Electronic Engineering

EDITORIAL, ADVERTISING AND PUBLISHING OFFICES, 28, ESSEX STREET, STRAND, W.C.2.

TELEPHONE CENTRAL 6565 (10 LINES) MANAGING EDITOR: H. G. FOSTER, M.Sc., M.I.E.E. TELEGRAMS: LECTRONING, ESTRAND, LONDON

Monthly (published last Friday of preceding month) 2/- net.

Subscription Rate: (Home or abroad) Post Paid 12 months 26/-, or $3.75 (U.S.).

Vol. 22. No. 269. July, 1950

CONTENTS

Television Aerials

CONSIDERABLE uncertainty has existed for some time past as to the rights of landlords and tenants in the matter of erection of outdoor television aerials, and this uncertainty has been increased by recent action on the part of several local councils in banning aerials from their housing estates.

The reasons behind these restrictions do not appear to have been based on very sound argument, and more than welcome, therefore, is the publication this month of the booklet "Television Aerials—Outdoor or Indoor?" issued jointly by the Radio and Electronic Component Manufacturers and the British Radio Equipment Manufacturers' Association.

This admirably prepared booklet is intended "primarily for the benefit of municipal authorities and other property owners whose judgement their tenants may have to accept for the choice of television aerial they may employ".

Apart from architectural objections, the technical arguments are put forward mainly on the grounds of risk of damage to property—usually the chimney stack—and the increased hazards of lightning. The booklet states, however, that television aerials are made by reputable manufacturers who have gained sufficient experience to prove that the erection of an outdoor aerial can actually strengthen the chimney stack and that the increased lightning danger is negligible.

The booklet goes on to show that, assuming that the restrictions on the erection of outdoor aerials can now be removed, there are still a number of factors to be considered in the selection of the most suitable type, and that the choice of the familiar "H" type in strong signal areas can be (and often is) based on sounder reasons than mere "snob" value.

If television is to be as much a part of our everyday life as the broadcast service, it is, perhaps, not inappropriate to consider the possibilities of supplying that service by other than the existing methods.

One possible method is, of course, by the use of microwaves, which has the advantage of requiring smaller receiving aerials, but adequate coverage, particularly in a built-up area, would be problematical; reflexion troubles from buildings would be increased and interference might not necessarily be reduced.

The alternative is perhaps the relaying of television by means of cable, and this method raises a number of interesting possibilities.

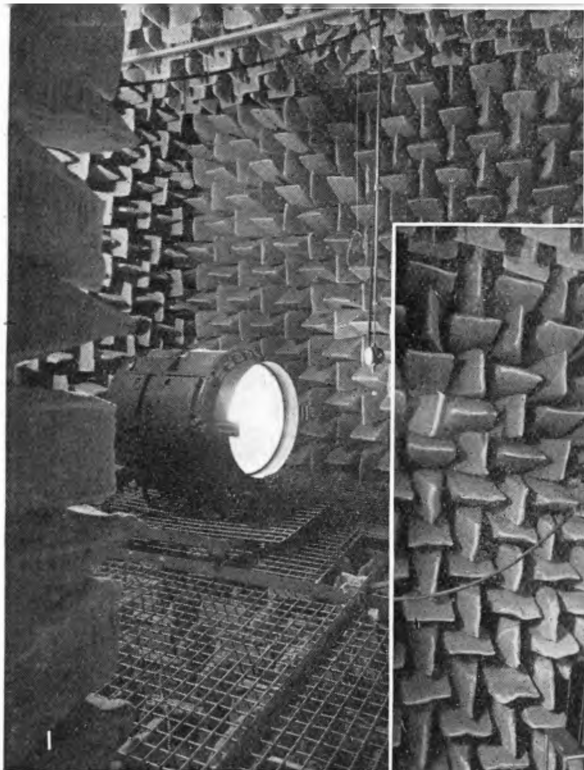
There is only a single programme at the moment, a second programme is not due for several years to come and the prospects of alternative or commercial transmissions are so remote as to be ignored. The viewer, therefore, is compelled by the very nature of things to be restricted in his choice of programme and he cannot roam the ether in search of other programmes. Since we are to be tied to a single source of programme, it might not seem so far fetched as it first appears to relay television (and radio) programmes by cable. It might, at the same time, remove the bugbear of direct reception—namely, interference, particularly from motor cars. Nor need the cost be excessive if carried out on a large enough scale.

No one can pretend that the sight of all these aerials perched on our roofs is anything but an eyesore, as a train journey through our suburbs will demonstrate.

When we have reached that particular aspect of our all-electric all-atomic energy era in which all our electricity, radio, telephone and television services (and gas, too, if still required) are piped into our homes by a single coaxial cable, aerials should be as much an anachronism as the chimneys to which they are attached.

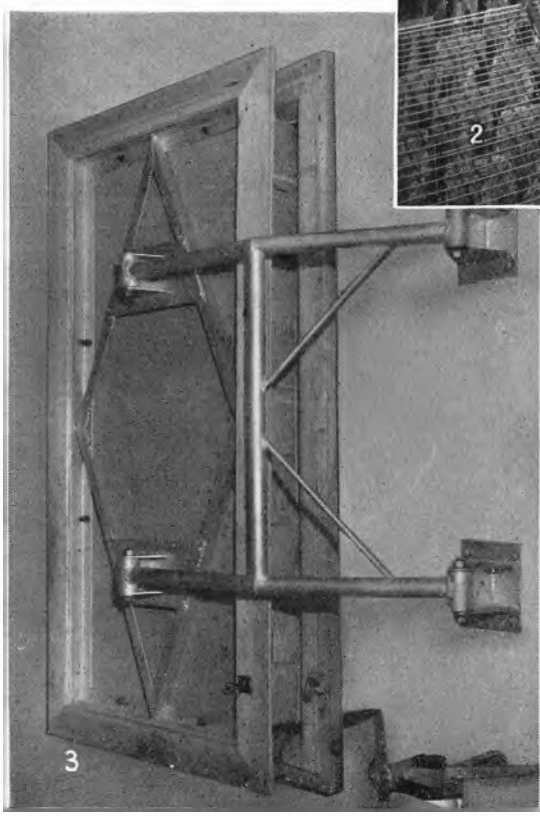
AN ECHO-FREE ROOM

at G.E.C. Research Laboratories, Wembley

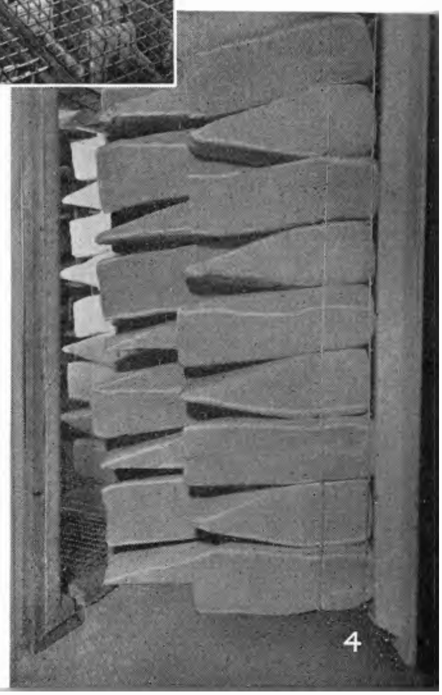


1.—Measuring the arc-noise from a G.E.C. 150 ampere double-negative high intensity studio arc lamp

2.—Measuring the overall audio fidelity of a G.E.C. television receiver. This illustration shows the sound transparent floor particularly well



3.—Outside wall of the room, showing the construction of the door hinges which allow initial parallel motion of the door. The handle below the hinges is used for rotating the turntable within the room



4.—Side view of partly-open door, showing how the "Fibreglass" wedges are held in position in a wire mesh

The Single-Sideband System of Radio-Communication

By H. D. B. KIRBY, M.A.

(*Late of Standard Telephones & Cables, Ltd., London*)

THIS article describes the single-sideband system of communication and shows the advantages it possesses over the normal double-sideband method of operating a radio link.

The method was invented by J. R. Carson and used for several years for wire transmission before it was applied to radio. It was used on the first transatlantic telephone circuit, which operated on long waves, and in 1930-31 tests carried out between Paris and Madrid, and Madrid and Buenos Aires showed that it could be applied with equal success to short waves. Five years later a single-sideband short wave circuit was opened between London and New York. Now a large number of point-to-point radiotelephone links throughout the world are being changed over from D.S.B. to S.S.B. operation and the use of the system for the long-distance ship-shore services has commenced with the installation of S.S.B. equipment in the Cunard White Star liner "Caronia."

What is S.S.B.?

In order to appreciate what the single-sideband system involves it is necessary to consider first what happens when a speech wave is used to modulate a radio-frequency carrier. To take a simple example let us suppose that at some particular time the speech wave consists of a 1,000 c/s. sine wave, together with its second harmonic (2,000 c/s.) at half the amplitude. Two cycles of such a wave are shown in Fig. 1 (a, b and c). If this is used to modulate a carrier of 10 kc/s. (d) by the ordinary method of amplitude modulation the waveform shown at (e) is produced. Without going into mathematics it is possible to state that this waveform may be analysed into the following sine waves: 8 kc/s., 9 kc/s., 10 kc/s., 11 kc/s. and 12 kc/s. The first two are known as the lower sideband, the third as the carrier and the last two as the upper sideband. The amplitudes of the components of the sidebands are proportional to the components of the original modulating signal, so that in this case the 8 kc/s. sine wave which corresponds to the original 2,000 c/s. has half the amplitude of the 9 kc/s. which corresponds to the original 1,000 c/s. Similarly, the 12 kc/s. has half the amplitude of the 11 kc/s. When this wave is demodulated in the ordinary receiver a large number of frequencies are produced, among them the original 2,000 c/s. and 1,000 c/s.

It will be appreciated that the carrier frequency, amplitude and phase remain the same whatever the modulation and that it therefore plays no part in conveying intelligence. It is, however, necessary for the ordinary process of demodulation. If the 10 kc/s. carrier is removed from the waveform shown at (e), the wave shown at (f) is produced. If this is applied to a square-law detector the audio-frequencies which are produced will be twice those in the original signal. If it is demodulated by the addition of a locally generated carrier it is necessary for this carrier to be not only exactly the same frequency as the original but also the same phase. This type of signal, therefore, though very much more easily produced than the single sideband signal is of little use for communication.

However, since there are two sidebands present in the waveform shown at (f) the information is in effect being sent twice, and it is therefore possible to remove one of the sidebands. This will give a waveform as shown at (g). This can be demodulated by mixing with a locally generated carrier which must be very nearly the same frequency as the original carrier, but need not be in phase with it.

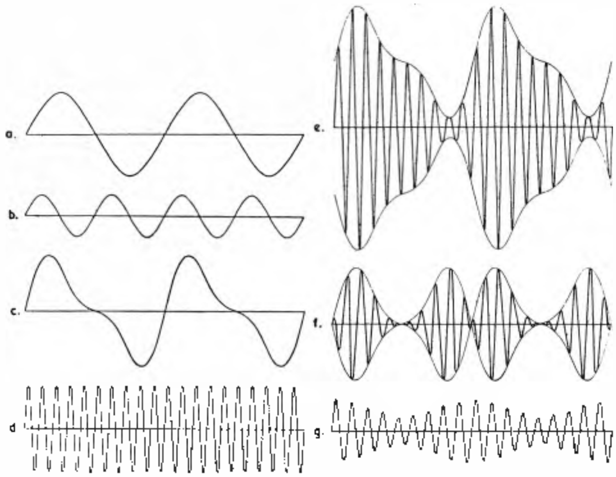


Fig. 1.

Waveforms: (a) 1,000-c/s note. (b) Second harmonic of a. (c) Combination of a and b. (d) 10-kc/s carrier. (e) Double-sided signal. (f) Double-sideband signal with carrier suppressed. (g) Single-sideband signal

Advantages

The advantages of using the S.S.B. system may be summarized as follows:

(1) Improved signal-to-noise ratio at the receiver for a given transmitter power.

(2) Less bandwidth occupied per channel.

(3) Freedom from the distortion normally caused by selective fading and multi-path propagation.

One reason for the improved signal-noise ratio follows from the explanation of the nature of the S.S.B. signal. It can be shown that if a double-sideband transmitter is fully modulated by a sine wave signal the peak amplitude of the output is twice that of the unmodulated carrier. The maximum voltage of each sideband is therefore one-quarter of the peak amplitude of the output wave. If single-sideband operation is used on a similar transmitter, since the limit of output is normally the peak emission of the power amplifier valve, the maximum value of the sideband is the same as that of the total output, or four times that of each sideband in the double-sideband transmission. At the receiver both the sidebands of the double-sideband transmission contribute to the output of the demodulator, so that the voltage output from the single-sideband receiver is twice that from the double-sideband receiver. The power output from the receiver is therefore four times as great, which is equivalent to an improvement of signal-to-noise ratio of approximately 6 db.

The second advantage of the single-sideband method of transmission (less bandwidth occupied per channel) also follows from the preceding discussion. In our example the double-sideband transmitter generated frequencies between 8 and 12 kc/s., whereas the single-sideband transmitter generated only 8 and 9 kc/s. or 11 and 12 kc/s. There will be similar results, whatever the modulating frequencies so that less space in the radio frequency spectrum is occupied by a S.S.B. transmission. In these days when the utmost economy in the use of frequency space is essential, this is a very important advantage. Apart from pulse code modulation, the practical use of which may be some years away, the single-sideband system is more economical of bandwidth than any other.

Arising out of this reduction in bandwidth, there is also a further possible gain in signal-to-noise ratio at the receiver. Since a smaller band of frequencies is being transmitted it is possible to use filter circuits in the R.F. and I.F. stages of the receiver which have a narrower pass-band than that used in a D.S.B. receiver. As a result the random noise picked up with the signal is reduced in proportion, since this noise may be assumed to be distributed evenly through the frequency spectrum. Thus, if the passband of the I.F. transformers of a receiver be halved the random noise energy passed to the audio-stages will also be halved.

Thus the signal energy is quadrupled and the noise energy is halved, resulting in a total improvement in the signal-to-noise ratio of eight times or 9 db.

These figures assume sine-wave modulation, whereas the normal signal will be a complex wave. However, the actual operation of commercial S.S.B. circuits has shown substantial agreement between the figures obtained in practice and those predicted by theory. Comparative tests on long-distance circuits also show a considerable improvement in quality with the S.S.B. system, which is due to the absence of selective fading effects. The reason for this can be understood from a consideration of the causes of selective fading. Fig. 2 shows a transmitting Station A and a receiving Station B outside ground-wave range. Signals from A are therefore received at B after one or more reflexions from the ionosphere. In general the various possible paths between the stations will be of different length and consequently the signals received there will not necessarily be in phase with each other. When the signals arrive out of phase they tend to cancel each other and cause the effect generally known as fading. This in itself may not necessarily be very important, since changes of level at the receiver can be overcome by automatic gain control systems. Such systems can, however, be effective only when the carrier and sidebands retain their original relative strengths. Unfortunately, since the bending effect of the ionosphere is dependent on frequency the different components of the transmitted wave will come back to the earth at different angles. For instance, it is possible that at some particular moment the frequencies in the lower sideband might follow the path indicated by the full lines in Fig. 2 and the upper sideband might follow the broken lines. In this case the lower sideband reaches the receiver after two reflexions and the upper sideband after one reflexion. Now the audio frequency output of a demodulator to which a D.S.B. signal is applied is made up of two parts for each audio frequency: one produced by the upper sideband beating with the carrier and the other produced by the lower sideband beating with the carrier. And it is one of the requirements for distortionless transmission that the phase-shift between the carrier and each frequency in one sideband should be equal to the phase-shift between the carrier and the corresponding frequency in the other sideband. This requirement is normally

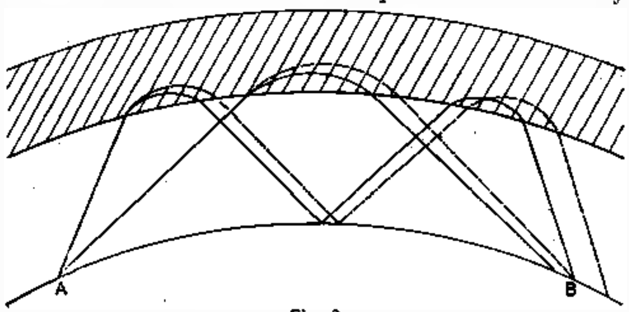


Fig. 2
Multi-path propagation

fulfilled, but when the two sidebands arrive by different paths the phase-shift will no longer be proportional to frequency, and the relative phase between the audio outputs contributed by each sideband will vary with the audio frequency.

In the single-sideband system on the other hand the second component for each frequency is missing in the detector output, so that this form of delay distortion cannot occur.

Another form of selective fading caused by multipath propagation occurs when the carrier fades out and leaves most of the two sidebands at their normal amplitude. This causes the detector output to consist largely of second harmonics of the components of the original signal wave caused by the beating of one set of sidebands against the other. This form of distortion again cannot occur in S.S.B. systems, since only one sideband is present and the carrier is either locally generated at the receiver or else is amplified separately and recombined with the sideband frequencies in the demodulator.

When non-linear distortion is severe in a D.S.B. system intelligibility will be very low, and especially if a privacy system is in use the circuit may be unworkable. Freedom from this type of distortion is one of the chief reasons why S.S.B. working is preferred for long-distance radiotelephone circuits.

Transmission

There are a number of possible ways in which the single-sideband signal can be produced, but we do not yet know how to produce it directly. Instead it is necessary to produce a D.S.B. signal and then remove the unwanted components.

Removing the carrier is comparatively simple since the familiar balanced modulator can be used. The circuit of such a modulator is shown in Fig. 3. The carrier is applied to the grids of the two valves in parallel and the signal is applied in push-pull. A transformer with a centre-tapped primary is used in the anode circuit and it can be seen that the carrier will be balanced in the two halves of this winding and therefore will not appear at the secondary. The two sidebands on the other hand will add in this transformer and produce a waveform similar to that shown in Fig. 1(f).

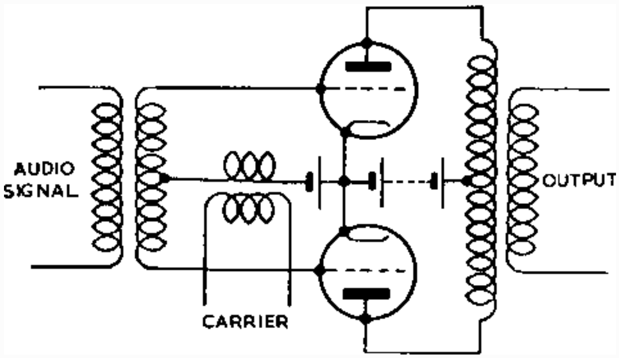


Fig. 3.
Simplified schematic of balanced modulator

It is possible to arrange similar circuits so that one sideband is cancelled out, but it is difficult to make the balance hold over the audio range required. A more satisfactory way in the present state of the art is to use crystal filters to suppress the unwanted sideband. In the example shown in Fig. 1 it would be comparatively easy to design a suitable filter since, assuming that the lower sideband is to be suppressed, the filter is only required to pass frequencies of 11 kc/s. and 12 kc/s., and to suppress frequencies of 8 kc/s. and 9 kc/s. Thus it is not necessary for the characteristics of the filter to have particularly steep sides. If, however, it is desired to pass a band of frequencies extending down to 100 c/s. on a carrier of 10 Mc/s., the filter will have to discriminate between frequencies of 9.9999 Mc/s. and 10.0001 Mc/s. It can be seen that the design of such a filter presents much greater difficulty. Therefore, when high-frequency carriers are used it is necessary to use a slightly more complicated circuit arrangement in order to simplify the design of the filters.

The block diagram of a simple single-channel single-sideband transmitter, shown in Fig. 4, illustrates the method normally adopted. It can be seen that the single-sideband signal is produced at a comparatively low frequency and is then converted to the higher frequency required for radiation.

The line amplifier and volume indicator is provided to enable the audio input to the first modulator to be set to the correct level. This modulator is of the balanced type shown in Fig. 3, the audio signal being applied to the grids of the two valves in push-pull and the output of a 100 kc/s. oscillator in parallel. The output of this modulator therefore consists of two sidebands, the upper being 100 kc/s. plus the band of audio frequencies and the lower being 100 kc/s. minus the audio frequencies. Either one of these is selected as required by means of the crystal filters, one of which is designed to pass a band of frequencies 100.2-106 kc/s. and the other 94-99.8 kc/s.

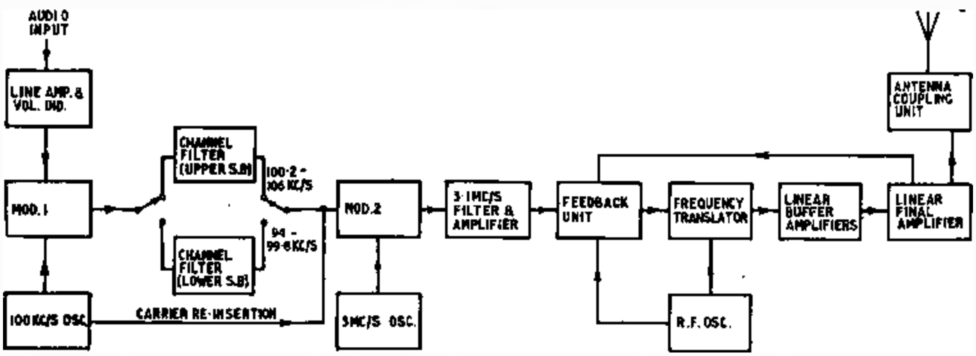
As will be seen in connexion with the receiver, it is convenient to have a small residual carrier in the received signal. As the 100 kc/s. component is almost completely suppressed in the balanced modulator, it is reintroduced at the required level after the filters.

The signal is then mixed with the output of a 3 Mc/s. oscillator in the second modulator which is again of the balanced type. Two sidebands are produced in the output of this modulator, one comprising a band of frequencies in the region of 3.1 Mc/s., the other in the region of 2.9 Mc/s. These two bands are sufficiently widely separated for the 2.9 Mc/s. band to be suppressed by means of ordinary tuned circuits.

The 3.1 Mc/s. signal after amplification is mixed with the output of a variable R.F. oscillator to produce the required operating frequency. When the operating frequency is below 10 Mc/s. it is normal practice to make the oscillator frequency 3.1 Mc/s. above the operating frequency; when it is above 10 Mc/s. the oscillator frequency is 3.1 Mc/s. below the operating frequency.

Thus if the upper sideband is being used an audio signal of 1,000 c/s. is first translated to 101 kc/s. and then to 3.101 Mc/s. If the final carrier frequency is 5 Mc/s., this is then converted to 4.999 Mc/s.; if the final carrier frequency is 15 Mc/s., it is converted to 15.001 Mc/s.

