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The B.B.C.'s Annual Report

IGNORING the sociological and the many controversial issues contained therein, the B.B.C.'s annual report* which has just been published, should have considerable appeal to all those engaged professionally in the radio and television industry.

It is a document of some 90 pages covering all aspects: broadcasting, television, engineering, administration and finance involved in the working of a corporation which serves 11,898,400 sound and 345,000 television licence holders.

The emphasis in the B.B.C.'s report is, unquestionably, on television—there was nearly a three-fold growth in the number of television licence holders during the year ending March 31, 1950—and the plan for the development of television is here unfolded. Approved by the Government and first announced by the Postmaster-General in the House of Commons on November 9, 1949, the scheme provides for five high-power and five low-power television transmitters to be completed by the end of 1954 to bring 80 per cent of the population of the United Kingdom within reach of television.

Along with the development of transmitters is the plan for the 13-acre site at White City for a permanent building for the television headquarters and studios, but limitation of capital expenditure make it impossible to say when the plan will even be started.

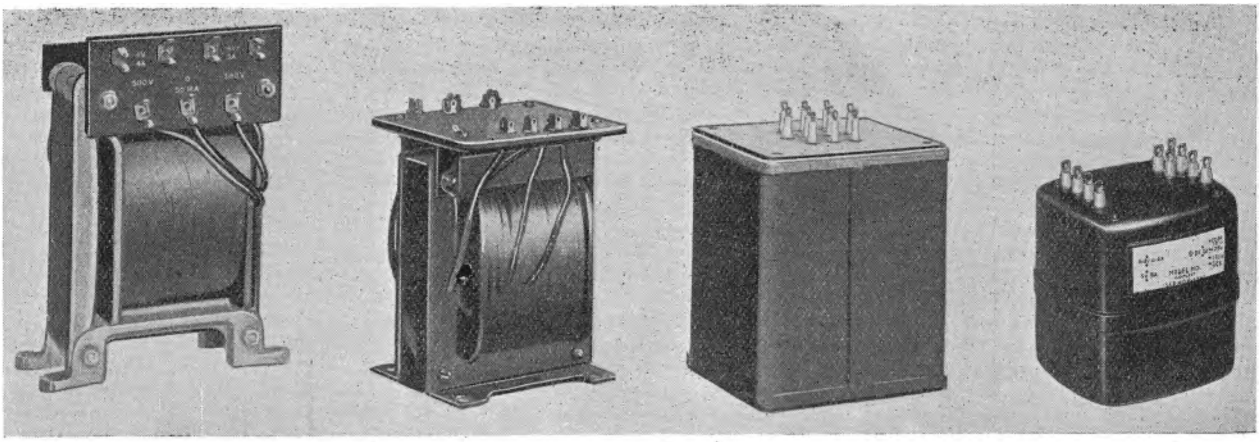
It is realized that in giving this high priority to television development some curtailment is inevitable in other directions, and, one of these directions is, unfortunately, the V.H.F. Station at Wrotham, Kent. To quote from the report itself "The B.B.C. attaches great importance to the future of V.H.F. Broadcasting as a means of eventually overcoming the difficulties arising out of the shortage of long and medium wavelengths for broadcasting" and it is apparent that in order to give adequate coverage in this country resort will have to be made to the shorter wavelengths.

The station at Wrotham is complete and contains amplitude-modulated and frequency-modulated transmitters so that a direct comparison can be made between the two systems and the "AM versus FM" question settled, at least as far as broadcasting is concerned.

We understand that full-scale field trials are shortly to be started from Wrotham, but the B.B.C. is, however, most anxious not to commit itself to maintaining a regular programme schedule at the onset, nor indeed is this likely until the television programme is completed. While those out of range of Alexandra Palace and Sutton Coldfield still await the coming of television listeners in the Home Counties cannot expect this additional service in the way of high fidelity transmissions.

There is no doubt whatsoever about the marked improvement in quality of the present V.H.F. transmissions. Whether these will be A.M. or F.M. depends on a number of factors, apart from considerations of quality.

* Annual Report and Accounts of the British Broadcasting Corporation for the year 1949-50 (Cmd 8044 H.M.S.O. Price 3s.).



Developments in Small Chokes and Power Transformers*

Four Transformers all having the same electrical characteristics.
Reading from left to right: 1936 design; 1940 design; 1946 design—fabricated case, oil filled; 1949 design — 'C' core, deep drawn case, oil filled.

BEFORE 1938, small power transformers used in ships' wireless equipment were normally of the open type of construction and were designed to withstand very heavy overloads. The insulation that was specified had a very big margin of safety, and the transformers, although somewhat large and heavy, gave reliable service and had a long life.

With the introduction of radar and more complicated wireless equipment the demand for transformers having a smaller physical size for the same electrical characteristics became insistent, and they were required in very large numbers to meet the wartime demand. This demand was met by providing transformers designed and manufactured to the commercial standards that then prevailed in industry. Their size was smaller than the original type used in Naval Service, this reduction being achieved by reducing, to some extent, the margin of safety and also by adopting core-clamping frames of lighter and less bulky construction. Generally speaking, these transformers were unreliable, but as at the time radar was an adjunct to, rather than an integral part of, the fighting equipment of ships, this unreliability was accepted as a matter of course and little adverse comment reached the designers to spur them on to producing better and more reliable transformers.

By 1944 equipments were envisaged where the radar played a far more important rôle in the efficiency of the fighting equipment of ships than had been the case before, and an investigation was made into the causes of unreliability of the existing transformers. This investigation proved to be difficult, but the main causes of trouble were tracked down to:—

Solvent.—During the baking of an ordinary varnish-impregnated coil, solvent is driven off, some of this being trapped in the windings. This solvent will attack the insulation and eventually cause breakdown.

Moisture.—The penetration of moisture into a winding reduces the insulation resistance and results in premature failure.

Non-homogeneous Insulants.—Where adjacent layers of insulation comprise materials of widely different dielectric constants, this causes a non-uniform distribution of the voltage stress in the insulation, and eventually causes breakdown.

In 1945 the Admiralty introduced a new process of wax impregnation followed by a bitumen coating. This process produced an immediate improvement in reliability, but had a number of disadvantages. The process was expensive, and difficult to use because noxious materials were involved. Also the bitumen coating was liable to be damaged and so allow the ingress of moisture.

At the same time investigations were started with a solvent-less varnish. This varnish has now been used in some experimental transformers which have been in continuous use since 1945 and no deterioration has yet been detected. It is therefore considered that this varnish is satisfactory and it has now been adopted as a standard impregnant. It cannot, how-

* Compiled from data kindly supplied by the Admiralty

ever, be considered as ideal as it is expensive and needs very careful processing to produce reliable results. This is particularly important as it has not yet been possible to devise a test that will detect an unreliable transformer directly after it has been impregnated. It does, however, appear that, if very careful quality control is used at all stages of processing, satisfactory results can be achieved.

Hermetically Sealed Transformers.

It has been realized for a long time that one solution to the problem of producing a small reliable transformer was to house the core and winding in a hermetically sealed case filled with oil or other insulating material. Experimental work in this field was carried out by several service establishments and also in the commercial research laboratories.

In 1946 the Admiralty placed the first large production contract for hermetically sealed transformers for service use.

Many difficulties were encountered. The metal cases were fabricated and it proved very difficult to make them leak-proof; also, being rectangular in shape, they were very liable to damage at the corners, a knock being sufficient to split a seam and cause leakage. The electrical connexions had to be brought out through glass or ceramic terminal seals which were fragile and proved to be a continual source of trouble.

Transformers of this type are extremely reliable from the electrical point of view, but are usually too fragile for general service use, besides being bigger and heavier than the original open type, because of the extra bulk and weight of the case and its oil filling.

Hermetically Sealed " C " Core Transformer

The next big advance came in 1949 with the advent of the hermetically sealed "C" core transformers. This advance was made possible by the simultaneous availability of three new developments:—

(a) Terminal seals made of fused aluminium oxide.

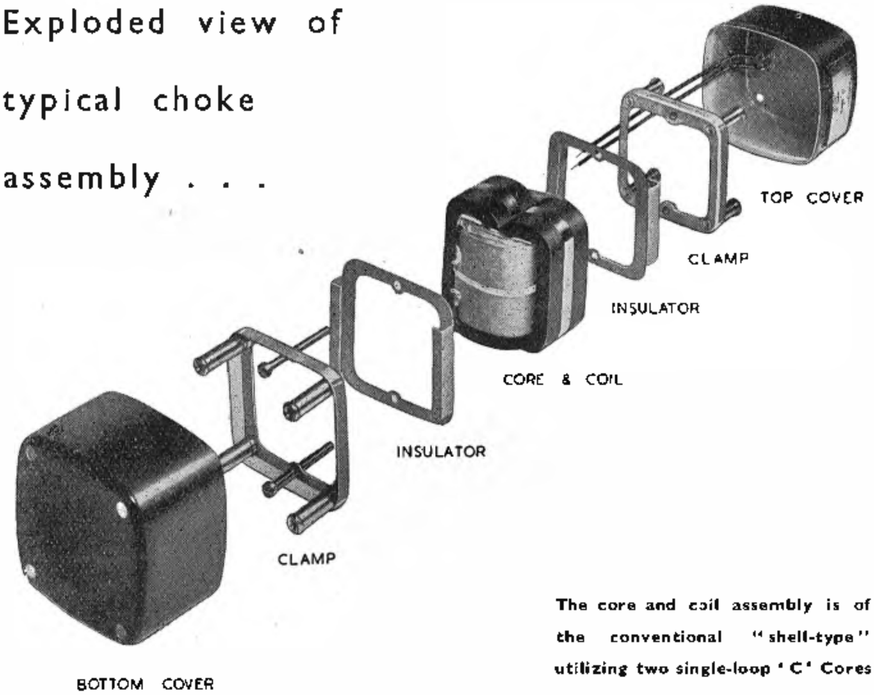
(b) "C" type transformer cores.

(c) Deep-drawn two-piece metal cases.

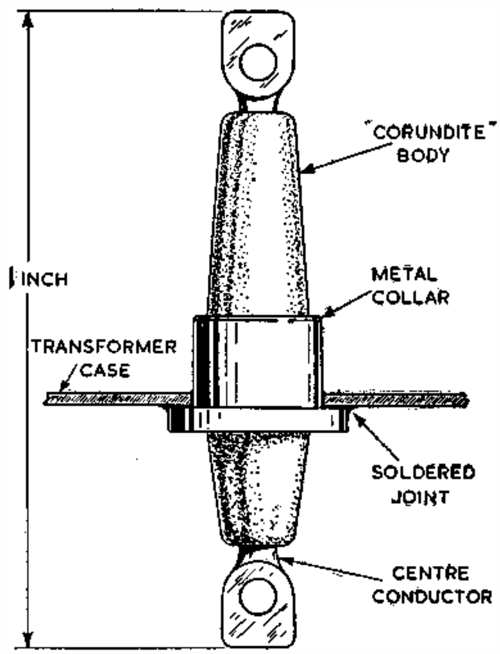
These developments are described more fully below:—

(a) Fused aluminium oxide is a material of the synthetic ruby type, nearly as hard as diamond and possessing very good electrical insulation properties combined with mechanical properties far superior to those of glass or ceramic. It is moulded like clay into the required shape and then fired. The resulting substance can be ground, when accurate dimensions are required, and glazed to give it a smooth surface. The usual technique is to provide a slight taper on the insulator to accommodate a flanged collar which is pressed on. The inside of

Exploded view of typical choke assembly . . .



The core and coil assembly is of the conventional "shell-type" utilizing two single-loop ' C ' Cores



Typical terminal seal, with insulant of fused aluminium oxide the collar is tinned to make a layer of soft metal which will make a very intimate contact with the insulator and prevent any possibility of leakage. The lead through the centre of the insulator is sealed in with glass of the "Pyrex" type. The design of these insulators was sponsored by the Admiralty and the product has proved to be extremely robust and satisfactory.

(b) High efficiency "C" cores were originally made in the U.S.A., but are now being produced in bulk in this country. They are made from a low silicon content grain-orientated iron, produced in the form of flat strip insulated on both sides. The strip is rolled in such a way that the individual crystals are aligned with their cube edges parallel to one another, so that, provided the strip is fluxed in the direction of rolling, extremely good magnetic properties are obtained. The strip may be slit to any desired width, and may be wound tape-like on to a mandrel, the number of turns depending on the size of core required. The core is impregnated with a bonding material to hold the turns together, and then cut in two. The resultant pieces are shaped like the letter C and it is from this that the name is derived. The two halves are always kept together as a pair, in order to ensure that when re-assembled the butt surfaces will make as perfect a contact as possible. Either two or four C's can be used for each transformer. It has been found that it is best to use four C's forming a core of the shell type for normal sizes of transformer and two, which will form only a single loop, for the smaller ones.

Advantages of the "C" core when compared with orthodox silicon iron stampings include:—

(i) A reduction of as much as 75 per cent in the magnetizing VA required to produce a given flux density.

(ii) A reduction of as much as 45 per cent in the iron losses at a given flux density.

(iii) An increase of as much as 30 per cent in the permissible flux density.

(iv) Improved incremental permeability, which is of value in power chokes and audio-frequency transformers.

(v) Reduced flux leakage due to concentration of the magnetic field.

(vi) Improved space factor of up to 95 per cent due to the thin coating of insulation.

(vii) Reduced time for assembly of core and coil.

In practice these advantages mean that it is possible to design "C" core transformers and chokes that are as much as 40 per cent smaller than their counterparts using orthodox laminations. In order to cater for the many different applications of "C" cores they are being produced in 32 different sizes, each available in three thicknesses of strip, as follows:—

(i) 0.013 in. for power transformers operating at 50 c/s. and 60 c/s., also for vibrator transformers and chokes.

(ii) 0.005 in. for power transformers operating at 400 c/s. and above.

(iii) 0.002 in. for pulse transformers.

(c) Experience had shown that the early type of fabricated case was unreliable in service and therefore work on an entirely new series of cases was started by the Admiralty early in 1949. By May of that year contracts had been placed with industry for the development of these designs on a production basis. Advantage was taken of the commercial experience already gained in the use of cylindrical deep-drawn cases for orthodox laminations. These had been highly satisfactory in performance, but had occupied excessive chassis space due to their circular plan. It was therefore decided to combine the advantages of deep-drawn cases with a shape following the core contour more closely. The contour which was finally adopted gives the minimum plan area consistent with the high degree of strength required. The material chosen for the cases was steel, which provides a magnetic screen around the core. High permeability nickel-iron alloy is used,

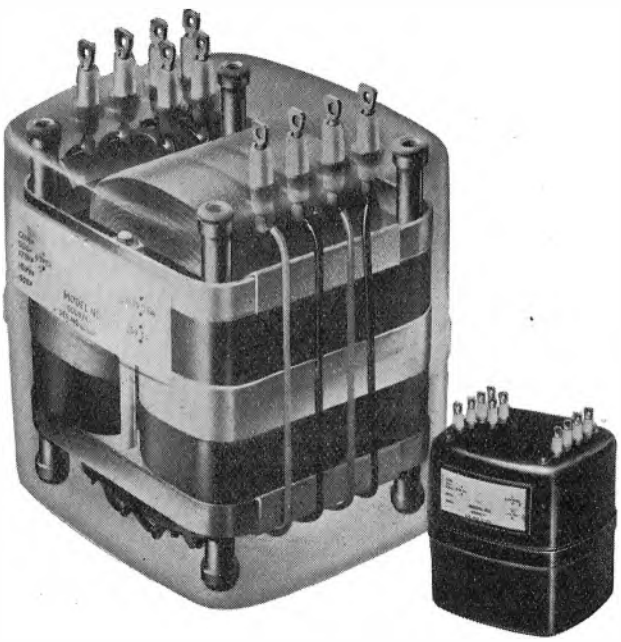
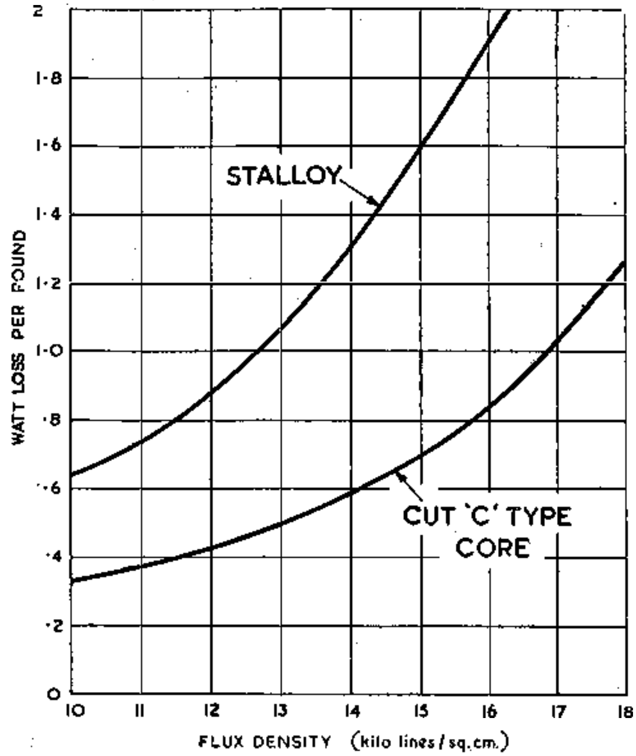


Illustration showing construction of typical 'C' Core Transformer

however, where a higher degree of screening is required. These cases have an important advantage over the fabricated types, in that the solder seam is reduced in length by about two-thirds with a corresponding reduction in the risk of oil leakage.

Furthermore, where extremes of temperature are experienced, the pressure inside a transformer case will vary, and if a fabricated case is employed, each side will act as a diaphragm and will be flexed accordingly. Thus the solder holding the sides in position will also be subjected to continual flexing. Over a period of time this can cause the solder to become crystalline which, in turn, will result in split seams and oil leakage. However, with the deep-drawn case the one seam required is located round the girth of the transformer so that whatever flex-



Curves comparing the iron losses of Stalloy and 'C' type Cores

ing of the case may occur there is no relative movement between the two joint faces.

On account of the high cost of each set of tools required to draw the cases the number of tools was limited to 12, but by allowing for different depths of draw, 32 different sizes of case are available.

A range of internal clamping frames has been designed which hold the cores securely under the worst conditions of shock and vibration that are likely to be met in practice. Here again, deep-drawn pressings have been used in preference to the conventional folded construction.

In building up a "C" core hermetically-sealed transformer, the requisite number of insulators with tails attached are soldered into position in the top half of the case. The upper framework is secured by hank bushes to the case, and the coil and core assembly is then clamped to this framework. The tails from the insulators are then connected to corresponding terminal tags on the face of the coil. The bottom half of the case is then fitted to the assembly and secured by means of further hank bushes inserted from the outside. The lap-joint is then soldered and the case filled with oil, an air bubble being left to allow for expansion.

By using the hank bushes at either end, the transformer may be secured to the chassis in either an upright or inverted position. This arrangement ensures that the weight of the coil and core assembly is transmitted directly to the chassis and not through the case. The strain imposed on the case is thus reduced to negligible proportions.

Extensive tests on this type of transformer have been satisfactorily completed as follows:—

(i) Type-approval tests to Inter-Service Specification RCS.214.

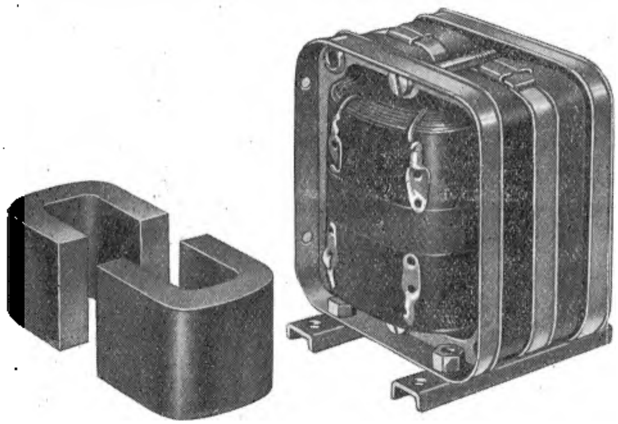
(ii) Each size of case has been tested hydraulically to withstand a pressure of 100 lb. per square inch.

(iii) A number of transformers have been tested to destruction by imposing overloads of up to 400 per cent on all secondary windings. Other tests, in order to obtain a hot spot, have been made by a dead short circuit on one secondary winding only. Results of these tests show that often a temperature is reached which remains steady and, provided this is below 200° C., the only result has been a slight bulging of the container top and bottom, the transformer itself being undamaged. Where the temperature has exceeded 200° C. the solder at the seam has melted sufficiently to release excess pressure, and in some instances a small quantity of filling material has seeped out; this was insufficient to cause material damage to surrounding components. Had similar overloads been applied to open type transformers the risk of fire could not have been excluded.

Standardization

In view of the rigorous tests imposed it is not surprising that these transformers are the first and only type to qualify as fully Inter-Service Type-Approved Standards. Although the design of this range is Admiralty property, it may be used freely for commercial purposes by any manufacturer who so desires. Where firms do not wish to tool the range for themselves, they can obtain supplies of cases, internal fittings, etc., through commercial sources. More than twelve firms have already taken advantage of this arrangement, and, as their production increases, so a greater flow of these standardized transformers and chokes will reach the home and export markets.

A single-loop 'C' core, and a typical open-type 'C' Core Transformer for applications where a performance slightly below that of the hermetically-sealed type can be accepted



The Video Output Stage

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Part II

The Anode Bend Demodulator Video Output Stage

THE use of a separate demodulator may be avoided by employing the video valve as an anode bend demodulator. The advantages which might be claimed for this are as follows:—

(i) a valve structure (the vision demodulator) is saved,

(ii) the damping on the last R.F. (or I.F.) circuit may be reduced,

(iii) the use of an anode bend detector results in a higher gamma for the receiver.

Advantage (i) may not necessarily be realized, since the gain of an anode bend demodulator video stage is not high, so that with certain receivers an extra R.F. (or I.F.) stage may be necessary. Advantage (ii) may not necessarily be realized if the anode-grid capacitance of the valve is high in value since the effective input conductance and capacitance may be increased. This point will be dealt with subsequently. Advantage (iii), although possibly desirable from the point of view of its effect on the picture quality (although this is open to doubt), may be a disadvantage in circuits where the synchronizing pulse amplitude required is greater than 3-4 volts peak.

In connexion with the effect of the video valve on the overall receiving system gamma, the majority of picture tubes have a gamma of 2.5-3.0. This gamma combined with a transmitted gamma of 0.4 from the Emitron (or from the C.P.S. camera with gamma degradation) results in an overall system gamma (televised scene to receiver picture) of unity or slightly greater than unity. An overall system gamma slightly greater than unity is desirable since the receiver picture is black and white, and so the loss of apparent contrast due to the absence of colour contrast is somewhat restored.

However, an overall system gamma very much greater than unity (say greater than 1.5) results in accentuation of the highlights and cramping of the blacks (and incidentally the synchronizing pulses). Although this may result in obtaining acceptable pictures when the televised scene is badly lighted (such as an O.B. on a dull day) in general, for well lighted televised scenes the results may be unacceptable.

