogen-compound impurities and their characteristics were investigated. The effect of fluorine and fluorine compounds on the conductivity was nil, that of alkali halides small and that of bromine pronounced. Thallium bromide, copper bromide, tellurium and arsenic have a deleterious effect on the rectifier properties. Results also show that useful rectifiers can be made using very pure Se.

621.314.653 : 537.311.33 **272**

**Semiconducting Ignitors for Ignitron Rectifiers.**—V. V. Pasynkov. (*Zh. tekh. Fiz.*, March 1955, Vol. 25, No. 3, pp. 377–396.) A detailed report on an experimental investigation is presented, as a result of which methods are proposed for the design and manufacture of ignitors based on silicon carbide.

621.316.722.1 **273**

**On the Design of Rectifiers with Electron Stabilization.**—G. I. Levitan. (*Radiotekhnika, Moscow*, Feb. 1955, Vol. 10, No. 2, pp. 40–49.) The design of a stabilized rectifier should be divided into two stages: (*a*) calculation of the limits of stability, i.e. of the limiting values of the mains voltage and load current for normal operation of the rectifier, and (*b*) determination of the coefficient of stability. Suitable methods for dealing with stage (*a*) are proposed. Details of the calculation are set out, and the case of the regulating valve with a shunt is discussed separately.

## TELEVISION AND PHOTOTELEGRAPHY

621.397.2 : 621.397.8 **274**

**The European Television Transmission Network and 'Eurovision 1954'.**—F. Kirschstein & H. Bödeker. (*Fernmeldetech. Z.*, Aug. 1955, Vol. 8, No. 8, pp. 419–425.) Details are given of the network and its operation. The observed transmission faults are described and illustrated by means of oscillograms and photographs. An indication is given of improvement work in hand.

621.397.24 : 621.372.552 : 621.375.2 **275**

**A Balanced Equalizer-Amplifier for transmitting Video Signals over Telephone Lines.**—J. B. Sewter & D. Wray. (*Electronic Engng*, Oct. 1955, Vol. 27, No. 332, pp. 422–429.) The response of ordinary balanced telephone lines is equalized by controlling the common cathode feedback network of balanced valve pairs. One equalizer unit gives up to 70 dB lift at 3 M/cs. Five corrected telephone links have been connected in tandem to give a video-signal line about 8 miles long.

621.397.242 + 621.395.44 + 621.315.212 **276**

**The Position of the Italian Coaxial Network at 31st December 1953.**—Niccolai. (See 262.)

621.397.5 **277**

**Scientific and Technical Conference on Television Broadcasting.**—(*Radiotekhnika, Moscow*, Feb. 1955, Vol. 10, No. 2, pp. 75–77.) Brief report on a conference held in Leningrad in December 1954. Twenty-six papers were read dealing with the operation of television centres of the country, colour television, transmitting and receiving tubes, commercial applications of television, long-range reception, etc.

621.397.5 278

**2½ Years of German Television.**—W. Nestel. (*Fernmeldetech. Z.*, Aug. 1955, Vol. 8, No. 8, pp. 414–418.) An illustrated review.

621.397.61 : 621.372.55 279

**Differentiating Equalizer for Television Studio Equipment.**—A. Krolzig. (*Fernmeldetech. Z.*, Aug. 1955, Vol. 8, No. 8, pp. 426–429.) Advantages and limitations of equipment based on principles described by Gouriet (1936 of 1953) are discussed. Cases of successful use in film transmissions are instanced.

621.397.611.2 280

**Basic Problems in Film Pickup for TV Broadcasting.**—E. M. Gore. (*Elect. Engng, N.Y.*, July 1955, Vol. 74, No. 7, pp. 600–604.) Includes a description of a commercially available system for colour films using three vidicon tubes.

621.397.62 281

**New Vision A.G.C. System.**—S. E. Gent & D. J. S. Westwood. (*Wireless World*, Nov. 1955, Vol. 61, No. 11, pp. 542–544.) A gated a.g.c. system for television receivers is described in which the gating pulse is obtained by differentiating the line synchronizing pulse and using the positive half of the resultant waveform.