Fig. 4.
Block schematic of simple S.S.B. transmitter



The output of the third modulator (or frequency translator) valve comprises the required single-sideband signal, but its power at this point is insufficient for radiation. Since the signal is modulated, linear amplifiers have to be used to increase the power to that required for feeding the aerial. To ensure good linearity negative feedback is used between the output stage and the frequency translator.

Reception

The chief problem in a single-sideband receiver is the provision of a suitable carrier to demodulate the received signal. The frequency of this carrier must be within 10-15 cycles of the carrier which was suppressed at the transmitter. Naturally the higher the operating frequency the more difficult it is to keep down the frequency difference between the two carriers, and at the normal frequencies used it is not practical to maintain two independent oscillators within these limits. At 1 Mc/s., for instance, the frequency of a crystal oscillator may vary by as much as ± 10 c/s. There is thus a possible variation of 20 c/s. between two similar oscillators even at this frequency. In addition the grinding accuracy of a crystal is not usually better than 100 c/s. It is therefore normal practice to transmit a pilot carrier at a very low level, to amplify this separately at the receiver and use it either as the demodulating carrier (re-constituted carrier) or to control the frequency of a local oscillator (locally-generated carrier).

The block diagram of a simple S.S.B. receiver is shown in Fig. 5. The signal is first amplified at radio frequency and then mixed with the output of a tunable oscillator, whose frequency differs by 3.1 Mc/s. from that of the received signal. For frequencies below 10 Mc/s. it is tuned above the incoming signal, and for frequencies above 10 Mc/s. below the incoming signal. In this way the correct relationship between the audio frequencies is maintained.

The output from the first frequency-changer is amplified at 3.1 Mc/s. and then mixed with the output of a 3 Mc/s. oscillator. The 100 kc/s. modulation product is selected and after amplification is applied to a crystal filter which passes either the band 94-99.8 kc/s. or the band 100.2-106 kc/s. The output from the filter after amplification is mixed with the output of a 100 kc/s. oscillator so that the original audio frequencies are produced. To ensure that these audio frequencies are the same as those transmitted, it is necessary that the frequency difference between the outputs of the filter and the 100 kc/s. oscillator shall be the same as that between the carrier and the sideband in the transmitted signal. To achieve this the output of the 100 kc/s. amplifier is passed to a narrow crystal filter which is peaked at 100 kc/s., so that it selects the carrier which was transmitted at a low level. After amplification the frequency of this reconstituted carrier is compared with that of the 100 kc/s. oscillator. Any difference between these frequencies produces a voltage which is used to vary the frequency of the 3 Mc/s. oscillator until this voltage is reduced substantially to zero. Various methods of automatic frequency control are used involving either reactance tubes or motor-driven trimmers. The latter are usually preferred.

Since the carrier is transmitted at a constant level, whereas the sidebands are continually varying, the former is used for the production of the automatic gain control voltage.

Under certain conditions good results may be obtained by using the output of the carrier filter and amplifier instead of the locally-generated 100 kc/s. wave to demodulate the signal.

Examples of Equipment

The simple equipments so far described have been designed for single-channel working, but it is clearly economical to use the two sidebands independently for the transmission of intelligence. Consequently most practical transmitters and receivers are designed for double-channel working.

A number of transmitters and receivers for S.S.B. operation have been developed and manufactured by Standard Telephones & Cables during the last few years. The transmitters have output powers ranging from 300 watts to 40 kilowatts, and the receivers vary in complexity according to the facilities required.

The simplest transmitter in the range has a circuit very similar to the block schematic shown in Fig. 4. It has a frequency range of 4-22 Mc/s. and its output power is 300 watts (peak envelope). It is self-contained and built in a single cabinet measuring 5 ft. 6 in. by 2 ft. 8 in. by 2 ft. 1 in. and weighs approximately 600 lb. A transmitter of this type is installed in the new Cunard White Star Liner "Caronia."

The larger transmitters make use of a drive unit which contains all the circuits required to convert two incoming signals to two sidebands of a reduced amplitude 3.1 Mc/s. carrier, together with monitoring facilities for the output and important points in the circuit.

A 4 kW model is built in a single assembly requiring front access only. An interesting feature

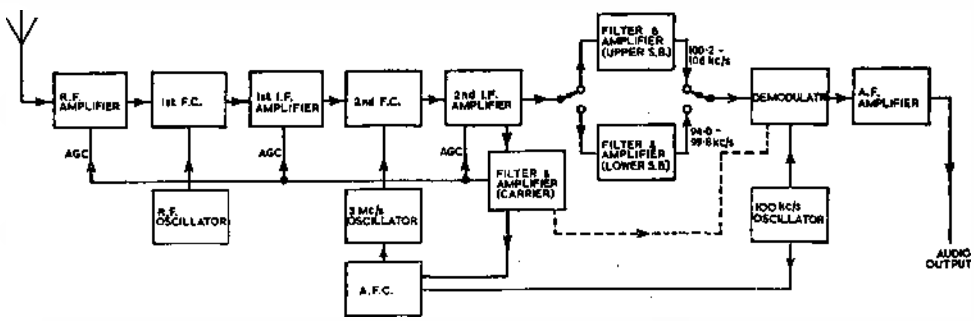
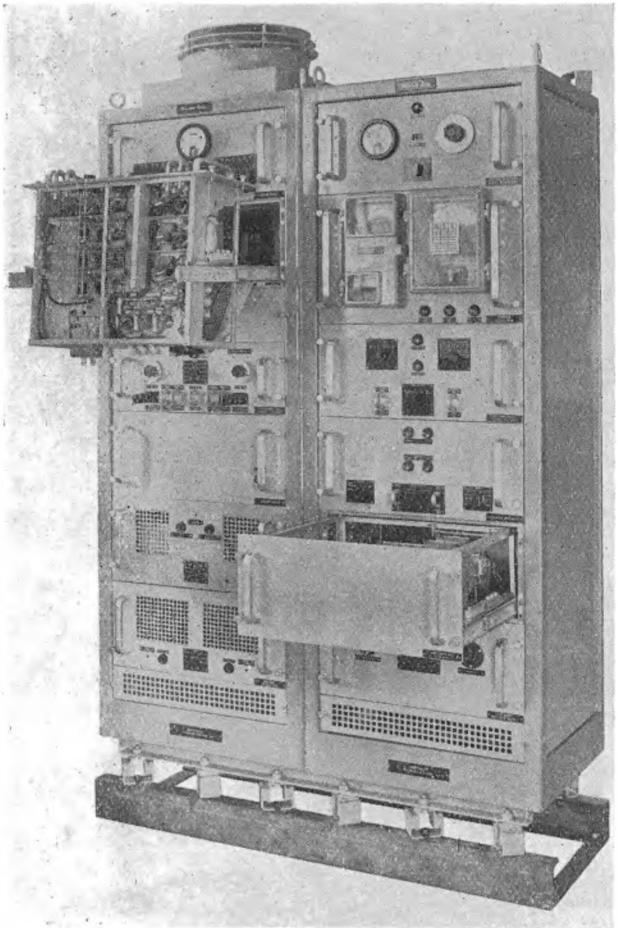


Fig. 5. Block schematic of simple S.S.B. receiver



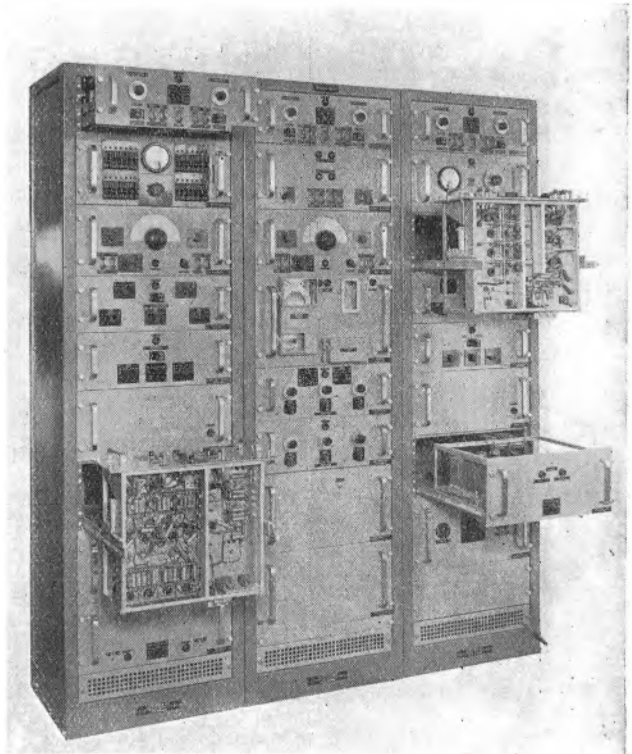


Fig. 6 (*left*).
S.S.B. Receiver as installed in the "Caronia"

Fig. 7 (*right*)
Triple Diversity S.S.B. Receiver type RX.1

is the use of an air-blast cooled pentode output valve. The larger transmitter, together with its cooling plant is built in an enclosure, which can be entered through a door at the side. Two grounded-grid water-cooled valves are used in the final amplifier stage. Negative feedback is used in both models to give a high degree of linearity and reduction of noise and unwanted phase modulation.

The receiver installed in the "Caronia" is shown in Fig. 6 and a more complex one for land operation in Fig. 7. The "Caronia" receiver is designed to provide the best possible facilities consistent with the restriction on size and weight introduced by ship-board operation and may be used for the reception of one audio signal on each sideband. It is continuously tunable over the range 4-25 Mc/s. and motor-driven automatic frequency control is incorporated. The first oscillator is normally crystal-controlled and six crystal positions are provided in a thermostatically-controlled oven.

The receiver shown in Fig. 7 provides full space-diversity reception facilities using three separate aerial systems and receivers. The outputs of the three receivers are combined in such a way that the audio output is derived substantially from the strongest received signal. In this way the effects of fading are very considerably reduced, since it is probable that at any moment a reasonable signal will be picked up by at least one of the aerials.

Conclusion

The advantages of using S.S.B. operation have been discussed above. The chief disadvantages are the increased complexity of some of the equipment required and the comparatively high degree of stability required in power supplies, oscillators, crystal filters and some of the components. Thus the receiving equipment tends to be more expensive than comparable D.S.B. receivers, but against this must be offset the possibility of using transmitters with lower power output. It has been found that these considerations, together with the extra operating time available owing to the higher quality of the circuit, make the S.S.B. system not only theoretically but also practically the most suitable method of operating commercial long-distance telephone links. No reference has been made to telegraph operation, but the audio-frequency signals transmitted may take any form and can therefore comprise one or more voice-frequency telegraph channels. Satisfactory operation has been obtained with six telegraph channels on a single S.S.B. link.

Microwave Lenses

By J. BROWN, M.A., and S. S. D. JONES, M.A.*

Part IV

MICROWAVE lenses are, like their optical counterparts, subject to a variety of defects. The principal defects of optical lenses, or aberrations, as they are usually called, may be conveniently listed as follows:—

(1) Spherical aberration,

(2) chromatic aberration,

(3) coma,

(4) surface reflexions.

Similar aberrations occur with microwave lenses, but because of the differences between optical and microwave lenses, the most important aberrations are not the same in the two cases. The most important property of a microwave lense is generally the gain, and the defect which has the most influence upon this is the last of the above list. Surface reflexions will therefore be discussed in detail in the next section, together with another defect known as spillover, which is usually neglected in optics.

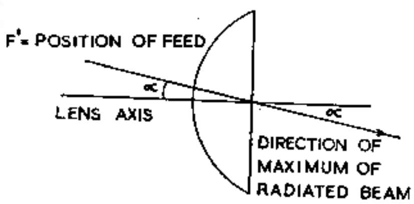


Fig. 27. Movement of the beam caused by moving the feed

In optical lenses spherical aberration arises from the assumptions that such lenses are thin, and that the surfaces may be approximated by sections of spheres. In the calculation of microwave lens contours by the methods described in previous articles, however, no approximate assumptions have been made and so these lenses will be free from spherical aberration, at any rate so long as the lens feed is situated at the design focus. Movements of the feed along the axis will result in some aberration, which shows itself as a curved instead of a plane phase front in the wave emerging from the lens. The permissible movement of the feed before this aberration causes any appreciable deterioration of the radiation pattern is, however, quite large, so that the position of the feed may be preserved to the required degree of accuracy without any mechanical difficulties. In some applications of lenses, it is desired to move the feed off the axis in order to scan the radiated beam over a range of angles: in Fig. 27, for example, if the feed is moved to the position F', the radiation pattern will be such that the direction of the maximum will make an angle α with the lens axis. It follows that the direction of the beam may be altered by moving the primary feed only, the lens itself remaining fixed. It is obviously a much simpler mechanical problem to move only the feed than to rotate the complete system to the required direction. The maximum angle through which the beam may be swung is limited by the spherical aberration, and also by the coma. The scanning properties of microwave lenses will be discussed later in this article.

The only remaining aberration is chromatic, which is the limitation imposed on the range of wavelengths over which the lens may be used. This has already been discussed for each of the dielectrics described in the previous articles and will not be considered further.

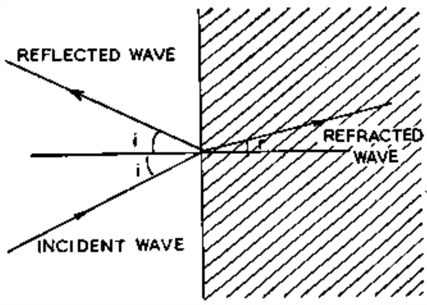


Fig. 28. Reflexion at a dielectric interface

Surface Reflexions

One of the most serious defects of a lens is the occurrence of reflected waves, caused by the discontinuities in refractive index at the surfaces. Such reflexions may be studied theoretically,[11] the simplest case being that occurring at the interface between free space and a solid dielectric as shown in Fig. 28. It is assumed that the interface is infinite in extent and that the incident wave is plane polarized. There will be in the dielectric a refracted wave, whose direction is given by Snell's Law, Equation (1), and there will also be a reflected wave. This reflected wave will be in the same direction as if the interface were a perfect reflector. The power reflexion coefficient, R, defined as the ratio of the power reflected from any area of the interface to the power incident on the same area, may be calculated from the boundary conditions which must be satisfied at the interface. These conditions are that the tangential components of both electric and magnetic fields must be continuous across the interface. Two distinct cases arise: if the electric

* Radar Research and Development Establishment, Ministry of Supply.

vector lies in the plane of Fig. 28 the power reflexion coefficient is

$$R_1 = \frac{\tan^2(i - r)}{\tan^2(i + r)} \quad \text{.................} \quad (50)$$

while if the electric vector is normal to the plane of Fig. 28 then

$$R_2 = \frac{\sin^2(i - r)}{\sin^2(i + r)} \quad \text{.................} \quad (51)$$

where i and r are the angles which the directions of the incident and refracted waves make with the normal to the interface. The values of these angles must, of course, satisfy Snell's Law. In the limiting case of normal incidence, both expressions tend to

$$R = \left(\frac{n - 1}{n + 1}\right)^2 \quad \text{...............} \quad (52)$$

The only case when the reflexion coefficient becomes zero occurs when $i + r = \pi/2$. Then $\tan (i + r)$ becomes infinite and so $R_1 = 0$. The value of i for which this happens is known as the Brewster angle, as in optics. The power transmission coefficient T, defined similarly to the reflexion coefficient, has always the value $1 - R$. It may also be shown that the phases of the incident and refracted waves are continuous at the interface. The results of the above paragraph apply only to solid dielectrics, and modifications are necessary before they can be applied to the various artificial dielectrics discussed in previous articles. differ in two important respects from those for solid dielectrics: firstly, for values of the angle of incidence exceeding a certain value, a second reflected wave occurs. This additional reflected wave is caused by the edges of the plates acting as a diffraction grating. The overall transmission coefficient is appreciably reduced when this occurs. Secondly there is a storage field set up at the interface, the effect of which is to cause a phase change at the interface between the incident and refracted waves. This phase change varies from point to point over the surface of a lens and so causes small phase errors. Carlson and Heins have ignored the effect of finite plate thickness in their calculations: it appears that if allowance is made for this plate thickness, larger values of the reflexion coefficient will be obtained. It is therefore important in practice to keep the thickness of the plates in a metal plate lens to the minimum acceptable from the point of view of mechanical strength.

So far only one surface has been considered, whereas lenses have two surfaces. The effect of this upon the nature of the reflexions may be seen by studying the case of the plane sheet of Fig. 30. It will be sufficient to consider a plane wave incident normally on to the sheet, and it may be shown that the reflexion coefficient for this case is

$$R' = \frac{2R(1 + \cos 4\pi d/\lambda')}{1 + 2R \cos 4\pi d/\lambda' + R^2} \quad \text{........} \quad (53)$$

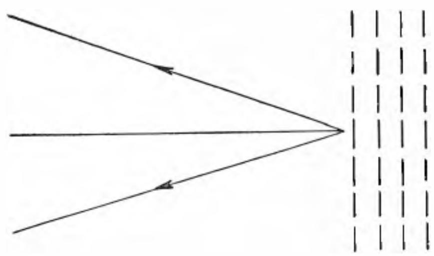


Fig. 29. Reflexion from a strip dielectric

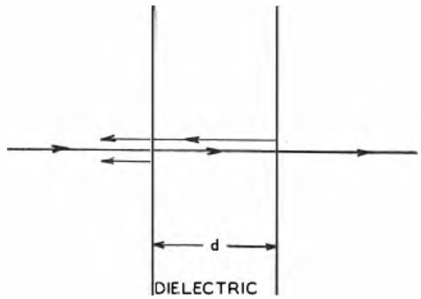


Fig. 30. Reflexion by a dielectric sheet

Details of the modifications required will only be given here for the cases of strip delay dielectrics and metal plate dielectrics, as these are the only two which have as yet been studied in detail. As was shown in Part II, strip delay dielectrics operate only when the electric vector is perpendicular to the direction of the strips. If the strips lie as shown in Fig. 29, it is natural to suppose that the reflexion coefficient should be given by Equation (50). Such is, in fact, the case, provided that the wavelength used lies in the range for which the refractive index is constant, and that the direction of propagation within the dielectric is nearly normal to the planes of the strips. This case is the one which occurs most frequently in lens design, and other cases have not yet been investigated.

The reflexions from the stack of metal plates which forms the basis of a metal-plate lens have been examined theoretically by Carlson and Heins.[22] Their results are very complicated except in the simplest cases, because of the constraint imposed by the plates on the direction of the refracted wave. The results

where R has the value given in Equation (52)

d is the thickness of the sheet

and λ' is the wavelength measured within the dielectric.

The value of R' oscillates between 0 and $4R/(1 + R)^2$ as the thickness d varies, while the transmission coefficient T will show corresponding oscillations between 1 and $(1 - R)^2/(1 + R)^2$. One complete cycle of this oscillation occurs for each change in d of $\lambda'/2$.

The discussion so far given in this section has assumed interfaces of infinite extent and plane waves. The results may, however, be applied at any point on a lens surface provided that the curvature of the surface is not excessive. In a lens, the position is more complicated than for a simple plane sheet, but a fuller analysis shows that the transmission coefficient of the lens also oscillates from point to point as the lens thickness changes. The mean value at any point is approximately the product of the transmission coefficients of the two interfaces.

The occurrence of reflected waves has several effects upon the performance of a lens. Firstly, since much of the reflected energy is thrown backwards, there will be a reduction in the amount of the power radiated in the forward direction and so a loss of gain. For a lens of refractive index 0.6, the value of R given by Equation (52) is 0.0625 and so the average transmission coefficient through the lens surfaces will be approximately $(0.9375)^2$, i.e., 0.88 approximately. The variation of the transmission coefficient has been neglected in arriving at the above value, and it is found that when this variation is taken into account the average value over the whole surface is less than 0.88. Even this conservative allowance of the effect of reflexion corresponds to a drop in gain of 0.5 db., which is sufficiently large to be of importance in many applications.

Some of the reflected energy will be picked up by the primary feed, and causes a mismatch of the oscillator supplying the radiated power. Such a mismatch has an adverse effect on the oscillator, particularly if it is a magnetron, as will often be the case. The amount of energy picked up in this way by the feed is much smaller than for a paraboloid, but in one respect it is more troublesome. The focal length of a lens is much greater than for a paraboloid of the same aperture size; for a given change in wavelength, the phase difference between the outgoing and reflected waves will therefore alter more for the lens than for the paraboloid. This means it is more difficult to match the feed over a given range of wavelengths for the lens than for the paraboloid.

The last effect of the reflexion arises from the oscillations of the transmission coefficient caused by the varying lens thickness. It has been shown theoretically that these oscillations are responsible for the far-out side-lobes mentioned in Parts II and III. Besides the alteration in the shape in the radiation pattern, these side-lobes represent a further loss of power from the main beam.

Reflexions in optical lenses have been widely studied and methods of removing them have been developed. A widely used technique is known as "lens blooming" and consists of coating the lens surfaces with a layer of material of suitably chosen refractive index and thickness. When the analogy between transmission lines and free space propagation is invoked, it is found that this coating process is essentially the same as inserting a quarter-wave transformer to achieve a match between two transmission lines of different characteristic impedances. An application of this process to microwave lenses would remove all the undesirable effects of reflexion which have been discussed above. The way in which it may be done for the various types of lenses will be described below.

With dielectric lenses, the process may be carried out directly by using two materials whose refractive indices satisfy the required relation that that of the matching layer must be the square root of that used for the lens proper.[8] An alternative procedure is to machine slots of approximately a quarter-wavelength depth in each face of the lens; this slotted portion will behave similarly to the matching layer.

The refractive index of a strip dielectric has been shown in Part II to depend on the strip dimensions and spacings, so that by altering one of these dimensions the refractive index may be adjusted to the value required for the matching layer. The most convenient method of doing this is to increase the gap between the strips, while keeping the lattice dimensions constant. Cohn[11] has confirmed experimentally that a delay lens, matched in this way, produces smaller reflexions than an unmatched lens. A similar technique may be used to match a rodded lens, the diameter of the rods being reduced in the matching layer.

Metal plate lenses are more difficult to match than any of the above types, and so far no satisfactory practical solution has been obtained. One possible method, similar to that of slotting the surfaces of a dielectric lens, is to serrate the edges of the plates. Other possible methods make use of the analogy to the method of matching the junction between two transmission lines by a suitably positioned reactance. Ways of obtaining this reactance are to insert irises within the plates, as is done in wave guides, to place wires of suitable diameter parallel to the edge of each plate, or to insert dielectric blocks or wedges within the plates. One solution, which is attractive mechanically, is to use the spacing rods, which are required to maintain the necessary distance between the plates, as the reactance.

It is thus theoretically possible to match any of the types of lenses discussed in previous articles, but doing so considerably complicates the design and construction of the lenses. In certain cases it may be sufficient to avoid only one of the effects of reflexion.