Thus care must be taken in incorporating in any part of the receiver a stage which might increase the overall gamma of the system to greater than 1.4 if acceptable pictures are to be obtained.

From the point of view of an analysis of operation, it is convenient to distinguish two modes in which the system may perform. Mode I is distinguished by the steady bias being at, or in the region of cut-off, while Mode II is distinguished by operating the valve with a standing anode current, so that the anode current is never cut off even on signals corresponding to peak white. These two modes are demonstrated in Fig. 5a.

Mode I

We shall assume the I_a/V_g characteristic to be approximated by the law

$$I_a = kV_g^n \qquad (11)$$

where V_g is measured from cut-off.

Then if an R.F. (or I.F.) signal $A \sin\omega t$ is supplied to the grid of the valve, the mean signal current output is given by

$$\bar{I}_a = kA^n/2\pi\int_0^\pi \sin^n\omega t \; d(\omega t)$$

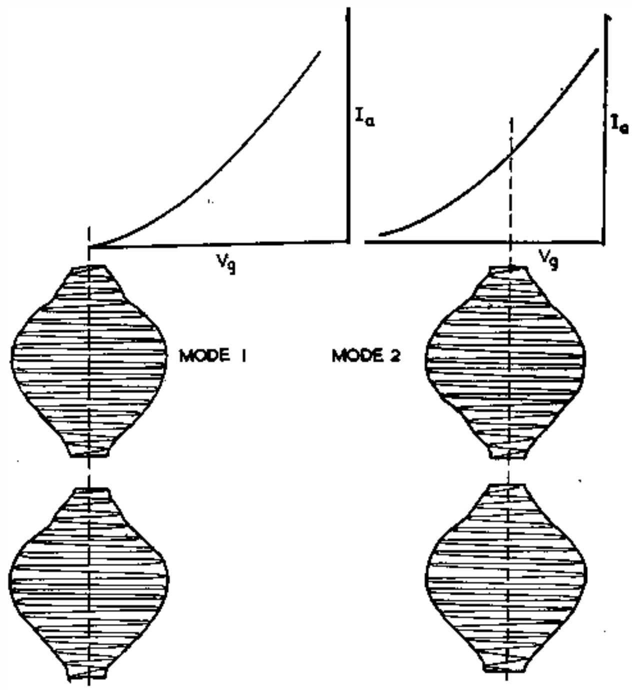


Fig. 5 (a)
Operating modes for the anode bend demodulator video output stage

$$= kA^n/2\sqrt{\pi} \times \frac{\left(\frac{n-1}{2}\right)!}{(n/2)!} = kA^n\psi \qquad (12)$$

The function ψ is plotted in Fig. 5b against n. It will be observed that the video current output for the same input swing is greatest for a linear I_a/V_g characteristic provided k is the same.

Expression (12) may be used to calculate the efficiency curve for any valve. The resulting curve will be most in error for very small input swings, since the law (11) will fail to hold in the region of cut-off.

Assuming $n = 2.3$ and $k = 0.151$ for the Type UL46 Fig. 5(c) has been constructed.

It will be noted that for the UL46 operated under these conditions, the effective gamma is 2.3, so that if the ratio of synchronizing signal to peak picture signal at the input is 30 per cent this ratio is reduced to 6.3 per cent at the output.

Using the UL46 with an anode load of 1.5 kΩ, a video current swing of 20 mA is required for 30 volts video output. This requires an input swing of 10.9 volts R.M.S., and the amplitude of the output synchronizing pulses is only 2 volts peak.

Mode II

We shall again assume the I_a/V_g characteristic to be given by

$$I_a = kV_g^n \quad \dots\dots\dots\dots\dots\dots\dots \quad (13)$$

while the applied signal is $B + A \sin \omega t$, corresponding to a standing bias (measured from cut-off) of B volts, with an applied R.F. (or I.F.) signal of peak value A volts. We shall further assume that the valve is not swung to cut-off so $A < B$.

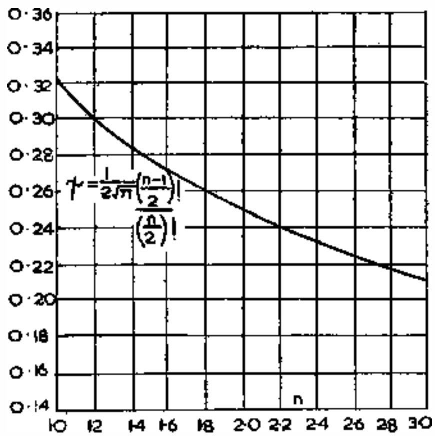


Fig. 5 (b)
Graph of function γ

The output current is then

$$k\,(B + A \sin \omega t)^n$$

$$= kB^n\left\{1 + n(A/B)\sin \omega t + \frac{n(n-1)}{2!}\cdot (A^2/B^2)\sin^2 \omega t + \dots\right\}$$

and picking out the D.C. terms yields

$$kB^n\left(1 + \frac{n(n-1)}{2.2!}(A/B)^2 + \frac{n(n-1)(n-2)(n-3)}{2^2 . 4!}\cdot (A/B)^4 + \dots\right)$$

This series converges rapidly for normal values of n, and putting $I_{ao} = kB^n$ where I_{ao} is the standing current we find for the video signal current output,

$$\bar{I}_a = \frac{n(n-1)}{4}(A/B)^2 I_{ao} = \eta A^2 \quad \dots\dots\dots \quad (14)$$

It would appear from (14) that with values of $n > 2$, highest efficiency is obtained with a high standing current, while for values of $n < 2$, a small standing current results in highest efficiency.

For the UL46, taking $n = 2.3$, we obtain the following table:

Bias volts ...	−17.5	−15.0	−12.5	−10.0	−7.5	−5.0
B (volts) ...	5	7.5	10.0	12.5	15.5	17.5
η (mA/volt²) ...	0.180	0.200	0.218	0.240	0.250	0.258

Using these values and Equation (14) the theoretical efficiency curves shown in Fig. 5(d) were calculated. The vertical strokes at the ends of the curves indicate where the grid would be swung to $V_g = -1$ and so grid current would be introduced, while the dotted portions show regions where A > B and so the foregoing theory does not apply.

Comparison of these curves with experimental curves shows that the theory holds to a reasonable degree of accuracy, so that Equations (12) and (14) may be used confidently in the assessment of the operation of any valve as an anode bend demodulator video valve.

It will be noted that the effective gamma is 2 (in-

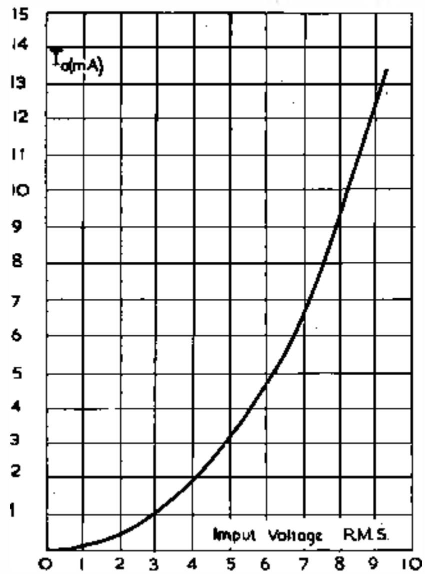


Fig. 5 (c)
Efficiency curve for the UL46 valve

dependent of the valve used) so that if the ratio of synchronizing pulse to peak video signal input is 30 per cent, the ratio at the video valve anode will be 9 per cent.

Operating the UL46 with a standing bias of 15 volts and an anode load of 1.5 kΩ, the required input is 7 volts R.M.S. for 30 volts video output and the synchronizing pulse output is 2.7 volts peak.

It will be seen that operation in Mode II is more efficient, but at the price of the standing current (15 mA in the above example) which is required.

As previously noted, the effect of the anode-grid capacitance should be considered. If the anode load consists of a resistance R and capacitance C in parallel the effective input resistance and capacitance due to anode-grid capacitance feedback at any frequency are given by:

$$R_i = \frac{1 + \omega^2C^2R^2}{\omega^2C^2R^2} \times \frac{C}{g_m\, C_{ag}}$$

$$C_i = C_{ag}\left(1 + \frac{g_m\, R}{1 + \omega^2C^2R^2}\right) \quad \ldots\ldots\ldots \quad (15)$$

The value of g_m actually varies as the input swing varies, so that if a large input resistance is required, the anode-grid capacitance should be very small, so that $1/R_i$ and C_i are small, and their effect is swamped by other circuit elements.

It will further be noted that operation in Mode I results in a higher effective value of R_i than that obtaining in Mode II since the g_m over the operating range is smaller.

If small input swings are being considered the g_m at the operating point should be used in the above formulæ. If large swings are being considered the average g_m over the portion of the grid base employed may be used.

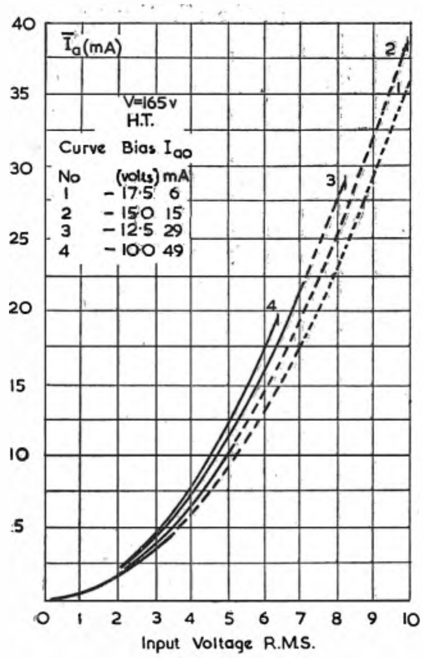


Fig. 5 (d)

Theoretical efficiency curves for the UL46

If the R.F. (or I.F.) is very large so that $\omega CR >> 1$ formulæ (15) tend to

$$R_i = C/(g_m\, C_{ag})$$
$$C_i = C_{ag} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (15a)$$

while if the I.F. is very low (about 4 Mc/s. is the lowest ever used) ωCR tends to about 4/3 yielding

$$R_i \rightarrow 1.5\, C/(g_m\, C_{ag})$$
$$C_i \rightarrow C_{ag}\,(1 + 0.35\, g_m R) \quad \ldots\ldots\ldots \quad (15b)$$

Thus, although the input resistance due to feedback is increased slightly by operating at a lower I.F., the input capacitance rises rapidly so that it is not advisable to work with a low intermediate frequency.

Microphony in Video Output Stages

The video valve employed in cathode compensated or anode compensated stages for direct-view receivers is usually a high slope R.F. pentode in which the grid-cathode clearances have been reduced to a minimum. For the anode-band demodulator video valve for direct view receivers and for the anode compensated stage for projection receivers, the valve employed may be specially designed for the purpose, or it may be an output pentode valve which is made for normal broadcast receivers. These valves may show no sign of microphony in the R.F. and I.F. stages or in the sound output stage, but be quite unusable in the video output stage. This follows in the case of an R.F. or I.F. amplifier, since the low-frequency microphonic disturbances are not transmitted by the inter-valve couplings and they can only become of importance if of sufficient magnitude to cause modulation of the R.F. signal.

The effect of microphony on the picture is twofold. It can directly modulate the picture tube, producing horizontal black-and-white bands across the raster. It can also interfere with the horizontal synchronizing producing a tearing of the picture, or with the vertical synchronizing causing the frame lock to slip. Usually these effects occur simultaneously.

The sensitivity of a particular circuit to microphony will depend on the gain from the grid of the valve under consideration to the picture tube grid.

A synchronizing pulse of about 1 volt in amplitude represents the lowest acceptable signal, and this should not vary by more than 10 per cent if synchronizing is not to be affected. This gives 0.1 volt output as the maximum interfering microphonic voltage amplitude tolerable, corresponding (taking an average video L.F. gain figure of 10) to 10 mV $p-p$ at the grid of the video stage.

Mechanical excitation of the valve may be due to sound waves from the loudspeaker, which may be propagated through the air or transmitted by the chassis and cabinet. As this is the normal condition, tests for microphony should be carried out in the cabinet. Alternatively, microphonic excitation may take place by shocks accidentally applied to the cabinet.

Something may be done by choice of circuit to reduce sensitivity to microphony. D.C. restoration can materially reduce sensitivity to microphony since frequencies lower than 500 c/s. need not be reproduced by the coupling components, the D.C. and L.F. components being reinserted later. This is of particular value in the synchronizing separator, as D.C. insertion is usually employed at the limiter grid. A low value of the grid-leak and capacitor can produce a marked improvement to the insensitiveness of the synchronizing separator to microphony, the limit being set by the amount of synchronizing pulse chopping that can be tolerated. Unfortunately, the synchronizing pulse is already small in the case of the anode-bend demodulator video stage, which partially accounts for the comparative unpopularity of this type of circuit.

Other Considerations in the Design of Video Output Stages

There are a number of secondary considerations which have to be investigated in the design of video stages.

Valve Bottoming

When the anode current is a maximum the anode

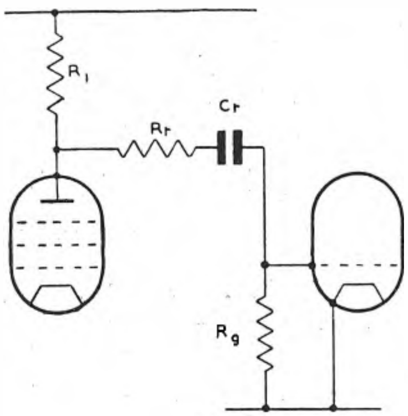


Fig. 6 (a)
The feed to the synchronizing pulse separator

voltage will be a minimum and care must be taken that the anode does not bottom. Bottoming results in distortion of the peak whites and further may result in excessive screen current with a danger that the screen dissipation rating may be exceeded. In certain cases, the stage has been designed so that the valve just does not bottom on peak white but does so on interference pulses, the stage then providing its own vision interference limiting. This practice cannot be recommended unless an exhaustive investigation is made into tolerances on the knee of the valve used and on the resistors employed.

Negatively-going Transitions

When employing cathode compensation, care must be taken that a negatively-going transition at the grid is followed reasonably well at the anode. The higher the feedback factor, the longer will be the anode time constant. If a sharp negatively-going input wave is applied, the fastest rate of rise of anode potential is V/CR where V is the potential difference between the H.T. rail and the anode at the instant the transient is applied, and CR is the anode circuit time constant. With cathode compensation, however, V is usually large (since the anode load is high) and so in practice, circuits may be designed which deal adequately with the negatively-going transients encountered in television.

Feed to the Synchronizing Separator

If the video signal is fed directly from the video valve anode to the grid of the synchronizing valve, the grid-cathode space of the latter functioning as the first limiter, care must be taken with the circuit design to avoid a bad transient response at the video anode, or "pulling after white." The circuit usually employed is shown in Fig. 6a.

To provide a sufficient source impedance for the limiter action the resistor R_r is incorporated. If a high feedback factor is employed in the video stage, with a resulting high load resistance R_l this resistor R_r may not be necessary. If R_r is made too large the phenomenon of "pulling after white" occurs.

From the point of view of the response at the video valve anode, the time constant formed by R_r and the input capacitance of the synchronizing separator may have a profound effect. The effect is usually manifested by a drag on the indicial response as shown in Fig. 6b. With careful design this effect may be eliminated.

Tolerances

As with all circuit design the tolerances on circuit and valve parameters must be investigated. In particular, it should be appreciated that the tolerance allowed on the cathode capacitor used in a cathode compensated stage becomes smaller as the feedback factor is increased.

Cost

Finally, in the assessment of a circuit, its cost should be considered. Unless a negative bias line is available the anode compensated stage is usually dearer than the cathode compensated stage due to the need for an electrolytic capacitor by-passing the cathode resistor. Further, in the case of video stages to yield large outputs the cathode compensated stage may often employ a smaller-current valve than an anode compensated stage. This may be an important factor in the design of an A.C.-D.C. receiver.

Dissipation Ratings

It is, of course, understood that the valve dissipation ratings should not be exceeded. In any given design this should always be checked. It may be

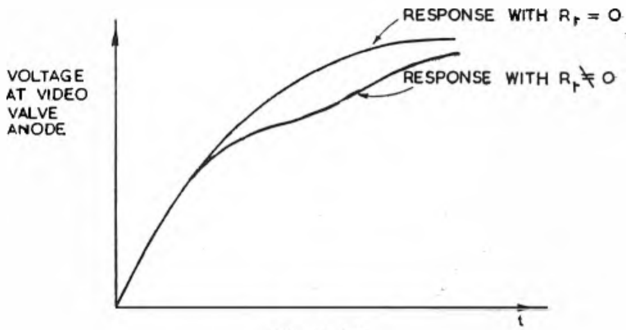


Fig. 6 (b)
The effect of the synchronizing separator input circuit

noted here that the anode rating will never be exceeded if the value of the load resistor is greater than $(V_{H.T.})^2/4\ W_a$ where W_a is the maximum anode rating.

In many cases the screen rating will be exceeded before the anode rating. A great deal of trouble may be avoided by ensuring at the commencement of a design that the screen rating is not exceeded. This is illustrated below.

The Design Procedure for Anode Compensated Video Output Stages

As an example of the procedure to be followed we shall attempt the design of a first order anode compensated video output stage which

(a) delivers 30 volts p-p output signal,

(b) has a build-up time of 0.130 μS,

(c) has the fastest rate of rise without overshoot,

(d) has a total anode loading capacitance of 30 pF.,

(e) has a positively-going signal (black to white) at the input.

As already seen, the first order anode compensated stage has a build-up time, with condition (c) satisfied, of $1.57RC$ so that inserting the conditions (b) and (d) above we obtain

$$R = 2{,}760 \text{ ohms.}$$

The anode load resistor will therefore provisionally be taken as 2.7 kΩ. The *p-p* current swing required for 30 volts *p-p* output is 11.1 mA.

The valve used will be the EF80 which has W_a = 2.5 watts and W_{g2} = 0.6 watt.

Using a 170 volt H.T. line $(V_{H.T.})^2/4R$ equals 2.68 watts so that we shall have to check at the end of the design whether the anode rating is exceeded.

The maximum steady screen current that may be passed is given by $W_{g2}/V_{H.T.} = 0.6/170 = 3.53$ mA. We shall take the peak permissible screen current to be 4.0 mA since the picture signal is never completely at peak white; the presence of synchronizing pulses decreases the mean screen current to about 85 per cent of the mean screen current which would result from continuous operation at peak white.

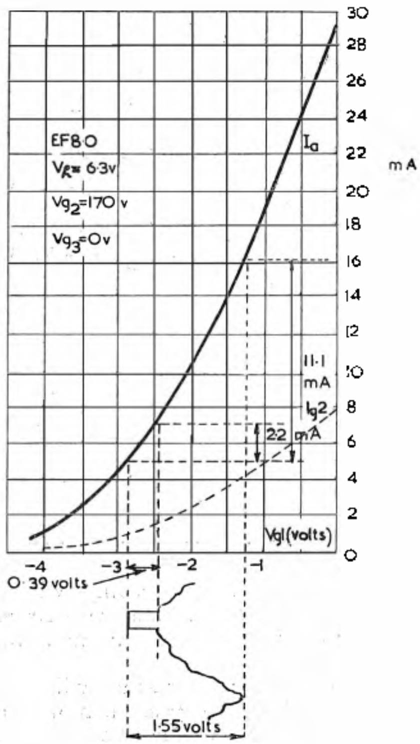


Fig. 7 (a)

The EF80 in an anode compensated video stage

Considering the I_a/V_g characteristics of the EF80 operated from a 170 volt H.T. line (Fig. 7a) it will be seen that under peak white conditions we may choose:

$$V_{g1} = -1.3 \text{ volts}$$
$$I_a = 16.1 \text{ mA}$$
$$I_{g2} = 4 \text{ mA}$$

Since we required a current swing of 11.1 mA, the anode current under no signal conditions will be 5.0 mA giving the no signal conditions

$$V_{g1} = -2.8 \text{ volts}$$
$$I_a = 5.0 \text{ mA}$$
$$I_{g2} = 1.0 \text{ mA}$$

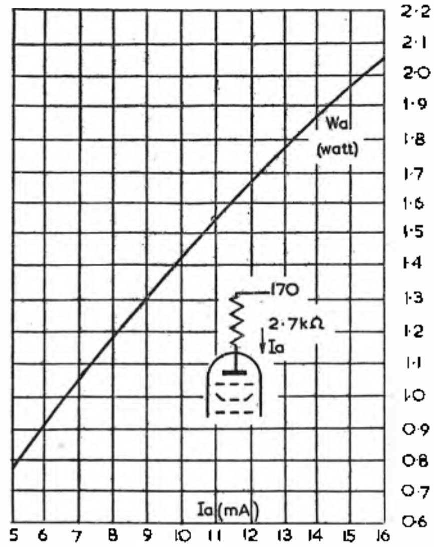


Fig. 7 (b)

The anode dissipation of the EF80 video stage

The input swing required is 1.5 volts *p-p* resulting in a stage gain of 19.3 (25.7 db). Assuming a synchronizing percentage input of 25 per cent (it will not be the full 30 per cent due to cramping at the demodulator) the synchronizing pulse at the output is 19.8 per cent of the total output signal. The synchronizing pulse is thus 5.95 volts *p-p* with full picture tube modulation which is perfectly satisfactory for most synchronizing separators. At reduced contrast settings it is to be expected that the synchronizing pulse output will be sufficient for most synchronizing separators.

The anode wattage must now be investigated. This is given at any anode current by

$$W_a = I_a\,(V_{H.T.} - RI_a)$$

Using this expression the anode dissipation at any given current may be plotted as in Fig. 7b. It will be noted that the rated maximum of 2.5 watts anode dissipation is not exceeded.

At peak white signal the anode potential will be 126.5 volts which is well above the knee of the I_a-V_a characteristic of the EF80, so that we may safely say that no trouble with bottoming will be experienced.

The value of the compensating inductance required is given (from Table II) by

$$L = 1/4\,CR^2 = 54.6\ \mu\text{H}.$$

This may conveniently be achieved by winding 100 turns of 42 S.W.G. D.S.C. wire on a ½ watt Eric resistor, winding width = 3/16 in.; Douglas wave-winder gears A50, B37, C38, D50, E40, F80.

The complete design is shown in Fig. 7(c).

The Design Procedure for Cathode Compensated Video Output Stages

As an example of the procedure to be followed we shall attempt the design of a cathode compensated stage which:—

(a) delivers 30 volts *p-p* output signal,

(b) has a build-up time of 0.130 μS,

(c) has the fastest rate of rise without overshoot,

(d) has a total anode loading capacitance of 30 pF,

(e) has a positively going signal (black to white) at the input.

As seen in Part I the cathode compensated stage has a build-up time, with condition (c) satisfied, of $2.19RC/F$ so that inserting the conditions (b) and (d) above, we obtain:—

$$R/F = 1.97 \text{ k}\Omega \quad (16)$$

The value of feedback factor, F, obtaining with various values of cathode resistor R_k must now be found by examination of the EF80 characteristics. This is shown in the following table:—

$V_{g2} = 170$ volts.