621.397.62 : 621.373.43 282

**A Self-Oscillating Line Timebase Circuit.**—A. Boekhorst. (*Philips tech. Commun., Aust.*, 1955, No. 2, pp. 12–18; *Electronic Applic. Bull.*, June 1954, Vol. 15, No. 6, pp. 69–75.) A simple circuit for television receivers uses a single Type-PL81 pentode, with a feedback coil in its control-grid circuit, operating simultaneously as saw-tooth generator and output valve. The synchronization is highly insensitive to noise on the synchronizing signal.

621.397.7 283

**The Synchronization and Delay Correction of Scattered Television Picture Sources.**—A. B. Shone. (*Electronic Engng*, Oct. 1955, Vol. 27, No. 332, pp. 454–457.) The system used in B.B.C. outside broadcasts is described. All synchronizing-signal generators are synchronized to a common tone distributed over telephone circuits. A phase-shifting goniometer in the feed to each generator is used to ensure that the pictures associated with the different generators reach the central switching point in correct relation.

621.397.7 284

**Operating Experiences at the Lime Grove Television Studios of the British Broadcasting Corporation, London.**—D. C. Birkinshaw. (*Arch. elekt. Übertragung*, July 1955, Vol. 9, No. 7, pp. 311–325.) See 1504 of 1955.

621.397.7 : 621.396.11 285

**Television Coverage of Great Britain.**—R. A. Rowden. (*J. Telev. Soc.*, July–Sept. 1955, Vol. 7, No. 11, pp. 457–461, 464–476.) Standards of field strength necessary for satisfactory reception are indicated; properties of the transmitter and of the propagation path affecting coverage are considered. A full account is given of the distribution of field strength in the British television service as at February 1955; planned and possible developments are discussed.

## TRANSMISSION

621.376.32 : 621.396.41.029.6 286

**A Frequency Modulator for Broad-Band Radio Relay Systems.**—I. A. Ravenscroft & R. W. White. (*P. O. elect. Engrs' J.*, July 1955, Vol. 48, Part 2, pp. 108–109.) An i.f. modulator based on the ladder-network type of *RC* phase-shift oscillator is described, in which the dynamic input resistances of two grounded-grid valves serve as variable resistance elements. The performance is adequate for monochrome television transmission, or for 240-channel telephony signals.

621.376.4 287

**A New Method of Phase Modulation.**—A. D. Artym. (*Radiotekhnika, Moscow*, Jan. 1955, Vol. 10, No. 1, pp. 53–60.) A method is proposed in which nonlinear modulation of three h.f. voltage vectors is transformed into phase modulation. The theory of the method is discussed and a circuit is described providing phase deviations up to $\pm 300°$; for deviations up to 45° the coefficient of nonlinear distortion does not exceed 0·4%. The parasitic anode modulation is of the order of 2%.

621.396.61 : 621.372.542.2 288

**A Ten-Kilowatt Low-Pass Filter.**—Broad & May. (See 47.)

## VALVES AND THERMIONICS

537.5 : 538.56 289

**Plasma Oscillations in Electron Beams.**—K. G. Hernqvist. (*J. appl. Phys.*, Aug. 1955, Vol. 26, No. 8, pp. 1029–1030.) "Self-sustained ion and electron plasma oscillations have been observed simultaneously in an ion neutralized electron beam traveling in a transverse magnetic field. The observed frequencies of oscillation agree with the Langmuir-Tonks theory for plasma oscillations. Electron plasma oscillations occur when the electron plasma frequency equals the gyro frequency of a circular orbit in the magnetic field."

621.314.63 : 621.383.5 : 621.396.822 290

**Low-Frequency Noise in Illuminated Germanium Diodes.**—R. E. Burgess. (*Proc. phys. Soc.*, 1st Aug. 1955, Vol. 68, No. 428B, pp. 569–571.) It is shown theoretically that the current increase in a Ge diode under a given reverse bias when illuminated is independent of the bias voltage and is accompanied by an addition of shot noise which is negligibly small at low frequencies. The comparison of shot-noise behaviour between a dark and an illuminated diode at a given reverse current is inappropriate.