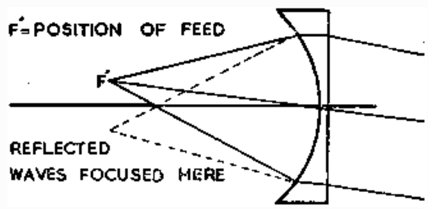


Fig. 31. Reduction of mismatch of feed : first method

Frequently, the second effect of those discussed above, i.e., the mismatch of the source caused by the reflected energy picked up by the feed, is the most objectionable and fortunately it is the one which can be most readily diminished.[5] There are two simple ways of doing this. Firstly the feed may be situated slightly off the axis of the lens as shown in Fig. 31, so that the reflected waves are focused at a point different from the position of the primary feed. Secondly, one half of the lens may be displaced a quarter-wavelength relative to the other as shown in Fig. 32. This displacement will have a negligible effect on the shape of the radiated beam, but will cause the reflexion from the two halves of the lens to be exactly out of phase at the primary feed and so to cancel each other out. In neither of these two methods is there any reduction in the total amount of energy reflected by the lens: the object of the methods is to redistribute the reflected energy in the least objectionable way.

It is always simpler to match a plane surface than a curved one and this may be done for the types of lenses which have one plane face. Matching one surface will reduce the total amount of energy which is reflected, thus increasing the gain of the lens system. It will also remove the oscillation in the transmission coefficient across the outer surface of the lens, since there are no longer two reflected waves which may interfere with each other. The far outside lobes in the radiation pattern, which are caused by the presence of such an oscillation, will therefore be removed.

Ideally, all the energy radiated by the primary feed is focused by the lens, but in practice some of the energy will propagate round the edges of the lens. The energy lost in this way is known as spillover.

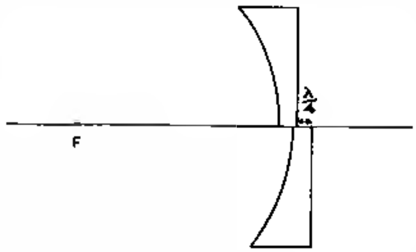


Fig. 32. Reduction of mismatch of feed : second method

It represents a loss in the gain of the lens and will also cause a slight increase in the far-out side-lobe level. This spillover may be reduced by making the primary feed more directive, but if this process is carried too far, the primary feed will become so directive that only a portion of the lens will be illuminated. Some compromise must therefore be found, and the most satisfactory one in practice is to make the directivity of the feed such that the amplitude distribution across the outer face of the lens is so tapered that it is 10 db. lower at the edges than at the centre of the lens. With such a tapered distribution, the spillover loss will not be excessive and the level of the largest side-lobe in the radiation pattern will not exceed about 22 db. below the level of the main beam. Spillover loss may be avoided altogether if the lens is inserted at the mouth of an electro-magnetic horn, whose aperture size must, of course, be the same as that of the lens. When this is done it follows, from the properties of horns, that the distribution across the outer face of the lens will be uniform in the direction of the electric vector and tapered in the direction of the magnetic vector. A system of this type is known as a lens-corrected horn and is superior to an ordinary horn in that larger flare angles are possible, thus reducing the total length of the system. Lens-corrected horns have effective radiating apertures of about 80 per cent of their physical apertures[6] and are among the most efficient aerial systems at present available.

Wide-angle Scanning

This section will commence with a discussion of the aberrations arising when the feed of a plano-concave metal plate cylindrical lens is moved off the axis. Only the type of lens in which the waves within the lens are constrained by the plates to travel parallel to the axis will be considered for the moment. This type has been selected as the calculations involved in finding the aberrations are much simpler than for any of the other types. Suppose then that the feed is moved to the position F′ as shown in Fig. 33. The phase difference Φ degrees between any point P on the outer surface of the lens and the central point O′ is given by the expression

$$\Phi = 360\left[\left\{(f\cos\alpha - z)^2 + (x - f\sin\alpha)^2\right\}^{\frac{1}{2}} + nz - f\right]/\lambda \quad (54)$$

where x, z are related by the equation for the lens contour, Equation (32).

The expression on the right-hand side of Equation (54) may be expanded in powers of x giving

$$\Phi = -360\left[x\sin\alpha + \sum_{n=2}^{\infty} A_n x^n\right]/\lambda \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (55)$$

where the coefficients A_n may be determined from

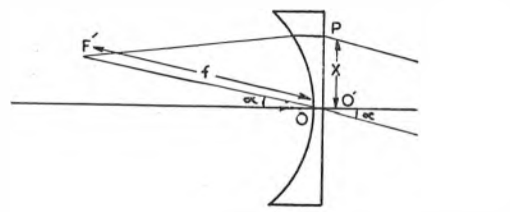


Fig. 33. Scanning aberrations of a constrained cylindrical metal plate lens

Equation (54). The first term of this, i.e., $-360\ x\sin\alpha/\lambda$, is the phase distribution required to move the maximum of the main beam radiated by the lens to a direction making an angle α with the axis. The remaining terms are unwanted errors and may be separated into two sums, which involve only even and odd powers of x, respectively. The first of these sums gives the magnitude of the spherical aberration, while the second gives the magnitude of the coma. Spherical aberration causes a reduction in the gain of the lens, an increase in the beam-width of the radiation pattern and an increase in the size of the side-lobes relative to the main beam. The coma term is asymmetric, i.e., the value for a negative value of x is equal in magnitude but opposite in sign to that for the corresponding position value of x, and it will therefore cause the radiation pattern to be unsymmetrical about the direction of the maximum.

The magnitude of both errors increases as the angle is increased and will eventually become so large that the radiation pattern will no longer be acceptable. In other words, the spherical aberration and coma place a limit on the maximum value of the angle α, which the beam may be moved off axis. As a reasonably accurate guide, it may be said that the errors at the edge of the lens must be less than 60° for the spherical aberration and 90° for the coma term. If the primary feed is moved on the arc of a circle of radius equal to the focal length and centre the point O in Fig. 33, the spherical aberration will reach its maximum permitted value before the coma term. It may be shown that the maximum value of α depends on the focal number **F** of the lens and the refractive index, typical values for a lens of the type so far discussed being 15° for $F = 1.5$ and $n = 0.6$. For a given value of n, the maximum angle of scan will increase as the focal number increases. The nature of the radiation pattern at the maximum value of α is shown in Fig. 34, for the 8-ft. metal-plate lens discussed in Part III.

In view of the advantages to be gained in moving only the feed to scan the radiated beam, it is of importance to obtain the design of a lens which will

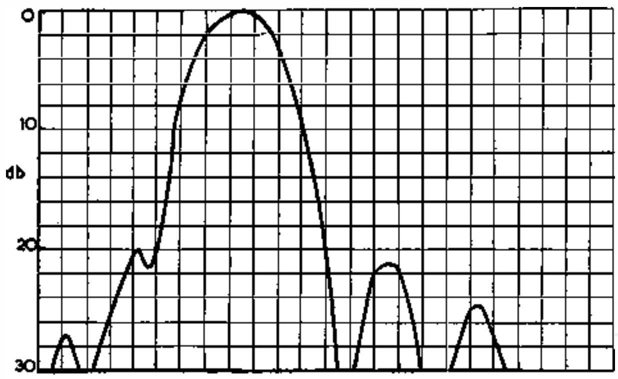


Fig. 34. Radiation pattern when the feed is moved off axis

produce the maximum scanning angle. As was pointed out in the previous paragraph, the scanning angle is limited by the spherical aberration when the feed moves on the circular arc. It is possible to reduce the spherical aberration by moving the feed along the line FO, i.e., by refocusing the lens. In the particular case so far discussed, the scanning angle may be slightly increased in this way. Although the discussion so far has been limited to the constrained type of cylindrical metal plate lens, it may be shown that other types of cylindrical lens have similar scanning properties. The maximum aberrations in spherical lenses occur at the corners, and so the maximum scanning angle is in general less than for the corresponding cylindrical lens of the same F number.

The second way in which the scanning properties

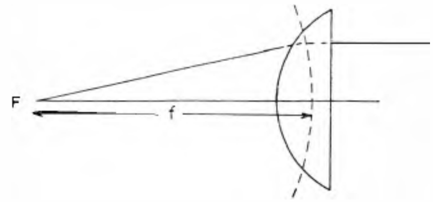


Fig. 35. Design of a dielectric lens with optimum scanning properties

of a lens may be improved is to redesign the lens contour, allowing both surfaces to be curved. This problem has been tackled by Friedlander[13] for a dielectric lens, by making use of a theorem in optics, known as the Abbé sine principle. This states that the lens contour, with the best scanning properties, is such that the ray paths from the focus to the back surface will intersect the ray paths emerging from the front surface on a circle centre the focus and radius the focal length, as shown in Fig. 35. The calculation of the lens profile, which will satisfy the sine principle, is very complicated and involves the numerical integration of a differential equation. Friedlander has shown, however, that if the refractive index of the lens is approximately 1.6, a plano-convex lens, which has its plane surface on the same side as the feed, is a sufficiently good approximation to the ideal solution.

(To be continued)

References

[11] J. Stratton, : " Electromagnetic Theory," McGraw-Hill Book Company, 1941, p. 490.
[12] J. F. Carlson and A. E. Heins : " The Reflexion of an E.M. Plane Wave by an Infinite Set of Plates," *Quart. Applied Math.*, 4, p. 313, Jan., 1947.
[13] F.G. Friedlander : " A Dielectric Lens for Wide-Angle Beam Scanning," *J.I.E.E.*, Vol. 93, Part III A, No 4. 1946, p 658

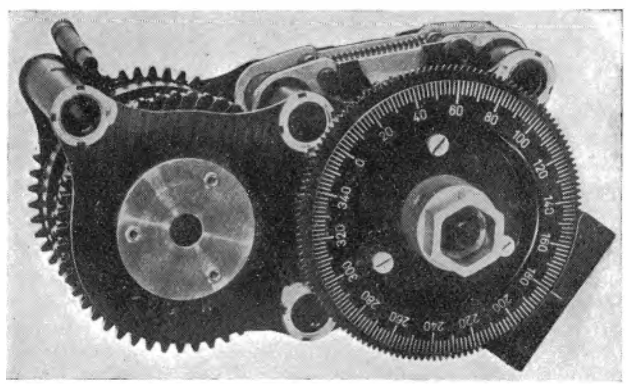
"Autoselector" Automatic Tuning Mechanism

THE Equipment Division of Mullard Electronic Products, Limited, Century House, Shaftesbury Avenue, London, W.C.2, has secured from Messrs. Philips Telecommunications Industries the U.K. sales, design and manufacturing rights of a new "Autoselector" automatic tuning and selecting mechanism.

These autoselectors are used to turn shafts repeatedly to pre-set positions with speed and accuracy. Automatic electrical remote control is employed, and the setting accuracy is such that the positions selected never differ by more than $\pm$ 0.050° from those which have been pre-set in the mechanism.

There are many applications of these new autoselectors, both in radio engineering and in industry. For example, they enable the frequencies of radio transmitters and receivers to be speedily and accurately selected. In industry, they may be a valuable asset where accurate machine tool operations have to be repeated many times, for example, on lathes and on grinding and cutting machines.

The speed of operation of autoselectors is such that, using small transmitters, the system reaches the required setting within two seconds of the operation of the selector switch. Using larger transmitters, the setting time is less than ten seconds. As a result of this, tuning adjustments are much more rapid than with manual tuning.



Two types of autoselectors are at present available. These are similar in design but are of different capacities. The small type SZT.101 can transfer a maximum torque of 2.4 lb.-in. (3 kg.-cm.) and is intended for small apparatus, such as aircraft radio equipment. It can stop a shaft in any one of 12 pre-set positions. The large type SZT.102 can transfer a maximum torque of 12 lb.-in. (15 kg.-cm.) and is used in larger apparatus. This type can stop a shaft in any one of 11 pre-set positions, and has additional facilities for manual adjustment of the controlled spindle. Autoselectors capable of transferring a maximum torque of 80 lb.-in. (100 kg.cm.) will soon become available.

Noise in Variable Resistors and Potentiometers

By FRANCIS OAKES

VARIABLE resistors and potentiometers are widely used in all branches of electronic engineering, for many different purposes and in a great number of circuits. But while the resistance range for normal applications extends from a few ohms up to several megohms, the majority of such variable resistors and potentiometers are of similar physical size and essentially the same construction.

Apart from the wire-wound rheostats and potentiometers used for special purposes, the carbon track potentiometer with an approximate size of $\frac{3}{4}$ in. to 2 in. diameter and a load-carrying capacity of the order of one watt is used to cope with the majority of electronic applications. To use such a potentiometer as a variable resistor, one of the extreme terminals is simply left disconnected. The most common design is characterized by a carbon track applied on a paper or plastic base in the form of an arc of about 270-300° (sometimes more) with a metal sliding contact moving over the arc. Another design makes use of a "squash plate" of thin metal in the shape of part of a conical surface, brought into contact with the carbon track along one moving radial line by means of a movable wiper arm. A recent German design makes use of a small drop of mercury held in a cup and moved along the carbon track as a liquid contact of low-noise and non-wear characteristics.

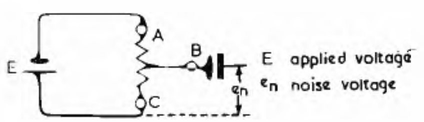


Fig. 1.
Circuit for noise testing

Resistance-Rotation Characteristic and Rotational Noise

Whether or not a current is passed through the terminations, noise voltages appear across any two of the three terminations, and so long as the rotor is stationary, the noise phenomena observed do not differ from those of fixed resistors. When the rotor is turned and the moving contact passes over the resistance track, additional noise is produced different from the stationary noise in magnitude and frequency distribution. The magnitude, for normal potentiometers, is of the order of about one or two millivolts per volt applied across the extreme terminations A and C in a typical case such as illustrated in Fig. 1, and for a speed of rotation of approximately one full rotation per second. This magnitude chiefly depends on the type, design and condition of the potentiometer, and also on its resistance value, the speed of rotation and the loading current.

The characteristic of a variable potentiometer most significant with respect to its influence upon noise characteristics is the resistance-rotation characteristic. This characteristic depicts the percentage variation of resistance between the "anti-clockwise" terminal C and the moving terminal B as a function of percentage of clockwise total rotation. Fig. 2 shows the resistance-rotation characteristic of a linear-law potentiometer, i.e., of a potentiometer where the voltage across B-C is proportional to the amount of clockwise rotation when a constant voltage is applied across A-C.

For the use as a volume control or gain control, however, it is desirable that the voltage across B-C should vary approximately as an exponential function of the amount of clockwise rotation, therefore the logarithm of the resistance between the "anti-clockwise" terminal C and the moving contact terminal must be approximately proportional to the percentage rotation. This is in consequence of Weber-Fechner's logarithmic law of sense perception, according to which the intensity of sense perception varies as the logarithm of the amplitude of the stimulus. The resistance-rotation characteristic of such a logarithmic-law potentiometer is shown in Fig. 3. A significant parameter for the classification of potentiometers as to the nature of their resistance-rotation characteristic is the percentage of total resistance across B-C for 50 per cent rotation. For a linear-law potentiometer this is 50 per cent and for a logarithmic-law potentiometer such as shown in Fig. 3 it is of the order of 10 per cent.

The gradient at any point of the resistance-rotation characteristic is a measure of the change of resistance per change of setting, and also a measure of the voltage gradient along the track at this particular point. This is of great importance for the noise behaviour of a potentiometer, in fact, the noise produced at any point of the track when passed by the moving contact will naturally be approximately proportional to the voltage gradient at this point, because of the dependence of noise generation upon the voltage load across the resistance element. For example, in the case of Fig. 3, only 10 per cent of the voltage applied across A-B appears across the first 50 per cent of the resistance track, whereas the remaining 90 per cent appears across the remaining half of the track.

It follows from these considerations that the rotation-noise characteristic, i.e., the characteristic showing the distribution of noise over the rotational range, will be uniform for a linear-law potentiometer; for a non-linear potentiometer, the noise produced over the first half of the rotational range will be much smaller than for the remainder of the range. A rotation-noise characteristic is shown in Fig. 4 for a linear-law and in Fig. 5 for a logarithmic-law potentiometer.

Such characteristics can be obtained by photo-

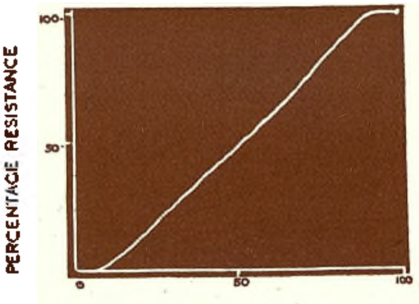


Fig. 2.
Resistance-rotation characteristic of linear-law potentiometer

graphic recording of an oscilloscope trace. The potentiometer is connected with its terminals *A* and *C* across a source of D.C. such as a 4.5 volt battery and the terminals *B* and *C* to the grid and earth terminals, respectively, of a high-gain wide band amplifier. The output of the amplifier is connected to a cathode-ray oscilloscope. No time-base is used, and the trace is recorded by means of the rotary drum camera shown in Fig. 6. Another simpler arrangement which, however, does not permit the recording of long extended traces as does the rotary drum camera, makes use of an ordinary camera, with the Y-plates coupled to a wire-wound precision potentiometer ganged to the spindle of the potentiometer on test. The arrangement is shown in Fig. 7. This arrangement offers the advantage of simplicity in operation and of an accurate co-ordination of noise and the area where this noise is produced.

The long strips of photographic record obtained by the first method have been cut in several sections for ease of reproduction. The timing marks on the top edge of the film strip denote 1/50 sec. and have been obtained by means of 50 cycle A.C. on the second electron beam of the double beam tube used for recording. The most striking feature in comparing the two types of potentiometers is the very small amount of rotational noise in the lower half of the logarithmic-law potentiometer. This fact is not generally realized, but should be obvious from the following considerations: As only about 10 per cent of the voltage is dropped over the first half of the track, the voltage gradient and the noise are only 1/10 of that for a linear-law potentiometer. The remaining half, taking approximately 90 per cent of the load, has approximately double the voltage gradient and double the noise level as compared with a linear potentiometer. This is a decided advantage for such applications as for volume control, because in a radio receiver the first half is normally used on low volume level and for local stations where little background noise is expected. When turned further clockwise into the "steep" region, the increase in noise level in case of distant station reception or the increase in output for local stations will retain a high audio-signal to noise ratio.

Causes of rotation noise

The causes for noise generation in a potentiometer during rotation are manifold. Close inspection of the resistance-rotation characteristic will reveal minute deviations from a straight or smooth line, and a certain amount of low frequency noise will be due to these irregularities alone when the contact is moved along. When the mechanical characteristics of the track surface and of the contact or squash plate are investigated under the microscope, these surfaces appear to have far from ideal smoothness,

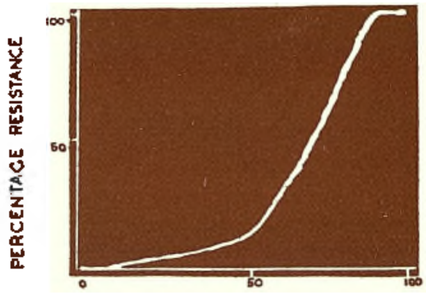


Fig. 3.
Resistance-rotation characteristic of logarithmic-law potentiometer

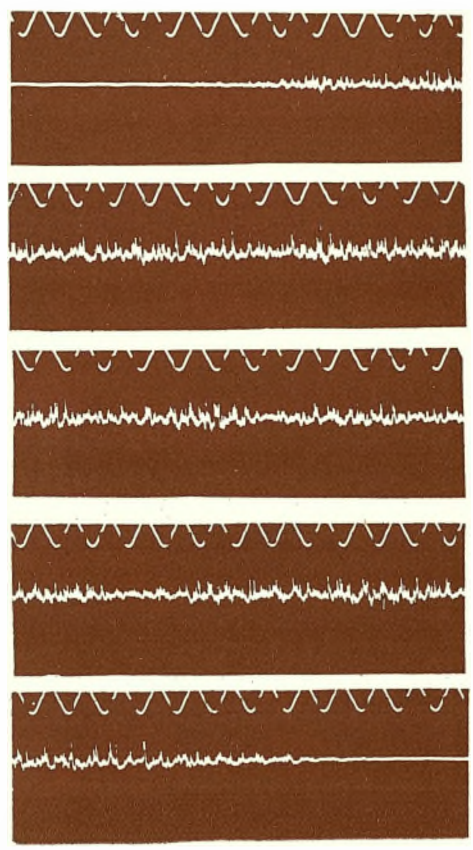


Fig. 4.
Rotation-noise characteristic of linear-law potentiometer

and unless the potentiometer is absolutely unused, a certain amount of scoring along the line of contact can be observed. Suitable design of the sliding contact is likely to reduce scoring to a tolerable amount and to counteract undesirable effects of surface roughness of the resistance track. In some cases this is done by a rounded contact of approximately cylindrical shape in the neighbourhood of the contact region, which will make contact with the track along a radial line. Other designs make use of a flat contact disk flexibly mounted so as to have a large number of "points" of contact with the resistance element. In each case, the contact pressure will have to be controlled between suitable limits because excessive contact pressure is always responsible for undue wear, whereas a certain minimum pressure is necessary to keep noise and contact resistance as low as possible under this circumstance of compromise. It is often unjustified to conclude that noise is necessarily produced by surface roughness, or that the potentiometer with the rougher track surface is always the one with greater noise. In actual fact, it is always a combination of surface and contact, and in the case of track roughness, of the particular *type* of roughness. On the other hand, it is usually correct to assume that a coarse surface of the sliding contact will result in noisy performance and undue wear.