R_k(ohms)	V_{g1}(volts)	I_a(mA)	I_{g2}(mA)	gm(mA/V)	gk(mA/V)	F
150	−1.95	10.2	2.6	7.3	9.15	2.37
180	−2.10	9.3	2.3	6.9	8.60	2.55
220	−2.25	8.3	2.0	6.5	8.08	2.78
270	−2.45	7.0	1.7	5.9	7.34	2.98
330	−2.57	6.5	1.5	5.5	6.78	3.24
390	−2.72	5.5	1.2	5.1	6.21	3.42
470	−2.85	4.8	1.0	4.7	5.67	3.66
560	−2.98	4.2	0.95	4.2	5.15	3.89

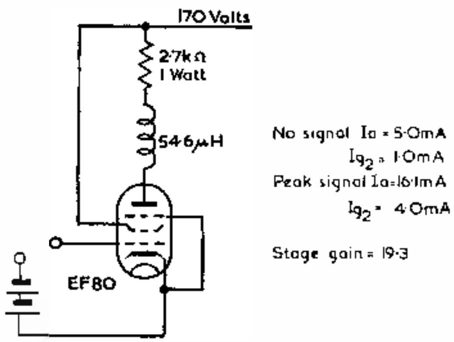


Fig. 7 (c)
Complete design of an anode compensated video stage using the EF80

Using these values of F the anode load required may be calculated from (16). The current swing required, and the gain of the stage (g_mR/F) may further be calculated as in the following table:—

R_k(ohms)	R(kΩ)	Current Swing (p-p. mA)	Peak Anode Current (mA)	Stage Gain
150	4.66	6.43	16.63	14.4
180	5.02	5.98	15.28	13.6
220	5.48	5.47	13.77	12.8
270	5.87	5.12	12.12	11.6
330	6.38	4.70	11.20	10.8
390	6.73	4.46	9.96	10.1
470	7.21	4.16	8.96	9.3
560	7.65	3.93	8.13	8.3

The maximum anode dissipation for the EF80 is 2.5 watts so that if the anode load is greater than $(V_{H.T.})^2/4W_a = (170)^2/4 \times 2.5 = 2.9$ kΩ, this rating can never be exceeded. Thus all the cases contained in the table above are satisfactory from the point of view of anode dissipation.

The maximum screen dissipation for the EF80 is 0.6 watt which is realized when the H.T. line is 170 volts at a screen current of 3.53 mA. The anode current corresponding to this screen current is 14.0 mA. It will thus be seen that for the case above where R_k = 150 or 180Ω there is a possibility of exceeding the screen wattage rating

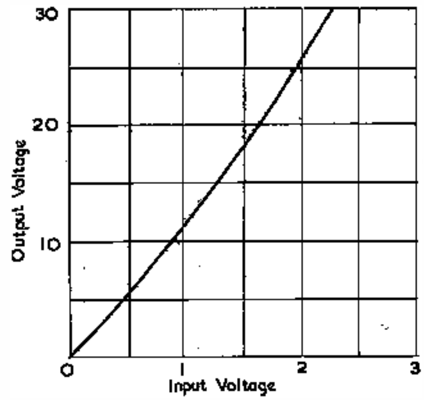
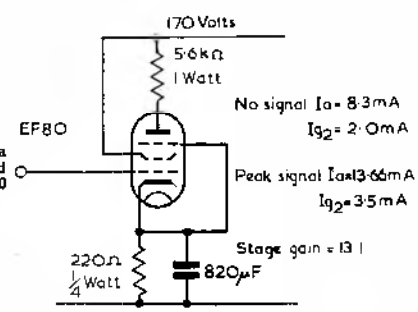


Fig. 8 (a)
Linearity curve of cathode compensated stage

(in actual fact the case for R_k = 180Ω is probably satisfactorily due to the fact that no television waveform is 100 per cent peak white because of the presence of synchronizing pulses). We may reasonably take the attitude that we shall choose the case where most gain is obtained without the possibility of exceeding the valve electrode wattage ratings. This is given by R_k = 220Ω. We shall take as the anode resistor 5.6 kΩ which indicates that for 30 volts p-p voltage output swing, the current swing required is 5.36 mA. This indicates a peak anode current of 13.7 mA and a peak screen current of 3.5 mA (the peak screen dissipation being 0.6 watt). The gain of the stage is 13.1 (22.35 db), and the anode potential swings between 123.4 volts and 93.4 volts so that no trouble with bottoming will be experienced.

The input voltage required for 30 volts p-p output is $30/13.1 = 2.29$ volts p-p. Assuming that the synchronizing pulse percentage at the input is 25 per cent (the demodulator produces some synchronizing pulse cramping) the input corresponding to the

Fig. 8 (b)
Complete design of a cathode compensated stage using the EF80



synchronizing pulse is 0.57 volt peak. Using the I_a-V_g characteristic of the EF80, the input-output linearity curve may be drawn.

This is most easily done as follows:—

(a) choose an anode current,

(b) from the I_a/V_g curve determine V_{gk},

(c) calculate V_k by multiplying R_k by $(I_a + I_{g2})$,

(d) subtract (b) from (c) to determine V_{in}.

The result is shown in Fig. 8a. It will be seen that the linearity is very good, and that the synchronizing percentage will not be appreciably reduced. The synchronizing pulse output will be of the order of 7.5 volts peak for a fully modulated picture.

The value of cathode capacitor required is calculated from

$$C_k = RC/R_k = 762 \text{ pF}.$$

A small amount of overshoot is usually tolerable on the indicial response, and so a cathode capacitor of 820 pF will be included.

The complete circuit is shown in Fig. 8b.

L.F. Considerations in the Design of Video Stages

As explained in the introduction, there are two ways in which the amplification can be maintained uniformly down to zero frequency. The first is the straightforward method of using a directly-coupled stage. This method is often employed when the C.R.T. cathode is fed from the video valve anode, since there is little chance in this arrangement of the bias being removed from the C.R.T. under fault conditions. However, this method has its attendant problems. The first arises if the H.T. supply has an appreciable impedance. This may result in changes of picture level as the average brightness of a scene varies. The inclusion of appropriate screen decoupling elements may be utilized to neutralize this effect. Another point which must be considered in any design is the heater-cathode voltage rating of the C.R.T. Direct feed of the signal to the cathode of the C.R.T. means that the cathode may be at a considerable potential above earth, in which case the heater-cathode rating may be exceeded. Usually this problem may be overcome by a separate heating winding for the C.R.T. (this winding being connected to the cathode by about 20 kΩ) or in the case of series-connected heaters by appropriately choosing the position of the C.R.T. in the valve heater chain. In addition, when using a tetrode gun C.R.T., the first anode potential has to be supplied. If the cathode is already, say, 150 volts above earth, and it is required to operate the first anode at a potential of 200 volts relative to the cathode, this implies that a potential of 350 volts relative to earth must be applied to the first anode. Occasionally such a potential is available from an H.T. line in the receiver, but more usually it must be obtained either by bleeding the E.H.T. or by rectification of an appropriate pulse voltage in the receiver. For example, the supply to the first anode has been obtained from the pulse appearing across the line deflector coils, the rectification being effected by a non-linear device such as Metrosil.

In the case of anode compensated stages directly feeding the C.R.T., special problems arise in obtaining the bias voltage for the video valve without introducing L.F. degeneration. If the design is such that the grid of the C.R.T. is fed, then under no-signal conditions the video valve passes a high current; application of the signal swings the grid more negative. Thus the value of cathode resistor required for auto-bias is quite small (usually of the order of 50 ohms) and it is convenient to leave this unbypassed, tolerating the small loss in gain incurred, but eliminating any L.F. degeneration. If the cathode of the C.R.T. is fed, the video valve passes a small current under no-signal conditions and the cathode resistor required for auto-bias operation is sufficiently large to decrease severely the gain if left unbypassed. There are a number of ways in which circuit compensation may be introduced in the anode circuit to compensate for the L.F. degeneration caused by bypassing the cathode resistor, but all of these methods involve the use of at least two additional components. An alternative scheme is to bias the grid of the video valve from a negative bias line, the cathode of the video valve being tied to earth. It is important that such a negative bias line should not drift. These problems do not arise with cathode compensation.

The second method of maintaining the D.C. component is to use D.C. restoration. When this method is employed no great care need be taken with decoupling time constants in the cathode or screen of the video valve, provided that these time constants are sufficiently long to make the sag in the output waveform small over a line scanning period. D.C. restoration requires the use of a further valve structure (or equivalent device), and has the additional disadvantage of slightly reducing the amplitude of the synchronizing pulses.

Acknowledgments

Acknowledgments are due to the Mullard Radio Valve Company, Ltd., and to Dr. C. F. Bareford, manager of the Mullard Electronic Research Laboratories, for encouragement in this work and permission to publish the article.

References

[1] Kallmann, Spencer & Singer, *Proc. I.R.E.* 33, 169, 1945.
[2] Thomson, W. E., *Wireless Engineer*, 24, 20, 1947.

Manchester to Edinburgh Micro-wave Television Link

THE contract has been placed and work has commenced on the micro-wave television link which will further extend television to Scotland. The new radio link, which will be the longest of its kind in Europe, is being designed and installed by Standard Telephones and Cables Limited who, since the first demonstration of a micro-wave link which they gave between Dover and Calais in 1931, have been responsible for many micro-wave developments, including some of the portable links at present being used by the B.B.C. Television Service for "outside broadcasts."

A Standard portable microwave link was used to span the Channel during the recent television transmissions from Calais.

The radio-relay will follow a route from Manchester northward on the east side of the Pennine Chain to Kirk o' Shotts; an overall distance of 245 miles. The contract includes the provision of seven unattended repeater stations spaced at about 30-mile intervals, to receive, amplify and re-transmit the requisite signals over successive line-of-sight transmission paths. Repeaters will consist of main and reserve receivers and transmitters with four paraboloid aerial assemblies mounted on steel towers ranging in height from 28 to 200 ft.

An "Ideal" Post Deflexion Accelerator C.R.T.

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(G.E.C. Research Laboratories, Wembley)

THE two main advantages to be gained in the use of commercially available post deflexion accelerator (P.D.A.) cathode-ray tubes as compared with normal electrostatic tubes are:

(i) The attainment of higher screen brightnesses with only a relatively small increase in the deflecting forces required to deflect the beam, and

(ii) The problem of insulation in the glass pinch and the base are eased in view of the fact that the final H.T. is applied to a contact on the side of the tube.

These points, which have been discussed in greater detail in previous articles,[1,2] appear to be very attractive to the tube user, but, of course, the adoption of this procedure introduces certain disadvantages, the chief of which is scan distortion. This distortion can become serious when one attempts to utilize the full screen diameter. It is known that this defect is due to the non-uniform fields introduced in the vicinity of the accelerator bands, and can be improved by careful tube design. With tubes at present available, however, these distortions are not completely eliminated. Fig. 1 represents an outline drawing of a P.D.A. tube, with the accelerators formed by bands of graphite on the internal wall of the tube.

W. G. White[2] has referred to a form of P.D.A. tube in which the electrons, after travelling through the deflecting system and the customary field free space are given a final increase in their axial velocity by an infinitely thin electrode at the back of the fluorescent screen. This electrode would be maintained at a voltage higher than that of the final anode of the electron gun. A tube of this nature, however, would suffer from defects similar to those outlined for the normal type of P.D.A. tube, since the accelerating field would still penetrate down the tube, thus introducing scan distortions. Tubes of this type have been shown to offer little in the way of advantages over the normal P.D.A. tube. Figs. 1 and 2 indicate the general form that the equipotential surfaces assume in the two special instances outlined above.

A special design of tube employing the P.D.A. technique has been investigated however, which is completely free from scan distortions. Furthermore, this tube can give increased brightness at higher screen potentials with no increase in the deflecting voltages required. Naturally, this achievement is obtained at the expense of another important characteristic of the tube, namely, resolution, and it is certain that this latter quality would, at the present moment, have to be carefully considered in any proposed use of this type of tube.

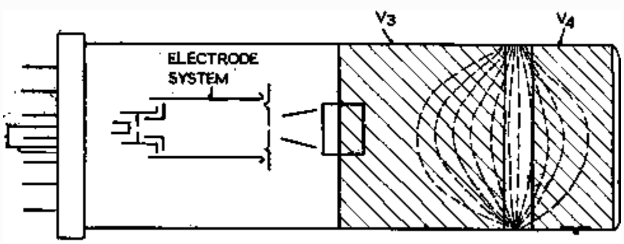


Fig. 1.
Form of equipotential surfaces in normal P.D.A. tube

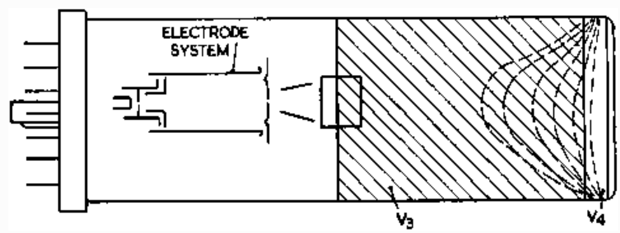


Fig. 2.
Form of equipotential surfaces in W. G. White's proposed P.D.A. tube

The "Ideal" P.D.A. Tube

In achieving the "Ideal" form of P.D.A. tube, the final accelerating field was introduced between two parallel planar electrodes placed very near and parallel to the fluorescent screen. In this arrangement, the equipotential surfaces effectively form parallel planes between the accelerating electrodes.

(a) Demountable Apparatus

In preference to making completed cathode ray tubes, the demountable form of tube was used. This allowed greater flexibility in the design and construction, and also enabled the results to be obtained readily. A normal three-stage mercury vapour diffusion pump was used, which allowed the experiments to be performed in a gas pressure of 10^{-5} mm. of mercury, or less. The pump head consisted of a 16 pin socket surrounded by a 90 mm. diameter ground glass flange, to which could be affixed any size of cathode ray tube bulb having a similar flange. (The sizes of bulb used for the experiments ranged from 1½ in. to 6 in. in screen diameter.) A further feature of this bulb was that it possessed another ground glass flange at the screen end, thus allowing the accelerating electrodes to be easily inserted. Through this second flange electrical contact was made to the accelerating electrode in contact with the screen. The electrode system, mounted on a 16 pin plug, was held in the pump head by the 16 pin socket, the electrical contacts to which were brought out through a normal 16 wire pinch. Fig. 3 represents the complete demountable system.

The property of the parallel electrode nearest the gun is such that as well as being continuous it must allow the electrons to pass through. For this purpose, wire gauze was used, initially of 60 mesh per inch, having a wire diameter to space ratio of 1 : 3. For the other parallel electrode, which should be in contact with the fluorescent screen, a similar gauze mesh was used. Alternatively, for this latter

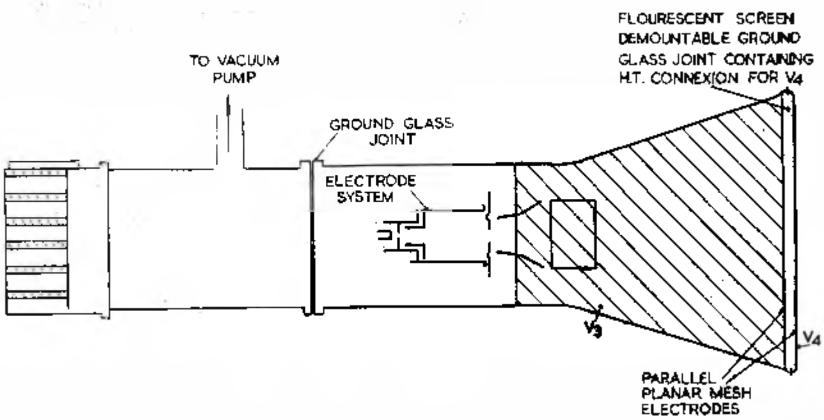


Fig. 3.
Demountable C.R. Tube

electrode, a metal backing of aluminium could have been employed as is now becoming a common feature with the magnetic type of television tube.

The spacing between the parallel electrodes was fixed at 4 mm., although this could have been readily adjusted to any other value, by the method of supporting the electrode nearest the gun in the body of the bulb. With these assemblies it was possible to assess spot quality, screen brightness and deflector plate sensitivity.

Test Procedure and Results

Direct comparison between the characteristics of a normal tube and the "ideal" P.D.A. tube was easily obtained by merely varying the potential of the electrode in contact with the screen. The deflector plate sensitivities are given in graphical form in Figs. 4, 5 and 6. In these, Figs. 4 and 5 refer to the 1½ in. diameter tube, with particular reference to the following experiments:—

Curve A. Normal tube with no P.D.A. ($V_3 = V_4$).

Curve B. Customary P.D.A. tube which has two bands of graphite.
(V_3 = 1.5 kV, V_4 = 1.5-7.0 kV).
(G.E.C. Ref. No. 401 C.A.H.A.)

Curve C. P.D.A. tube as proposed by W. G. White.
(V_3 = 1.5 kV, V_4 = 1.5-7.0 kV).

Curve D. P.D.A. tube having parallel planar accelerators
(V_3 = 1.5 kV, V_4 = 1.5-7.0 kV).

Fig. 6 refers to the 6 in. diameter tube, with particular reference to the following experiments:—

Curve A. Normal tube with no P.D.A. ($V_3 = V_4$).

Curve B. Customary P.D.A. tube, having two bands of graphite.
(V_3 = 2.0 kV, V_4 = 2.0-7.0 kV).

Curve C. P.D.A. tube as proposed by W. G. White.
(V_3 = 2.0 kV, V_4 = 2.0-7.0 kV).

Curve D. P.D.A. tube having parallel planar accelerators.
(V_3 = 2.0 kV, V_4 = 2.0-7.0 kV).

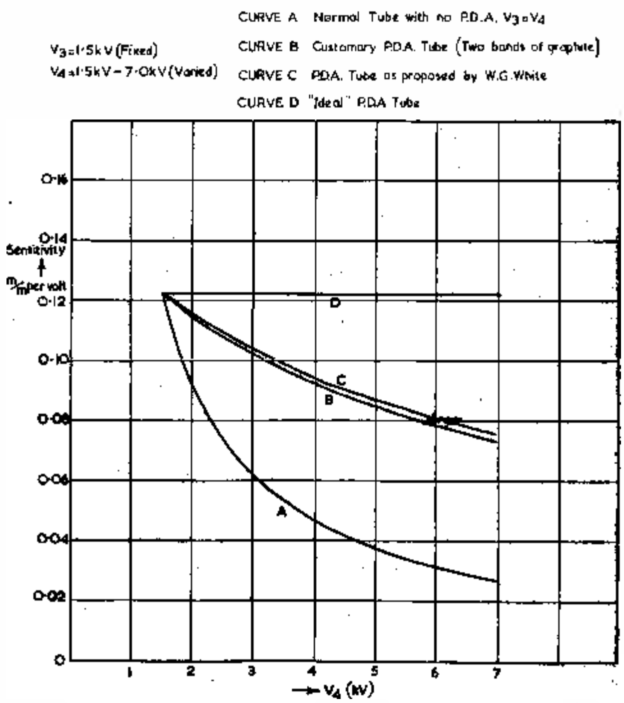
From these results it can be seen that the introduction of the parallel planar mesh P.D.A. electrodes enabled the screen potential to be increased without sacrificing deflector plate sensitivity.

The range of voltages applied to the final accelerator (V_4) in these experiments was much greater than that normally applied in the operation of P.D.A. tubes. It is customary to restrict the ratio V_4/V_3 to a value of approximately 2, otherwise very severe scan distortions could be introduced. However, this ratio was extended to ascertain the gain in sensitivity obained when using parallel planar mesh electrodes. It should be noted that the form of curve B in Figs. 4, 5 and 6 depends on the configuration of the graphite bands acting as the accelerators, and for these experiments, a well established design of accelerator bands was used. In addition to this gain in sensitivity, it was observed that the scan distortions were also eliminated due to the final accelerating field not being allowed to penetrate down the tube. The increased screen voltage naturally resulted in a higher brightness, which followed approximately the customary square law relationship.

However, some disadvantages were also apparent, for, as the beam passed through the first mesh electrode, secondaries were liberated which were then accelerated towards the screen. One result of these secondaries was that the contrast range on an intensity modulated raster was reduced.

Secondly, the resolving properties of the tube were

Fig. 4.
Back plate sensitivity measurements on 1½in. diameter P.D.A. tube



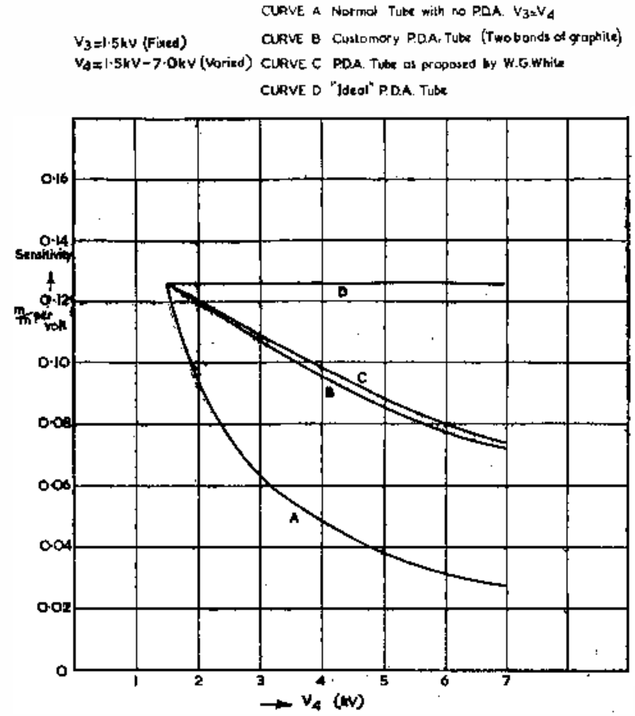


Fig. 5.
Front plate sensitivity measurements on 1½ in. diameter P.D.A. tube

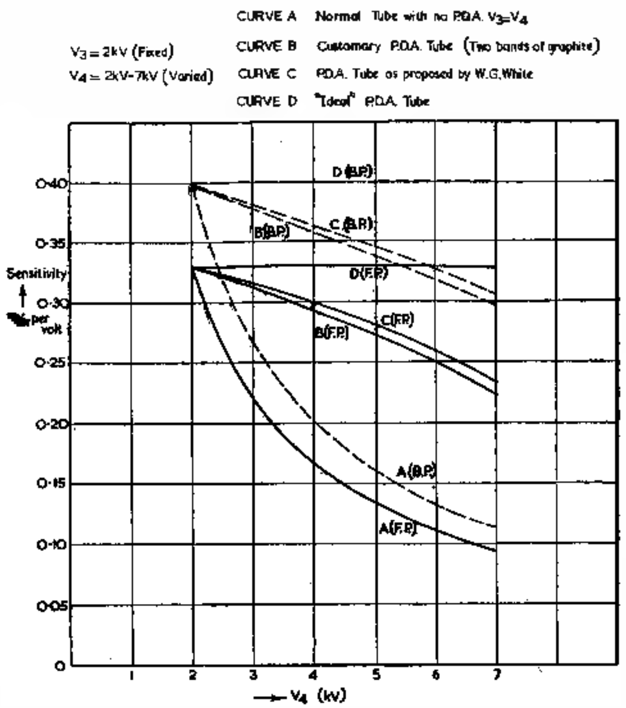


Fig. 6.
Back and front plate sensitivity measurements of 6 in. diameter P.D.A. tube

also affected, for these secondaries gave an apparent increase in spot size in the central region of the tube face, while at the edges they formed a second trailing spot. The displacement of this second spot from the main spot was dependent upon the spacing between the parallel planar mesh electrodes. Attempts were made to reduce the secondary emission by suitably coating the mesh electrodes, for example with carbon, but this did not eliminate the trouble. The use of a wide spaced mesh (approximately 10 mm. between wires), to limit the number of secondaries, was also tried. With this, however, a lens action was introduced between the mesh and the screen, resulting in a magnified image of the mesh structure appearing on the screen. This effectively reduced the useful screen area available.