621.314.632 : 546.289 291

**Forward Characteristic of Germanium Point Contact Rectifiers.**—M. Cutler. (*J. appl. Phys.*, Aug. 1955, Vol. 26, No. 8, pp. 949–954.) The problem is treated by combining the flow equations for the barrier region, derived from emission theory (771 of 1955), with the corresponding equations ('spreading' or 'diffusion' equations) for the flow of carriers in the region beyond the barrier. Numerical solutions are obtained for typical values of the contact parameters. Swanson's assumption (2536 of 1954), that the flow of holes in the barrier region is controlled by diffusion, is discarded; in consequence, the value found for $\gamma$ at large currents is smaller, the hole concentration at the contact approaches a maximum value proportional to the equilibrium concentration of holes at the surface, and the spreading voltage becomes a linear function of the current.

621.314.632 : 546.289 : 537.312.6 292

**Thermal Effects in Point-Contact Rectifiers.**—R. E. Burgess. (*J. appl. Phys.*, Aug. 1955, Vol. 26, No. 8, p. 1058.) Critical comment on notes by Armstrong (1242 of 1954 and 881 of 1955). See also 1948 of 1954 (Armstrong).

621.314.7 **293**

**Alloyed-Junction Avalanche Transistors.**—S. L. Miller & J. J. Ebers. (*Bell Syst. tech. J.*, Sept. 1955, Vol. 34, No. 5, pp. 883–902.) Minority carriers accelerated in the high-field strength-region of a reverse-biased junction produce hole-electron pairs by collision with atoms. The action may be made cumulative, resulting in avalanche multiplication, which may be controlled by an emitter junction placed close to the multiplying collector junction. The theory of the process is presented and applications to circuits for switching and transmission are discussed. See also 2781 of 1955 (Kidd et al.).

621.314.7 **294**

**Variation of Junction-Transistor Current-Amplification Factor with Emitter Current.**—L. J. Giacoletto. (*Proc. Inst. Radio Engrs*, Oct. 1955, Vol. 43, No. 10, p. 1529.) Using suitably chosen constants, the theory presented by Webster (2798 of 1954) can be experimentally verified for *n-p-n* as well as for *p-n-p* transistors.

621.314.7 : 537.311.33 **295**

**The Theoretical Temperature Coefficient of a Junction Transistor.**—E. Groschwitz. (*Z. angew. Phys.*, June 1955, Vol. 7, No. 6, pp. 280–282.) The *p-n-p* junction transistor is considered on the basis of Shockley's theory. If the collector voltage is chosen so that the collector operates in the saturation region, the emitter and collector currents are functions of the emitter voltage only and hence the temperature coefficient of the resistances can be determined. Calculations lead to results similar to those obtained for the *p-n* junction (3298 of 1955), where the principal term in the expression for the temperature coefficient is negative and voltage dependent, and the second term is positive and dependent on the carrier lifetime and the width of the base zone.

621.383.27 + 621.385.15 **296**

**Statistical Problems of the Electron Multiplier.**—L. Yanoshi (Jánossy). (*Zh. eksp. teor. Fiz.*, June 1955, Vol. 28, No. 6, pp. 679–694.) Two problems are considered: (*a*) the probability $P_N^{(k)}(\nu)$ of $\nu$ electrons being emitted by the $N^{th}$ electrode if $k$ primary electrons are incident on the first electrode, and (*b*) the distribution of $P(\nu)$, the probability of $\nu$ secondary electrons being emitted in response to the incidence of one primary electron, required to obtain a given distribution $P_N(\nu)$ after $N$ steps.

621.383.27 **297**

**The Characteristics of Photomultiplier Tubes and their Use for Absorption Measurements.**—K. W. Keohane & W. K. Metcalf. (*J. sci. Instrum.*, July 1955, Vol. 32, No. 7, pp. 259–260.) The cathode uniformity of Type-1P21 photomultipliers was examined by using a calibrated electrical attenuator to adjust the output of a comparison photomultiplier to equality with that of the tube under test, which was moved past a slit. The signal strengths used were sufficiently low to avoid fatigue effects. Variations in the characteristics were found to be within the experimental error (of the order of 0·5%); the area of maximum sensitivity of the photocathode was determinable to the same accuracy and the input/output characteristics, in the range 2–0·05 $\mu$A, were linear. The greater nonuniformity indicated in the results obtained by Edels and Gambling (*J. sci. Instrum.*, 1954, Vol. 31, p. 121) is considered due to the experimental methods adopted.