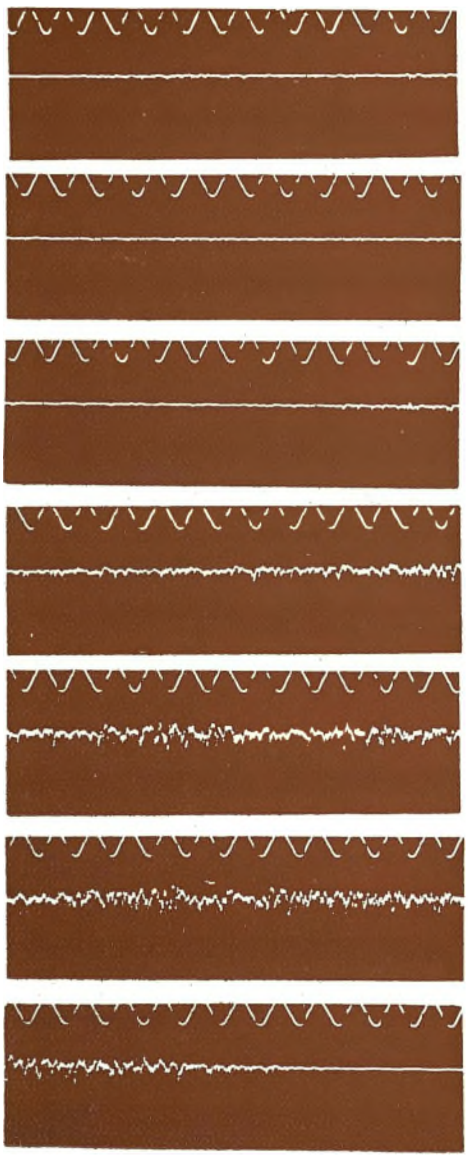


Fig. 5.
Characteristic of logarithmic-law potentiometer

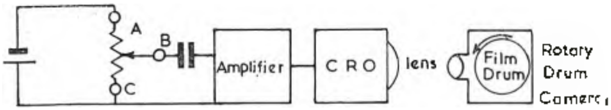


Fig. 6.
Recording noise traces with a rotary drum camera

The squash-plate type of moving contact is designed to avoid all lateral movement of surface against surface, thereby avoiding sliding action and reducing wear. The plate is distorted by a wiper moving on top of the plate and pressing it down on to the resistance track in the fashion of a cone rolling about its apex on a plane. Contact is made along a moving radial line. Noise is generated by virtue of the fact that parts of the track are suddenly short-circuited by the squash plate descending on to a slightly protruding part of the surface further along the track, by small particles of dust on the plate or resistance-element and by the irregularities of the resistance-rotation characteristic. Whereas the problem of wear is considerably reduced by this type of moving contact, very great care has to be taken to eliminate surface flaws, and to prevent dust particles from location between the contact surfaces.

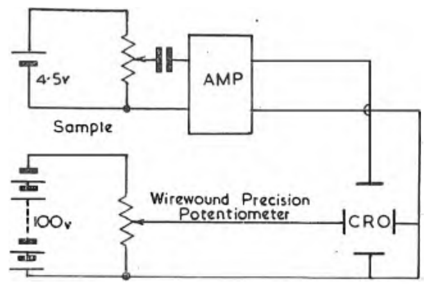


Fig. 7.
Recording noise traces with a stationary film and X deflexion by a standard potentiometer

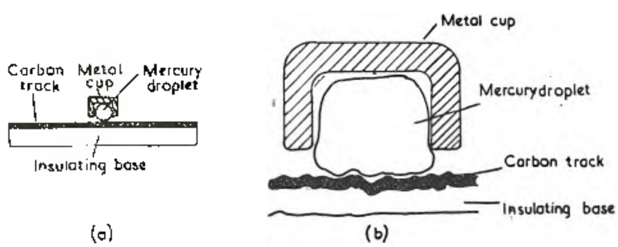


Fig. 8.
Mercury contact type of potentiometer

The mercury contact illustrated in Fig. 8 is capable of reducing wear to a minimum as well as avoiding undue noise generation by surface roughness. A certain amount of noise is generated, however, by the advance of the mercury drop, which naturally must not wet the cup or surface, and therefore in its advance acts in a fashion combining that of the wiper and that of the squash plate. A schematic view of the advance of the mercury drop is shown in Fig. 8(a).

The phenomena described so far refer to potentiometers in good condition and of efficient design. In the case of faulty components, very greatly increased noise can often be found due to aggravation of surface defects by scoring, wear or climatic and chemical action, sometimes resulting in momentary open circuits, which cause considerable noise voltages. Other causes of noise are bad connexion of the extreme terminations or poor contact in the slip ring or other device used to connect the moving contact to the stationary centre terminal.

Frequency Distribution

The frequency distribution of the noise produced by a potentiometer on rotation varies greatly with the cause of noise. In distinction to the noise produced by fixed resistors, this distribution is very irregular and usually contains a far greater percentage of lower audio frequency contributions. This is illustrated in Fig. 9, of which the top trace was taken with a potentiometer on fast rotation and the bottom trace with a fixed resistor and increased amplifier gain setting.

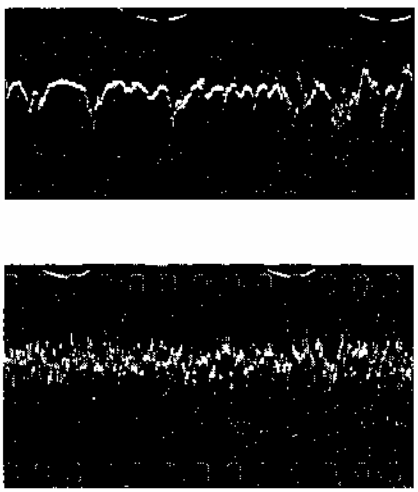


Fig. 9.
Noise traces of fast moving potentiometer (top) and fixed resistor. Amplitudes not to scale

The potentiometer oscillogram in Fig. 9 is a centre portion representing one-fifth of the total length of a record made when turning the rotor through its full course in about 1/5 second. The amplitude and frequency distribution of the noise produced on fast rotation is different from that produced on slow rotation. Normally, the amplitude is greater and the distribution more uniform on fast rotation.

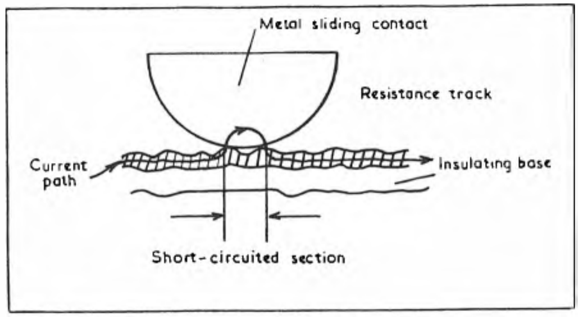


Fig. 10.
Noise generation by intermittent short-circuiting of track elements

It is interesting to observe that repeated operation of a potentiometer in good condition yields fairly consistent noise records, but that the record taken for clockwise operation is often greatly different from that for anti-clockwise operation, both in amplitude and distribution of frequencies.

Certain wiper-type sliding contacts show distinctly non-symmetrical noise traces. Fig. 4 is a good example of this. The direction of sharp peaks is seen to be upwards in this case, and would be reversed for reversed direction of sliding contact movement. An explanation of this phenomenon can be found in the mechanism of contact movement. "Sticking" action due to surface friction and elasticity of the moving contact assembly and "jumping down" from high spots due to spring pressure can be responsible. "Saw-tooth" roughness of the surface cannot be made responsible for this type of asymmetrical noise trace because this would not account for reversibility of noise peaks on reverse rotation of the potentiometer spindle.

Mechanical Considerations

A number of mechanical considerations are involved in the investigation of rotational noise generation. Most important of these is the fact that contact is never made on a single point, along a line or on a single integral surface. Therefore, in the advance of the moving contact, parts of the resistance path are continually short-circuited and also "switched" into circuit. This can be understood from inspection of the schematic drawing Fig. 10. While the moving contact slides over two "hilltops", the "valley" is short-circuited and, at least for this area, no resistance change or voltage change takes place for a small incremental movement. When a "hilltop" is left or when another "hilltop" is touched in the advance of the moving contact, a large resistance and voltage change is caused by a small incremental movement. Simultaneously, the current carried by the resistance element in the case of constant voltage across its terminals (or the total voltage across the resistance element in the case of constant current) undergoes a corresponding variation in consequence of the variation of the total resistance caused by the mechanism of switching parts of the resistance in and out of circuit. With a smaller surface of contact this phenomenon is reduced; this, however, is done at the expense of increased specific contact pressure and consequently increased wear of the surface. It is for that reason that extension of the contact surface is generally sought in a radial direction. In this way the switching phenomena are carried out in parallel with one another and their detrimental effect in producing voltage fluctuations is effectively reduced. Another reason for reducing the specific contact pressure as far as is compatible with a reasonably high absolute contact pressure is to avoid undue local deformation of the track by the passage of the moving contact. The latter considerations are more important for sliding solid contacts than for plate or mercury contact types.

An important problem is that of surface lubrication. Efforts are sometimes made to improve performance by the application of grease lubricants, but their effect, though temporarily beneficial, usually has its adverse consequences in shortening the life of the resistance element by chemical or physical action on the lacquer or base.

Wear is normally the largest single factor impairing the noise characteristics of a variable potentiometer. Consequently, careful consideration must be given to this problem in manufacture and development of new types. The common practice is to test samples of potentiometers by means of a cycling test. This is carried out in turning the potentiometer spindle from extreme starting to extreme stopping position and back in a suitable testing machine capable of carrying out tens of thousands of such operations within one or two days. The noise produced is measured at suitable intervals, and in this way the wear-noise characteristics of a potentiometer type can be investigated.

Three regions on the track of a potentiometer warrant particular consideration from the point of noise generation. These are the starting region at the anti-clockwise extreme, the stopping region at the clockwise extreme and the juncture of the low-resistance and high-resistance sections in the case of a logarithmic-law potentiometer. The first two mentioned are characterized by short areas of low-resistance coating which serve to provide a reliable contact between the extremes of the resistance track and the terminals *A* and *C*. Here, with a well-designed potentiometer, very little noise is generated. Bad design, on the other hand, can be responsible for excessive noise generated in this region alone, even where the remainder of the resistance track gives satisfactory operation. For a potentiometer with as widely different resistance changes per degree of rotation as is necessary for a logarithmic law, the resistance element or the resistance coating is made of two or more sections of different specific resistance. The region of juncture normally is characterized by noise generation half-way between that of the adjacent sections.

For use in electronic circuits, potentiometers are often made fixed in order to ensure low-noise performance. On the other hand, modern variable potentiometers provide a means for smooth control and much can be done by careful selection of the correct position within the circuit of application to ensure satisfactory operation. For use where current is passed through the moving contact more care must be taken as the amplitude of fluctuations is greater for such cases. Where a voltage rather than a current is controlled, a potential being tapped at a variable ratio by means of the moving contact, current should not be passed through the potentiometer resistance element unnecessarily. In many cases it is possible to block the D.C. when only an A.C. potential is required, such as in the case of audio-amplifier volume control, where low noise is essential. Another method sometimes helpful in obtaining a reduction of noise is to shunt a potentiometer by a fixed resistor carrying the greater portion of the current while the variable proportion of voltage is obtained from the potentiometer at a low noise level. Care must be taken for correct estimation of signal-to-noise ratio for negative feed-back applications as shown in the circuit Fig. 11. This circuit is sometimes used to provide gain control in high-gain wide-band amplifiers. When calibrating to the required gain it is often found that the deflexion on the output meter is erratic due to noise produced by this variable resistance while it is moved. This severely hinders the process of calibration. The difficulty can be overcome by replacing the variable resistor by a fixed resistor (or several steps of fixed resistors) and replacing the fixed resistor *R* by a variable resistor. In this way, only a fraction of the standing current will pass through the variable resistor, and the generation of noise will be reduced considerably. For application in tone control circuits similar precautions are advisable where a high ratio of amplification follows the tone control stage. In radio-frequency circuits where the use of variable potentiometers and resistors is normally restricted to parts of the circuit not carrying a signal current, as, for instance, in the control of variable-mu valves, rotation noise is of little consequence.

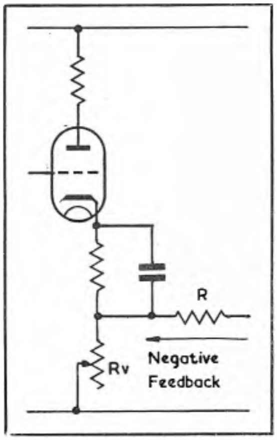


Fig. 11.
Variable negative feedback circuit, sometimes unsuitable because of noise

For many applications, as, for instance, that of volume and tone control in radio receivers, the problem of rotation noise has still to be solved. While in new equipment trouble is rarely experienced from this source, noise often appears as a consequence of wear. This trouble does not only arise when the potentiometer is actually operated by hand, but in many cases vibrations transmitted mechanically and acoustically to the elements of the potentiometer can give rise to audible noise or distortion. With the new types of potentiometers now in production, or in a state of development, it is hoped that these phenomena will be reduced to such an extent that eventually, at least for radio receiving and television equipment, their noticeable effects will be eliminated.

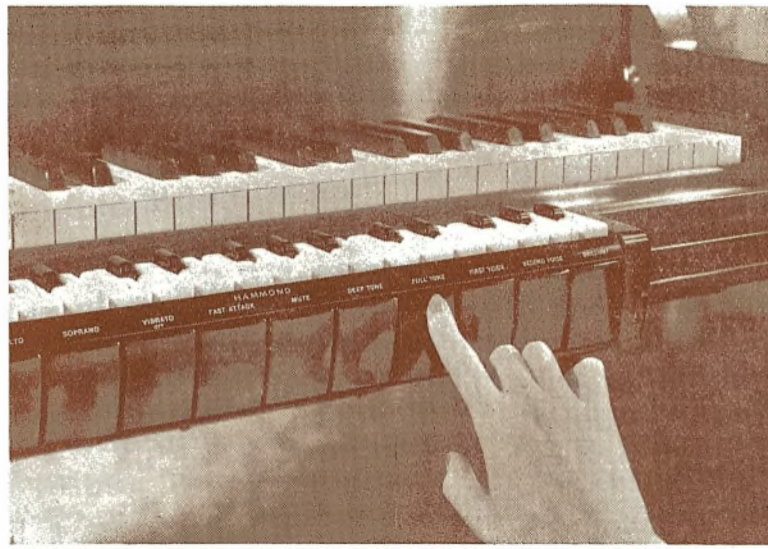
RESULTS ON A MIDLANDS "TELEVISOR", AT 35 MILES RANGE

Snapped by a Home Constructor

The Solovox

A Novel Electronic Musical Instrument

By ALAN DOUGLAS*



THIS device is one of the products of the Hammond Instrument Company of Chicago. It is a purely melodic instrument, that is, one on which only one note at a time can be played, although more than one octave of any one note can be sounded simultaneously. In consequence, it requires that some other musical instrument should accompany it, and since the most probable choice is a piano, the Solovox is designed to attach directly to the front of a standard piano frame.

Playing keys and controls

The compass of the playing clavier is three octaves, terminating on top B; to extend it to C would require extensive circuit changes. The keys are of standard width and compass, but somewhat foreshortened so as not to interfere with a normal playing position for the performer. Each key is hollow, and contains the contact mechanism for closing the necessary circuits. The keys are moulded from Lucite, which is a plastic having the same coefficient of friction and the same moisture absorption as ivory; the "feel" is, therefore the same to the player.

In front of the playing keys is a series of rocking tablets (above). These control the various pitches and tone-colours which are available from the circuit. The intensity or volume level is controlled by a pivoted knee pedal, similar to that on an American organ. An initial displacement of this operates the starting switch.

Oscillation generating circuits

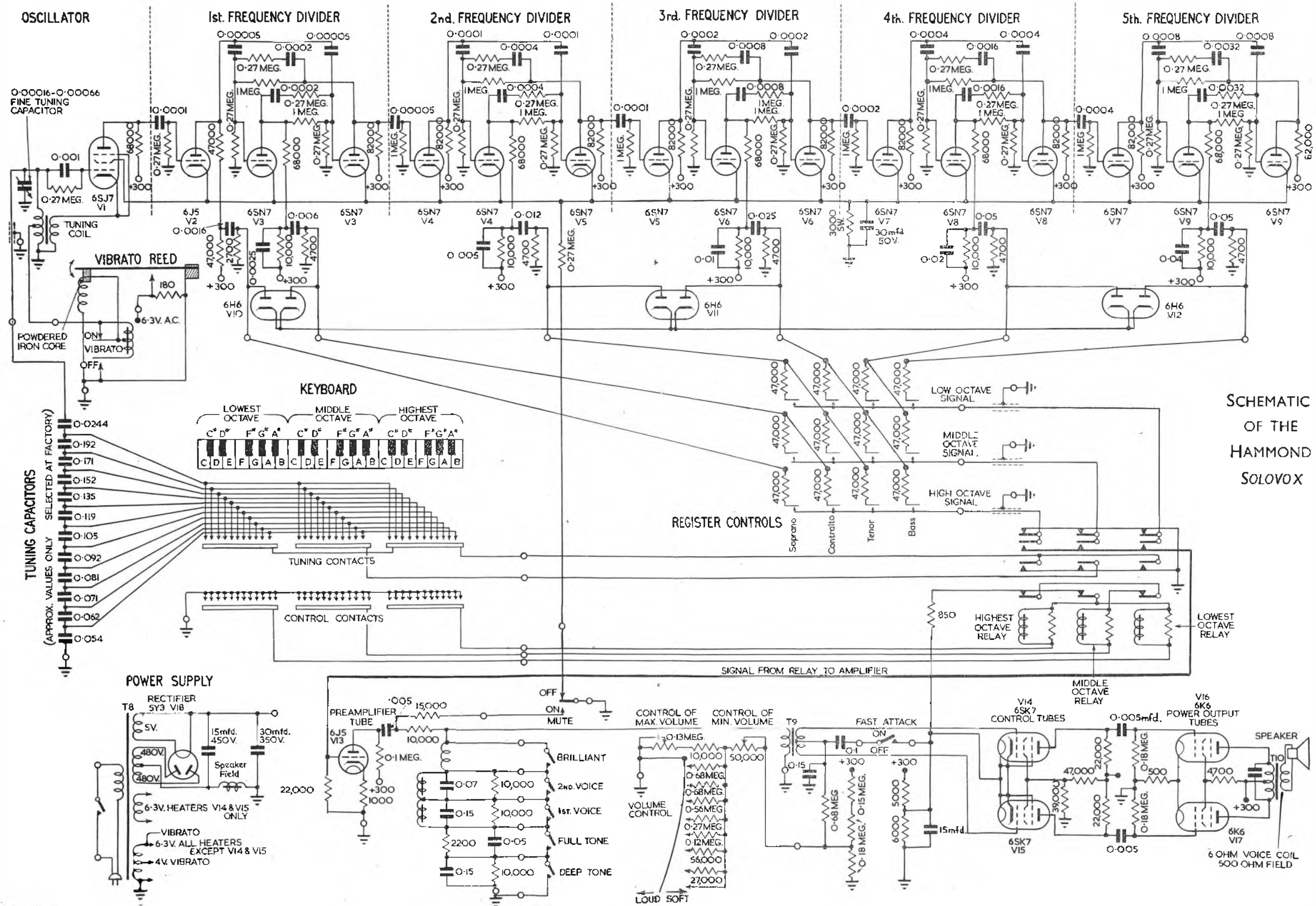
The circuit is shown complete overleaf. The instrument functions in the following manner. A single valve oscillator is normally tuned to the highest note obtainable on the Solovox. (B = 3,951 c/s., A being standard pitch 440 c/s.) When any playing key is depressed, the oscillator is re-tuned to that note frequency by alteration of the capacitor stack shown in the schematic. The oscillator output is fed to a series of cascaded frequency dividers, which produce notes in successively lower octave relationship. Thus for any one playing key fundamental, five other lower octaves are simultaneously available, so that, although only one note can be played at a time, all five octaves of that note can be made to sound at once, adding considerable boldness to the tone. Any or all of these octaves can be selected by some of the tablets shown.

There are a number of circuits for obtaining one output pulse from two input pulses.* Some of these are characterized by poor waveform, but the harmonic content of the Solovox dividers is entirely acceptable from a musical standpoint. This frequency divider consists of three triode valves; two of these are combined in one envelope in all stages except the first. The feedback circuits connecting the two triodes are of such a nature that only one valve can conduct at a time, for by drawing anode current it keeps the other valve cut off.

If, for example, the first triode conducts and the second is cut off, then a negative pulse on the grids of both triodes cannot affect the second one, which is already cut off, but will cut off the first. This produces a positive pulse at the anode of the first triode, which is applied to the grid of the second through the feedback network. The second triode then conducts, producing a negative pulse at the anode; this latter pulse is fed back to the first grid and thus cuts this valve off. This is the reverse condition to that which we imagined at the start of the cycle. Thus two input pulses produce only one output cycle, so halving the frequency. The output signal comes from one anode, the other being used to drive the next divider.

So far we have seen that if any note key is depressed, the oscillator is re-tuned. In order to prevent an off-pitch tone being generated if two keys are accidentally pressed together, the register selecting relays in the circuit diagram do not permit more than one circuit at a time to close. The series arrangement of the tuning capacitors automatically controls the tuning so that always the higher of the two notes will sound.

* *The Electronic Musical Instrument Manual*; A. Douglas. (Pitman).



SCHEMATIC OF THE HAMMOND SOLOVOX

Methods for forming tone-colours

The waveform from the dividers contains a wide series of even and odd harmonics, and is thus susceptible to tone-forming circuits. There are three means of altering the shape of the envelope of the generated oscillations. It can be seen that across each divider output a diode valve is shunted. If this is connected, both even and odd harmonics will be passed; if it is disconnected, the even harmonics are suppressed, thus favouring a hollow and rather clarinet-like sound.

In addition, a series of resonant and low-pass filter circuits is provided in the output circuit of the pre-amplifier. The formants thus available can be combined in many ways to produce novel effects. The inductance in the "Brilliant" circuit removes all the bass and produces a biting tone-colour.

Attack and delay circuits

The third form of control is the capacitor in the control valve cathode circuit marked "Fast attack". This modifies the rate of build-up of the signal.

In all circuits in which keying of the signal takes place, transients will occur in the loudspeaker unless means are taken to suppress them. A familiar American method is to use steep cut-off pentodes, but in this case the control valves are of the variable-mu pattern and volume control is effected by varying the grid bias. The grid voltage range for complete cut-off during keying must therefore be much greater than usual.

This circuit functions as follows. The cathodes of the control valves are connected to the voltage divider as shown. This cut-off voltage is of the order of +175 volts. When a playing key is pressed, one of the selector relays must operate, and one relay coil is thus put in parallel with the lower resistor of the voltage divider. This causes the cathode voltage to drop to +55 volts. The valve then conducts. The large capacitor shown removes any trace of abruptness in attack or decay.

Expression control and vibrato

The volume is adjusted by means of a series of fixed contacts which eliminates the wear encountered with a rotating wire rheostat. There are seven steps, together with adjustable limiting resistances for maximum and minimum gain, and the control valve grid potential is varied by this resistance arm from about +45 volts to almost earth potential (minimum volume).

It will be seen that a vibrato is provided by an A.C. operated tuned reed which inserts or removes a small part of the main oscillator tuning inductance at about 7 c/s. Adjustment of the tuning inductance is instantly carried out by means of a coin slot in the oscillator coil core, which is on the top of the chassis. The instrument keeps in tune for very long periods without attention.

By suitable combination of the various controls, some interesting effects are possible. For example, soprano, deep tone and brilliant produce a creditable string tone; tenor, vibrato off, mute and full tone give a good clarinet; contralto, vibrato off, fast attack and second voice, a rather pungent trumpet; tenor, vibrato off, second voice and brilliant, a horn-like quality; and so on. The possible combinations are, indeed, practically limitless.