Some detection of the wire mesh structure was observed with the first type of mesh that was used, but it is felt that this could have been eliminated by using a much finer mesh (i.e., where the mesh size is small compared with the spot size).

Conclusion

Ideally for this system to work satisfactorily, a very fine mesh, or, in the limit, a very thin continuous metallic film is required for the first parallel electrode. This must allow the beam to pass through and yet liberate no secondary electrons. Since this condition has not yet been achieved, the deterrent features of loss of contrast and reduced resolution have been introduced. However, the experiments have indicated that the " ideal " form of P.D.A. tube having parallel planar mesh electrodes is capable of achieving the desired characteristic, i.e., increased screen brightness without the expense of requiring higher deflecting forces, and also without the introduction of scan distortion.

It is considered that, in view of the performance at present obtained with commercially available P.D.A. tubes, the disadvantages introduced with the " ideal " P.D.A. tube outweigh its advantages, thus rendering its practical usefulness of small value.

References

1 " After Acceleration and Deflection," J. R. Pierce, *Proc. I.R.E.* Jan. 1941, p. 28.

2 " Cathode Ray Tubes with Post Deflection Acceleration," W. G. White, ELECTRONIC ENGINEERING, March 1949, p. 75.

Fire Risk Due to TV Lenses

IT is a well-known fact that the sun's rays, if concentrated through a magnifying lens and directed upon readily combustible material, can cause fire. An example of this has been quoted where the sun's rays, passing through a transparent plastic door handle, ignited a bath robe. Heath and forest fires have been attributed to the concentration of the sun's rays through broken glass and bottles and, today, a new medium has been introduced into daily life by means of which " insolation " (as it is called) can occur.

Radio dealers are displaying in their show windows new types of magnifying lenses for attachment to television sets. Already some incidents have occurred where insolation has set fire to the contents of radio dealers' windows where lenses are displayed.

—From the *Fire Protection Association Journal*, July, 1950.

The Measurement of Noise in Resistors

By FRANCIS OAKES

NOISE is produced in every resistor, fixed or variable, and whether carrying a current or not. This noise is due to themal agitation, and, usually, a number of other causes contributing to the total of noise fluctuations in varying degree. As the noise characteristics of resistors, and especially such resistors as are used in the field of electronics, are an important though undesirable property, the measurement of this complex quantity, its theory and technique, has also become an important problem. The most important aspects of the subject of noise measurement are: the inspection of production processes of radio resistors; the testing of commercial radio resistors for the amount of noise they produce on load, and in the case of variable potentiometers on load and upon rotation, the noise measurements that have to be carried out as part of the development work leading to the design of efficient radio components; and the various types of noise measurement made necessary by research work in studying the nature of noise and investigating its causes.

The Quantities to be Measured

Thermal agitation noise and its measurement are, as such, not of importance from this aspect, because magnitude, frequency distribution and the physical nature of this phenomenon are well known, and their numerical values can be calculated from other data, without measurement. The position with all other types of noise is, however, quite different. Apart from the fact that noise from other sources cannot be calculated numerically from such simple data as resistance value and ambient temperature, the problem is complicated by the nature of the quantity to be measured.

The fluctuations of voltage, or current, normally referred to as resistor noise, consist of a series of transient phenomena of relatively small amplitude and very uneven distribution. As can be seen from the oscillogram shown in Fig. 1, being a noise trace of a

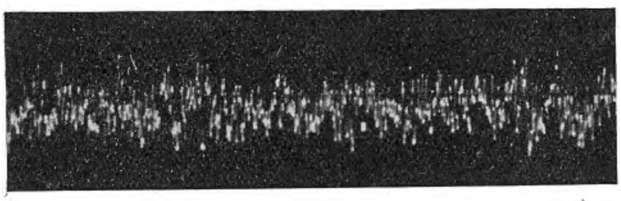


Fig. 1.

Noise trace of 1MΩ resistor, comparatively noisy sample but stable. Voltage across sample : 100V (10mV = ½ inch)

fixed resistor carrying a small current, the fluctuations do not occur cyclically, and their amplitude and frequency distribution change in an irregular fashion. This condition is vastly exaggerated in the case of a faulty resistor, an example of which is shown in Fig. 2. This resistor has a very unstable noise characteristic, showing only a few per cent of the amplitudes observed during "noise-bursts" for considerable

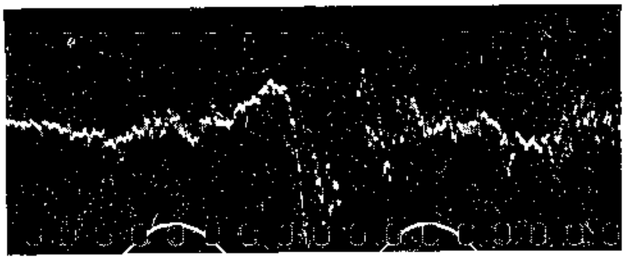
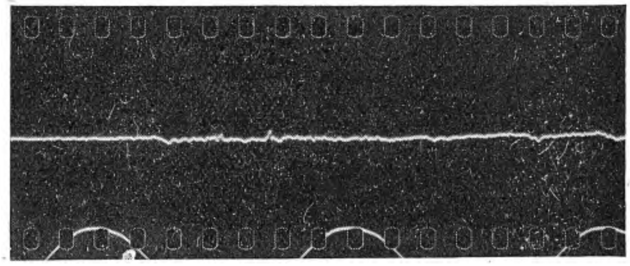


Fig. 2.

Noise trace of unstable ¼MΩ resistor

(*a*) *Above* : during noise bursts when noise peaks exceed 10 mV

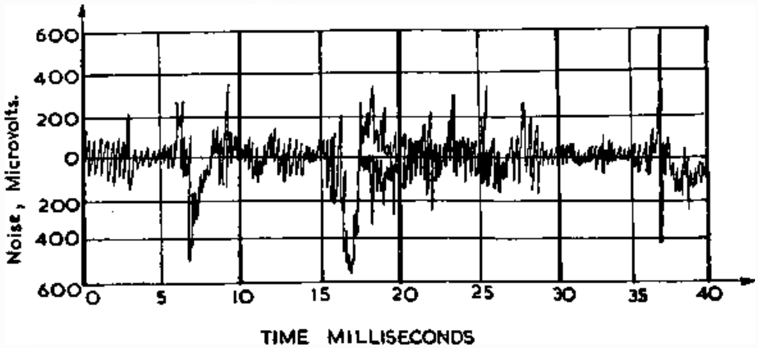
(*b*) *Below* : between noise bursts, recording to the same scale as (*a*) i.e. 10 mV = ¼ inch



intervals of time between them. It is therefore impossible to speak of a particular figure as "peak" voltage or of a particular figure as "average" voltage in describing the "magnitude" of these fluctuations. This is also the case for the noise characteristic of the resistor of which the noise trace is recorded in Fig. 1. Here, the visible "peaks" over certain intervals of time vary along the trace. In other words the peak value between the first two timing marks of 1/50 second would be different from the peak value between the next, and so on. Similarly, the R.M.S. value taken over the first interval would be different from the one taken over the next. If a long noise trace were to be investigated in this way, it would be seen that the noise peak values and the noise R.M.S. values over each fixed interval are, in turn, fluctuating quantities.

It is quite clear from these considerations that the terms peak value, R.M.S. value and also, analogously, average or linear mean value of noise voltage or current must be further specified, and that measurement of these quantities must be carried out according to more specific definitions.

With peak values, the question is normally that of determining the maximum voltage peak occurring during the period of observation, bearing in mind that during a second observation of equal duration this value would be different from that obtained first time. It can be observed, however, that with normal radio resistors, the difference in the values so obtained decreases as the interval of observation is lengthened. This is illustrated in Fig. 3.



Interval		Values	Maximum difference
5msec	Peak value	200 500 150 550 300 350 100 600	Maximum difference 500
	Difference	300 350 400 250 50 250 500	
10msec	Peak value	500 550 350 600	Maximum difference 250
	Difference	50 200 250	
20msec	Peak value	550 600	Maximum difference 50
	Difference	50	

Fig. 3.
Illustration of decrease in difference values of resistance noise measurements as the interval of observation increases

In the case of R.M.S. and average values, again, the value measured depends on the period of observation. And here, too, values measured during two periods of equal length will not be identical, because the occurrence of noise fluctuations is not a cyclic or recurring one. With normal resistances, the R.M.S. values measured over reasonably long periods will differ only by a small amount.

" Nuisance Value "

For practical purposes, it is of importance how disturbing the noise generation will prove to be when the resistor is in operation in a radio receiver or other circuit of application. High noise peaks of very short duration and which only occur rarely, will have little audible effect, and therefore be of lesser " nuisance-value " than lower noise peaks of longer duration occurring more frequently. Noise with a uniform fairly low R.M.S. value may show a lesser " nuisance-value " and yet a greater long-interval R.M.S. value than non-uniform noise, such as shown in Fig. 2, where bursts of momentary high R.M.S. value occur while the long-interval R.M.S. value is very low due to the low noise level between the bursts. It is therefore necessary to design noise measuring equipment in such a way that the noise readings obtained give the desired information. This is done by a careful study of the circuit time constants of the metering circuits and by a number of subjective tests with resistors in operation in amplifier circuits, carried out to determine and check the performance of noise measuring equipment. This is particularly important for industrial noise measuring equipment. For research purposes, on the other hand, peak and long-period measurements are of greater importance, and in distinction to the approximation to physiological factors, the accuracy of measurements and the repeatability of value obtained is in the focus of attention.

Measuring Equipment

Noise measuring equipment normally consists of the following fundamental parts:

1. Input circuit.
2. Amplifier.
3. Attenuator.
4. Filter.
5. Output stage.
6. Metering circuit.
7. Monitor C.R.O. or recording equipment.

Input Circuit and Amplifier

The input circuit contains the resistor on test, a standard resistor, and a current-supply to load the resistor being measured. Also, part of the input circuit is the grid circuit of the valve providing the first stage of gain. This is shown diagrammatically in Fig. 4. E is the voltage source, which is either a dry battery, accumulator battery, or a very carefully stabilized D.C. voltage from a specially designed rectifier circuit, which must also eliminate residual hum to a high degree. The standard resistor is a resistor of known resistance value and lowest possible noise. Usually, a non-inductively wire-wound resistor is used for this purpose, of a value equal to that of the resistor on test. In this case, the voltage loading the standard resistor is equal to that of the resistor on test, and equal to half the voltage of the battery supply. As the superimposed voltage can only be varied in steps in the case of batteries, and as it is not practically possible to equip measuring gear with a very large number of standard resistors, this condition is approximated as far as possible, and the loading voltage across the resistor on test is computed as follows:

$$E_L = R \frac{E}{R + R_S}$$

The noise voltage across the grid impedance of the first amplifier valve is not the total noise voltage produced, but depends on the combination of impedances in the grid circuit. Normally, the grid capacitances and the wiring capacitances and inductances can be neglected, but the resistance of the grid-leak must be taken into consideration. The portion of the noise voltage reaching the grid is therefore:

$$e_G = \frac{R' \cdot e_N}{R + R'} \quad \text{where } R' = \frac{R_S R_G}{R_S + R_G}$$

therefore, the total noise voltage e_N can be calculated from the input voltage e_G.

$$e_N = e_G \frac{R + R'}{R'}$$

This means that the noise voltage can be calculated from the constants of the input circuit if the performance of the amplifier is known, or if the output circuit contains a meter calibrated in terms of amplifier input voltage. Knowing also the voltage applied to the resistor on test, the noise of the resistor can then be expressed in terms of microvolts noise (R.M.S.) per volt load.

The amplifier, of which the valve mentioned above is a part, is a carefully designed and built wide band voltage amplifier of 100 to 120 db gain, sometimes even more. Resistance coupling, negative feedback and the use of wirewound resistors, as well as careful screening and shock-absorbing mounting are some of the features necessitated by the large gain. Also,

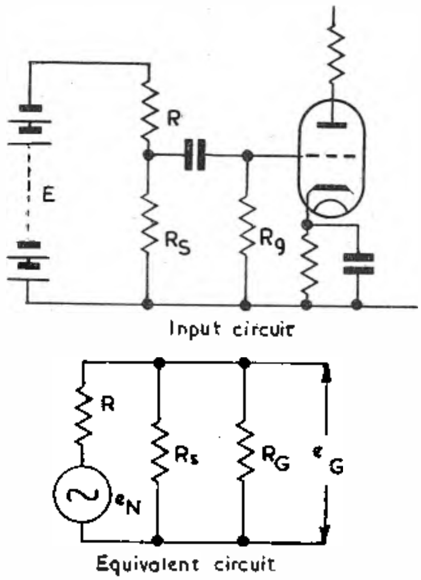


Fig. 4.
Input circuit for noise measuring equipment, with equivalent circuit

care must be taken to select valves, particularly for the first two stages of amplification, which have low-noise characteristics and a minimum of microphony. Perhaps one of the most difficult problems is that of avoiding instability, by means of careful layout—especially of all earth returns—and by screening and proper location of the various portions of the amplifier to prevent stage-to-stage coupling and common impedances.

Attenuator, Filter and Output Circuits

An attenuator is interposed between the second and third stage of amplification, in many cases this location provides sufficient pre-amplification to prevent the attenuator from increasing the level of background noise to any appreciable extent, while at the same time avoiding overloading of any stage preceding or following attenuation. A filter normally follows the attenuator so that the frequency response characteristic of the amplifier is given a convenient shape to allow easier interpretation of the measuring results. Ideally, the pass band of a noise amplifier should be a rectangle extending from a fairly low audio-frequency up to about 15 or 20 kc/s. for ordinary purposes. If the amplifier proper is substantially flat over this range, the filter will govern the frequency characteristics of the amplifier.

The stages following the filter are again of the voltage amplifier type, using negative feedback for the elimination of undue distortion which is much more important here than in the preceding stages, because of the larger magnitude of the signal voltages. The last valve is chosen so as to provide the correct characteristics required by the output circuit, and if several different output circuits are required, these are normally provided by a set of valves driven in parallel from the last stage of amplification. The following types of output circuit are used alternatively or in combination.

1. Thermo-couple and moving coil meter for measurement of R.M.S. value.
2. Moving coil meter driven by a diode detector circuit for measurement of linear mean value.
3. Diode detector circuit with fast-charging, slow-discharging characteristics provided by suitable *R-C* circuit and moving coil meter for measurement of peak value.
4. Diode detector circuit as above, but with somewhat faster discharge characteristics for measurement of " nuisance-value."
5. Diode detector circuit as described under 2, with the moving coil meter calibrated so as to give approximate R.M.S. values, based on an experimentally derived conversion factor to indicate an R.M.S. reading from a linear mean value measurement.
6. C.R.O. for monitoring or recording purposes. A normal electrostatically deflected tube is used for this purpose in connexion with an output tube of the voltage amplifier type suitable to handle the large voltages required, without overloading. Negative feedback is useful, and a double beam C.R.T. is also an advantage, because a timing signal can be applied to the second beam, as was done to obtain illustrations Fig. 1 and Fig. 2. The timing wave was a 50 c/s. alternating voltage applied to the second beam which was shifted up so that timing marks were just visible.

Stabilization Monitoring and Calibration

The power supplies for the heater and anode circuits of the amplifier are supplied by batteries or by a stabilized power supply unit. This is necessary because mains voltage fluctuations are likely to cause drift in the amplifier gain and will upset the calibration, producing an error in the noise value reading. It is sometimes sufficient to stabilize the H.T. supplies alone, which can be done by a conventional stabilized H.T. supply unit comprising a rectifying valve, a two-valve stabilizing circuit and a mains transformer to supply the alternating heater and high tension supply voltages. In many cases, however, such as for laboratory noise measuring equipment for development or research work rather than for comparative measurements for inspection of production purposes, the heater voltage supplies must also be stabilized. This is very important because the valve characteristics are dependent upon the heater voltage which controls the cathode temperature. Therefore, a change in heater voltage will produce an appreciable change in amplification when accuracy and constancy of noise measurement is important. In these cases, magnetic stabilizing equipment of the saturated transformer type is often used successfully, and the heater and anode supply sources are both stabilized in this way.

Sometimes a small loud speaker driven by a power valve is used instead of, or in addition to the C.R.O. monitor.

The C.R.T. as well as the loudspeaker is often useful in assessing the "nuisance value" of the noise measured. As already mentioned, it is also possible to choose the time constants of the detector and meter circuits so that the meter reading will correspond approximately to this value. Fig. 5 shows in block-diagram form the various parts of a noise measuring amplifier and their correlation.

The calibration of a noise amplifier measuring equipment which could also be called a noise valve voltmeter, is carried out by means of a calibrated

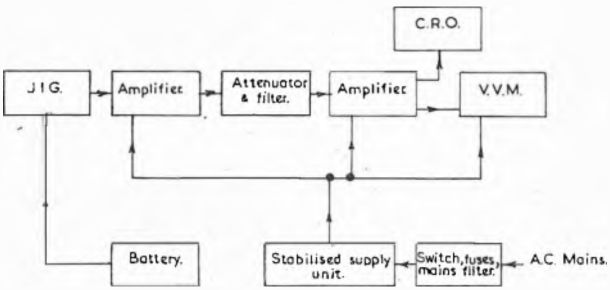


Fig. 5.
Block schematic of noise measuring amplifier

audio-oscillator and accurate attenuator. The audio-oscillator is set to frequencies within the flat range of the amplifier response curve, say to ½, 2 and 10 kc/s. respectively and output adjusted to, say 0.1 volt R.M.S. The attenuator is then set to obtain 50, 100, 200 microvolts, etc., fed into the amplifier input, and the amplifier attenuator is set accordingly to allow checking of the output readings. The first of these checks, say, with the input adjusted to 50 microvolt at 500 kc/s. is used to set the negative feedback control of the amplifier so as to give the correct output reading, all other steps should then be merely for checking purposes and require no further adjustment of the amplifier gain controls.

The frequency characteristic of the amplifier can be checked by the conventional methods, and is controlled by filters of conventional design. For both frequency and amplitude checks it is important to prevent direct radiation from producing errors in calibration. This can usually be prevented by careful screening and by keeping the input impedance of the amplifier and the output impedance of the oscillator and attenuator low during calibration.

Measurement of Amplitude and Frequency Distribution

The measurement of noise can then be carried out by direct reading of the output voltage, taking into account the setting of the amplifier attenuator and possible non-linearity of the metering circuit. With a more elaborate attenuator incorporating fine steps, measurement can be carried out in terms of insertion loss, with the meter adjusted to read as during calibration, i.e., full scale, or another fixed setting. This method is a little more cumbersome in use but offers far greater accuracy.

A noise valve voltmeter as described in the foregoing paragraphs provides a means of making measurements of absolute noise figures, peak, linear mean and R.M.S. based on calibration and knowledge of the frequency response curve. The latter can easily be established by point-by-point computation from measurements at variable-frequency, constant-amplitude inputs. If only comparative measurement is required, such as for factory test purposes, a much simpler type of equipment is often used. A representative example is shown in the block diagram, Fig. 6. Filter and monitor are omitted, although the latter, in form of a small C.R.T. is always very useful, and the low frequency response is kept low artificially by the choice of small coupling capacitors in the first two stages. In this way pick-up hum is reduced and the design of the jig for holding the resistor is simplified. This jig normally takes the form of a

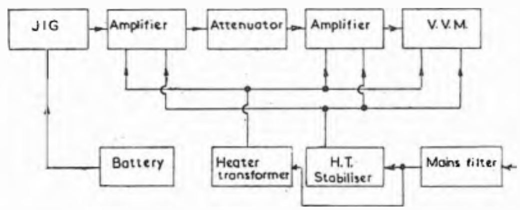


Fig. 6.
Block schematic of noise comparator

small screened box to accommodate the sample on test and often also the exchangeable standard resistors. A screened door is equipped with a gate-switch so that the input is short-circuited to safeguard the amplifier from overload during manipulation. When the resistors are inserted and the door is closed, the input circuit is again completed. It is important to design the mechanical components of the jig and input leads so as to be of non-microphonic characteristics. This can be done by aiming at high rigidity and at the same time including shock absorbing mountings and acoustical damping. The higher the gain the more necessary is the greatest care in the design of the input circuits, mechanically, because at amplifications of more than 100 or 110 db normal co-axial cable, ordinary wiring, terminals and similar elements act as microphones, especially when having a direct potential.

Neither the extremely low audio-frequencies nor the high frequency end of the noise spectrum is included in the range of normal noise measuring equipment. So far as the low end is concerned, the band is not very wide, and is not of importance as a source of interference, generally. The upper end, on the other hand, is still in the region where severe disturbances might be caused. Fortunately, the distribution of noise amplitude decreases with frequency, and the high frequency noise amplitudes are comparatively small, when resistors are carrying a current.

The measurement of the frequency distribution of resistor noise can be carried out by means of the equipment shown in Fig. 7. If necessary after pre-

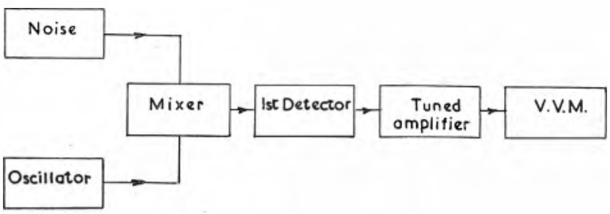


Fig. 7.
Circuit to measure the frequency distribution of resistance noise

amplification, the noise voltage is mixed with a voltage of constant amplitude and known frequency. A beat will be produced with slightly different frequency components of the noise voltage, and after rectification only those frequencies which are in the two corresponding narrow bands to either side of the added wave of known frequency will contribute to the output from an amplifier containing a tuned low frequency filter. A similar system of frequency analysis can be based on an intermediate frequency method, using a crystal filter and operating in a manner somewhat similar to the superheterodyne radio receiver. By variation of the frequency of the added voltage over the frequency spectrum, and observation of the change of output, a frequency characteristic of any desired width can be obtained within the flat part of the amplifiers and attenuators used. With the low frequency method and a low frequency filter resonating at 30 c/s. with only a few cycles band width the side bands are only 60 cycles apart and need not necessarily be separated. With the intermediate frequency method, on the other hand, one of the sidebands has to be suppressed.