621.383.27 **298**

**A Method of estimating Photomultiplier Pulse Durations and Shapes.**—B. Meltzer, J. A. Lodge & P. L. Holmes. (*J. sci. Instrum.*, July 1955, Vol. 32, No. 7, pp. 270–271.) The spectral output of a photomultiplier being given by the Fourier transforms of the current pulses initiated by individual photons incident on the photocathode, measurements of the output frequency spectrum may be used to show the average shape and duration of the pulses. Some experimental tests are described.

621.383.27 **299**

**A Photomultiplier Tube for Scintillation Counting.**—R. Champeix, H. Dormont & E. Morilleau. (*Philips tech. Rev.*, March 1955, Vol. 16, No. 9, pp. 250–257.) The tube described has ten stages of secondary emission, giving an overall current amplification of 500 000. The area of the $SbCs_3$ photocathode is 5 cm²; a Wehnelt cylinder focuses the photoelectrons on the first dynode; a staggered arrangement of curved dynodes is used. Steps are taken to keep the inherent noise level low. The tube is suitable for spectroscopy as well as scintillation counting.

621.385.029.6 **300**

**Power Flow in Electron Beams.**—L. R. Walker. (*J. appl. Phys.*, Aug. 1955, Vol. 26, No. 8, pp. 1031–1033.) An alternative approach to that used by Louisell & Pierce (2145 of 1955) is indicated for determining the flow of microwave power in an electron beam in a weak field.

621.385.032.216 **301**

**Measurement of Excess Ba [and/or Sr] in Practical Oxide Coated Cathodes.**—L. A. Wooten, G. E. Moore & W. G. Guldner. (*J. appl. Phys.*, Aug. 1955, Vol. 26, No. 8, pp. 937–942.) Over 300 cathodes of various designs were aged for periods up to 50 000 h, tested for emission and then analysed chemically. The complexities and limitations of the available methods of analysis are discussed. The measure of Ba content significant for emission is taken as the amount of $H_2$ formed on 16 h exposure.

621.385.032.216 **302**

**Excess Ba Content of Practical Oxide Coated Cathodes and Thermionic Emission.**—G. E. Moore L. A. Wooten & J. Morrison. (*J. appl. Phys.*, Aug. 1955, Vol. 26, No. 8, pp. 943–948.) Measurements on 125 indirectly heated and 15 directly heated cathodes are reported; methods described in 301 above were used. The evolution of $H_2$ by uncoated cathodes was also measured. No correlation was observed between the excess Ba content and the emission, and it is inferred that in practice cathode performance is seldom limited by excess Ba.

621.385.1 **303**

**On the Space-Charge-Limited Current between Nonsymmetrical Electrodes.**—G. D. O'Neill. (*J. appl. Phys.*, Aug. 1955, Vol. 26, No. 8, pp. 1034–1040.) "The validity of employing an equivalent symmetrical anode for expressing the space-charge-limited current between nonsymmetrical electrodes is examined. It is shown that when the equivalent anode is determined by the capacitance, the error in the expression for the current is usually quite small and its approximate magnitude may often be anticipated. Exact expressions for the capacitance of some basic electrode arrangements are given, followed by a brief résumé of corrections required for interpretation of experimental data."

621.385.832 : 621.374.4 **304**

**Generation of Harmonics of a Given Fundamental.**—Katz & Rau. (See 61.)

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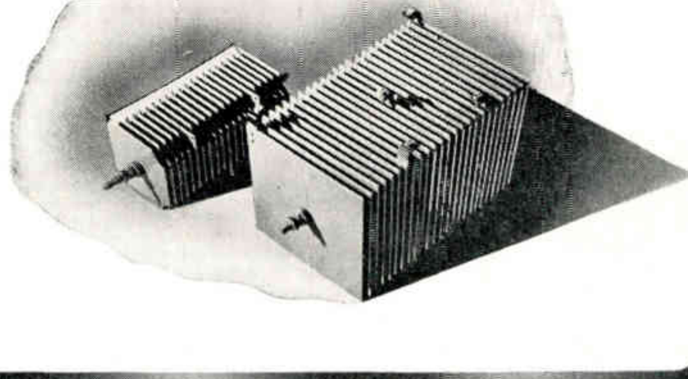