The present dollar exchange position precludes the sale of this ingenious little instrument in this country, but a very similar device of French origin, known as the Clavioline, will shortly be produced here.

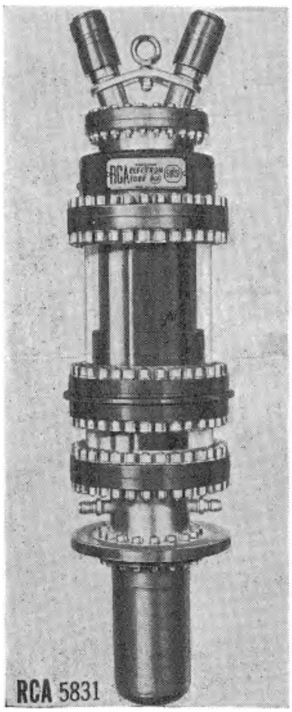
Super-Power Beam Triode

WHAT is believed to be the world's most powerful electron tube, a high-vacuum "super-power beam triode" capable of 500 kilowatts of continuous output and tested at one million watts input, has been announced by RCA.

The new valve, RCA Type 5831, represents the first commercial attempt to build power output of this magnitude into a single valve. Despite its enormous power input capabilities—at least four times that of any previous RCA type—the new 5831 is of compact design, measuring 38¾ in. in length and weighing only 135 lb.

In unmodulated class C service, the 5831 has a maximum plate voltage rating of 16,000 volts, a maximum plate input of 650 kilowatts, and a maximum plate dissipation of 150 kilowatts.

The new valve generates its power at high efficiency and with exceptionally low driving power. To provide 500 kilowatt output in unmodulated class C service, under typical operation condition, the 5831 requires only about 900 watts grid drive. The tube can be operated with maximum rated plate voltage and plate input at frequencies throughout the broadcast band, and much higher. The limitations for operation of the tube at high frequencies and at higher power have not yet been determined.



The outstanding feature of the 5831 is an array of 48 independent unit electron-optical systems arranged cylindrically in the tube. The great power capabilities of the new tube are due largely to the successful achievement of this design, which, in effect, concentrates 48 triodes in a relatively small space.

In detail, each of the independent electron-optical systems consists of a filament in a slot in the beam-forming cylinder, grid rods and the copper anode. Electrons leaving the emitting surface of the filament are beamed between a pair of grid rods to the anode by the focusing action of the beam-forming cylinder. Even though the grid may be positive, relatively few stray electrons strike it.

The Cable Transformer

By F. J. CHARMAN

A coupling network, made from lengths of standard radio-frequency cable, which is designed to couple a balanced 70-ohm feeder to a quarter wave aerial

It is well known that a quarter wavelength of transmission line will transform between Z_1 and Z_2 where the line surge impedance Z_0 is $\sqrt{Z_1Z_2}$.

It may be inconvenient to have two kinds of cable in use, but a transformation can always be made using a length of cable plus two reactors to form a kind of π coupler.

Working in admittances and normalizing (i.e., dividing out by y_0 to refer to a "one mho cable") we have in Fig. 1 a load g_2 (the aerial) which is shunted

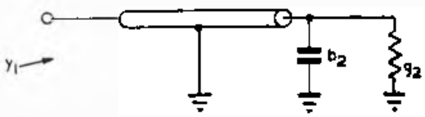


Fig. 1.
Cable terminated by complex admittance

by a susceptance b_2. The length of the cable is θ electrical angle and the admittance y_1 at terminal 1 will appear complex. The phase angle is corrected by the addition of a susceptance b_1, leaving a desired admittance $y_1 = g_1$.

The problem is to proportion b_2, θ and b_1 so that $g_1 = 1$, whereupon the line feeding end 1 will be correctly terminated.

The cable equation normally appears as

$$Z_1 = \frac{Z_2 \cosh \gamma l + Z_0 \sinh \gamma l}{Z_0 \cosh \gamma l + Z_2 \sinh \gamma l} \quad \dots\dots\dots \quad (1)$$

Dividing by Z_0, inverting, to get the admittance, normalizing, and dividing by $\sinh \gamma l$ we arrive at

$$y_1 = \frac{1 + y_2 \coth \gamma l}{y_2 + \coth \gamma l} \quad \dots\dots\dots \quad (2)$$

Assuming the line loss to be negligible, so that $\gamma l = j\theta$, we get

$$y_1 = \frac{1 - jy_2 \cot \theta}{y_2 - j \cot \theta} \quad \dots\dots\dots \quad (3)$$

Putting $y_1 = 1 - jb_1$, we get

$$(1 - jb_1)(y_2 - j \cot \theta) = 1 - jy_2 \cot \theta \quad \dots\dots\dots \quad (4)$$

Expanding and equating real and imaginary parts gives

$$\text{(Real)} \quad g_2 + b_1b_2 - b_1 \cot \theta = 1 + b_2 \cot \theta \quad \dots \quad (5)$$

$$\text{(Imag.)} \quad b_2 - \cot \theta - b_1g_2 = -g_2 \cot \theta \quad \dots\dots\dots \quad (6)$$

from (6) $b_1g_2 = (g_2 - 1) \cot \theta + b_2$

$$\text{or} \quad b_1 = \frac{(g_2 - 1) \cot \theta + b_2}{g_2} \quad \dots\dots \quad (7)$$

Substituting (7) in (5) will give $\cot \theta$ and θ.

Example. A quarter wave earthed vertical aerial is to be fed through cable which is normally made for

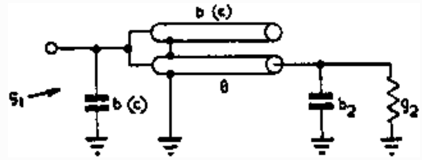


Fig. 2.
Cable transformer, showing alternative methods of phase-correction

feeding dipoles and which has a characteristic impedance of 70 ohms. The aerial impedance is 35 ohms. As a matter of interest let the frequency be 160 Mc/s.

In this case $g_2 = 2$. The value of b_2 is arbitrary, but, as an intelligent guess, let us make $b_2 = 1$. (At 160 Mc/s., this is equal to 14.3 micro-microfarads.)

Then, by (7), $b_1 = \dfrac{\cot \theta + 1}{2}$

and, in (5),

$$4 + \cot \theta + 1 - (\cot \theta + 1) \cot \theta = 2 + 2 \cot \theta$$

giving the quadratic $\cot^2 \theta + 2 \cot \theta - 3 = 0$

whence $\cot \theta = 1$ or -3 and $b_1 = 1.0$.

The length of line for $\cot \theta = 1$ is $\frac{1}{8}$ wavelength, and for $\cot \theta = -3$ is 0.45 wavelength.

The shorter length will be chosen since the 0.45 wave line would require an inductance at end 1 so that, at 160 Mc/s., using polythene cable, the length is therefore about 6 in. The element b_1 can be a fixed capacitor or another $\frac{1}{8}$ wavelength of open circuited line. The element b_2 can be made to include the aerial terminal-bushing stray capacitance and may be adjusted, possibly by shortening the line a little.

Now let us investigate the performance over a band of frequencies, assuming that the transformer is set up correctly at 160 Mc/s. and that the aerial resonance is very broad or, alternatively, that the aerial can be re-tuned. If a whip aerial is employed, it will be necessary to tune it by adjustment of its length since the necessary small diameter prevents it from having broad-band characteristics.

Fig. 2 shows the alternative versions of b_1, i.e., capacitor or line section.

Then at $f_0 = 160$ Mc/s.

(c) $b_1 = 1.0$ $\quad \cot \theta = 1.0$

(l) $b_1 = 1.0$ $\quad b_2 = +j1.0 \quad g_2 = 2.0$

and at $f_1 = 176$ Mc/s.

(c) $b_1 = 1.1$ $\quad \theta = 49.5 \quad b_2 = +j1.1$

(l) $b_1 = \tan 49.5° = 1.171$ $\quad \cot \theta = .854 \quad g_2 = 2.0$

$$g_1 = \frac{1 - jy_2 \cot \theta}{y_2 + j \cot \theta} + jb \quad \dots\dots\dots \quad (8)$$

using b_1 (c) $\quad y_1 = 1.06 + j0.34 = 1.12\angle -17\tfrac{1}{2}°$

and b_1 (l) $\quad y_1 = 1.06 + j0.41 = 1.13\angle -21\tfrac{1}{2}°$

The capacitive phase corrector is thus slightly better than the line equivalent, but in any case the resulting mismatch S.W.R. at the band extremities would not exceed 1.5, which would satisfy many practical cases. Thus with a suitable broadly-tuned aerial the range 160 Mc/s. ±10 per cent could be covered by a simple fixed device comprising two one-eighth wavelengths of cable and a suitable mounting bush capacitance for the aerial.

A Method for the Measurement of Capacitance and Resistance

By G. T. BAKER, M.Sc.
(British Telecommunications Research, Ltd.)

An electronic chronometer is used to measure the time taken for a capacitor/resistance combination to discharge between given potential limits. Alternatively, the discharge time is compared with two fixed intervals to determine whether a component is within a given range of values.

A WIDELY used method for the measurement of short time intervals involves the discharge of a capacitor through a resistance.[1,2] The time is given by:—

$$t = CR \log E_0/E_t \qquad (1)$$

where E_0,E_t are the boundary potentials. Electronic technique provides alternative methods for measuring these times and it is possible to use Equation (1) to determine one of the other parameters, e.g., capacitance or resistance. By fixing E_0/E_t and either R or C the equation becomes:

$$C = kt \text{ or } R = k_1 t \qquad (2)$$

If the values of k and k_1 are suitably chosen, the time measurements will correspond directly to capacitance or resistance units, e.g., microfarads or ohms. Two types of measurement are described. In the first the actual time period is determined to give precision measurements of capacitance and resistance. In the second, the time is compared with pre-set values to determine whether the capacitor or resistor under test is within a specified range.

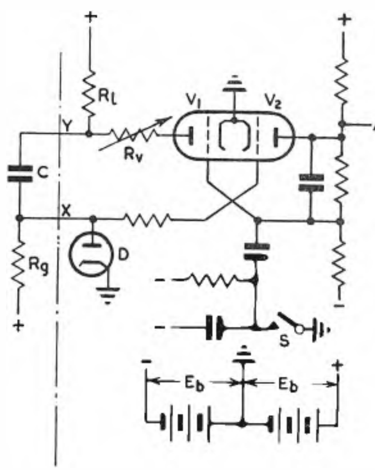


Fig. 1
Single-shot multivibrator circuit used for measurement

The requirements of the measurement circuit can be obtained from Equation (1). At $t = 0$, potential E_0 must be applied to the capacitor and a signal extended to the timing equipment to initiate the measurement cycle. When the potential on the capacitor falls to E_t, a "stop timing" signal must be generated. The capacitor potential tends to move through the operative point E_t comparatively slowly and trigger circuits should be included, so that the beginning and end of the timing cycle are clear and well-defined. The potential measuring circuit should constitute no load on the capacitor, otherwise the discharge rate is modified. Preferably E_t should be a fixed fraction of E_0, so that variation in applied potentials does not affect their ratio.

A number of circuits was investigated and the "single-shot" multivibrator shown in Fig. 1 was found to be the most useful. Valve $V2$ is normally conducting, since its grid is held at earth or cathode potential by the diode D. Part of the potential fall in the anode load resistor is transferred by a divider system to the grid of $V1$, holding this valve in a non-conducting condition. A sharp positive pulse applied to the grid of V_1, e.g., by closing switch S, causes the system to move rapidly to the condition wherein $V1$ is conducting and $V2$ cut off. C corresponds to the unknown capacitance in Equation (1) and the disturbing potential E_0 is provided by the sudden shift of the anode of $V1$ towards the negative. The capacitor discharges through R_g (equivalent to R in (1)) until the potential of the point X reaches the lower cut-off potential of $V2$. Then the system reverts rapidly to its original condition. The purpose of R_v is discussed later.

The circuit fulfils the main requirements mentioned above. During the discharge period, the grid of $V2$ is considerably negative and does not constitute a load on the capacitor. Moreover, the on-and-off pulses are sharp and clean. It is difficult to measure the actual values of E_0, E_t, but this is unimportant as long as their ratio remains substantially constant. It is essential that the system should be stable and should have a linear response. The extent to which these criteria are realized is discussed in the next section.

Scale Shape

The operative elements of the circuit are shown in Fig. 2. R_a represents the internal resistance of $V1$ including R_v; other symbols are as Fig. 1.

$$E_b = R_a i_a + R_l(i_a - i_g) \qquad (3)$$

$$E_b + v = R_a i_a + R_g i_g \qquad (4)$$

$$\text{Writing } R_s = R_a R_l/(R_a + R_l) \qquad (5)$$

$$R_s E_b/R_a + v = i_g(R_s + R_g) \qquad (6)$$

$$= C v'(R_s + R_g) \qquad (7)$$

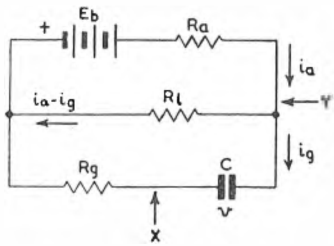


Fig. 2

The measuring element of the circuit shown in Fig. 1

$$= K\,EXP. -t/C\,(R_s + R_g) \quad\ldots\ldots\quad (8)$$

If the circuit values are so chosen that the measured discharge period is of the order of 0.1 sec. or more, the switching time can be neglected. At $t = 0$, $v = E_b$.

$$\text{Hence } K = E_b\,(R_s + R_a)/R_a \quad\ldots\ldots\quad (9)$$
$$= E_0 \text{ (Equation (1))}$$

The system reverts to normal when the potential at X reaches the cut-off of $V2$.

$$\text{i.e., } R_g i_g = E_b + E_{co} \quad\ldots\ldots\quad (10)$$
$$R_s E_b/R_a + v_t = (E_b + E_{co})(R_g + R_s)/R_g \quad (11)$$
$$= E_t \text{ (Equation (1))}$$

The cut-off potential of the valve is largely proportional to the applied anode potential. Hence if E_{co} is written $p\,E_b$, Equation (1) becomes:

$$t = C(R_s + R_g)\log\left(\frac{1}{1+p}\cdot\frac{R_s + R_a}{R_a}\cdot\frac{R_g}{R_g + R_s}\right) \quad (12)$$

The capacitance scale is therefore of the required form, i.e., with R_g constant, $C = kt$. This is confirmed by the measured values shown on A, Fig. 3. Closer observation reveals a slight curvature due to strays which have been neglected on Fig. 2.

The resistance scale suffers from two defects. As long as R_g is large compared with R_s, the relationship is useful and linear. The R_s term in the first bracket displaces the line, so that its projection does not pass through zero. Moreover, as R_g is decreased until it becomes comparable with R_s, the final log. term diminishes, generating a curvature towards the origin. These details are illustrated by the observed values, shown in B, Fig. 3, and in the enlargement of the curved portion shown in C.

Precision Measurements

Very accurate measurements of capacitance and resistance can be obtained by measuring the discharge time by means of an electronic chronometer.[3]

It is doubtful whether the measurements described herein would, of themselves, justify the installation of a chronometer. It is believed, however, that this instrument will become a standard feature of most laboratories in the near future, because of its value in other directions. The capacitance and resistance measuring unit would then be in the nature of an accessory.

The chronometer consists of three parts; an accurate pulse source, which is usually a 100 kc/s. crystal controlled oscillator; a digital type electronic counter, and a gate which controls the passage of pulses from the oscillator to the counter. The timing circuit is connected to the gate in such a manner that pulses are allowed to pass only during the switched period of the multivibrator. The reading on the counter then indicates the required discharge time. Pentagrid valve $V3$ in Fig. 4 constitutes the gate. The point A is connected to the normally conducting anode of the multivibrator by a potential divider system, so that the inner grid of $V3$ is held at a substantially negative potential while the system is at rest. During the switched period the potential of this grid is made more positive, the capacitor causing it to follow the movement of the multivibrator anode. Pulses from the oscillator applied to another grid of $V3$ are freely transmitted to the counter until the system reverts to the normal condition.

For the measurement of capacitance, the multivibrator must include fixed resistances corresponding to R_g of such value that k in Equation (2) becomes unity. Several values can be included, each representing a different capacitance range. The optimum number of ranges depends on the number of significant figures available on the counter. With an instrument of five decades, two ranges, 0.001-0.1 μF and 0.1-10 μF, have proved sufficient. For a lesser number of digits it is convenient to limit each range to a 20 db. coverage. With these limits, correction

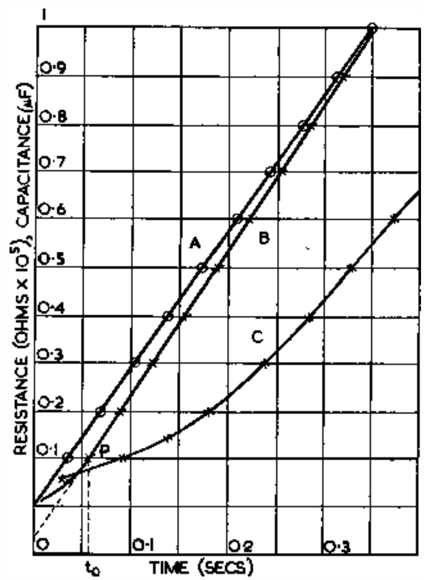


Fig. 3

Defects observed in resistance measurements

A: Time/Capacitance (R_g = 100kΩ)
B: Time/Resistance (C = 1μF)
C: Lower part of B $\times$ 10

can be made for the slight lack of linearity in the characteristic mentioned earlier.

The resistance values can be determined experimentally. A standard capacitor of the required full scale value is connected to the unit, and the resistance value adjusted until the desired reading is obtained on the counter. The valve compensation mentioned later should be approximately central. If the checking capacitor is of the decade type, the unit can be tested over the full range and errors due to non-linearity balanced. The range switch is of the wafer type with connexions as shown in Fig. 5. A 10 kΩ resistor is in series with one pole of the switch and this must be in circuit when the range resistances are being determined.

For resistance measurements the lower bend of the characteristic shown in Fig. 3 must be obviated. This is effected by giving R_g an artificial zero in the form of the 10 kΩ resistor mentioned in the previous paragraph, which value must be deducted from the counter reading. Alternatively, an additional "single-shot" multivibrator can be inserted to delay the beginning of timing by an amount shown as t_0 on Fig. 3. The effect is to shift the origin to the point P so that resistance readings are given directly. Two resistance ranges are suggested, one 100 kΩ full scale, the other 10 MΩ. The internal capacitors are determined experimentally and wired as shown on Fig. 5.

Allowance must be made for valve ageing and replacements. In Equation 12 the valve affects two terms, R_a and p. The former represents a significant variation, and compensation has been inserted in the form of variable resistor R_v in series with the anode of $V1$. A range of about 2 kΩ is sufficient to cover most valves of the specified type. Variations in p, representing changes in the grid cut-off potential of the valve, are not so important, since p is very small and is largely masked by the unity in the same term. A single compromise adjustment on R_v is sufficient to cover both variations.

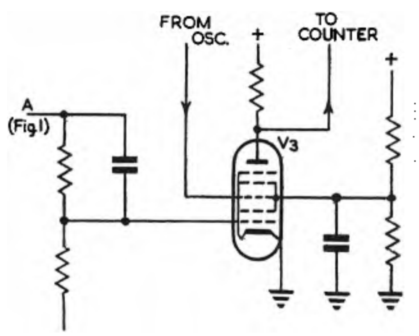


Fig. 4
"Gate" circuit between oscillator and counter

One position of the range switch is used for adjusting R_v. The operation virtually consists of measuring the capacitance of one of the included capacitors. This is a fixed and known value, and if a valve has been changed, it is a simple matter to adjust R_v until the original reading is obtained, since it should be constant for any given instrument. Another position of the range-change switch can be used for adjusting the zero delay when this is fitted. The external terminals are short-circuited within the instrument, and the time delay is reduced until the counter just reads zero.

Performance

The system includes no moving parts, and the accuracy depends entirely on the stability of the components used. Temperature changes are the main disturbing factor, but with suitable choice of equipment a variation less than about 4 parts in $10^6/°C$. can be obtained.

The assumption that p (12) is a constant is not completely justified. The return switching action does not take place until the overall amplification of the system exceeds unity. Due to the lower bend in the valve operating characteristic, this does not occur at the true cut-off point and the switched time is dependent to some extent on supply potentials. For precision readings, therefore, these must be stabilized. Elaborate control is unnecessary since the time variation is only a very small proportion of the potential change. Mains ripple leads to erratic results, and the smoothing should be as efficient as possible. Other fluctuations were traced to the filament supply, although indirectly heated valves were used. The heater current should be derived from accumulators for the highest accuracy.

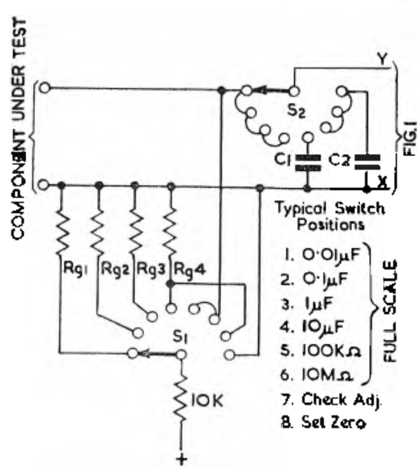
The effect of potential changes and random fluctuations has been considerably reduced by connecting the charging resistances R_K to the positive potential instead of to the cathode as is usual in multivibrator practice. The slope of the discharge curve at potential E_t is increased by a factor of about 4 or 5; the time change for a given potential fluctuation is decreased in the same ratio.

It was found difficult to obtain constant readings on a valve taking appreciable grid current, probably due to heating of the grid spiral. A series grid resistor is added to overcome this difficulty, but has the effect of allowing the point X (Fig. 1) to become positive. The diode D is introduced to hold this potential and to still further reduce grid current.

With the plate potential stabilized and heaters taken from the mains (via transformer), no difficulty was experienced in maintaining an accuracy of 0.1 per cent. With accumulator supply to the heaters, the maximum variation over a run of several hours was ± 0.03 per cent. The residual variation was probably due to direct pick-up, since it appeared to coincide with the working of adjacent apparatus which was not entirely spark-free. The effect is fundamental to all trigger type circuits, and the only solution is complete shielding, together with adequate smoothing on all incoming wires.

The above results were taken on a switched time of 0.35 sec. As this period is reduced, the percentage variation is increased and the finite switching time becomes important. Experiments are being

Fig. 5
Correction circuit for the non-linearity shown in Fig. 3



conducted with an H.F. pentode pair to determine the minimum switching time.