Measuring Noise in Variable Resistors and Potentiometers

The measurement of noise produced by variable potentiometers when these are moved from one setting to another while carrying a D.C. load, (in order to assess their behaviour when carrying a signal) is much simpler than the measurement of noise produced by fixed resistors or stationary potentiometers, because the magnitudes of noise produced are sufficiently large to avoid extremely high gain. The chief problems are the determination of the most suitable output circuit and meter characteristics and the design of a jig to hold the potentiometer on test and to facilitate rotation of the spindle at a constant predetermined speed. The determination of metering circuit constants is governed by the approximation of "nuisance value" indication. This is done by a diode detector circuit designed to indicate average noise voltages, but with a lengthened discharge time constant so that noise bursts lasting a fraction of a second will yet be indicated. With a meter of relatively short time-constants, which is able to reach full-scale reading from zero in one-quarter or one-half second, the maximum deflexion for the three noise performances shown in Fig. 8, *a*, *b*, and *c*, will have

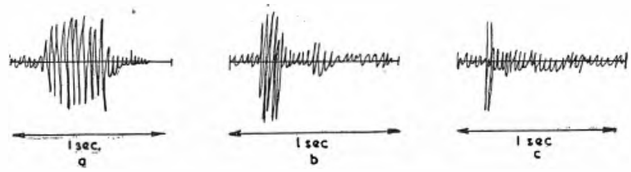


Fig. 8.
Types of response which the noise meter has to be capable of indicating

the same maximum reading, for *b* and *c* of the same and for *a* slightly longer duration. In other words the peak values of averages taken over periods of about 0.1 second, but extended by a discharge-half-time of about one-quarter second are indicated. A meter with a slower time constant would not be able to follow the rapid changes as the rotor of the potentiometer is turned, and a short burst of noise (which is very noticeable when the potentiometer is in use and a noisy spot is passed) would not be indicated. On the other hand, a faster discharge time would not allow even a rapidly indicating meter to follow up quickly enough. The compromise is governed by suppressing bursts of durations much shorter than one-tenth second, which will not be very noticeable. It is very useful to include also a peak value indicator, which can be done by switching the detector circuit, using the same diode and meter as for the average, with different *R-C* members. A comparison of peak and average noise figures often yields interesting information for the interpretation of noise measurement results. Even more useful is the observation of noise traces during noise measurement as this often permits the location of the noisy area on the resistance track and even allows tentative diagnosis of the cause of rotation noise in some cases of excessive magnitude.

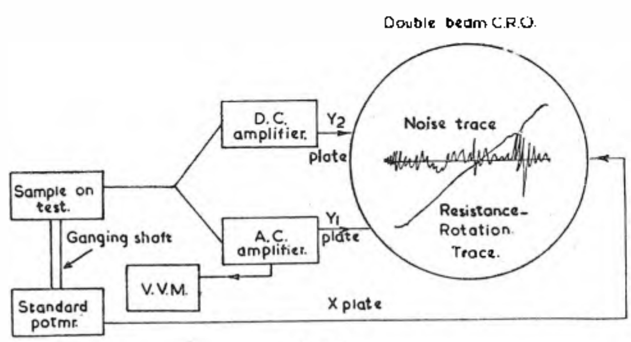


Fig. 9.
Circuit to give resistance rotation characteristic superimposed over noise trace

A useful combination of noise measurement, noise observation and observation of the resistance rotation characteristic making the location of the noise area possible is provided by a circuit such as shown in Fig. 9. A double beam cathode-ray tube is used as a

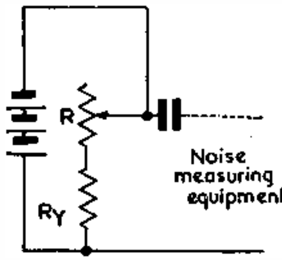


Fig. 10.

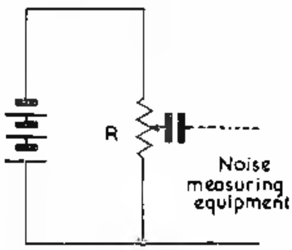


Fig. 11.

Circuits for measuring noise in a potentiometer whose sliding contact carries current (Fig. 10) and where all the current is carried by the resistance element (Fig. 11)

monitor, with an X-shift provided by a standard potentiometer carrying a D.C. deflecting voltage large enough to provide a sufficient horizontal deflexion. The Y_1 plate is connected to a noise amplifier, the Y_2 plate to a direct coupled amplifier connected to the potentiometer on test. On the screen, the resistance rotation characteristic is seen superimposed over the noise characteristic, and a transparent graticule can be used, which is calibrated in terms of degrees, or per cent of full rotation. The spindles of the standard and tested potentiometers are connected by a connecting sleeve, and turned simultaneously.

Another noise test which is sometimes carried out, and which needs a greater amount of amplifier gain is for microphony of a stationary variable potentiometer. The potentiometer is inserted in the jig and tapped or vibrated slightly, and the increase in noise produced is measured. It is natural that certain types of potentiometers are more prone to microphony than others, according to the design of their constituent parts.

In all these tests, it is important to distinguish between such applications where the contact carries an appreciable current, and those where the resistance element carries an appreciable current, but the sliding contact does not. Noise measurements serving to assess the suitability of variable potentiometers to operate under these conditions must be made with equipment designed accordingly. For the first case, therefore, the measurement is carried out as shown in Fig. 10. The series resistor R_Y takes care of limiting the current and must be included, otherwise serious damage will result to the potentiometer on test when the potentiometer which is connected as a variable resistor approaches zero-setting. The second case, representative of measuring noise from receiver gain controls (volume controls) is illustrated in Fig. 11. No series resistor is necessary as the current is constant and only small amounts of A.C. pass through the moving contact.

It is characteristic for the subject of noise measurement that a number of conventional techniques are applied in a considerably modified manner. In applying these techniques it is important to realise that noise is a complex quantity of small magnitude, and that interpretation of figures obtained by measurement can only be reliable if the effect of all parts of the measuring equipment, electrical and mechanical, their properties and influence upon the resulting numerical meter indications is fully taken into consideration.

Pulse Amplifier for Fast Ionization Chambers and Proportional Counters

A two-unit equipment particularly suitable for coincidence work and counting in the presence of a high background.

SPECIFICATION 120 issued by the Atomic Energy Research Establishment, Harwell, describes the pulse amplifier Type 1049B, which has been designed primarily for use with fast ionization chambers and proportional counters. The upper half-power frequency of the amplifier is 2 Mc/s., thereby making the equipment particularly useful for coincidence work and counting in the presence of high backgrounds. There are two units: a head amplifier which may have its maximum gain set to 100 for ionization chamber work or 5 for proportional counter work and which also contains an E.H.T. filter unit; and a main amplifier with a maximum gain of 10,000 and which contains all the necessary power supplies.

The technical details of the two units are as follows:—

Main Amplifier

Maximum gain 10,000 with provision for a reduction to 100 in minimum steps of 2 db.

Input impedance 100 ohms.

The time constants of differentiation and integration, separately adjustable on two switches, may be set to any of the following values: 0.08 μs, 0.16 μs, 0.32 μs, 0.8 μs, 1.6 μs, 3.2 μs and 8 μs. These figures correspond to an overall frequency coverage for the amplifier of 2 Mc/s. to 20 kc/s. respectively.

The main amplifier will deliver a maximum position output pulse of 50 volts with a linearity over this range not worse than 1 per cent.

The maximum signal ever likely to be fed into the main amplifier will be approximately 1 volt (10 mA into 100 ohms) and this will give rise, on full gain, to an overload of 200 times. The amplifier is designed to handle this overload factor without appreciable paralysis. Overswings on the pulse, which may also cause overloading of the later stages in the amplifier, are similarly catered for.

There is no provision for phase reversing in the main amplifier and for a positive output pulse the nominal input is negative.

Head Amplifier

Maximum gain is 100 for ionization chamber use, and 5 for proportional counter use. The head amplifier has a gain frequency characteristic substantially flat over the range 50 c/s. to 2 Mc/s. There is also provision in this amplifier for phase reversing, this being accomplished by a simple interchange of leads inside the unit.

An E.H.T. filtering circuit is built into the head amplifier, and this may be used with both ionization chambers and proportional counters.

Full details can be obtained on request to A.E.R.E., Harwell, Berkshire, quoting Specification 120.

Precision A.C. Voltage Stabilizers

By G. N. PATCHETT, Ph.D., B.Sc.(London), A.M.I.E.E., M.I.R.E.

(Bradford Technical College)

Part III

The Controller

IN electronic A.C. stabilizers the controller usually consists of saturated reactances or high vacuum valves. The former[13,22,41,42-47] were tried by the author and abandoned due to the following disadvantages:—

1. Serious distortion is caused due to the non-linear property of the reactance producing a highly distorted voltage across the reactance and therefore distortion of the stabilizer output voltage.
2. The maximum voltage drop across the reactance depends on the current, so that the range of input voltage covered by the stabilizer depends on the load.
3. A large drop is required across the reactance to cause appreciable changes in output voltage since, at unity power factor, this voltage is in phase quadrature with the supply voltage.
4. The time constant tends to be large, causing "hunting" and a large response time.

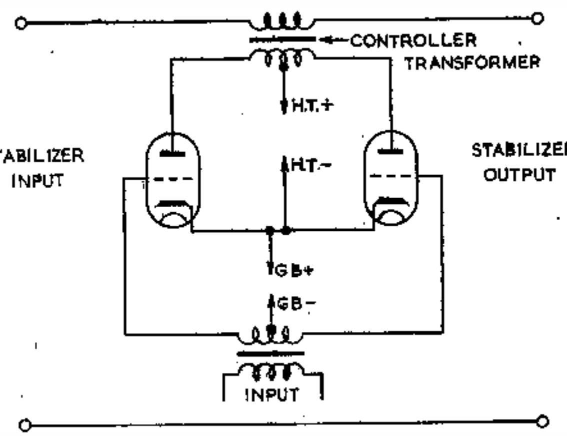


Fig. 15.
Circuit of controller (Arrangement A)

Valves may be operated in a number of ways, but, since the power to be dealt with in the controller is usually large it is important that they are used efficiently, and also in such a manner that the distortion produced is a minimum. Two methods have been used by the author and these will now be considered in detail. The first method is the more universal in application, but the second can deal with larger powers, for a given valve dissipation.

Arrangement A.

The circuit of the controller is shown in Fig. 15 and from this it might appear that the circuit operates as a conventional push-pull output stage, as used in audio frequency amplifiers. As will be shown later, this is a long way from the truth. Most authors[16] dealing with valve controllers dismiss the operation of the circuit in a few lines, assuming that it operates in the normal manner. As no detailed account of the methods of operation of valves in controllers appears to have been given previously, the author will consider these in some detail.

For simplicity, it is convenient to consider first a single valve operating under Class A conditions, fed in a similar way to that shown in Fig. 15. In normal stabilizers, since the change of input voltage is a small percentage of the input voltage (say 5 to 10 per cent), the maximum voltage across the controller is also a small percentage of the supply voltage. The current through the load (and through the controller, assuming the controller in series with the load) is practically independent of the magnitude of the controller voltage. Thus the controller current, for a fixed load, may be assumed constant independent of the grid drive to the valve. This is the main difference compared with the normal method of operation where the valve current (A.C. component) is approximately proportional to the grid drive.

Considering Fig. 16, let A be the static operating point determined by the H.T. supply (V_{ad}) and the grid bias on the valve (-20 volts). If the D.C. anode current is I_{ad}, then the static power input, and therefore anode dissipation, is $I_{ad} \times V_{ad}$. If the valve were operated in the normal manner feeding a resistance load, the operating line or load line would be such as PQ, and the valve would be generating A.C. power, so that the anode dissipation would be reduced by the amount of the A.C. power output. The maximum anode voltage and current would be limited in value and depend (for a fixed resistance load) on the magnitude of the grid drive. In the case of valves operating in the controller of a stabilizer the peak anode current is determined by the load on the stabilizer and, as has been shown, is practically independent of the grid drive. The peak anode current is shown by the two horizontal lines in Fig. 16, representing a peak anode current of I_a. For simplicity, it will first be assumed that the load on the stabilizer is resistive and that the grid drive is in phase or antiphase with the load current. If the grid drive is zero, then the valve will operate along the fixed grid bias line R-S and the A.C. voltage drop across the valve will be small. If the grid drive is in phase with the anode current the operating line becomes such as T-U or, with greater grid input, V-W, in all cases the anode current going between the fixed limits set by the horizontal lines. Since the anode voltage and anode current are in antiphase the valve is generating power, the voltage across the secondary of the controller transformer assisting the supply. Since the valve is now generating A.C. power, the anode dissipation will be reduced compared with the static case.

If the phase of the grid drive is reversed so that

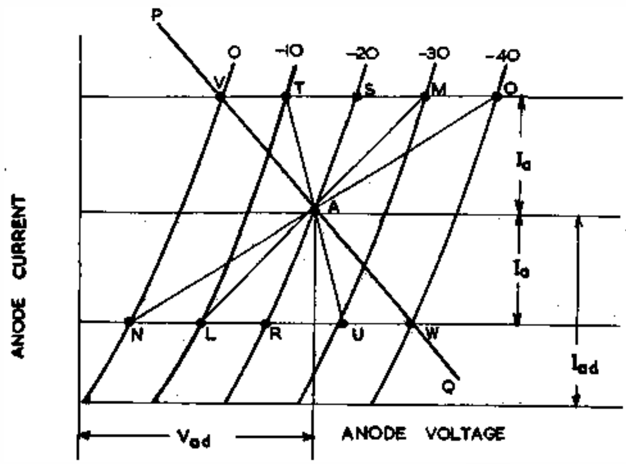


Fig. 16.
Operation of controller (Class A)

the anode voltage and anode current are in phase, the operating line becomes such as *L-M* and *N-O*, the valve absorbing power and the voltage across the transformer secondary now opposing the supply voltage, so reducing the stabilizer output voltage. The anode dissipation is now increased above the static value by the amount of the A.C. power absorbed. In normal output stages the static power dissipation is usually made about the same as the maximum anode dissipation and, if this procedure is used for valves operating under the above conditions, the valve's safe anode dissipation will be exceeded in the case where the controller is bucking the supply voltage. Further, since I_a is more or less constant, if the grid drive of the valves increases above normal (due to the input voltage to the stabilizer increasing so that it is outside the normal range of operation) the operating line becomes more and more horizontal and, since it must also go between the fixed current limits, the anode voltage swing will increase and very excessive anode voltages may result. This may either damage the transformer or valves. If the stabilizer is operating on a reactive load it can be shown that the operating lines become ellipses, but the same general remarks apply.

In practice it is preferable to use two valves operating Class B push-pull, as the efficiency can then be increased, and this is extremely important due to the large powers normally involved. Considering now the operating conditions under Class B, point *A* of Fig. 17 is the static operating point and the horizontal line is again the limit of anode current I_a, as fixed by the load. With a resistive load on the stabilizer and no grid input, the operating line, for one valve for one half-cycle, will be *A-Q*. Under the conditions of the controller boosting the supply, the operating line becomes a line such as *A-P*. Under the conditions of bucking the supply, the operating line becomes a line such as *A-R*. If the load on the stabilizer is reactive the operating line becomes half an ellipse, one case being shown dotted.

Let ϕ be the angle between the A.C. anode voltage and the A.C. anode current. (This is not exactly the same as the angle between the grid drive and the anode current due to the impedance of the valve, but depends on the power factor of the load in a similar way and is not very different under the maximum voltage conditions.)

$$\text{Let } i_a = I_a \,.\, \sin\theta$$

$$\text{and } v_a = V_{ad} + V_a \sin(\theta + \phi)$$

where I_a is the peak A.C. anode current

V_a is the peak A.C. anode voltage

Then

$$\text{instantaneous power } w = [V_{ad} + V_a \sin(\theta+\phi)][I_a \sin\theta]$$

$$= I_a V_{ad} \sin\theta + I_a V_a \sin\theta \,.\, \sin(\theta + \phi)$$

$$= I_a V_{ad} \sin\theta + I_a V_a/2 \cos\phi - I_a V_a/2 \cos(2\theta + \phi)$$

Then the average power per valve (no current flowing on the negative half-cycle of anode voltage)

$$= 1/2\pi \int_0^{\pi} [I_a V_{ad} \sin\theta + I_a V_a/2 \cos\phi - I_a V_a/2 \cos(2\theta + \phi)]\, d\theta$$

$$= 1/2\pi\, [2\, I_a V_{ad} + (I_a V_a \pi \cos\phi)/2]$$

$$= I_a V_{ad}/\pi + (I_a V_a \cos\phi)/4 \quad \ldots\ldots\ldots\ldots\ldots (1)$$

This will be a maximum when the controller is bucking the supply, i.e., the valve receiving A.C. power, which is when $\phi = 0$ and a minimum when the controller is boosting the supply, i.e., $\phi = 180°$.

In order to operate at all power factors without distortion, V_{ad} must equal V_{amax} where V_{amax} is the maximum A.C. anode voltage allowable and it is convenient to make $V_{ad} = V_{amax}$. Under these conditions the anode dissipation of the valve under the conditions of maximum A.C. anode voltage is

$$I_a V_{amax} (1/\pi + \tfrac{1}{4} \cos\phi)$$

when $\phi = 0$, the condition of maximum anode dissipation, this becomes:—

$$0.57\, I_a\, V_{amax}$$

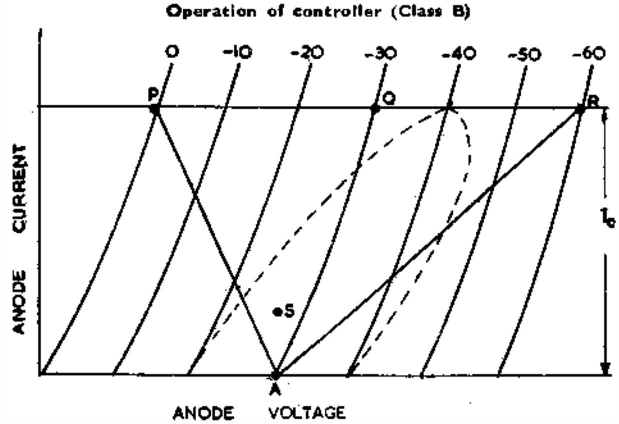
or, $I_a V_{amax} = 1.76 \times$ maximum anode dissipation per valve

$= 0.88 \times$ maximum anode dissipation of both valves.

Now the maximum volt-ampere input to the controller is $I_a/\sqrt{2} \times V_{amax}/\sqrt{2}$ from which we obtain the result that the:—.

Maximum controller volt-ampere input is $0.44 \times W_{amax}$ where W_{amax} is the maximum anode dissipation of both valves. The maximum volt-ampere output of the controller will be somewhat less than this, since the anode swing in the opposite direction will not be V_{amax} but probably $\frac{3}{4}\, V_{amax}$. Therefore the volt-

Fig. 17.
Operation of controller (Class B)



ampere output of the controller becomes $\frac{3}{4} \times 0.44 \times$ anode dissipation of the two valves.

The total change of volt-amperes produced by the controller is therefore

$$(1 + \tfrac{3}{4}) \times 0.44 \times W_{amax}$$
$$0.77 \times W_{amax}$$

This method of operation results in a larger change of controller volt-amperes than if the valves were operated under power output conditions only, as in a normal amplifier. It is extremely important that the grid drive is not allowed to exceed the value which makes the anode swing equal to V_{amax} or the anode dissipation will be exceeded and excessive anode voltages will be produced.

It is easily shown that the power from the D.C. source is $I_a V_{ad}/\pi$, the first term of Equation (1). The second term represents the power from the A.C. source. Thus the mean D.C. anode current is proportional to I_a, the peak A.C. anode current, which is proportional to the load current of the stabilizer. This variation of the mean anode current causes difficulties, since if this method of operation were used it would be necessary to design the H.T. supply with a good regulation from no load to full load. To overcome this difficulty it is possible to use automatic (cathode) bias for the valves. The cathode resistance (suitably bypassed by a capacitor) is designed to give the required bias corresponding to point *A* (Fig. 17) when the full current is taken from the stabilizer. When the load is reduced, the A.C. anode current decreases. This would result in a decrease in bias, but this causes the operating point to change to a point such as *S*, so that a steady anode current now flows (i.e., the valve changes towards Class A operation) and the mean anode current is approximately constant. The loss of efficiency due to this change towards Class A operation is not important, because with the reduced load current of the stabilizer the maximum power to be handled by the controller is reduced in proportion, the maximum change of controller voltage being fixed.

This arrangement of the controller has the advantages that it will operate equally well at all power factors from zero lag to zero lead, and from all loads from no load to full load. It has the disadvantage that the power change that can be controlled is only about three-quarters of the anode dissipation of the two valves. For greater power outputs a number of valves may be operated in parallel.

Arrangement B

This arrangement is based on the fact that if two resistors, one fixed and one variable, are connected in parallel and fed with a constant current, maximum power is dissipated in the variable resistor when it has a value equal to that of the fixed resistor. The voltage drop is then one-half of the maximum value, obtained when the variable resistor is made infinite. The maximum power dissipated in the two resistors is equal to four times the maximum value dissipated in the variable resistor. Thus, if this arrangement is used as the controller the power change of the controller will be four times that of the rating of the variable element. The variable resistor is replaced by two valves, acting as a variable resistor, so that the total controller power will be theoretically four times the dissipation of the valves. Owing to the fact that the valve resistance cannot be reduced to zero, the practical figure is about three times the anode dissipation, a considerable improvement in this respect on arrangement A.