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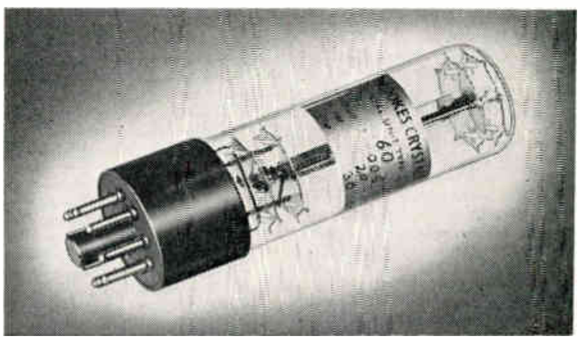

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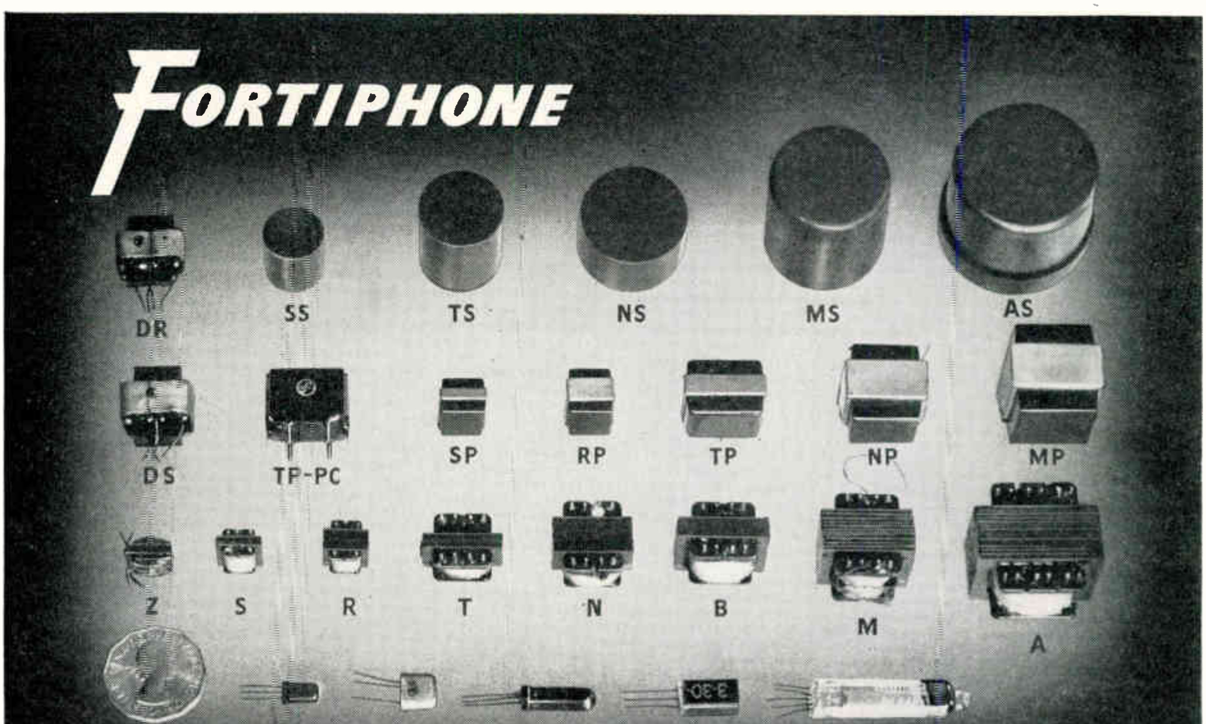


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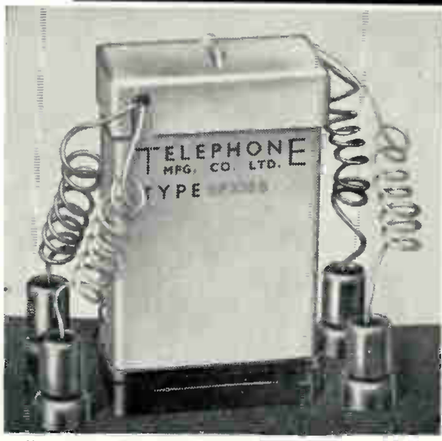

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**APPOINTMENTS—The engagement of persons answering these advertisements must be made through the local office of the Ministry of Labour and National Service, etc., if the applicant is a man aged 18-64 or a woman 18-59 inclusive, unless he or she or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

HER MAJESTY'S OVERSEA CIVIL SERVICE

Applications are invited for the following post:

**Telecommunications Engineer,**

**Posts and Telegraphs Department, Sierra Leone**

Duties cover all general branches of telecommunications engineering, maintenance and construction of overhead and underground line plant, all types of telephone exchanges, telegraph apparatus, and point to point radio networks.