Tolerance Measurements

Precision measurements of the type described are required in comparatively small numbers. The majority of capacitors produced are for such purposes as smoothing, speech transmission, spark-quenching, etc., where the capacitance tolerance is wide, ±10 per cent being quite common. For such measurements, a modification of the instrument has been designed which gives a rapid, direct indication whether a capacitance is within a given range, usually expressed as a negative and positive percentage tolerance on a specified nominal value.

It is unnecessary to determine the actual switched time of the multivibrator, so that a gate circuit and chronometer are not required. Instead, the time is compared with two known values, T_n and T_p, representing the negative and positive time equivalents of the capacitance tolerances. The two preset time devices, in effect, constitute an electronic "go/no-go" gauge, so that for a good capacitor $T_p > t > T_n$.

In Fig. 6 the three lower units marked *N, O, P* represent multivibrators similar to that shown in Fig. 1. The nominal time T_o can be made any value, a convenient figure being about 0.35 sec., corresponding to a 1 μF capacitor associated with a 100 kΩ resistor. The R_n resistance forming part of the *N* unit is in the form of an 80 kΩ fixed and 20 kΩ variable; that of *P* is 100 kΩ and 20 kΩ variable. Over the range covered, resistance is substantially linear with time, so that the dials on the variable resistors can be scaled directly as percentage decrease and increase of time T_o. A selector switch associated with the *O* unit enables a number of resistances to be connected, so that the time T_o corresponds to a different nominal value of capacitor for each switch setting. 100 kΩ corresponds to a capacitance of 1 μF; other resistors might give 0.1 μF, 2 μF, etc.

The three multivibrators are switched simultaneously, the order of release being determined by the upper elements *A, B, C*. These are Eccles-Jordan toggles of conventional design,[1] with two stable positions. An external switch (not shown) restores all three to the normal or "reset" position prior to each test. The neon-tube associated with each toggle glows to indicate the operated or "set" condition. The arrowed connexions represent pulse operation paths, the pulse being emitted as the originating element returns to the normal condition.

Consider the sequence when a capacitor below the permitted tolerance is connected to *O*. The lower units release in the order *O, N, P*. *A* receives its "reset" pulse before being "set" by *N*, and therefore remains in the "set" condition. The glowing of the associated neon-tube indicates "capacitor below tolerance." If the capacitor under test is above the range permitted, the sequence terminates with *B* "set," the glowing of the neon-tube giving the appropriate signal. Only if the component under test lies between the specific values, i.e., if the units release in the order *N, O, P*, can the *C* toggle be "set" to indicate a useful capacitor.

Acknowledgment

This investigation was carried out in the Strowger Laboratories, Liverpool. The author wishes to thank the Chief Engineer of the Automatic Telephone & Electric Co., Ltd., for permission to publish this material.

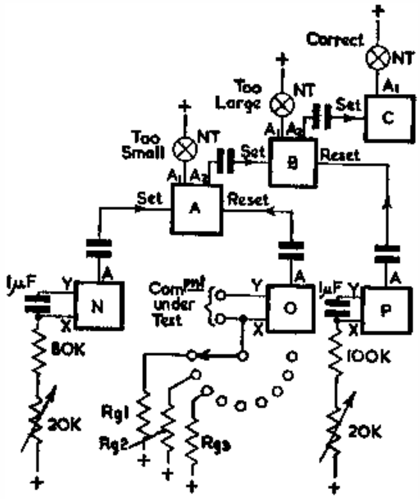


Fig. 6

Elements of the comparator circuit

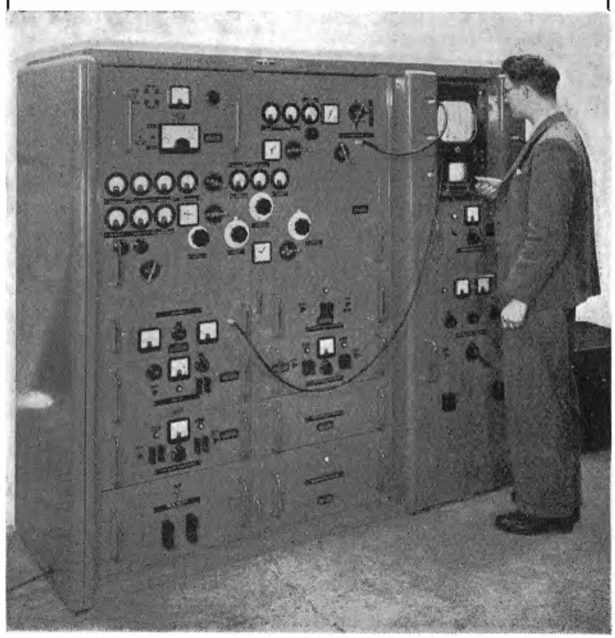
References.

[1] R. S. J. Spilsbury and A. Felton: *J. Inst. Elect. Eng.*, *94*, p. 284 (June 1947).
[2] S. S. West and L. C. Bentley: *Electronic Engineering*, *19*, p. 207 (July, 1947).
[3] G. T. Baker: *J. Scientific Instruments*, *25*, p. 194 (June, 1948).
[4] G. T. Baker: *Ibid.*, p. 128 (April, 1948), Fig. 4.

New O.B. Television Transmitter

Television outside broadcasts from places not served by the special television cable will be considerably facilitated when this new transportable television transmitter, designed and produced by Marconi's Wireless Telegraph Co., Ltd., is brought into service by the B.B.C. Due to its small size and modest power requirements it can be installed in a single vehicle, which will carry also the fire-escape type extensible ladder for supporting the aerial.

The new double-sideband transmitter is designed to radiate 350 watts within the frequency range 174 to 216 Mc/s.



Time Base Expansion and Compression Circuits

for varying Trace Speed during time of sweep

By D. M. MACKAY, B.Sc.

(*Wheatstone Laboratory, King's College, University of London*)

THE Miller time base circuit, using a single valve, has already been described many times and in particular in *Electronic Engineering* for August 1948.[1] During the war the writer devised a number of methods[2] for varying the speed of sweep of the Miller time base at selected points during the sweep period, thus producing either an expansion or compression of part of the trace. A description of some of the experimental circuits may be of interest, the principle being equally applicable to all capacitor-controlled time bases.

In Fig. 1, C is the "Miller" capacitor, and the time of sweep is controlled by the values of R_g and C, with C_1 and R_1 in series across C. The value of R_1 is preferably some two or three times that of R_a, the anode resistance, (for reasons which will be seen later), and must be small compared with R_g. From the junction of R_1 and C_1 a diode is connected to earth through a potentiometer P which determines the diode anode voltage V_1.

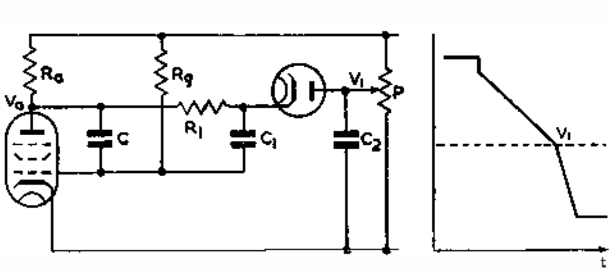


Fig. 1

Simple speed changing circuit producing an increase in speed at a selected voltage

Typical values (with EF50):

$R_a = 47k$, $R_g = 3.3M$, $R_1 = 100k$.
C, C_1 according to speeds required
$C_2 = 8\mu F$, $P = 100k$

When the anode potential V_a falls to that of the diode (V_1) the potential across C_1 is clamped by the action of the diode and the effect of C_1 on the sweep velocity practically ceases. R_1 then becomes an additional anode load across R_a (hence the requirement that it should be two or three times as great), and the sweep continuous at a speed determined only by C and R_g. The effect is shown by the waveform adjoining the circuit diagram, which gives a sudden increase in speed of sweep beyond the point marked V_1.

A fast sweep changing to a slow one is provided by the circuit of Fig. 2. C_1 is connected across C through a diode D_1, the anode potential of which is determined by R_1 and P. When D_1 is non-conducting, C_1 has practically no effect on the sweep speed, but when the potential V_a falls below V_1, C_1 discharges through D_1 in parallel with C and thus slows down the sweep. At this stage R_1 is again effectively in parallel with R_a, and the value should therefore be kept high as before. The waveform is shown adjacent to the diagram.

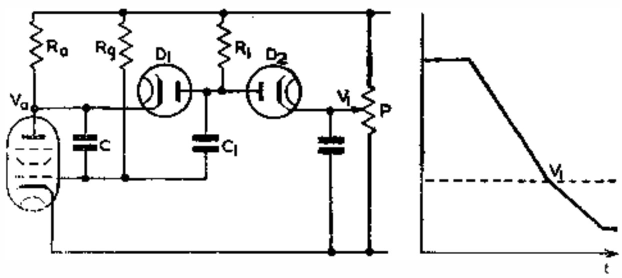


Fig 2

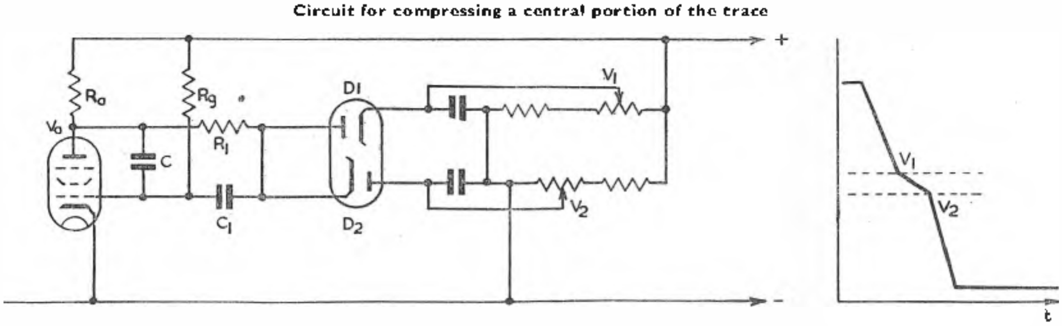
Double-diode circuit giving a decrease in speed at the required voltage

If the diode of Fig. 1 is reversed in connexions, a single diode variant of Fig. 2 is obtained, which has the same effect on the sweep speed. R_1 in this case is present initially as an additional anode load.

For compressing a portion of the scan, the circuit of Fig. 3 may be used. This is a combination of the principles of the preceding circuits. C_1 has no effect until V_a falls in potential to V_1, when D_1 ceases to conduct. C_1 then discharges through R_1 (which is small compared with R_g) until the potential V_a falls below V_2. At this stage D_2 conducts and clamps C_1 once

Fig. 3

Circuit for compressing a central portion of the trace



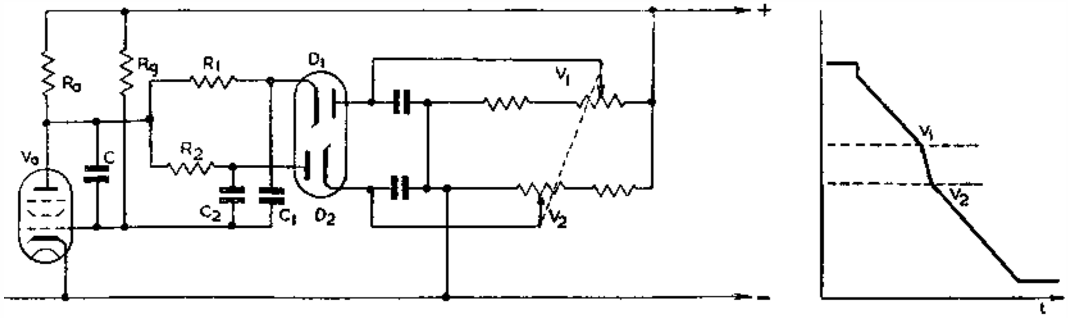


Fig. 4

Circuit for expanding a central part of the trace. $C_1 = C_2$ if the time-base is to resume its original speed

more, enabling the sweep to be resumed at its original speed.

Finally, the circuit of Fig. 4 can be used to expand a portion of the sweep. In this, C_1 slows down the sweep until V_a falls in potential to the value V_1, when D_1 clamps C_1. The sweep then falls at a rate governed by C and R_g. When V_a reaches V_2, D_2, which has been conducting, ceases to clamp C_2, and the sweep is again slowed down. If $C_1 = C_2$ and $R_1 = R_2$ the slope will resume its original value. The expanded portion may be moved up and down the time base by adjustment of V_1 and V_2 with a ganged potentiometer.

The foregoing, of course, represents a much-simplified description of the action of the circuits, but is adequate if the approximations mentioned are valid.

References

[1] Briggs B. H. " The Miller Integrator ", *Electronic Engineering* 20, 243, 1948.

[2] MacKay D. M. " Improvements in or relating to Timebase Generators and Cathode-ray Oscillographs." Admiralty Patent No. 592372

The Videograph

A New Instrument for Home Entertainment

UP to now instruments for home entertainment have been produced as separate articles, with little or no attempt to incorporate them one in the other, and there is thus much waste of space, efficiency and money, if one purchases the units separately, and they rarely seem to go well together artistically. The possible exception is the radiogram which at least combines two uses in one, with considerable saving.

This view of the prototype model of the " Videograph " shows a promise that the futuristic interior will be encased in a smart exterior. The shoulders of the cabinet hold the "shorts" (both film and cocktail)

The logical extension of the radiogram idea was the addition to the television receiver of facilities to view recorded programmes, i.e., films; and thus was formed the " Videograph."

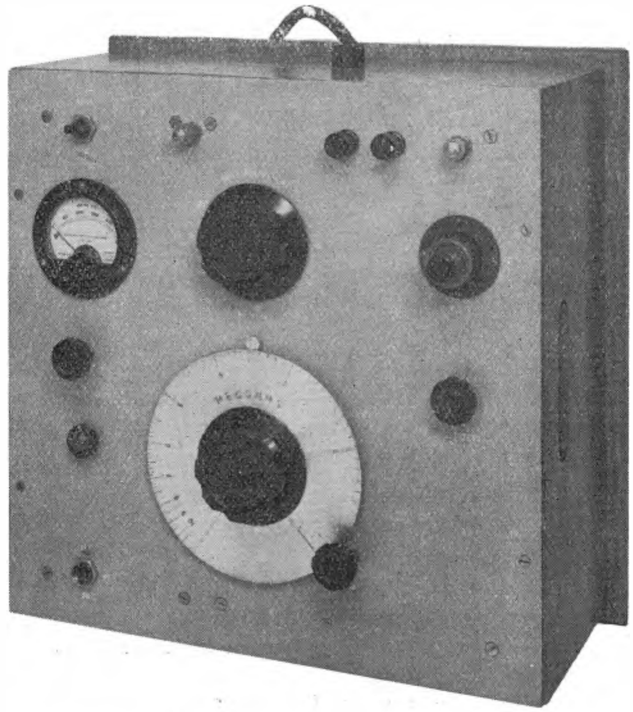
A typical model is shown, and consists of a normal television receiver, with the addition of a simple mechanism for running 16 mm. talkie film in front of a small electrostatic C.R.T. which scans through it on to a photoelectric cell and electron multiplier, the output of which can be switched into the video output of the receiver. The film runs smoothly, thus eliminating the complicated part of normal projectors, and is pulled by a small synchronous motor. The scanning potential for the electrostatic tube is obtained from the grid of the line output valve, ensuring automatic line lock, and the frame is locked by a novel method to the mains, ensuring automatic frame lock. Sound is picked off as usual, and is reproduced through the sound receiver. Later models will probably pick up sound by means of an iron dust, or wire, thread let into the film. This will make sound pick up much simpler.

All the film equipment is housed in one of the " shoulders " of the cabinet, which hinge open for loading in film, etc. Incidentally the other shoulder is a well stocked cocktail cabinet, surely a necessary part of a home entertainment device! Several record companies are planning to market recorded programmes, with subjects ranging from simple cabaret acts (dancing, singing, conjuring, etc.) to travel items, plays, instructional records, such as music lessons, engineering training, carpentry demonstrations, gardening, etc., and it is expected that public libraries will handle them on a loan basis, as they do at present with the (now) old fashioned printed books.

Our visionary contributor concludes with the regretful comment that there is, of course, no such thing as a Videograph—yet.

An improved
Insulation Resistance Bridge

By A. F. BOFF, B.Sc.
(Late of Standard Telecommunication Laboratories, Ltd.)



A robust self-contained instrument is described which permits rapid measurement of resistance in the range 10^6-10^{14} ohms. The test voltage may be varied continuously up to 500 volts, thus enabling voltage co-efficients of resistance to be determined

THE development of new insulating materials and the investigation of the electrical properties of textiles require measurements of resistivities approaching 10^{17} ohms-cm. For a sample 1 mm. thick having an electrode area of 100 cm.^{2}, the resistivity figure quoted corresponds to an actual resistance of 10^{14} ohms. Measurement of such a high value of resistance presents special difficulties which are not encountered when measuring lower values. The present methods in general use include: (a) Measurement of the steady current through the sample, when subjected to an electric stress, by means of a supersensitive galvanometer, and (b) ballistic measurement of the residual charge in a capacitor which has been allowed to discharge through the sample for a measured time. In most practical cases the sample may be subjected to a potential difference of 500 volts. When it is calculated that 500 volts across 10^{14} ohms cause a current of 0.000005 μA, the delicacy and limited range of method (a) is evident. In addition, it is impossible to protect the galvanometer effectively from current surges due to breakdown of the specimen. Method (b) requires a first-class capacitor and a sensitive galvanometer. It is laborious and measures only the average value of resistance over a voltage range. Voltage coefficients of resistance (often very considerable at high ohmic values) must therefore be estimated by measurements over successive voltage ranges as close together as sensitivity requirements will permit. The Wheatstone Bridge circuit would be suitable for high resistance measurements if the detector impedance were sufficiently high (preferably high compared with the greatest value of resistance to be measured), and had adequate sensitivity. Since the advent of the electrometer valve, several investigators have employed bridge methods, but owing, perhaps, to the numerous separate battery supplies used and the usually rather intricate setting-up procedure for devices designed to compensate for supply fluctuations, bridges for very high resistance measurement do not appear to have been universally adopted.

The method described here involves no new prin-

Fig. 1.
Schematic of the bridge

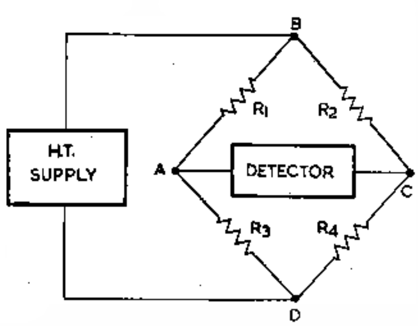
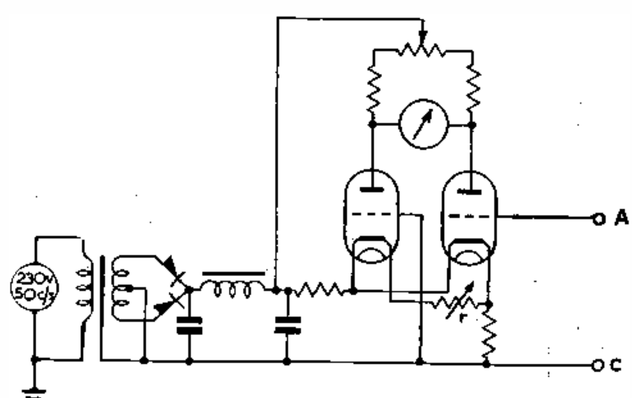


Fig. 2.
Balance detector (Terminals isolated from earth)



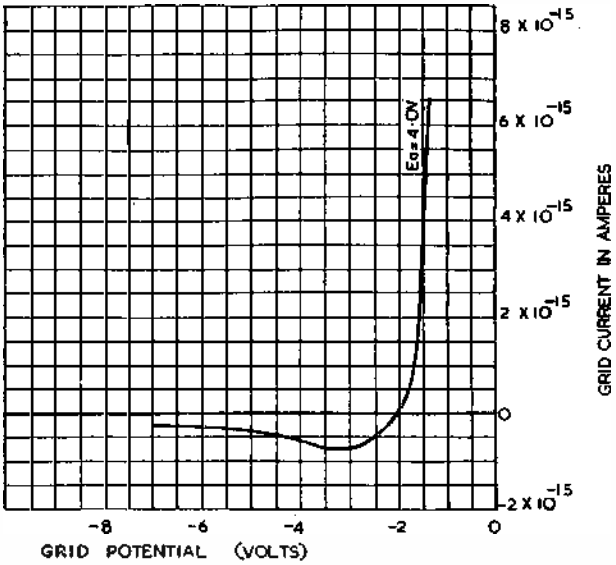


Fig. 3.
Grid current characteristic of average electrometer valve type ET 1

ciples, but represents a considerable simplification of technique without sacrifice of stability or accuracy, and permits rapid measurement of resistance up to 10^{14} ohms (corresponding for a convenient sample to a resistivity of 10^{17} ohms-cm).

Description of Apparatus and Circuits

The difficulties peculiar to the measurement of very high resistance are, in general, due to:—

Instability of the electrical circuits forming the detector.

Grid current flowing in the detector input circuit.

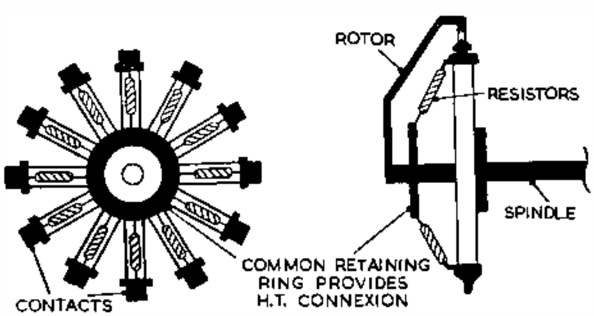
Time constants of the bridge arms.

Leakance across the high resistance arms of the bridge.

The following sections deal with these difficulties together with the means by which they have been overcome or reduced in this new instrument.

It is desirable that the high voltage supply should have one pole at earth potential and also that the resistance under test should be earthed at one end to facilitate screening. Since variable high resistances are not practicable, two arms of the bridge must be low. Maximum sensitivity will be obtained if the detector is connected between the junctions of the two high resistances and the low resistances. The basic circuit, therefore, becomes that of Fig. 1 where the point D is at earth potential, $R3$ is the unknown high resistance, $R1$ is the internal standard and $R2$, $R4$ are low resistance arms.