The actual circuit used by the author is shown in Fig. 18 where the two valves V_3 and V_4 are operated under conditions of no D.C. bias or D.C. anode voltage. The valves are connected in parallel with the two resistors R_3 and R_4. For simplicity, assume that the valves are triodes having the characteristics shown in Fig. 19. Assuming that the load on the stabilizer is constant, the line representing the peak anode current is *P-Q*, being zero when the valve has an infinite resistance as at *Q*, and being the total current when the valve has zero resistance as at *P*. With no grid drive the bias is zero and the operating line is therefore *O-S*, with only a small voltage drop across the valve. As the grid drive is increased (in opposite phase to the anode voltage and current) the operating line becomes such as *O-R*, *O-T* and finally *O-Q*. Maximum power is dissipated in the valve when it operates along the line *O-R*, *R* being the centre of

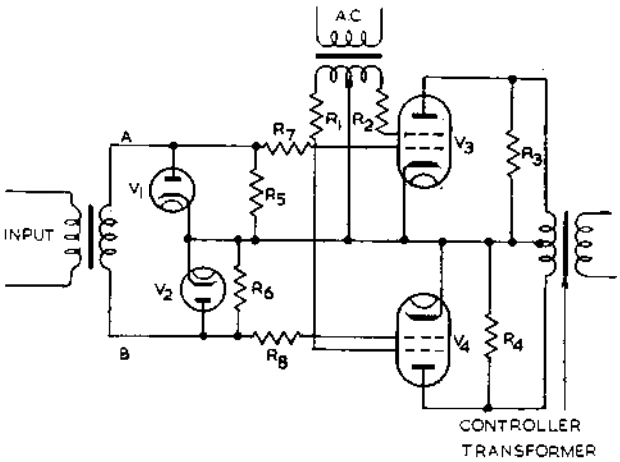


Fig. 18.
Circuit of controller (Arrangement B)

P-Q. For efficient operation, *S* should be as near *P* as possible and the valve should therefore have a low A.C. resistance, but, at the same time, the valve must withstand the high anode voltage corresponding to point *Q*. No suitable triode appears to be manufactured, so the author had to turn to pentodes or beam tetrodes. A normal pentode characteristic is quite unsuitable for the purpose, but a pentode can be made to have a dynamic characteristic similar to that of a triode by supplying the screen grid with a constant A.C. voltage, in phase with the anode voltage, from a separate supply, as is shown in Fig. 18. The resistors R_1 and R_2 are included in order to obtain a suitable characteristic and also to prevent the screen dissipation becoming excessive when the anode voltage is low, as at point *S* or if the load current is reduced. In order to make use of the full anode dissipation of the valves, the anode voltage at *Q* must be high and also the peak anode current at *S* must be high and a suitable valve must be chosen that will stand this method of operation. Beam tetrode valves with an anode dissipation of 35 watts have been operated in this manner in a complete stabilizer, using a peak anode voltage of 1,400 and a maximum

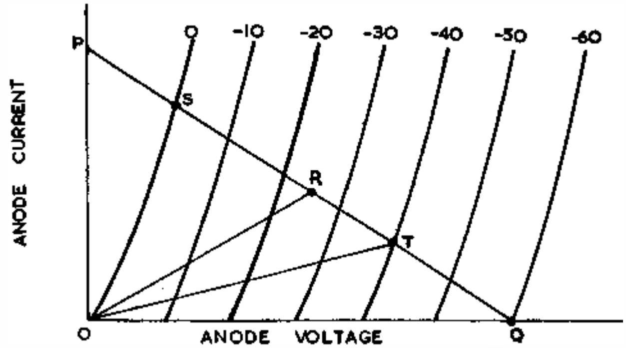


Fig. 19
Operation of controller (Arrangement B)

controller power of 240 watts for over 1,200 hours without the valves showing any marked deterioration.

As there is no bias on the valves it is desirable to feed only the negative half-cycle to the grid or grid current will flow. One method of doing this is shown in Fig. 18. When point A becomes positive with respect to B the diode valve V_1 will conduct and the full half-cycle will be developed across the resistor R_6 and fed through the grid stopper R_5 to the valve V_4. On the other half-cycle V_2 will be conducting and the full negative half-cycle will be fed to the valve V_3. In this way the full transformer voltage is fed to the valves on the negative half-cycles.

This type of controller has the following advantages:—

1. Large power control relative to the maximum anode dissipation of the valves.
2. No need to limit the grid drive as the circuit is self-protecting. The anode voltage cannot rise above that corresponding to point Q.
3. No H.T. or grid bias supplies are required,

but has the following disadvantages:—

1. Will only operate on a power factor near unity as the anode voltage must be approximately in phase with the fixed screen grid voltage supply, and, at low power factors, since the voltage drop across the valve must be in phase with the current (since they can only act as resistors) it has a small effect on the output voltage.
2. Only operates successfully on approximately constant load as the maximum voltage drop across the controller is proportional to the current. This may be partly overcome by using a tapped controller transformer.
3. The stabilizer output voltage must be less than the input voltage. This can be overcome by using a boosting transformer in the input to the stabilizer.

This type of controller enables a rather simplified stabilizer to be produced for large power outputs, but is of limited application.

Having considered the indicator and controller in some detail a complete stabilizer will be described in the next part.

[7] T. A. Ledward: "A. C. Voltage Stabilizer," *Wireless World*, 49, p. 166, 1943.

(To be continued)

Positive Feedback in A.F. Amplifiers

By C. H. BANTHORPE *(Central Equipment, Ltd.)*

GENERALLY speaking, positive feedback in A.F. amplifiers is avoided as far as possible, but it can be used to compensate for loss in gain due to negative feedback, and effect a saving in cost and space.

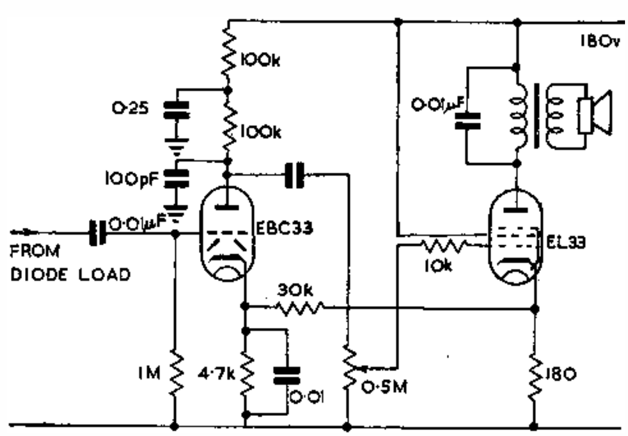
One such application, used by the writer, is shown, and will be seen to be in the A.F. stages of a broadcast receiver.

The cathode decoupling capacitors, usually 25 μF, are bulky and, of course, expensive. They also deteriorate, particularly in sets which run fairly hot; for instance, universal receivers in plastic cases.

If they are omitted, there will be negative feedback on both stages, and considerable loss of gain. If, however, positive feedback is introduced by means of a suitable resistor between the two cathodes, this will restore the gain. It is possible to calculate the value of the resistor, knowing all the factors, but it is probably easier to use a variable type, set it to the value which restores the gain, measure the value and substitute a fixed resistor.

As the cathode resistor of the double-diode-triode is in the diode load circuit, it must be decoupled by a small capacitor. In the arrangement shown, the volume control is in the grid circuit of the output valve. This is a good place for it as any inherent noise will be much less noticeable. It cannot usually be fitted in this way as there is a danger of the double-diode-triode being overloaded on strong signals, unless there is very good A.V.C., or unless A.V.C. is applied to the triode. Not many double-diode-triodes have suitable characteristics, however. With the arrangement shown, the over-load point is high as it is the output stage which overloads first if the gain control is at maximum, and when the gain control is turned down, the triode has more and more negative feedback applied, and can thus handle larger grid swings, as in a cathode follower.

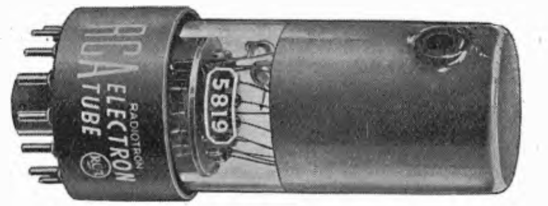
To reduce still further the chance of trouble due to the volume control becoming noisy, the output valve could be A.C. coupled to the volume control. This prevents any grid current flowing between the slider and track, which is one cause of noise in controls.



Multiplier Phototubes . . .
in Scintillation Counters

By R. E. B. HICKMAN

RCA Photophone, Ltd.



ONE of the most important recent advances in devices for the detection of nuclear radiations is the scintillation counter. While at present it is rather complicated and requires more elaborate circuitry than the Geiger-Muller counter, it is for many purposes superior to the older type of detector.

Principle of Operation

In its simplest form the scintillation counter[1,4] is illustrated in Fig. 1. *A* is a transparent phosphor crystal, which emits a flash of light when it absorbs a high energy nuclear particle such as an alpha, beta or gamma ray or a neutron. Many crystals have this property. Organic phosphors which have proved useful are naphthalene, anthracene and phenanthrene. Inorganic phosphors are slower than the organic ones, but more durable. Calcium tungstate, potassium iodide, thalium activated sodium iodide and silver activated zinc sulphide have all been used. The flash of light so produced falls on the multiplier phototube *B* producing photoemission from its photocathode and a current pulse in the output. This pulse is amplified at *C* and passed through a pulse height discriminator *D* to a count rate indicator *E*.

Suitable phosphor crystals can be found with a flash duration time of 10^{-8} or 10^{-9} seconds. These times are long compared to the processes involved in the multiplier tube, so that the resolving time of this type of counter can be made a small fraction of a microsecond, which compares with the 50 to 100 microseconds resolving time of a Geiger counter. This very short resolving time is one of the major advantages of the scintillation counter. Another important advance is that crystals are available which can convert nuclear radiation, and especially gamma-ray radiation, to which a Geiger counter is very insensitive, with many times greater efficiency than a Geiger counter.

Dark Current Pulses

The pulse height discriminator is necessary because a multiplier phototube even in darkness emits many random pulses which must be excluded from the pulse rate counter. By choosing a suitable phosphor and employing a wide aperture optical coupling to the photocathode it is possible to make the pulses due to scintillation larger than the majority of the dark current pulses. Within limits, the dark current itself does not interfere with scintillation counting. It is the fluctuations in this current which are troublesome.

The dark current consists mainly of thermionic electrons emitted from the photocathode and multiplied by succeeding dynodes. Other contributory causes are ionization of residual gas, cold discharge from irregularities on electrodes and ohmic leakage. An average value of dark current for a multiplier at room temperature and operating at 100 volts per stage is one microampere, while exceptionally good tubes may have as little as 0.05 microampere.

Using commercially available multipliers such as type 931A, it has been found that there are approximately 600 pulses per second of height equal to one unit of charge or greater. The current represented by these pulses is less than 1 per cent of the total dark current of the phototube, the remainder consisting of very small pulses which are below the amplifier noise level.

Factors Affecting Pulse Rate

The dark current pulse rate depends to a marked extent on the temperature of the phototube. Cooling the tube decreases the pulse rate considerably so that at temperatures in the region of the freezing point of water, the pulse rate is reduced to 5 to 10 per second.

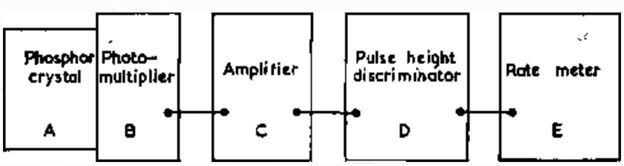
When the gain of the multiplier is raised by stepping up the overall voltage on the tube, the number of pulses per second of a given height increases. When the voltage exceeds a certain value, the number of large pulses increases owing to cold discharge phenomena.

Multipliers having a high pulse output may have their pulse characteristics modified by placing an external electrostatic shield around the glass envelope. With the shield at potentials up to 800V negative relative to the collector of the multiplier, the pulse rate decreases. Making the shield more negative causes an increase in the dark current pulse rate. This shield effect is not appreciable with the type 1P28 multiplier phototube which has an envelope of high resistance glass. It has been found that this expedient can be used to reduce the dark pulses from a particularly noisy multiplier tube, but that it does not greatly improve the performance of an inherently good tube.

Pulse Output Under Light Stimulation

The nature of the pulse output of a multiplier when stimulated by light is of equal importance with its dark current performance in determining its suitability as a scintillation counter. In general, it is found that the distribution of pulse heights obtained from a light stimulated multiplier shows considerably fewer large pulses than it does for the dark current. The distribution curves obtained from

Fig. 1.
Block schematic of the scintillation counter



emitting spots in different parts of the photocathode are almost identical, although considerably higher voltages are required for efficient collection of photoelectrons originating from the outer edges of the cathode.

It is possible from a curve of pulse height versus pulse rate to determine the background output of a detector. Given the maximum allowable background count for a given application, the number of photoelectrons required per scintillation to ensure detection may be determined from a curve of pulse height versus number of pulses per photoelectron.

Because of the small area of the photocathode and its distance from the glass envelope, it is difficult to achieve good optical coupling to a crystal phosphor.

Type 1P21 is identical in form and design to type 931A but is specially selected for very high gain, very low dark current and a high degree of stability and uniformity.

Type 1P22 is similar to type 931A but its maximum spectral response is at 4200 Angstroms instead of 4000 Angstroms.

Type 1P28 is similar to 931A but has a quartz glass envelope making it sensitive to ultraviolet radiation.

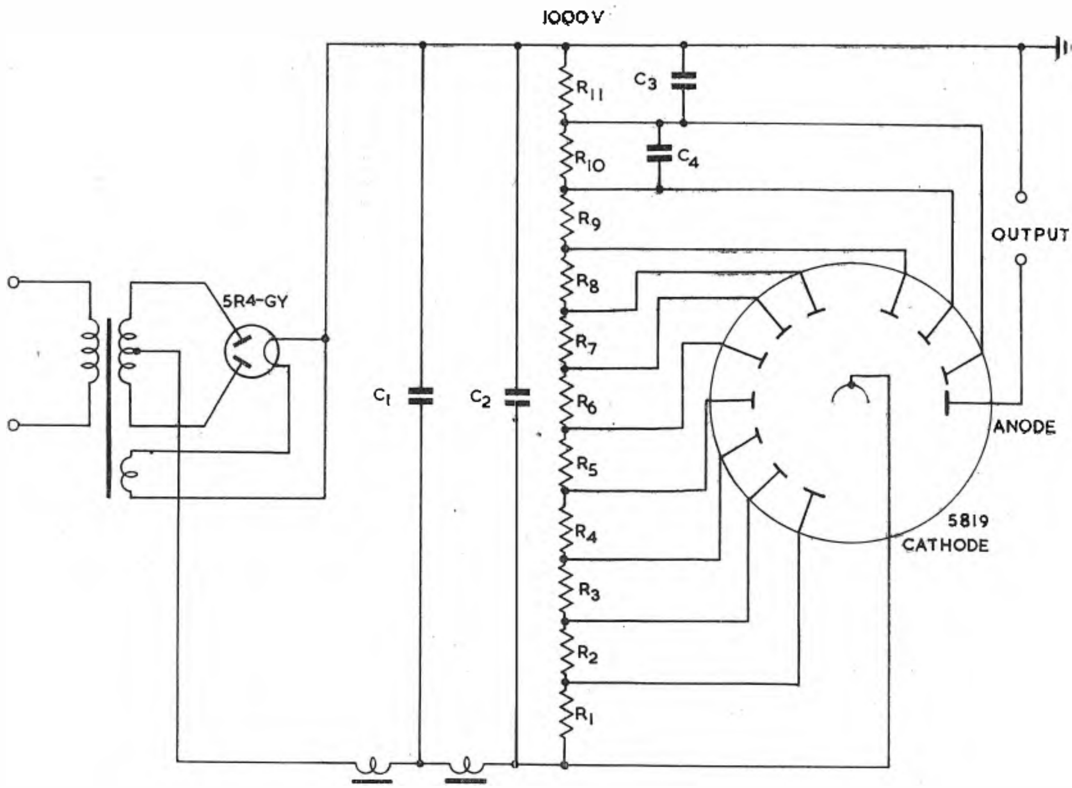


Fig. 2.

Circuit application of the multiplier phototube illustrated at the beginning of the article

R1—11 : 20kΩ, 1w. C1—2 : 2μF, 1000 V.D.C. C3—4: 8μF Electrolytic, 150 V.D.C.

Choice of Multiplier Phototube

Of the two principal types of multiplier phototubes, the static and the dynamic, only the former has been used to any extent in experiments in scintillation counters. Static multipliers may in turn be subdivided into several classes, including magnetically focused, electrostatically focused and unfocused. Of these, the most generally useful for scintillation counting is the electrostatically focused[2]. Commercially available are the types 931A, 1P21, 1P22, 1P28 and 5819.

Type 931A is a nine stage multiplier with a small internal photocathode. It has high gain, is normally stable in operation and of uniform characteristics.

Type 5819 is a ten stage head-on multiplier phototube having on the glass end of the tube a large area ($1\frac{1}{2}$ inch diameter) semi-transparent photocathode which is essentially flat. The internal structure is in general very similar to that of type 931A. The relatively large photocathode area permits a large optical coupling and very efficient collection of light from a large phosphor crystal. With a suitable phosphor it is possible to make the scintillation pulses larger in amplitude than the majority of the dark current pulses, giving a greater reliability of counting. The spectral response of type 5819 is in the range 3000 to 6400 Angstrom units, with a peak at about 4800. In this region many of the organic and

TUBE TYPE	931A	1P21	1P22	1P28	5819
Photocathode Area (sq. cm.)	1.9	1.9	1.9	1.9	11
Wavelength Maximum Response (Angstroms) ...	4000	4000	4200	3400	4800
Number of Stages	9	9	9	9	10
Volts Per Stage	100	100	100	100	90
Current Amplification (Average)	1×10^6	2×10^6	2×10^5	2.5×10^5	6×10^5
Luminous Sensitivity (Average—A/l)	10	80	0.6	5	7.5
Anode Current (Average—mA)	1.0	0.1	1.0	0.5	0.75
Anode Dark Current (Max.—μA)	0.25	0.1	0.25	0.25	0.05
Base	Sub-magnal — 11 pin				Diheptal 14 pin

inorganic phosphors respond efficiently to radio-active excitation.

A list of the most important characteristics pertinent to usage in scintillation counters of the above-mentioned multiplier phototubes is given in the table.

Circuit Arrangement

In a great number of cases circuits used in scintillation counters are unique to the experiment being performed, but some general considerations are given.

To satisfy conditions for electrostatic focus, the voltage steps, between successive dynodes, for multiplier phototubes of the types 931A, 5819, etc., should be sensibly uniform. Normally steps of 70 to 120 volts are used. Sometimes, however, a slightly higher step is used between the photocathode and the first dynode to ensure good collection, and higher voltages may be used on the later stages to obtain increased gain without an over-large corresponding increase in the dark current pulse rate. The voltage source should be of a high order of regulation, since a given percentage variation in overall voltage will produce approximately ten times that variation in gain. It has been found most practical to obtain the required voltage steps from a resistive voltage divider, which may take the form of small resistors soldered directly to the multiplier tube socket. The overall resistance of such a potential divider should not be less than one megohm nor more than about 100 megohms. It may be found advantageous to by-pass, using small capacitors, the resistors feeding the final stages of the multiplier. In most instances it is recommended that the positive high-voltage terminal be earthed rather than the negative terminal. Fig. 2 shows a full-wave rectifier power supply circuit which will give excellent regulation and minimum hum modulation.[3]

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[1] G. A. Morton and J. A. Mitchell, *RCA Review*, 9, 4, 1948.
[2] V. K. Zworykin and J. A. Rajchman, *Proc. I.R.E.*, 27, Sept., 1939.
[3] F. V. Hunt and R. W. Hickman, *Rev. Sci. Inst.*, 10, Jan., 1939.
[4] J. W. Coltman, *Proc. I.R.E.*, 37, June, 1949.

Optical Mapping of the Space-Charge Field in a Magnetron

AN accurate, sensitive technique for experimentally determining the electric-field distribution and space-charge density within a magnetron has been developed by D. L. Reverdin at the American National Bureau of Standards. The new method, which is also well suited to investigations of electron-optical lenses, gas discharge, and other space-charge problems, is a modification of the electron-optical shadow technique[1] recently developed at the Bureau for the quantitative study of minute electric and magnetic fields. A magnetic lens is used to produce shadow images of two fine wire screens placed at either end of the magnetron in the path of an electron beam. Then, from the distortion in the shadow network caused by deflexion of the electron rays as they pass through the magnetron field, the radial electric field is computed. These results are used to obtain the space-charge distribution.

The magnetron is a vacuum tube widely used for generating power at microwave frequencies. In its elementary form, it consists of a cylindrical cathode—ordinarily a straight-wire filament—and a coaxial cylindrical anode. The tube functions under the joint action of an externally applied magnetic field and the electric field created by applying a direct current potential between the anode and cathode.

The high space charge density within a magnetron is known to have an important bearing on performance. However, very little is actually known concerning the electric field distribution and space charge configuration within the tube. Although the problem has been investigated theoretically by many workers, the formidable mathematics involved has not permitted an exact solution, and the various simplifications of the theory that have been suggested have led to widely divergent results. Attempts at direct measurement have also proved unsuccessful, because the very critical symmetry of the field under study was disturbed. A promising approach to the problem has now been provided by the method developed at the Bureau. This technique has been used to map the charge distribution within a cut-off or steady-state magnetron. Further application to oscillating magnetrons should lead to a much better understanding of their operation and should yield information of considerable value to the engineer who is interested in designing improved types of magnetrons or in predicting the performance of existing types.

The Bureau's method uses an electron beam as a probe but keeps the charge density of the probe beam small compared to the space charge in the magnetron. Thus, the field under study is undisturbed. An electron gun sends the beam axially through the tube. Coaxial coils surrounding the magnetron provide a homogeneous magnetic field for the operation of the magnetron, and at the same time act upon the beam as a convergent magnetic lens, bringing it to a focus beyond the tube. Two fine wire screens are placed in the path of the electrons, one just in front of the magnetron, the other just beyond the back focus of the beam. A complex shadow pattern, due to the two wire screens, is then formed on a fluorescent screen. When the direct current potential across the magnetron is zero, the pattern is undistorted. However, when an electric field is applied to the magnetron, the shadow network on the fluorescent screen becomes quite distorted; and theoretical analysis of this effect has related the distortion of a given part of the pattern to the intensity of the electric and space-charge fields in the corresponding region of the magnetron.

In practice, photographs are taken of the shadow network, both in the undistorted and distorted form. The changes in the paths of the electron rays as they pass through the magnetron are then determined from measurements of the shadow patterns and the geometrical constants of the system, such as the positions of both wire screens, the magnetron, and the electron source, and the number of meshes per unit length of the wire screens used. From the deflexion of an electron ray entering the magnetron at a given radial distance from the centre, the strength of the electric field in the corresponding region of the magnetron is computed.

In comparison with previous methods using a pencil beam of electrons but no optical system, this method is much more sensitive and accurate. It also has the advantage of giving a complete field map in a very short time. The principal source of error lies in the uncertainty regarding the configuration of the electric fringe field at either end of the magnetron under space-charge conditions.