Appointment is either pensionable in the salary range £964-£1,652 per annum gross or on contract for two tours of 18-24 months each, in the salary range £1,042-£1,857 per annum gross, point of entry dependent on experience or war service. A gratuity of £37 10s. for each completed three months' satisfactory service is payable on contract appointments. Free first-class passages for the officer, his wife, and up to two children. Government quarters are provided, if available, at low rentals. Leave is granted at the rate of seven days for each month of resident service.

Candidates should be A.M.I.E.E. or have passed or be exempt from Parts A and B of the Institution's examination. A sound knowledge of the theory of telecommunications engineering is necessary and extensive experience of its application.

Apply in writing to the Director of Recruitment, Colonial Office, Great Smith Street, London, S.W.1, giving briefly age, qualifications and experience. Mention the reference number BCD 108/15/03.

ESTIMATORS

PHILIPS ELECTRICAL INDUSTRIES LIMITED

have a number of vacancies in the Estimating Department at their Mitcham plant. The posts available are for:—

(1) Senior Engineering Estimators.
(2) Trainee Estimators.
(3) Technical Clerks.
(4) Pricing and Analysis Clerks.

Applicants for the senior posts should have some experience in mass production techniques, particularly in the field of Radio and Television Receiver Manufacture. For the other positions, preference will be given to men who show evidence of willingness and ability to study Engineering Techniques and suitable applicants, after a period of service, may be considered for training as specialist Estimators.

All the above posts carry a progressive salary, in accordance with individual experience and qualifications. The Company provides good working conditions with a 5-day week, Pension Scheme, Sick Pay Plan, Sports and Social Club.

Applicants are invited to write, in confidence, to Mr. I. D. Shaw, Personnel Officer, Philips Electrical Industries Limited, New Road, Mitcham Junction, Surrey, quoting reference B.I.

**AIR MINISTRY and MINISTRY OF SUPPLY**

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Ambitious men can get first-class experience which will enhance their career prospects. Salaries range up to £1,095 p.a., according to age, qualifications and experience. The normal duties attaching to most of these appointments are designed to meet the requirements of professional responsibility for corporate membership of the professional institutions and applications are specially invited for these posts from candidates who possess the educational qualifications and practical training required for corporate membership of the institutions.

Applicants should have a sound knowledge of theory and practice of electronic or light electro-mechanical engineering. Experience in either design, development, maintenance, inspection, production or operation of electronic, electro-mechanical, radio or radar installations and equipment or in design of jigs and tools would be an advantage. All applicants should have served a recognized engineering apprenticeship or have had equivalent training in an appropriate trade. For higher grade posts applicants should either be Corporate Members of one of the Institutions of Civil, Mechanical, or Electrical Engineers or have exempting qualifications. Those lacking higher academic qualifications should have had several years' post-training experience in their trade and preferably have at least H.N.C. or equivalent qualifications.

Application forms from M.L.N.S., Technical and Scientific Register (K), 26 King Street, London, S.W.1, quoting D.558/5A/HS.

**ELECTRONIC TUBES LIMITED**

**AN ASSOCIATED COMPANY**

**of**

**ELECTRIC & MUSICAL INDUSTRIES LTD.**

have vacancies for several Experimental Officers and Assistant Experimental Officers to be trained, at full salary, on experimental thermionic valve projects. Successful candidates will be engaged upon the design, construction and electrical measurements of valves in the research and development stages. Qualifications required are:

**(a) Experimental Officers: Pass degree or higher national certificate in physics, or electrical engineering or equivalent.**

**(b) Assistant Experimental Officers: General certificate of education with advanced level in general science or ordinary national certificate.**

The above requirements may be waived in the case of applicants with experience in a suitable technical field.