Fig. 4.
Polythene switch (approximately quarter scale)

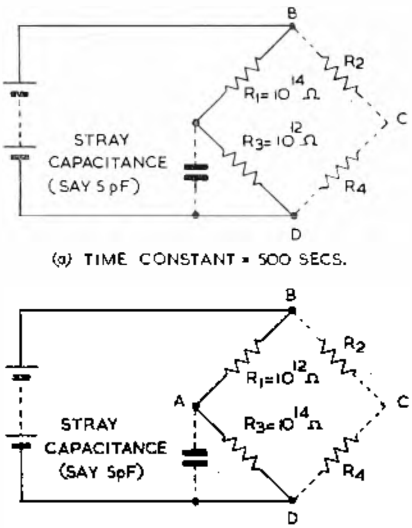


(a) SHOWING POLYTHENE SPOKES, COPPER CONTACTS & RESISTORS. THE ROTOR IS OMITTED

(b) SIMPLIFIED SIDE ELEVATION SHOWING POSITION OF RESISTORS & ROTOR

The Detector

The ideal detector would combine high sensitivity and inherent stability with respect to both electrical and mechanical disturbances. In addition, it would cause no shunting of the bridge arms nor pass current through them. It is well known that a thermionic valve, although presenting an impedance normally considered high, is limited to a few megohms by the existence of a finite grid current. This is not the familiar electronic current which flows when the control grid is made positive with respect to cathode, but a positive ion current existing when the grid is sufficiently negative to repel all electrons.



(a) TIME CONSTANT = 500 SECS.

(b) TIME CONSTANT= 5 SECS.

Fig. 5.
Showing how disposition of bridge arms affects charging time

The causes of this current have been listed by Metcalf and Thompson[1] as being:—

(1) Leakage over insulation.

(2) Ions formed by gas present in the valve.

(3) Thermionic grid emission due to heating of grid by the filament.

(4) Ions emitted by the filament.

(5) Photoelectric emission by control grid under action of light from filament.

(6) Electrons emitted by control grid under action of soft X-rays produced by normal anode current.

Item (1) may be effectively cancelled by means of guard rings and item (6) reduced by use of a low anode voltage,[2] but the remaining items depend solely on the valve design and manufacturing technique.

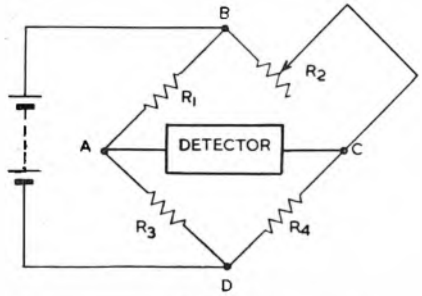


Fig. 6.
Conventional Wheatstone bridge

There are a number of specially constructed valve types now available, generally called electrometer valves,[3,4] in which the grid current has been reduced to a very low value. A typical curve showing grid current as a function of grid potential for one of these types is given in Fig. 3. It will be noted that at one value of grid potential the electron current and the positive ion current are equal, and there is zero resultant current. It has been found possible to work at this point over short periods, but the exact point is not sufficiently stable for general use and it has been found more satisfactory to accept the small residual positive ion current.

The circuit of Fig. 2 shows two electrometer triodes comprising a balanced detector system which is virtually independent of supply fluctuations. The valve filaments and anodes are supplied from a common D.C. line derived from the mains. If both triodes are identical, no out-of-balance results due to changes in supply. In practice the over-riding factor affecting the matching of the valves has been found to be the filament temperature. For this reason a variable resistance (r in Fig. 2) has been included. When correctly adjusted, the balance is undisturbed, apart from a slight relative thermal delay in the filaments by supply variations of up to ± 20 per cent. Changes due to mechanical disturbances are effectively avoided by use of soft rubber anti-microphonic mounting for the valves. It will be seen that terminal C of the detector has a path to earth by way of the insulation of the mains transformer. A good impregnated mains transformer has an insulation resistance greater than 500 megohms and C may, therefore, be connected to the low resistance junction without detectable effect.

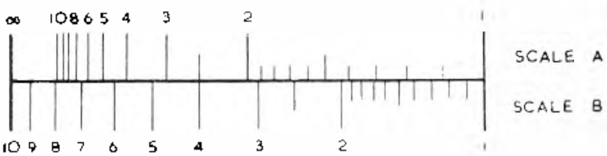


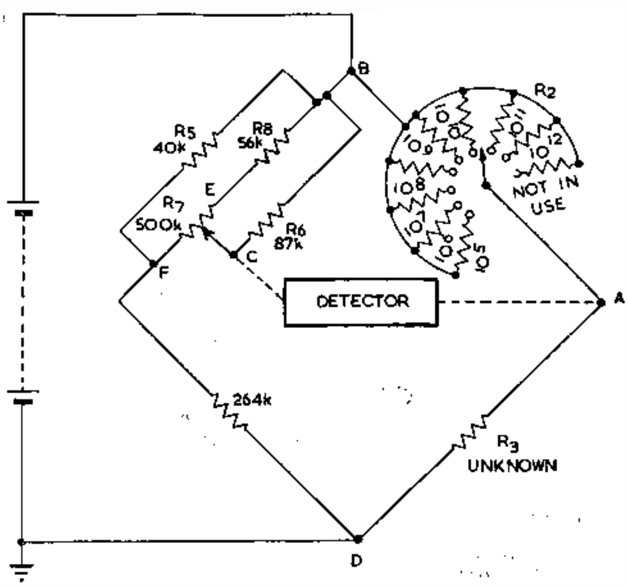
Fig. 7.
(a) Inverse scale shape
(b) Constant accuracy scale

The Bridge Arms

To provide adequate sensitivity, when testing at low voltages, without resort to a sensitive galvanometer, the maximum ratio of unknown to standard was limited to 100. This implied the use of internal resistors covering the range up to 10^{12} ohms. Special sealed resistors, available commercially, were found suitable for this purpose, especially since the maximum voltage to which they were subjected (50 V) fell within the range over which, it was claimed, they had negligible voltage coefficient of resistance.

To carry the resistors, without introducing leakance, a twelve position switch was constructed, having radial spokes of polythene and a rotary contact arm (Fig. 4). The spring contacts of the latter were designed to automatically centre the polythene spokes to avoid warping. All corners and sharp edges were smoothed to avoid unwanted ionization. An important factor in the disposition of the bridge arms is concerned with time constants. When it is considered that a capacitance of 5 pF associated with a resistance of 10^{14} ohms has a time constant of 500 seconds it is seen that the problem is not easy. A large time constant may completely mask the balance point, since fluctuations of the high voltage supply or movements of the operator (which slightly modify the stray capacitances) cause random displacement of the balance indicator, at a rate exceeding that at which they can be accommodated by the circuit. In the design of the high insulation terminal and the range switch, capacitance has been kept as low as practicable. However, in use considerable capacitance is often inherent in the specimen. The time constant resulting from a given capacitance is greatly influenced by the relative positions of the high resistance arms of the bridge. The effect is demonstrated in Fig. 5, where it is shown that a 100 to 1 improvement results if it is arranged for the stray capacitances to be charged through the lower resistance. The low resistance arms of the bridge are by no means critical, but to avoid unnecessary drain on the high voltage supply the resistances should be as high as practicable. For a continuously variable wire wound potentiometer this sets an upper limit of 500 kΩ which may be chosen as the low resistance arm. A circuit

Fig. 8.
Bridge ratio arms showing scale shape network



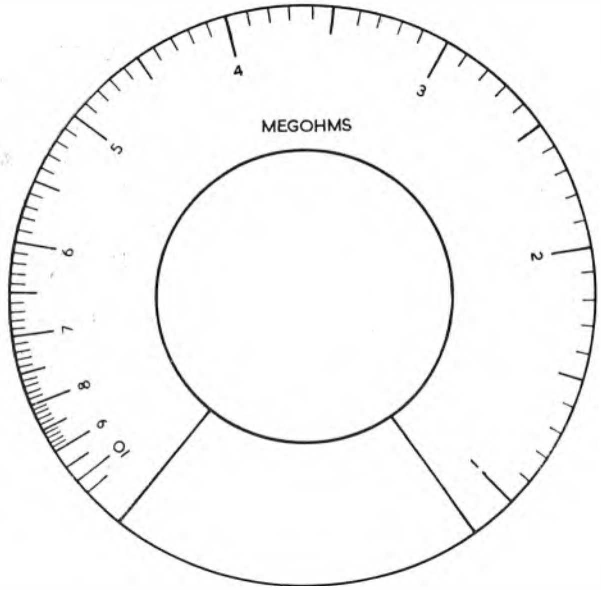


Fig. 9.
Actual scale shape in this instrument

of the type sketched in Fig. 6 has an inverse relation between the resistance $R2$ and the unknown $R3$, whereas, for constant accuracy of reading, a scale shape is required based on the law $R2 = Ae^{K\theta}$, where θ is the angular motion of the balance dial. Both scales are produced in Fig. 7 for comparison.

A close approximation to the ideal scale shape may be achieved by means of the network detailed in Fig. 8. Design of such a network is simplified by regarding BC and CD as a potentiometer across the high voltage supply. $R5$ is chosen to provide the maximum voltage required across arm CD of the bridge when the moving contact C is at E, while $R6$ must be selected so that when C is at F, the correct minimum voltage is obtained. Although the five resistors comprising BC and CD are interdependent, in practice the scale shape is governed, primarily, by the ratio $R7/R6$. Resistors $R8$ and $R5$ control the range and are finally adjusted during calibration. The technique is based on a method described elsewhere[5] and the resultant scale shape is shown in Fig. 9. The moving contact C is connected to terminal C of the detector and to the guard rings of the high insulation terminal and the electrometer valves.

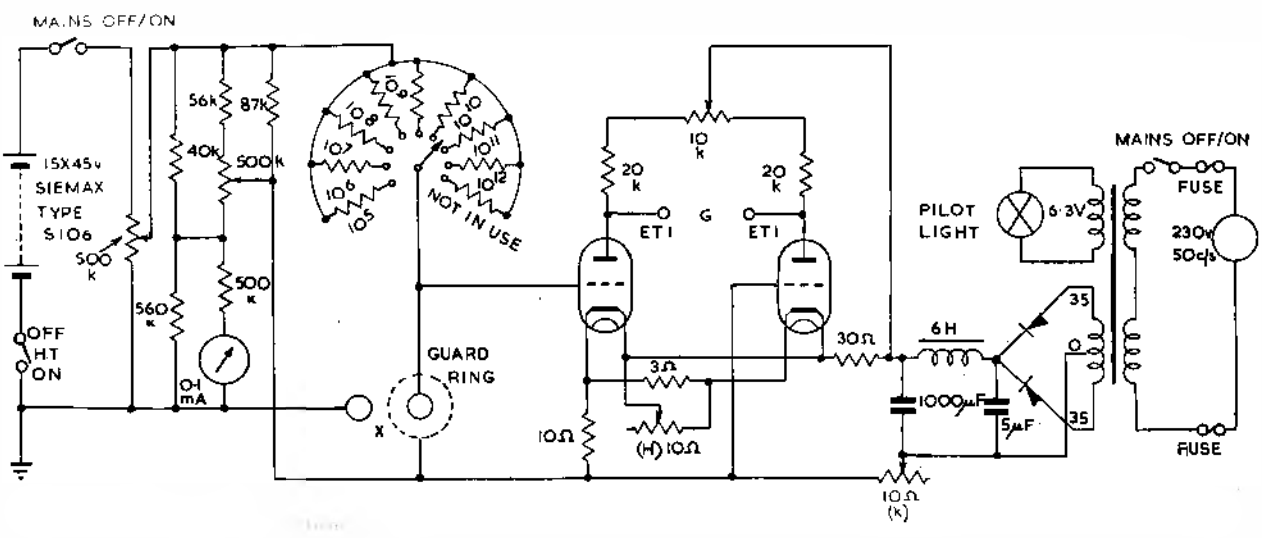
High Voltage Supply

Owing to the very stringent conditions arising out of the time constant problem discussed in the preceding section, a mains operated source for the high voltage was not considered practicable. A battery power pack was, therefore, built into the equipment. The pack consisted of 15 miniature batteries (nominally 45 V each), occupying a total space of approximately 4 in. by 4 in. by 10 in. The maximum load on the batteries is only 2.5 mA and at the time of writing the equipment has been in regular laboratory use for almost twelve months without requiring replacements. The voltage applied to the bridge circuit is continuously variable and is indicated on a voltmeter incorporated in the apparatus to facilitate measurements of voltage coefficients of resistance.

Calibration and Accuracy

The variable bridge arms BC and CD were accurately calibrated in terms of external standard resistances of relatively low ohmic value, but the very high resistances, utilized for comparison with the unknown arm, presented considerable difficulty in the absence of known standards of comparable values. When in use in the circuit, these resistances may be subjected to any potential difference up to 50 volts. A calibration method, therefore, should not apply a stress exceeding this value and should, in addition, check the voltage coefficient effects over this range. The procedure adopted was as follows: each resistor was measured in terms of resistors one and two decades lower, using the accurately known low resistance arms BC, CD. Values not exceeding 100 megohms were measured on a precision Wheatstone Bridge.

Fig. 10. Complete circuit diagram



A progressive error would apply to each succeeding decade, but it is estimated that all internal resistors are known to approximately 2 per cent. The measurements were checked by other methods and found to agree within the experimental errors associated with the particular method employed. The detector system has a sensitivity of 16 μA / volt. When a test potential of 500 volts is employed a change of 1 per cent in the unknown will produce a current of 0.8 μA. For normal measurements, therefore, a 0-50 μA moving coil instrument is sufficiently sensitive to permit balance within the accuracy of the internal standards. A portable galvanometer, having a sensitivity of 26 mm / μA has been used with success for measurements at very low test voltages.

REFERENCES AND BIBLIOGRAPHY

[1] G. F. Metcalf and B. J. Thompson. A Low Grid Current Vacuum Tube. *Phys. Review*, 36, 1489 (1930).

[2] C. E. Nielsen. Measurement of Small Currents: Characteristics of Types 38, 954, 959, as Reduced Grid Current Tubes. *Review of Scientific Instruments*, 18, Jan., 1947.

[3] G. W. Warren. The Electrometer Triode and its Application. *G.E.C Journal*, May, 1935.

[4] J. A. Darbyshire. A New Electrometer Valve. *Electronic Engineering*, September, 1946.

[5] A. F. Boff. The Use of Standard Linear Potentiometer to give Inverse or Reciprocal Scales. In press.

A Super-Sensitive Amplifier for Measuring Small Currents. F. J. Moles, *G.E.C. Review*, March, 1933.

One Tube Balanced Circuit for D.C. Vacuum Tube Amplifiers of very Small Currents. Walter Soller, *Review of Scientific Instruments*, August, 1932.

An Electrometer for Measurement of Voltage on Small Ionisation Chambers. F. T. Farmer, *Proc. Phys. Society*, Part 6, September, 1942.

Condenser Chamber Measuring Apparatus. Spiers and Hay, *British Journal Radiology*, May, 1946.

Procedures in Experimental Physics. J. Strong (1945). Prentice-Hall.

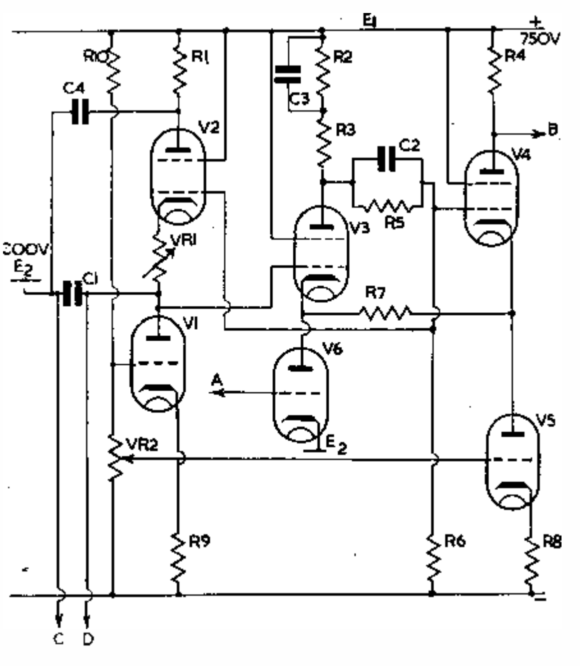
How humidity Affects Insulation. *General Radio Experimenter*, July and August, 1945.

A Sawtooth Generator for Oscilloscopes

ONE major difficulty in the design of sawtooth voltage generators for oscilloscopes is the development of a circuit which is linear at the commencement of the sweep. A further difficulty is in the design of a generator whose waveform at the commencement of the forward stroke is not influenced by the shape of the triggering pulse.

The circuit of a generator which has been found to overcome these difficulties is shown in the Figure.

The capacitor *C*1 is alternately charged and discharged to form the flyback and forward strokes respectively. The discharge is through valve *V*1 which has strong negative feedback via its cathode resistor *R*9 causing it to behave as a constant current device. Valves *V*3 and *V*4 form an unstable amplifier of wide band width using a constant current valve *V*5 (similar to *V*1) as the common cathode impedance, and cause *V*2 to conduct and charge *C*1 for flyback strokes. Synchronizing signals may be applied via *V*6. Output is taken across *CD*.



The operation of the circuit may be described as follows. Consider *V*2 and *V*4 to be cut off with *V*3 conducting. *C*1 will be steadily discharging through *V*1 producing the forward stroke of the scan. When the voltage on the anode of *V*1 has fallen to a sufficient extent it will cause *V*3 to stop conducting and its anode voltage will rise. This rise is transmitted by a D.C. connexion to the grid of *V*2 which starts conducting and quickly charges *C*1 (flyback stroke). It is likewise transmitted to the grid of *V*4 which also commences to conduct and so maintains *V*3 non-conducting. When, however, the voltage on *C*1 has risen sufficiently *V*3 again conducts, its anode voltage thus falling and returning *V*2 and *V*4 to the non-conducting stroke. The circuit has now reverted to the original condition with *C*1 steadily discharging.

The conditions of the valves may be summarized:

	V_2	V_3	V_4
Forward Stroke	OFF	ON	OFF
Flyback	ON	OFF	ON

*VR*2 adjusts the bias on *V*5 and therefore serves as a fine control of repetition frequency.

Synchronizing pulses are applied to the grid of *V*6, a pentode which is biased to cut-off except when caused to conduct by a pulse at *A*.

If sufficient resistance is introduced by *VR*1, the voltage of *C*1, due to the constant drain of current through *V*1, may not reach a large enough value to turn *V*3 on at the end of the flyback. The operation will therefore cease until a synchronizing pulse makes *V*3 conduct and thus commence the forward stroke. In this way the circuit gives a single sweep and flyback, leaving the spot waiting at the start position of the sweep. Brightening pulses for this mode of operation may be taken from *B* to intensify the trace during the forward stroke. Due to the control of *V*2 by the coupled pair *V*3, *V*4, the trigger pulse once having caused *V*3 to conduct has no further influence on the operation of the circuit and cannot interfere with the linearity of the sweep.

It was found that the charge and discharge of *C*1 has a tendency to "pull" the voltages of the power supplies *E*1 and *E*2. The decoupling combination *R*1, *C*4 may be arranged to eliminate this effect.

—Communication from E.M.I., Ltd.

NOTES FROM THE INDUSTRY

The Manchester Electronics Exhibition 1950

The Fifth Annual Electronics Exhibition organised by the North Western Branch of the Institution of Electronics will be held at the College of Technology, Manchester, on July 18, from 2.30 p.m. to 9 p.m. and on July 19 and 20 from 10 a.m. to 9 p.m. Part of the Exhibition will be devoted to the interests of the amateur constructor and there will be demonstrations of television reception on receivers designed for home construction.

Admission will be by ticket obtainable from Mr. A. Hickson, 205 Parrs Wood Road South, East Didsbury, Manchester, 20.

Award of the Albert Medal

The gold Albert Medal of the Royal Society of Arts for 1950 has been awarded, with the approval of Her Royal Highness The Princess Elizabeth, President of the Society, to Sir Edward Victor Appleton, G.B.E., K.C.B., M.A., D.Sc., Sc.D., LL.D., F.R.S., "for outstanding services to science and industrial research."

Sir Edward Appleton, who is now Principal and Vice-Chancellor of Edinburgh University, was Wheatstone Professor of Physics at the University of London from 1924 to 1936, and Jacksonian Professor of Natural Philosophy at Cambridge University from 1936 to 1939. From 1939 to 1949 he was Secretary of the Department of Scientific and Industrial Research. He was awarded the Nobel Prize for Physics in 1947.

Scottish Television Station

The contract for the site works and the construction of the approach roads for the main Scottish television transmitting station at Kirk o'Shotts has been awarded to Messrs. McLean and Company of Wishaw, Lanarkshire.

This station will have a 750-foot mast similar to those at Sutton Coldfield and Holme Moss.

Birthday Honours List

Mr. G. M. Wright, B.Eng., M.I.E.E., Engineer-in-Chief of Marconi's Wireless Telegraph Co., Ltd., has been awarded a C.B.E. in the Birthday Honours.

Mr. Wright originally joined the Marconi Company in 1912, and his work as a Research Engineer and Consultant gained for him an international reputation.

At the outbreak of war he became Assistant Director of Scientific Research and Chief Scientist at the Admiralty Signal Establishment. He returned to Marconi's in 1946, and in November 1948 was appointed a Member of the Radio Research Board of the Department of Scientific and Industrial Research.

Danish Radio Exhibition

The Danish State Radio and Denmark's Radio Industry are arranging an exhibition in Copenhagen from August 11 to 20, to mark the 25th anniversary of the State Radio.

The exhibition, which will be under the patronage of H.M. King Frederik, will include radio and electrical measuring equipment, radio components, radio and television apparatus, and will demonstrate working television and music studios. Most of the products on show will be of Danish manufacture.

The Forum, the largest exhibition hall in Scandinavia, will be used for this exhibition.

B.B.C. Orders for Television Equipment

The B.B.C. has ordered four sets of television camera equipment, two from Marconi's Wireless Telegraph Company, Ltd., and two from Pye, Ltd.

Two of these equipments, each comprising a mobile control room with three operational cameras and associated equipment, will be used initially at the Festival of Britain next year.

The other two equipments, each providing three operational cameras with their control equipment, will be used for the new television studios at Lime Grove, Shepherds Bush, and for outside broadcasts.

Government Department Electrical Specification

The Technical Panel of the Engineering Standards Co-ordinating Committee concerned with cables and wires (Secretary: Ministry of Supply, I.E.M.E., "Aquila," Golf Road, Nr. Chislehurst Station, Bromley, Kent) has recently issued a 1950 edition of GDES.25—Natural Rubber and Polychloroprene Sheathings for Electric Cables and Flexible Cords.

The main difference from the 1948 issue is in the introduction of a tear resistance test and increase of the tensile strength requirements of general purpose natural rubber and polychloroprene compounds. There are a number of other changes.

Copies can be obtained from H.M.S.O. price 3d.

Education and Training of Laboratory Technicians

In recent years there has been a growing recognition of the importance of securing increasing numbers of well trained technicians in science laboratories, including research laboratories in industrial establishments, research associations and academic institutions, as well as laboratories for instructional purposes in universities, technical colleges and schools.

The City and Guilds of London Institute has considered the whole question of the training and certification of laboratory technicians on the broadest lines and a national basis, having recently adopted regulations and syllabuses for the "Laboratory Technicians' Intermediate Certificate Examination," which come into force in the academic session 1950-51.