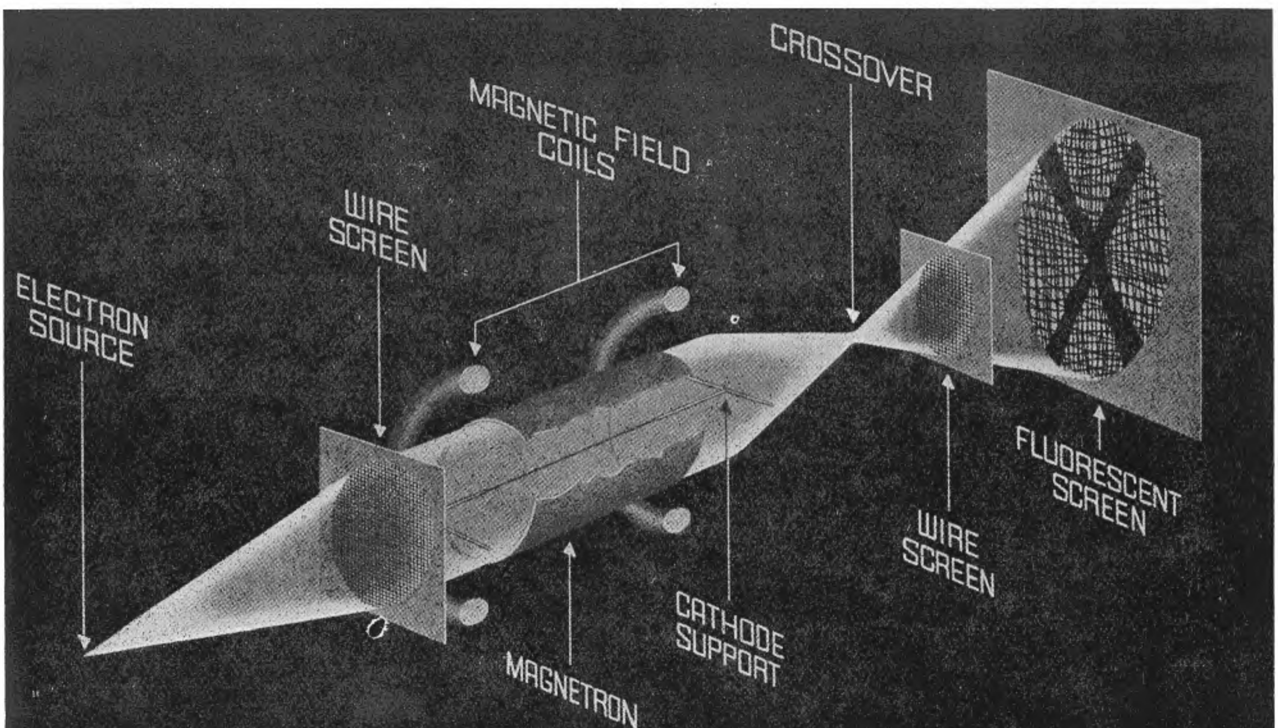
The Bureau's study of the field within a steady-state magnetron indicates that the actual space-charge distribution differs considerably from that predicted by the theorists. A number of different shapes of space-charge configuration were observed that are closely related to the symmetry of the magnetron. A certain lack of sharpness noted in the patterns gave a visual indication of the noise in the tube. This suggests further extension of the method to learn more about the problem of noise in an oscillating magnetron.

—*National Bureau of Standards Technical Bulletin*, 34, 57, 1950.

Reference

[1] *Electronic Engineering*, 22, 68, 1950.

The beam of electrons from an electron gun serves as a probe for mapping the magnetron field. A magnetic electron lens, the field of which is produced by two coaxial coils surrounding the magnetron, is used to form shadow images (on the fluorescent screen) of the two fine wire screens placed in the path of the beam. The same magnetic field serves for operation of the magnetron. Deflexion of the electron rays by the magnetron causes distortion in the shadow network, which may be measured, and from which the electric-field and space-charge distribution are computed.



The Grounded Grid Amplifier

An Analysis of the Essential Characteristics and Performance

By J. ROORDA

IN every amplifier stage the grid circuit, the anode circuit and the valve have one connexion point in common. Ordinarily this common point is grounded, or at least kept at a constant potential with respect to earth.

This consideration is the reason for classifying the amplifiers in three main types:

(a) The grounded-cathode amplifier; this is the classic type of amplifier circuit in which the grid circuit, the anode circuit and the valve have the cathode of the latter as common point.

(b) The grounded-anode amplifier; this type of amplifier circuit, having the anode of the valve as common connexion point with the input and output circuits, is more generally known by the name of cathode follower.

(c) The grounded-grid amplifier; in this type of amplifier circuit the control grid of the valve is the common point between the grid circuit and the anode circuit.

The grounded-grid amplifier has certain advantages over the other types, especially when used as a high frequency amplifier. It is the main purpose of this paper to study the essential characteristics of the grounded-grid amplifier as an H.F. amplifier. This will be done on the following presumptions: the amplifier valve is a triode (or, more generally, a multi-electrode valve of which only the cathode and anode potentials vary with respect to the control grid) of which the working point and the working conditions are so chosen that it is a linear amplifier, i.e., the internal resistance ρ, the amplification factor μ and the mutual conductance g are regarded as constants. Moreover, the interelectrode capacitances will be looked upon as being constant, i.e., the electron transit time in the valve and the impedance of the connecting leads to the electrodes of the valve are disregarded.

Basic Circuit of the Grounded-Grid Amplifier

The basic circuit of the grounded-grid amplifier is pictured in Fig. 1. In this diagram V_g represents the grid bias voltage and V_a the voltage of the H.T. supply from grid to anode. The total D.C. voltage, working between the anode and the cathode is the algebraic sum of V_a and V_g.

The load impedance Z_a is connected between the grid and the anode. Over this impedance the instantaneous anode voltage change e_a with respect to the control grid is developed under the influence of the instantaneous grid voltage change e_g, occurring between the cathode and the control grid. The voltage e_a works in this sense, that an anode current increase (corresponding with a grid voltage increase) decreases the instantaneous anode voltage with respect to the instantaneous grid voltage, and reversely, of course. From this it will be clear that the instantaneous voltage change, working between anode and cathode, equals $-e_a+e_g$. The instantaneous anode current change can therefore be calculated from:

$$i_a = g\,e_g - \frac{e_a - e_g}{\rho}$$

As e_a is related to i_a by $e_a = i_a Z_a$, the instantaneous anode current change proves to be:

$$i_a = \frac{(1+g\rho)\,e_g}{\rho+Z_a}$$

or:

$$i_a = \frac{(1+\mu)\,e_g}{\rho+Z_a} \quad \text{.......... (1)}$$

by using the well-known relation: $\mu = g\rho$.

From this we find for the voltage amplification of the valve:

$$A = e_a/e_g = \frac{(1+\mu)\,Z_a}{\rho+Z_a} \quad \text{.......... (2)}$$

Comparing this result with the classic formula for the voltage amplification of the grounded-cathode amplifier, we see that by changing to the grounded-grid connexion the voltage amplification is increased in the ratio $(1+\mu)/\mu$; it looks as if the amplification factor of the valve were increased in the mentioned ratio by the change of connexions.

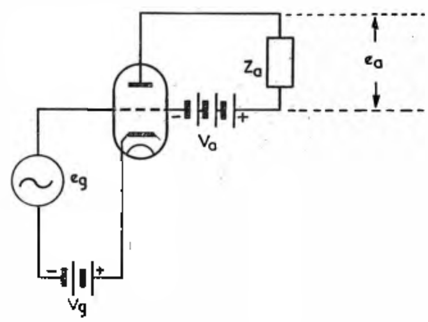


Fig. 1.
Basic circuit of the grounded-grid amplifier

Influence of the Interelectrode Capacitances on the Voltage Amplification

In the above paragraph the voltage amplification was calculated without considering the interelectrode capacitances of the valve. Neglecting these capacitances, however, is only allowable if the frequency of the alternating grid and anode voltages is low or very low. For high frequency amplification the influence of the interelectrode capacitances cannot be ignored. In this case the grounded-grid amplifier must be represented as depicted in Fig. 2, with the

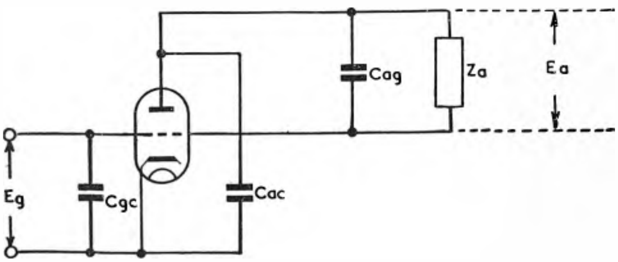


Fig. 2.
Grounded-grid amplifier with circuit "strays" included

omission of the supply batteries for V_g and V_a.

The alternating grid voltage with amplitude E_g and angular frequency ω gives rise to an alternating voltage across the load impedance Z_a with amplitude E_a and angular frequency ω.

From Fig. 2 we see that the load impedance Z_a is shunted by the anode-grid capacity C_{ag}, so that the actual impedance between control grid and anode is represented by the admittance

$$Y_a = 1/Z_a + j\omega C_{ag} \qquad (3)$$

On the other hand, the internal resistance ρ of the valve is shunted by the anode-cathode capacitance C_{ac}, so that the actual internal admittance is:

$$Y_i = 1/\rho + j\omega C_{ac} \qquad (4)$$

For the purpose of calculating the voltage amplification of the valve, the diagram of Fig. 2 can be redrawn as indicated in Fig. 3, in which the anode-

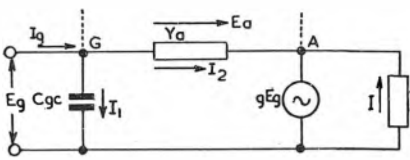


Fig. 3.
Simplification of Fig. 2 for calculating voltage amplification

cathode path of the valve is depicted as a constant-current generator, delivering the current $g\,E_g$, which generator is internally loaded by the admittance Y_i. The instantaneous direction of the voltages and currents is found easily by assuming that the alternating grid voltage is in the phase that drives the control grid positive with respect to the cathode. In the opposite phase the direction of all voltages and currents is, of course, reversed.

For calculating the voltage amplification, i.e., the ratio E_a/E_g, we find, by applying Kirchhoff's laws to the diagram Fig. 3:

$$\begin{cases} E_g = E_a - I/Y_i & (5) \\ g\,E_g = I_2 + I & (6) \\ E_a = I_2/Y_a & (7) \end{cases}$$

Eliminating I_2 from (6) and (7) gives:

$$I = g\,E_g - E_a Y_a$$

and this value, introduced in (5), yields:

$$A = E_a/E_g = \frac{g + Y_i}{Y_i + Y_a} \qquad (8)$$

or, by using the expressions (3) and (4), for Y_a and Y_i

$$A = \frac{g + 1/\rho + j\omega C_{ac}}{1/\rho + 1/Z_a + j\omega(C_{ac} + C_{ag})} \qquad (9)$$

Assuming for the moment that Z_a is a constant, we see from (9) that the voltage amplification is decreased as the angular frequency is increased. At very high frequencies, when the capacitive terms predominate, the voltage amplification becomes less than 1, i.e., there is no more a voltage gain, but a voltage loss.

For H.F. amplification the load impedance Z_a will generally consist of a tuned circuit. This circuit can be represented (see Fig. 4) by a resistance r shunted by a capacitor of capacitance C and by a coil of self-inductance L. With this configuration of elements the admittance of the circuit for an angular frequency ω is:

$$Y = 1/Z_a = 1/r + j\omega C + 1/j\omega L \qquad (10)$$

The voltage amplification is in this general case:

$$A = \frac{g + 1/\rho + j\omega C_{ac}}{1/\rho + 1/r + j\omega(C + C_{ac} + C_{ag}) + 1/j\omega L} \qquad (11)$$

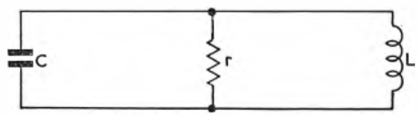


Fig. 3.
Representation of load impedance Z_a

The voltage amplification will be a maximum when the conditions are so chosen that the imaginary part of expression (11) equals zero. This will be the case when the angular frequency ω_0 or the constants of the circuit or both are so adjusted that the following condition holds:

$$(g + 1/\rho)\ \omega_0(C + C_{ac} + C_{ag}) - 1/\omega_0 L = \omega_0 C_{ac}(1/\rho + 1/r)$$

This yields

$$\omega_0^2 = \frac{1}{L(C + C_{ac} + C_{ag})} \cdot \frac{1}{1 - \dfrac{C_{ac}}{C + C_{ac} + C_{ag}} \cdot \dfrac{1/\rho + 1/r}{g + 1/\rho}} \qquad (12)$$

Expression (12) therefore describes the condition for which the amplifier is tuned to resonance, thus giving the maximum attainable voltage amplification. The latter then proves to be:

$$A_{max} = \frac{g + 1/\rho}{1/\rho + 1/r} = \frac{(1+\mu)\,r}{\rho + r} \qquad (13)$$

As A_{max} is generally much larger than 1, we find that the part

$$\frac{C_{ac}}{C + C_{ac} + C_{ag}} \cdot \frac{1/\rho + 1/r}{g + 1/\rho}$$

in expression (12) will be much less than 1, so that the condition for resonance is to a close approximation given by

$$\omega_0^2 = \frac{1}{L(C + C_{ac} + C_{ag})} \qquad (14)$$

showing that the actual tuning capacitance consists of the parallel connexion of the circuit capacitance C, the anode-grid capacitance and the anode-cathode capacitance.

The Input Admittance

The input admittance of the grounded-grid amplifier can be calculated from the diagram Fig. 3 by putting

$$Y_g = I_g/E_g \qquad (15)$$

As $I_g = I_1 + I_2$; $I_1 = E_g j\omega C_{gc}$; $I_2 = E\,Y_a = E_g\,.\,E_a/E_g\,.\,Y_a$

$=E_g \cdot AY_a$ we find

$$Y_g = j\omega C_{gc} + AY_a = j\omega C_{gc} + \frac{(g+Y_i)Y_a}{Y_i+Y_a} \quad \ldots\ldots\ldots \quad (16)$$

This general formula does not give very much information about the character of the input admittance, especially on the chief point of interest, viz., if and in which circumstances the input admittance can have a negative real part, so that the primary condition of the amplifier becoming a self-excited oscillator exists. As we are specially interested in the performance of the grounded-grid amplifier as an H.F. amplifier we will investigate this type more closely.

Combining (3) and (10) we get:

$$Y_a = 1/r + j\omega(C + C_{ag}) + 1/j\omega L$$

and from (4): $Y_i = 1/\rho + j\omega C_{ac}$, yielding

$$Y_g = j\omega C_{gc} + \frac{(g + 1/\rho + j\omega C_{ac})\left\{1/r + j\omega(C + C_{ag}) + 1/j\omega L\right\}}{1/\rho + 1/r + j\omega(C + C_{ag} + C_{ac}) + 1/j\omega L}$$

Introducing the abbreviation

$$X = \omega(C + C_{ag} + C_{ac}) - 1/\omega L \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (17)$$

we get

$$Y_g = j\omega C_{gc} + \frac{(g + 1/\rho + j\omega C_{ac})\left\{1/r + j(X - \omega C_{ac})\right\}}{1/\rho + 1/r + jX}$$

The real part of the input admittance, that is the input conductance, works out to

$$G_g = \frac{(g+1/\rho)X^2 - (g+2/\rho)\omega C_{ac}X + (1/\rho+1/r)\left\{1/r(g+1/\rho) + \omega^2 C_{ac}^2\right\}}{X^2 + (1/\rho + 1/r)^2}$$

$$\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (18)$$

The input conductance as a function of X has the following appearance:

$$G_g = f(X) = \frac{\alpha X^2 - \beta X + \gamma\delta}{X^2 + \gamma^2} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (19)$$

in which

$$\left.\begin{aligned} \alpha &= g + 1/\rho \\ \beta &= (g + 2/\rho)\omega C_{ac} \\ \gamma &= 1/\rho + 1/r \\ \delta &= 1/r(g + 1/\rho) + \omega^2 C_{ac}^2 \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (20)$$

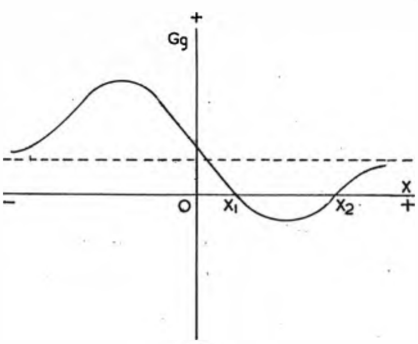


Fig. 5.

Graph of input conductance as a function of X

The general form of the graph representing G_g as a function of X is shown in Fig. 5. For certain positive values of X in the region from X_1 to X_2 the input conductance can have a negative value. In this region the amplifier therefore can become unstable, depending on whether the external grid-circuit conductance is smaller or larger in magnitude than the negative input conductance. Contrary, however, to the other types of amplifier, with the grounded-grid amplifier the working conditions can always be so chosen that the input conductance never becomes negative. Apparently this is the best safeguard for stable operation. The following analysis makes this clear and at the same time gives the condition for absolutely stable operation independent of the external grid circuit conductance.

The values X_1 and X_2, bounding the region of probably unstable working conditions are determined by

$$f(X) = \frac{\alpha X^2 - \beta X + \gamma\delta}{X^2 + \gamma^2} = 0,$$

giving

$$X_{1,2} = \beta/2\alpha \pm 1/2\alpha\sqrt{\beta^2 - 4\alpha\gamma\delta}$$

If, and only if, X_1 and X_2 are different, real values, the graph of Fig. 5 crosses the X-axis, showing that the input conductance may become negative. If $X_1 = X_2$ the curve just touches the X-axis, and if X_1 and X_2 are complex numbers the curve does not intersect the X-axis.

The condition for $G_g = f(X)$ never becoming negative is therefore

$$\beta^2 \leqslant 4\alpha\gamma\delta$$

or

$$(g+2/\rho)^2\omega^2 C_{ac}^2 \leqslant 4(g+1/\rho)(1/\rho+1/r)\left\{1/r(g+1/\rho) + \omega^2 C_{ac}^2\right\}$$

$$\ldots\ldots\ldots\ldots\ldots\ldots \quad (20)$$

From this we see that for a given valve and a given angular frequency the conductance $1/r$ of the tuned anode circuit can always be adjusted to such a value that the amplifier never can become unstable. The condition (20) is rather involved, but it can be simplified considerably for practical purposes and for estimating fairly accurately the necessary value of r.

First of all $1/\rho$ and $2/\rho$ are mostly very much smaller than g, so that they may be neglected with respect to g. With this (20) reduces to

$$g\omega^2 C_{ac}^2 \leqslant 4(1/\rho + 1/r)(g/r + \omega^2 C_{ac}^2)$$

or

$$\left\{g - 4(1/\rho + 1/r)\right\}\omega^2 C_{ac}^2 \leqslant 4g/r(1/\rho + 1/r)$$

The first term containing $\omega^2 C_{ac}^2$ is much larger than the second, so that the above condition is certainly fulfilled if

$$\omega^2 C_{ac}^2 \leqslant 4/r(1/\rho + 1/r)$$

From this we get

$$r \leqslant \frac{2\rho}{-1 + \sqrt{1 + (\rho\omega C_{ac})^2}} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (21)$$

as the condition for absolute stability of the grounded-grid amplifier.

Using the maximum allowable r according to (21) the maximum amplification which can be obtained with a grounded-grid tuned anode-circuit H.F. amplifier is, according to (13)

$$A_{max} = \frac{1 + \mu}{1 + \sqrt{1 + (\rho\omega C_{ac})^2}} \quad \ldots\ldots\ldots\ldots\ldots \quad (22)$$

This amplification is attained without the amplifier showing any tendency to instability.

Instantaneous Direction Finding

By J. RHYS JONES, M.B.E., M.I.E.E.

(The Plessey Co., Ltd.)

SINCE the earliest days of direction finding, there have been two incentives continually before designers: improved accuracy and greater speed of operation. In the ideal apparatus, the operation of a transmitter on the frequency to which the direction finding equipment is tuned immediately results in a pointer swinging round on a dial or a trace appearing on the face of a C.R.T. giving the direction of the transmitter from the receiver.

Consequently, in the late 'twenties, when a suggestion for an "instantaneous" instrument came from the Radio Research Board, considerable interest was aroused. This suggestion, in effect, recognized that if the N.S. and E.W. leads from a direction finding aerial system be connected to the deflecting plates of an exceptionally sensitive cathode ray tube, the plates being connected in the same pattern as the aerials, then a field will be created within the C.R.T. which is the counterpart of the field within the vicinity of the aerial system and, providing the face of the C.R.T. has been oriented correctly, the direction from which an incoming signal arrives will be correctly and instantaneously indicated on the face of said cathode ray tube.

As there is no C.R.T. of the order of sensitivity required, it became necessary to introduce amplification between the aerial system and the indicating C.R.T. The mechanical and electrical design of this amplifying system has to conform to rigid specification clauses to provide satisfactory performance, and the rigidity of these requirements put a brake on the adoption of this system of D.F. for general service.

The required amplification is carried out in two main parallel chains, one chain functioning between the N.S. part of the aerial system and the N.S. deflector plates in the C.R.T. and the other chain functioning between the E.W. part of the aerial system and the E.W. deflector plates in the C.R.T. In order that the field created within the C.R.T. by the deflecting plates shall be a true and exact replica of that existing within the vicinity of the aerial system, it will be appreciated that these two amplifying chains must have identical amplification and phase change characteristics and must hold this equality at all settings of the gain control, and over the entire tuning range covered by the equipment.

Should any change occur in either of these functions with the operation of any main control, means must exist to recognize the change, and to rectify the error. It is imperative, however, that any departure from the ideal be reduced to an absolute minimum, as the nearer the apparatus approaches the ideal specification, the greater the amount of traffic it can handle.

In the mid 'thirties the Radio Research Board team again stepped into the limelight and demonstrated a satisfactory and operative receiver functioning on this system. This effort—a milestone in D.F. technique—had two disadvantages; it was powered completely by batteries and all its tuned circuits had individual tuning capacitors. This lack of a ganged tuning capacitor made the receiver virtually a single frequency apparatus, since a change in frequency, unless small, was not a rapid operation. A considerable effort was applied to overcome these disadvantages, and the success achieved was demonstrated by the use made of this apparatus during the war.

The first successful commercially built apparatus of this type is shown in Fig. 1. This apparatus is of rack and panel construction and consists of two main racks, the larger carrying the receiver and having the operator's desk fitted in front, and the smaller carrying the necessary power supply and control circuits, this equipment being entirely mains operated.

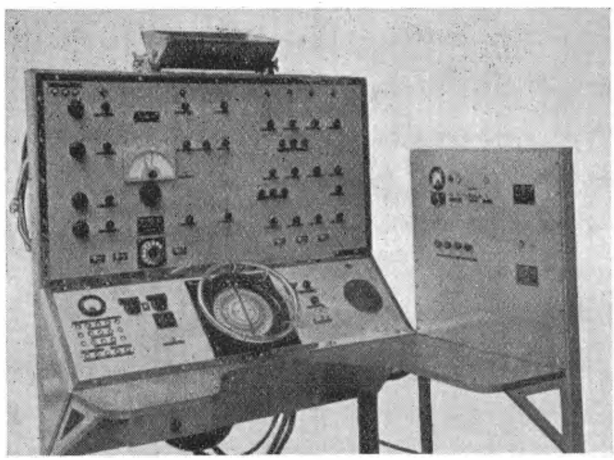


Fig. 1.
Pre-war Cathode Ray direction finding apparatus

The receiver part of this equipment was composed entirely of units, and as it had completely ganged tuning, the gang capacitor, which on this style of equipment is the heart of the system, was built as a very rigid structure and mounted direct on to the front panel. The individual tuning capacitors were placed in the most suitable geometric positions, and driven from the common tuning control shaft by steel belts, suitably spring loaded to avoid backlash.