Salaries are based on qualifications, experience and age, and are in keeping with present-day levels. General prospects within the organization are excellent and there is a generous pension scheme. Apply to **Personnel Officer, Electronic Tubes Limited, Kingsmead Works, High Wycombe, Bucks.**

**FERRANTI LIMITED, EDINBURGH,** are forming a new group for Research on Microwave Beam Tubes in their Valve Department, and have vacancies for:

1. A Graduate Physicist or Mathematician to co-operate in theoretical aspects of microwave valve operation and with experience in the design of relevant experiments. Experience in this field is essential.
2. A Graduate Physicist for work in electron gun systems. Previous experience is desirable but not essential. A non-graduate who is familiar with the use of electrolytic tanks will also be considered.

Salaries according to qualifications and experience. Staff Superannuation Scheme in operation. Apply to the Personnel Officer (37), Ferranti Limited, Ferry Road, Edinburgh 5, quoting Ref. VL.61.

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require

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Applications giving full details, qualifications and experience to be made to X-Ray Dept., 68 Ballard's Lane, Finchley, London, N.3.

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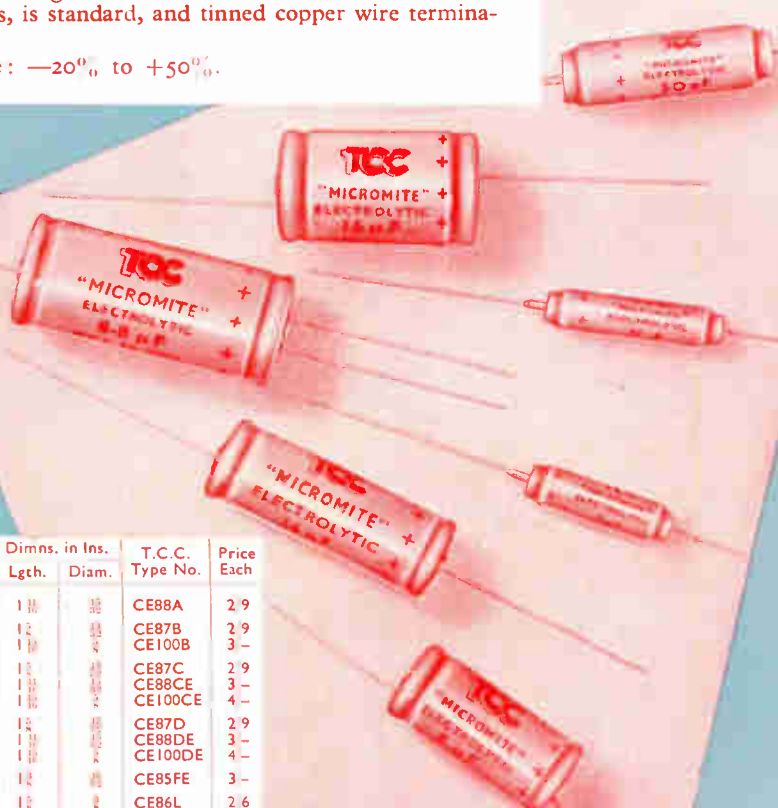
Capacity tolerance: —20% to +50%.