The scheme has been prepared by the Institute's Advisory Committee on Laboratory Technicians' Work, upon which serve representatives of laboratory technicians, professional institutions, industry, the Ministries of Education and Supply, and associations concerned with technical education.

Copies of the scheme are available on request from the Department of Technology, City and Guilds of London Institute, 31 Brechin Place, South Kensington, London, S.W.7.

British Standards Institution Annual General Meeting

The Annual General Meeting of the British Standards Institution will be held on Tuesday, July 11, 1950, at 3 p.m., in the Council Room of the Institution at 24 Victoria Street, London, S.W.1.

Changes of Address

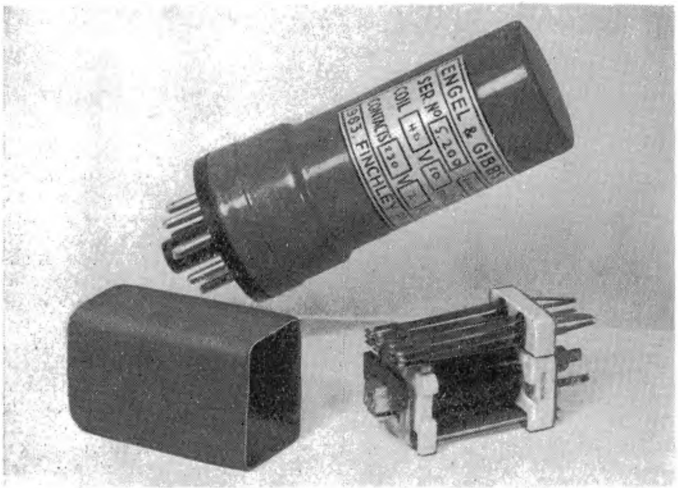
The Department of Scientific and Industrial Research has moved from Dorland and Rex House, Regent Street, to Charles House, 5-11 Regent Street, London, S.W.1. The telephone number remains Whitehall 9788.

New Electronic Products, Ltd., have opened new premises at Glasgow Terrace, Grosvenor Road, London, S.W.1, to which address all communications should be sent. The telephone number is Victoria 0834/5.

The offices of the Society of Relay Engineers have been transferred from 23 Dalkeith Place, Kettering, to Albion Road, Kettering, telephone Kettering 3422. The Secretary is Mr. T. H. Hall, M.Brit.I.R.E.

Electronic

A monthly record of British elec-
accessories, compiled from information



Plug-in Midget Relay

FITTED with aluminium cover processed to protect it from corrosion and adverse conditions encountered in the tropics, the relay is fitted with standard 8-pin international octal base. Rating and operating characteristics are shown on a side transfer. Circuit diagram and internal connexions are shown on a transfer fixed on top of the cover.

The relay operates on A.C. or D.C., the standard type requiring approximately 1.5W on D.C. and 4.5 V.A. A.C. to operate it. In the standard double-pole changeover version, the contacts are rated at 250 volts 3 amps. per pair non-inductive load A.C. 50 cycles. On D.C. the contacts are derated to 24V D.C. 3 amps or 200 volts D.C. 0.1 amp., where a suitable arc quenching circuit is arranged external to the relay.

Miniature Relay

THERE are two types, both the same physical size, the "dust cover" type and the "hermetically sealed" type.

This relay will operate on D.C. only and requires approximately 1W for standard operation. If A.C. only is available for control, a suitable rectifier and smoothing circuit has to be incorporated external to the relay. The contact arrangement is normally double-pole changeover type. The contacts are designed primarily for low voltage D.C. work, but will handle up to 300 mA, at 240V A.C.

The coil can be wound for direct operation on voltages up to 50V D.C. and for higher control supply voltages a series resistance is required.

The relay is suitable for chassis mounting and single-screw fixing is provided together with a series resistance if required.

Engel and Gibbs, Ltd.,
983/5 Finchley Road,
Golders Green,
London, N.W.11.

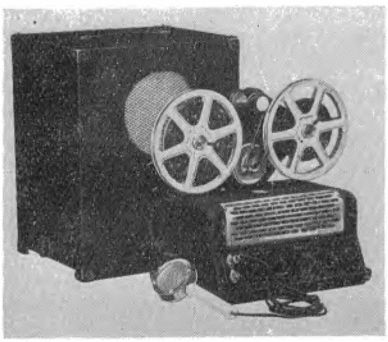
Magnetic Tape Recorder

(Illustrated below)

THE Cine-Soundmaster is a portable electronic sound recording/reproducing unit which, by means of a special flexible coupling, can be easily connected to most types of 8 mm., 9.5 mm., and 16 mm. sound or silent projectors. In this way a sound commentary with musical background and effects may be added to an otherwise silent film.

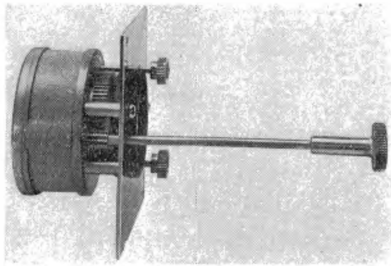
The recorder operates with spools of standard ¼ in. iron oxide plastic tape of sufficient length to provide 30 minutes of playing time. Errors in recording can be erased and the recording tape used many times over for fresh recordings. For the purpose of editing, the tape can be cut and rejoined without difficulty.

Scophony-Baird Ltd.,
Wells,
Somerset.



Valve Handling Tools

ILLUSTRATED right is one of two types of valve extracting tools which have been added to the "Spearette" range of miniature valve handling tools. The larger of the two tools, i.e., B8B size, has been modified for extracting the Mazda S.P. 41 and 61 valves. The B8A



Type Extractor is priced at 7s. 6d., and B8G, B8B and S.P.41 at 10s. 6d.

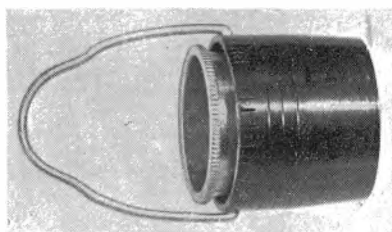
Spear Engineering Co., Ltd.
Titan Works,
Limpsfield Road,
Warlingham, Surrey.

Haynes P.M. Focusing Units

(Illustrated above)

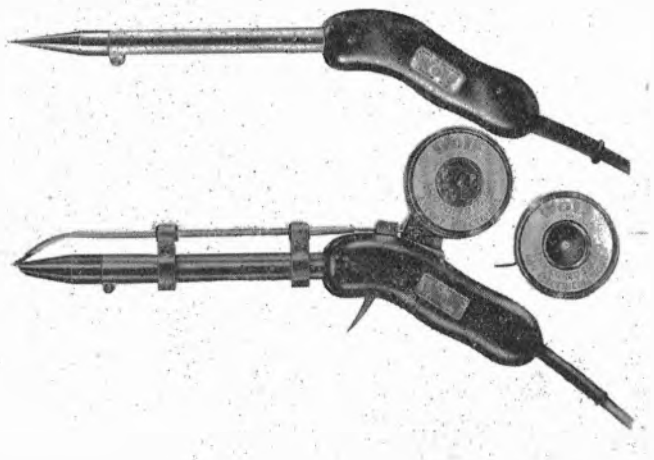
TWO new types have been introduced, the PM15A and the PM20A for use with tetrode and triode C.R. tubes respectively. These are fitted with vertical and horizontal shift controls conveniently placed at the back, as well as a focusing adjustment operating at the rear and beyond the base of the tube. Shift is produced by a slight movement of a "shuffle" plate which engages only a small part of the flux in the region of the gap, and the picture may be critically centred without derangement of focus. Resort to tilting is thus unnecessary, the focusing unit being firmly mounted with its axis aligned and concentric with the neck of the tube, for which support is provided. The focusing control by variable gap works through gears, and the control knob may extend out at the rear of the cabinet, affording the same ease of adjustment as that offered by the variable resistance used with focusing coils.

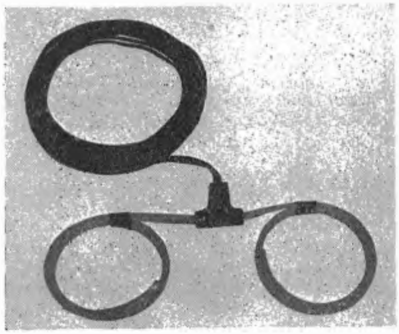
Haynes Radio Ltd.,
Queensway,
Enfield, Middlesex.



Equipment

tronic apparatus, components and supplied by the manufacturers.





Indoor Television Aerial

(Illustrated above)

THE "Flexible Dipole" is a full-size television aerial for indoor installation; when used in good or medium field-strength areas it will give similar results to those of a standard aerial. Coaxial feeder cable of 50-55 ohms impedance is used to couple the dipole to the receiver, and a transformer rejector unit gives correct coupling of the aerial to the feeder cable.

As a rule in locations up to about 10 miles from the transmitter the flexible dipole may be installed in a ground floor room with the receiver; at distances greater than 10 miles the dipole is better installed in an upper storey or loft.

The aerial can be fixed by the side of a window behind curtains, and is therefore unobtrusive. It works equally well with the majority of receivers. There are two models, one for London and one for the Midlands, and they are both supplied with 24 feet of coaxial feeder cable.

E.M.I. Sales and Service Ltd., Hayes, Middx.

Television Receiver Knobs

(Illustrated below right)

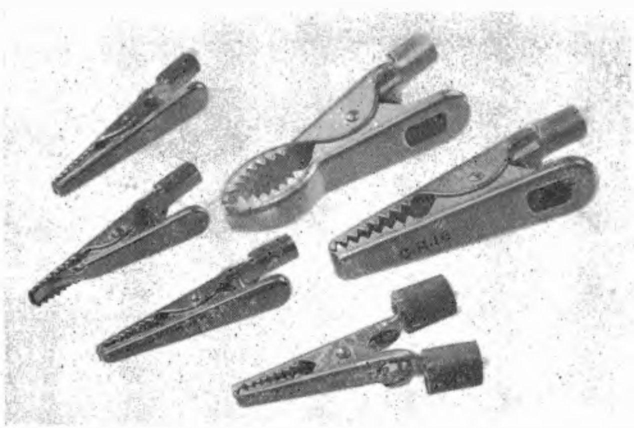
THE television receiver knobs illustrated bottom right, have been designed for better identification. The knobs' profile and straight-knurling gives a good grip and ensures fine adjustment and control, which is particularly necessary for focusing.

They all are $1\frac{1}{2}$ in. diameter by $\frac{5}{8}$ in. projection, having a standard $\frac{1}{4}$ in. diameter bore for $\frac{1}{4}$ in. shafts. All the knobs are obtainable in black or brown, with switch, focus, volume, contrast and brightness engravings.

Crocodile Clips

(Illustrated bottom left)

NEW additions to the range of Bulgin crocodile clips are the CR. 15-18 general purposes types of 25A rating, to supplement the original 5A size varieties. Others are the CR. 21-23 "Fine Wire" types of up to 5A rating (specially made for gripping flex-strands and fine wires) and the CR. 25-28 "Battery Charger" types, specially shaped to grip car battery lugs and rod or bus-bars. These clips will not turn or displace inadvertently, and are particularly safe and useful with accumulators.



A. F. Bulgin and Co. Ltd., Bye-Pass Road, Barking, Essex.

Electric Solderguns

(Illustrated above)

A new range of Electric Solderguns has been produced by Wolf Electric Tools, Ltd., which they claim will overcome hitherto common criticism of this type of equipment. Industrial soldering requirements have long called for soldering tools which will reach operating temperature quickly, will maintain a correct constant heat, prove economical in current consumption and are "easy in the hand" for long production periods.

The illustration pictures two irons from their range of six hand models; particular attention is directed to the off-straight hand grip which tends to much more comfortable control. By arrangement within the heating element, the heat is localized around the copper bit, and with the elimination of wasted heat areas considerable economy in current consumption has been effected. At the same time tendency towards excessive temperature rise is avoided and this in turn prolongs heating element life and saves oxydization of the copper bits.

Wolf solderguns can be supplied for the following voltages: 24, 50, 100/110, 115/130, 200/220, 225/250, and each gun is fitted with six feet of 3-core cable. Models are available for a wide range of purposes from fine instrument to heavy duty work.

Wolf Electric Tools, Ltd., Pioneer Works, Hanger Lane, Ealing, London, W.5.

BOOK

Elements of Electronics

By G. Windred. 197 pages, 100 figures, 6 plates. Chapman & Hall, Ltd., 1949. Price 15s.

THE subject of "Electronics" covers an extremely wide field, as readers of ELECTRONIC ENGINEERING are undoubtedly aware, ranging from atomic theory, circuit theory and wave propagation theory to practical engineering topics such as sound films, sound broadcasting and television. To cover this wide field within the scope of a single volume may well be considered an impossible task; nevertheless the author of this book, the late Mr. G. Windred, has been reasonably successful in giving the reader a bird's-eye view of the subject. Such a view is not without value since it enables the reader to see clearly how the fundamental researches of Sir J. J. Thomson, Lord Rutherford, Sir J. A. Fleming and many others laid the foundations for developments which affected directly or indirectly the lives of every human being in the civilised world.

The early chapters of the book survey in broad outline the discovery of the electron and the classical researches which enabled the charge and mass of the electron to be evalua'ted, the principles of thermionic emission and the behaviour of electrons *in vacua* and gases, and the photo-electric and photo-voltaic effects. The characteristics of thermionic valves are then discussed and the uses of valves as oscillators and in amplifiers are explained, including some reference to magnetrons and velocity-modulated valves, e.g., klystrons.

The later chapters cover the applications of electronic techniques; these range from mercury arc rectifiers for supplying thousands of kilowatts of power to electric traction systems, X-rays for medical and industrial purposes, cathode ray tubes, television, the recording and reproduction of sound on film, the electron microscope, cyclotrons and betatrons for producing high-voltage electron and ion beams, and finally radiolocation. Each chapter concludes with a list of book references for further reading, with comments on the treatment the subject is given.

The drawings are generally clear but there are a few errors, for example grid bias is omitted for the second stages of the amplifiers in Figs. 51 and 52, the grid-leak is missing in the quartz crystal controlled oscillator shown in Fig. 58 and the potentiometer P in Fig. 65 should be connected to the positive side of battery B_1. There are also some misconceptions, for example on p. 156 it is said that "the speed of propagation of radio waves is 186,000 miles per second and the entire process of signal production, transmission and reception is virtually instantaneous . . . this is an important factor which makes television possible." The important factor is not, of course, the speed of propagation but that all components of the signal should be transmitted with the same speed.

Perhaps the most important omission is the lack of an adequate account of the control of industrial processes by electronic techniques. It is in this field that electronics has made a most valuable contribution to mass production methods, notably in the United States of America, and it may well be that industry in this country would benefit from the wider use of such techniques.

However, the author has on the whole selected his material well and given it a clear and concise treatment, well suited to the needs of the general reader and the student of electrical technology who requires a survey of the subject. The specialist in any one of the many aspects of electronics covered by the book will probably find his particular subject dealt with all too briefly, but this is inevitable in view of the limited space available.

W. J. BRAY.

Electric and Magnetic Fields

By Stephen S. Atwood. Third Edition Chapman & Hall, Ltd., 1949. 475 pp. Price 44s.

THIS book, by the Professor of Electrical Engineering at the University of Michigan, is intended to serve as an introduction to the mathematical theory of electricity and magnetism for first and second-year students of physics and electrical engineering. It is in four parts dealing respectively with the electric field, the magnetic field, the ferromagnetic field, and combined electric and magnetic fields. The mathematical treatment is simple and concise and should be intelligible to anyone with an elementary knowledge of the calculus (there is a glaring mathematical error on p. 193). The physical principles underlying the mathematical results are clearly explained, and in the first chapter a statement of the way in which the theory is built up from experimental results is given.

Considerable emphasis is placed throughout the book on the plotting of electric and magnetic fields, and two chapters are devoted to the

REVIEWS

solution of Laplace's equation by field-plotting methods. There are a large number of carefully drawn field diagrams and many of these are for systems of practical importance, such as transmission lines and cables. An admirable feature is that there are frequent references to atomic and nuclear theories and the light they shed on electric and magnetic phenomena. An appendix entitled "Electrons in Atoms" gives the necessary background of atomic structure and quantum theory. Other appendices deal with the conversion of systems of units (in accordance with modern practice this edition has been rewritten using the rationalized M.K.S. system) and a brief historical survey.

This book will be found useful by the class of students for whom it is written, and it forms an excellent introduction to more advanced works on the subject. One minor criticism is that there is no mention of the very powerful method of solving Laplace's equation by conformal transformations.

J. BROWN.

Industrial Electronic Control

By W. D. Cockrell, pp. 385 + XV, 2nd. Edition. McGraw-Hill Book Company, Inc., London, 1950. Price 34s. net.

FIRST produced in 1944, this second edition of the book contains additional matter on electronic counting and switching, and closed-cycle systems. Primarily, the book is written for electrical engineers with little or no experience of electronic techniques, and accordingly starts with a section (six chapters) on valve fundamentals, with special reference to industrial types such as gas-filled rectifiers, ignitrons, thyratrons, and the like. Then follows five chapters on other circuit components, and not till nearly half-way through the volume does any discussion of complete circuits start; with five chapters on basic units, rectifiers, amplifiers, oscillators, timing and switching circuits, etc.

The added thirty page section on closed-cycle control systems is treated in a mildly mathematical manner and is adequate for its purpose, giving due deference to Nyquist and Bode, although perhaps the information is not sufficiently linked with practicalities. The book ends with what is probably the most useful section, on practical industrial systems, illustrated in fair detail from commercial examples. Especially covered are photo-electric systems; D.C. motor control; A.C. relays, timers and power control systems; and welder controls. The more "refined" industrial controls of temperature, pressure, moisture and pH, are not mentioned, presumably because they are so often better done by non-electronic methods, nor are many other points which while perhaps, strictly speaking, not being control problems, might have been included.

A lengthy appendix contains the inevitable colour codes and nomograms, optical data and some other miscellanea. The book is, of course, American, and has the advantages and disadvantages so implied, but will be of general use to the British reader. It will find its greatest service as an elementary introduction for the uninitiated or as a general reference volume for the practical man.

E. D. HART.

Principles of Electricity and Electromagnetism

By G. P. Harnwell. pp. 670, 2nd. Edition. McGraw Hill, Ltd. 1949. Price 36s.

THE important thing to realize about this book is that if you adopt it you are taking the plunge into the 20th century and leaving behind the 18th-19th century approach to "electricity and magnetism." (Most of us realize that this must be done some time, though we may shirk the effort on the pretext of fearing the effect on the student who is forced to change his system of units somewhere between matriculation and a university degree.) For the most obvious feature of this book is the use of rationalized M.K.S. units, which simplify the equation of electromagnetic fields as in transformers and radio waves—at the expense of complicating slightly the formulae for forces between charged pith balls and between permanent magnets. The other "modernisms" are the derivation of magnetic phenomena entirely from the effects of electric currents (independent of magnetic poles), and the choice of B rather than H as the basic magnetic quantity. This last feature has been over-done. To describe $H = B/\mu$ as a concept convenient for describing the properties of magnetic materials surely underrates the value in electromagnetism of the idea of "magneto-motive force"; and when, having found that B has the dimensions of volt-sec. per square metre, we are told that "The volt-second is called the *weber*, hence B is measured in units of weber per square meter," we stop to ask whether there ever was a man named Weber who was interested in ferromagnetism.

BOOK REVIEWS (Continued)

Considering the size of the book, one is at first surprised at the extent to which reference is made to other treatises for fuller details of almost every topic, though one is pleased to see British books represented in these references by such authors as Hague, Jeans, Mott and Moullin. The reason for limited treatment is that the scope of this book is really much wider than its title, and covers three fields:—

(1) The laws of electromagnetism, i.e., fields associated with charges and currents, leading to Maxwell's equations and radiation.

(2) The physics of materials used in the applications of electromagnetism, covering such features as thermo-electric and Hall effects, dielectric polarization, Fermi statistics in relation to semi-conductors and a particularly thorough section on the physics of magnetic bodies.

(3) An account of the applications of electromagnetic principles to electrical machines, radio, etc.

The treatment of the applications is necessarily introductory rather than comprehensive, but the standard reached is indicated by the inclusion of reference to the following: Johnson noise and shot noise (the latter without mentioning the effect of space-charge); the gyromagnetic frequency in the propagation of radio waves through the ionosphere; waves in a rectangular guide; the synchrotron particle accelerator; the Scott connected transformer for conversion between 2-phase and 3-phase power; the use of a synchronous motor as a capacitor; Nyquist's criterion for stability of a feed-back amplifier.

There are five appendices, but the weakest point is the index, which lacks cross-references: thus Nyquist's stability criterion is not to be found under "Nyquist" or "stability," but only by implication under "amplifier, feedback."

D. A. Bell.

Electrical Transmission of Power and Signals

By E. W. Kimbark, 461 pp. 229 diagrams. 23 tables. John Wiley & Sons, Inc., New York. Chapman & Hall, Ltd., London, 1949. Price 48s.

THIS book is intended for junior and senior students in electrical engineering and the author claims that it is the first book ever written that presents the basic theory of transmission lines together with applications to the three fields of power, telephony, and ultra high frequencies.

The first five chapters are devoted to describing and deriving the parameters of simple two conductor lines, multi-conductor lines, stranded and miscellaneous conductors and ground return lines.

In Chapter 6 the reader is introduced to the formulation of the differential equations representing the transmission of current and voltage along a transmission line with uniformly distributed parameters, and to the solution of the equations for the steady state case.

The representation of a smooth line as the extreme case of an infinite number of T or π networks in cascade is then dealt with, and the point is emphasized that by this means the use of differential equations may be avoided.

The next chapter is concerned with transient phenomena on transmission lines, and although presented in a clear pictorial manner the emphasis is far too strong on lightning surges and not enough on the equally important aspect of waveform distortion of signals in communication systems and the significance of this in television transmission, for example.

Chapter 12, dealing with telephone and telegraph transmission, is perhaps the least satisfying in the book and parts of it might well be omitted altogether. The statement on p. 263 that different groups of carrier frequencies are generally used for transmission in each of the two opposite directions, applies only to systems on open wire lines, which do not form the majority of carrier circuits in this country or in America.

Wave filters are discussed adequately in 26 pages, but it is surprising that other important networks such as attenuation and phase equalizers are dismissed in a few lines.

The book concludes with a chapter on Wave Guides presented in a simplified form, while the two appendices contain a lot of useful information on conductors and transmission lines, presented in tables and curves.

Each chapter contains a good bibliography, and the author has achieved a welcome consistency in the style and standard of presentation. This book may be recommended to all students of electrical transmission courses.

L. I. Farren.