| Cap. in µF. | Wkg. Volts | Surge Volts | Ripple in m A | Dimns. in Ins. Lgth. | Dimns. in Ins. Diam. | T.C.C. Type No. | Price Each |
|---|---|---|---|---|---|---|---|
| 100 | 6 | 7.5 | — | 1 [illegible] | [illegible] | CE88A | 2/9 |
| 50 | 12 | 15 | — | 1 [illegible] | [illegible] | CE87B | 2/9 |
| 100 | 12 | 15 | — | 1 [illegible] | [illegible] | CE100B | 3/- |
| 25 | 25 | 30 | — | 1 [illegible] | [illegible] | CE87C | 2/9 |
| 50 | 25 | 30 | — | 1 [illegible] | [illegible] | CE88CE | 3/- |
| 100 | 25 | 30 | — | 1 [illegible] | [illegible] | CE100CE | 4/- |
| 12 | 50 | 60 | — | 1 [illegible] | [illegible] | CE87D | 2/9 |
| 25 | 50 | 60 | — | 1 [illegible] | [illegible] | CE88DE | 3/- |
| 50 | 50 | 60 | — | 1 [illegible] | [illegible] | CE100DE | 4/- |
| 8 | 150 | 200 | — | 1 [illegible] | [illegible] | CE85FE | 3/- |
| 1 | 350 | 400 | 15 | 1 [illegible] | [illegible] | CE86L | 2/6 |
| 2 | 350 | 400 | 25 | 1 [illegible] | [illegible] | CE88L | 2/9 |
| 4 | 350 | 400 | 40 | 1 [illegible] | [illegible] | CE88LE | 3/- |
| 8 | 350 | 400 | 60 | 1 [illegible] | [illegible] | CE99LE | 3/3 |
| 16 | 350 | 400 | 90 | 2 [illegible] | [illegible] | CE91LE | 4/- |
| 32 | 350 | 400 | 150 | 2 [illegible] | 1 [illegible] | CE93LE | 6/- |
| 1 | 450 | 550 | 20 | 1 [illegible] | [illegible] | CE89P | 2/9 |
| 2 | 450 | 550 | 30 | 1 [illegible] | [illegible] | CE89PE | 3/- |
| 4 | 450 | 550 | 40 | 1 [illegible] | [illegible] | CE99PE | 3/3 |
| 8 | 450 | 550 | 60 | 1 [illegible] | [illegible] | CE90PE | 3/6 |
| 16 | 450 | 550 | 100 | 1 [illegible] | 1 [illegible] | CE92PE | 5/- |
| 32 | 450 | 550 | 160 | 2 [illegible] | 1 [illegible] | CE94PE | 7/6 |
| 40–40 | 150 | 200 | — | 2 [illegible] | 1 [illegible] | CE98FE | 7/6 |
| 32–32 | 275 | 325 | — | 2 [illegible] | 1 [illegible] | CE95HE | 8/6 |
| 8–8 | 450 | 550 | 80 | 2 [illegible] | 1 [illegible] | CE98PEA | 7/6 |
| 8–16 | 450 | 550 | 80 | 2 [illegible] | 1 [illegible] | CE98PEB | 8/- |
| 16–16 | 450 | 550 | 120 | 2 [illegible] | 1 [illegible] | CE95PE | 9/6 |

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W 97


**TYPE W97 MIDGET 'THERMETIC'**

| Capacitance Microfarads | List Number | Dimensions L (inches) | D |
|---|---|---|---|
| 200 Volts D.C. Working | | | |
| 0.002 | BM7 | 0.610 | 0.135 |
| 0.004 | BM8 | 0.610 | 0.135 |
| 0.004 | BM11 | 0.500 | 0.180 |
| 0.005 | BM9 | 0.610 | 0.135 |
| 0.005 | BM12 | 0.500 | 0.180 |
| 0.01 | BM13 | 0.500 | 0.180 |
| 0.02 | BM14 | 0.610 | 0.180 |
| 0.03 | BM15 | 0.610 | 0.260 |
| 0.04 | BM16 | 0.610 | 0.260 |
| 400 Volts D.C. Working | | | |
| 0.0004 | BM4 | 0.610 | 0.135 |
| 0.0005 | BM5 | 0.610 | 0.135 |
| 0.001 | BM6 | 0.610 | 0.135 |
| 0.002 | BM18 | 0.500 | 0.180 |
| 0.003 | BM19 | 0.500 | 0.180 |
| 0.005 | BM20 | 0.610 | 0.180 |
| 0.01 | BM21 | 0.610 | 0.260 |
| 600 Volts D.C. Working | | | |
| 0.00005 | BM25 | 0.500 | 0.180 |
| 0.0001 | BM26 | 0.500 | 0.180 |
| 0.0001 | BM1 | 0.610 | 0.135 |
| 0.0002 | BM2 | 0.610 | 0.135 |
| 0.0002 | BM27 | 0.500 | 0.180 |
| 0.00022 | BM28 | 0.500 | 0.180 |
| 0.00025 | BM29 | 0.500 | 0.180 |
| 0.0003 | BM3 | 0.610 | 0.135 |
| 0.0003 | BM30 | 0.500 | 0.180 |
| 0.0004 | BM36 | 0.500 | 0.180 |
| 0.0005 | BM31 | 0.500 | 0.180 |
| 0.001 | BM32 | 0.500 | 0.180 |
| 0.002 | BM33 | 0.610 | 0.260 |
| 0.003 | BM34 | 0.610 | 0.260 |
| 0.004 | BM35 | 0.610 | 0.260 |

If required, Type W97 can be supplied with a plastic insulating cover.

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