This range of receiver operated on the three amplifier principle, that is, there were three complete and separate receivers mounted horizontally on the main front panel, one above the other, one amplifying the N.S. component of the incoming signal, the second amplifying the E.W. component, and the third supplying the "sense" signal. The only common feature between the three amplifiers, apart from the tuning drive and gain control, lay in the fact that they used a common local oscillator for frequency changing purposes, which is usual with this class of apparatus.

The purpose of the "sense" amplifier is to remove the ambiguity which exists in D.F. installations of most systems, by indicating which end of the trace to take as the bearing of the transmitter. This is done by also receiving the signal being checked on an omnidirectional aerial, which is positioned in the middle of the D.F. aerial system, suitably amplifying it (in this case by the third chain), and then applying the resultant voltage to the control grid of the cathode ray indicator tube. Providing the relative phase of the sense signal and the D.F. signals is suitably adjusted, the result will be to add to the brightness of the required end of the trace on the face of the cathode ray indicator tube, and subtract from the brightness of the other end, thus removing the ambiguity.

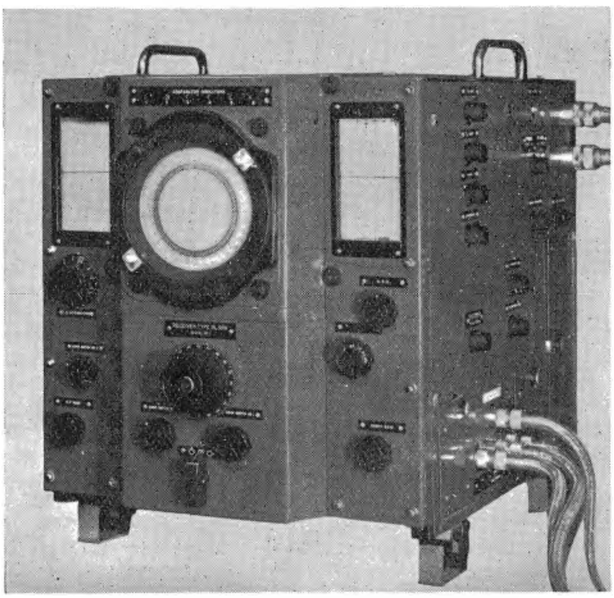


Fig. 2.
Improved Cathode Ray direction finding apparatus, of which compactness and operational convenience are features

This class of equipment gave excellent service, but owing to its bulk and weight, and its limited wave range, was superseded by a more compact and versatile equipment, the naval version of which has been very completely described in the *J.I.E.E.*, Vol. 9, Part IIIa, No. 15.

In this construction a seven gang capacitor is assembled on a horizontal base in a more or less orthodox layout, and sandwiched between the N.S. and E.W. chassis carrying the respective radio frequency amplifying and frequency changing portions of the circuit. The cathode ray tube is positioned just above the tuning capacitor, and the intermediate frequency amplifiers, circuit alignment facilities, and sense discrimination mechanism are suitably disposed around this basic arrangement.

This construction (see Fig. 2) presents great convenience operationally, as controls are well positioned, the cathode ray indicator tube is comfortably at eye level, and the circuits can be aligned sufficiently accurately and have adequate stability to permit watch being kept over a wide frequency band.

Novel features of this equipment are:—

1. The wide wave range covered—from long wave right through to 10 metres.
2. To achieve this coverage, a composite aerial system has been evolved, consisting of an Adcock four mast arrangement, having crossed loops at the centre. The required system can be selected at will.
3. The whole of the D.F. cabin is underground, symmetrically placed under the aerial system.
4. An earth screen has been fitted which ensures that the excavation under the aerial system is prevented from having an influence on the accuracy of the bearing obtained.
5. The receiver has been fitted with plug-in coils throughout, to permit coverage of the wide frequency range needed.
6. A test oscillator has been built into the control desk, to facilitate alignment of the D.F. receiver channels, to ensure maintenance of equality in gain and phase.
7. A second test oscillator has been installed in the field, and is remote controlled. This oscillator checks the whole station, including the aerial system, with a locally generated incoming signal, and thus ensures station accuracy.
8. This equipment, in common with the whole range of this type utilizes a modified sensing circuit, which obviates the need for the third amplifier which was a feature of the early receivers. A three-position switch is fitted to the front of the receiver having the positions—"N.S. sense"—"D.F."—"E.W. sense". In the middle or "D.F." position, the two chains operate normally, providing a bearing on the face of the C.R.T. complete with its ambiguity. On throwing the switch over to the "N.S. sense" position, the N.S. amplifier chain is left connected normally, but the E.W. amplifier chain has its input connected to a vertical aerial and its output connected to the control grid of the cathode ray tube indicator. This circuit change leaves the N.S. component of the signal on the face of the C.R.T., while, with correct phasing the E.W. output cuts off the ambiguous end of the trace, thus indicating whether the North or the South end of the D.F. trace should be accepted. Throwing the switch to the other direction reverses the roles of the two amplifiers, and indicates whether the East or the West end of the trace should be accepted.

One great advantage of this technique of direction finding lies in its ability to determine the separate directions of transmitters which are operating on a common frequency, and this feature led to the erection of a special D.F. station utilizing one of the equipments just described to keep a check on the activities of European broadcast stations at the latter part of the war. This station was built at the B.B.C. site at Tatsfield, and came into the news recently, when the Russians put up a radio jamming barrage.

LETTERS TO THE EDITOR

(We do not hold ourselves responsible for the opinions of our correspondents)

Planning V.H.F Mobile Systems

DEAR SIR,—I was interested to see the article on the above subject by E. R. Burroughes published in your August issue. It would seem, however, that the scheme shown in Fig. 3 of that article is wasteful in the use of the limited frequencies available. No less than five frequencies are used, one of these being duplicated with a 15 kc/s. shift.

A similar scheme for two frequency simplex as shown in Fig. A, using frequency modulation, needs two frequencies only, and the number of satellite stations may be increased without using any more frequencies. At each of the satellite stations the receiver and transmitter are linked so that the outgoing signal frequency is controlled by the incoming signal frequency. In this way all the satellite stations radiate at the same or approximately the same frequency and with the same frequency deviation.

No time delay circuits are used, as it has been shown that in the limited areas where the received signals are within 10 db of each other, intelligible speech is received even when a path difference of 30 miles exists.

Telephone land line links also are not required as would be the case in the author's scheme if the linking frequencies normally required were not available.

Regarding the choice of modulation and its dependence on the bandwidth available, with present techniques the system bandwidth is determined by the receivers, and not by the sideband components radiated by the transmitter. In current designs of F.M. mobile receivers the bandwidth is ±15 kc/s., and for A.M. receivers slightly greater. The technique to be adopted is surely to design the receiver to have the minimum bandwidth consistent with the components and quartz crystals available, and then occupy as far as possible this minimum bandwidth with useful intelligence required to be transmitted, hence giving the best possible signal-to-noise ratio. In this respect F.M. is to be preferred to A.M. as the minimum receiver bandwidth can always be utilized to the full.

Yours faithfully,

E. G. HAMER,

Wembley, Middx.

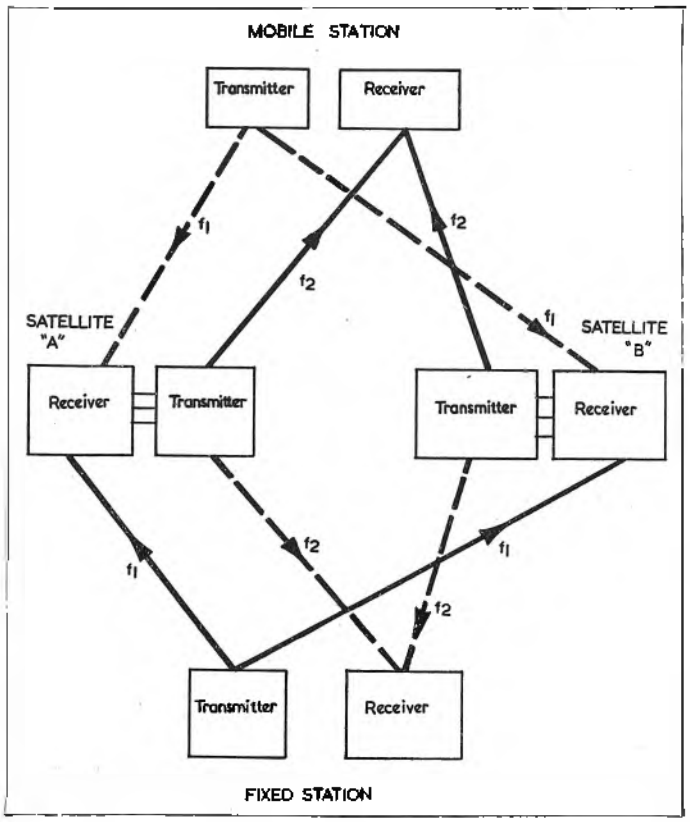


Fig. A. Scheme of Two Frequency Simplexworking proposed by Mr. Hamer in his letter on this page

DEAR SIR,—The twin diversity system in Fig. 3 was included primarily to show the operation of such a scheme as applied in practice. It will be recognized as being identical with the illustration of a multi-carrier system in Mr. Brinkley's article and is typical of a large number of police services in operation in England today.

The arrangement shown by Mr. Hamer can and has been used with amplitude modulation with the transmitter carriers staggered in frequency. No system of ensuring carrier synchronization is required with this system and, if necessary, the frequencies of the fixed-to-mobile go and return circuits can be nominally the same, whereas this is not practicable with the system shown by Mr. Hamer.

The point made concerning the receiver bandwidth is, of course, fundamentally correct and with existing frequency channelling there is no great problem. If, in the future, the frequency bands are re-allocated with a smaller frequency spacing between channels, it will be necessary to reduce the maximum deviation on F.M. in order to avoid interference, with a consequent reduction in its advantages. More stable circuits and crystals with a high order of frequency constancy will, of course, be required on both A.M. and F.M.

Yours faithfully,

E. R. BURROUGHES,

Johannesburg,

South Africa.

Electronic

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

Radio and Television Trimming Kit
(Illustrated below).

A NEW aid in the trimming and alignment of radio and television sets is the Newman Master Set Trimmer Kit. This kit, measuring 4 in. square by 1 in. deep is pocket size and fitted with the following :—

- 1 End Trimmer,
- 1 Side Trimmer,
- 1 Yaxley Switch Contact Adjuster,
- 1 Low Capacity Trimmer,
- 1 Screwdriver,
- 1 Set of feeler gauges,
- 1 Set of six Box Spanners from 1 to 8 B.A.
- 1 set of four Spanners from 0 to 8 B.A.

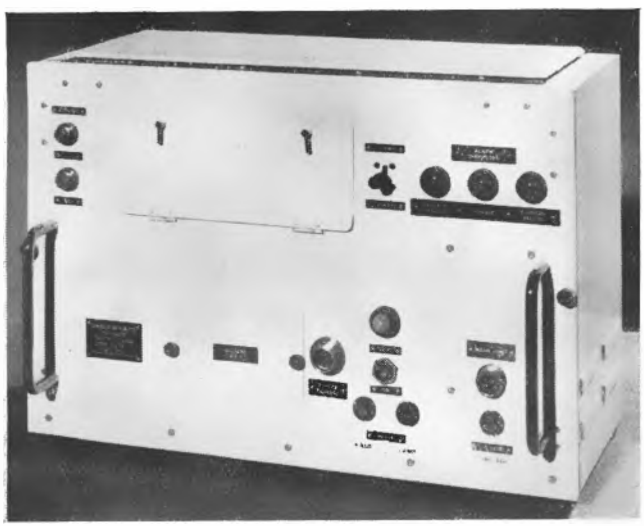
J. and S. Newman, Ltd.,
100, *Hampstead Road*,
London, N.W.1.

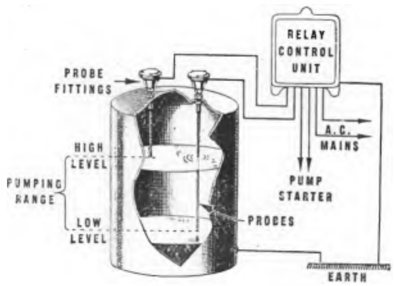
Door-Post Monitor
(Illustrated below)

THIS instrument has been designed to operate in conjunction with a 6 ft. G.M. tube mounted each side of a doorway leading into a laboratory where counting is taking place. The instrument will give visible and audible warning of any gamma-active material passing through the door.

The sensitivity level is adjustable, as are the individual counter voltages which are supplied by the instrument, and the ultimate limit of sensitivity is determined by the "background" of the laboratory in question. In an uncontaminated area a source of approximately 50 micro-curies carried at normal walking pace will operate the alarm. The instrument gives warning also of counter or mains supply failure.

Fleming Radio (Developments), Ltd.,
18-20 *Laystall Street*,
London, E.C.1.





Automatic Fluid Level Control

THE Elcontrol system of control for many types of electrical equipment replaces float switches, pressure bulbs, timers and pneumatic and magnetic float devices, etc.

This system comprises a simple relay control unit and an electrode or probe system consisting of one or more metal rods suspended from suitable insulated terminal fittings to which the control unit is connected. There are no stuffing boxes, floats or other moving parts to wear and fail in operation. This feature combined with high sensitivity makes the system particularly applicable for controlling a wide variety of fluids ranging from bauxite-caustic slurry to aerated substances, such as yeastfoam. The makers claim that reliable operation within close limits is independent of temperature or pressure, and is unimpaired by the presence of debris, etc.

The principle of operation is as follows: two metal electrodes, suspended from insulating bushes mounted above the liquid, project to the high and low levels between which a pump or other electrical auxiliary is required to operate. The probes are connected to the relay control unit by insulated wires contained in metallic conduit (see *the figure centre*). If the fluid in rising reaches the tip of the high level probe it completes the electrical circuit to earth. A minute current which flows in the circuit is amplified in the control unit to a magnitude sufficient to actuate the relay. The relay contacts control the pump motor either directly or through an automatic starter or contactor, and pumping is started or stopped as required. A second set of relay contacts changes over the probe connexions to prevent further action taking place until the fluid falls beneath the tip of the low level probe, when the original condition is restored. The cycle is repeated automatically. Thus the arrangement caters for the maintenance of the fluid level between pre-determined limits. Similarly the use of a single probe gives high or low level indication, which can be suitably signalled by lamp or alarm as desired.

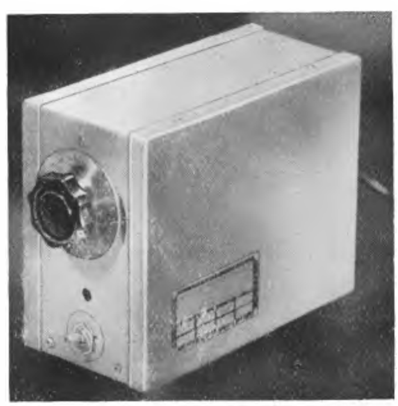
A doubly wound transformer isolates the equipment from the alternating current supply and reduces the voltage applied to the probes to a very low value.

Elcontrol, Ltd.,
10, *Wyndham Place*,
London, W.1.

"Sphere" Adaptor Unit, Type S.C. 88
(Illustrated below)

THIS unit has been produced to eliminate the necessity for troublesome, and often unsatisfactory, conversions of television receivers manufactured for the "London" frequency, to enable these to be used for the reception of the Sutton Coldfield transmissions.

The unit is inserted between the



Equipment

tronic apparatus, components and supplied by the manufacturers.

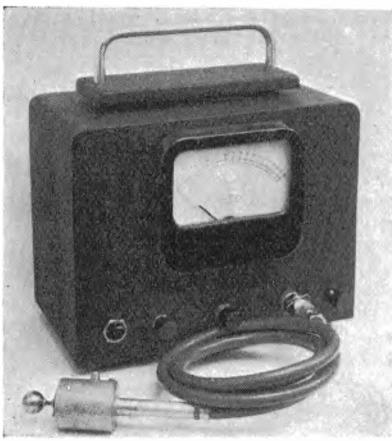
aerial and receiver, and has both input and output impedance of 80 ohms, the most popular di-pole termination impedance. Co-axial sockets are arranged at the rear of the unit, suitably marked, to take standard co-axial cable of 80 ohms impedance.

On front of the unit a control dial is fitted, to give a vernier discrimination between sound and vision channels. This should be adjusted to give a balance of sound and vision, with good sound quality, and full picture definition.

With the unit in use, the "London" set is left intact and untouched, retaining its full "design-efficiency" being at the same time adapted to the Sutton Coldfield transmissions and capable of first-class reception from the Sutton Coldfield station.

It has a power consumption of 15 watts, working on a voltage of 110/230/250 volts A.C. at 50/60 cycles.

Sphere Radio Ltd.,
Heath Lane,
West Bromwich.



M.P.J. Filmeter

A LIGHT, portable, robust mains operated meter for measuring electrically insulated coatings, such as cellulose, enamels, glasses, sprayed plastics, etc., on any metallic surface.

An important feature is the single contact head, in which the contacting electrode is a highly polished sphere, giving spot readings unaffected by moderate curvature of the surface under test. Readings are independent of alloy or thickness of base metal.

Various measuring heads, incorporating a built-in standardizing capacitor, can be supplied to suit individual applications, these heads being connected to the indicating unit by a flexible cable. The meter can be scaled for various thickness ranges up to about 0.05 in. of coating.

Once the meter is calibrated for a specific application, it is only necessary to contact the surface of the coating with the electrode, and press the button in the head to obtain a reading. There is little danger of damage to the most delicate finish, as the polished electrode only needs to contact it without pressure.

With specially designed heads, the meter may also be used to measure the thickness of rubber or plastic sheet during manufacture, or other measurement applications which may be resolved into small capacitance changes.

M.P.J. Gauge and Tool Co. Ltd.,
Hansons Bridge Road,
Birmingham 24.

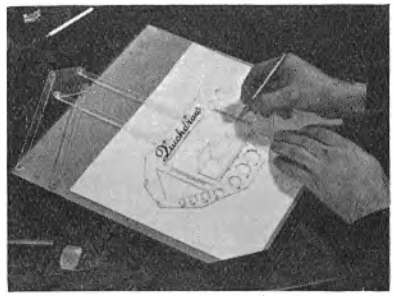
The "Quickdraw"

(Illustrated top right)

THE "Quickdraw" is a new device which enables persons with little previous experience of drawing to make sketches, plans, or other outlines rapidly and accurately to scale.

The pantograph, to which the template is necessarily attached, ensures that the principal lines are vertical or horizontal, whatever the position of the template. The shaped edges of the template enable lines to be drawn at any of the principal angles, including those for isometric "perspective" sketches. Circles of various sizes may be drawn quickly and accurately, and small perforations in the instrument enable a number of equally-spaced horizontal lines to be drawn at intervals of ⅛ in. The scales are in fractions of an inch and in millimetres. The template and pantograph are made of plastic and are mounted on a light board contained in a portfolio. One drawing-pin only is required to secure the paper in position. The largest size of drawing possible is 13 by 10 in., but, of course, smaller reproductions are easily effected.

The Quickdraw Co.,
127, Gunnersbury Avenue,
London, W.3.



New Hearing Aid with Automatic Volume Compression

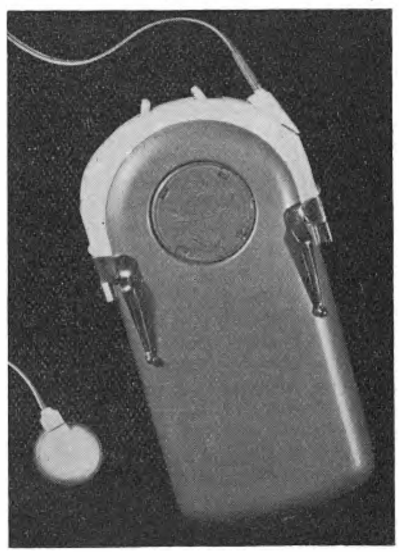
(Illustrated right)

SUPPLIERS of hearing aids have long been aware that there are many cases of deafness where the deaf person is able to tolerate only a lower maximum sound intensity than can a person of normal hearing. It is desirable in such cases to limit this maximum sound intensity to avoid discomfort, and this has been done by means of Automatic Volume Compression (A.V.C.). The same principle of A.V.C. can also be utilized to avoid over-loading of the output valve, the distortion of which will lead to a loss of hearing even for deaf persons who are able to tolerate the full output of the aid. In this latter case the A.V.C. is arranged to go into action at an output level just below the maximum undistorted output obtainable from the output valve. In both cases the A.V.C. serves the purpose of retaining the whole of the sound output within the dynamic range of the hearing aid and of its user.

The design of a hearing aid in which A.V.C. is to be really effective presents interesting electro-acoustic problems in addition to the design of the compression system itself. For instance, the microphone response must be flat, as otherwise there would be triggering of the A.V.C. system by relatively quiet sounds occurring at a frequency coincident with a microphone resonant peak. Additionally, the earphone must also have a flat frequency response free from resonant peaks, if an A.V.C. system operating only from the electrical input to it is to produce an acoustic output not exceeding a pre-set pressure level.

In the Model KA Amplivox a flat response microphone and earphone are used with A.V.C. applied at two alternative levels selected by means of a switch. In the minimum A.V.C. position output is limited at about 5 db below the maximum possible output, thus avoiding distortion due to overloading the output valve. In the maximum A.V.C. position the output is limited at 10 db lower still.

Amplivox Ltd.,
2 Bentinck Street,
London, W.1.



BOOK

The Structure of Matter

By Francis Owen Rice and Edward Telles. 361 pp. Chapman and Hall, 1949. Price 30s. net.

PROFESSOR Rice and Professor Teller have here set out to explain in physical terms the fundamentals of quantum theory and the way in which it is used to explain the structure of matter. Mathematical analysis has been kept to a minimum with the object of emphasizing the qualitative concepts which are set out. There is no doubt that the authors have succeeded in producing a most valuable book. Most university courses on quantum theory are designed by the established mathematical physicist to teach the next generation of mathematical physicists methods and technique rather than the applications. Very few systems can be treated quantitatively by the beginner so that the course ends with just those few systems treated. Even with these few systems the essentially mathematical treatment far too frequently leaves students capable—temporarily—of producing the quantitative analysis without understanding the qualitative meaning of their work. For experimentalists such a reversal of their normal order of things is disastrous. A section of Quantum theory half-learnt in this way is perhaps not useless, but its usefulness is like that of Latin to the non-classical scholar; it develops a dogged persistence in time of trial which must be of moral value, and develops a growth of mental capacity which might some time be applicable to something real, but is of little immediate advantage.

To such a problem this book provides an exceedingly satisfactory answer. The fundamentals of quantum mechanics are first dealt with, in a beautifully clear way. These, with their applications to the hydrogen atom, take only two chapters out of thirteen. The explanation of the periodic system and the energy-relations of electrons in complex atoms leads on to the interaction of electrons in different atoms and the structure of molecules. A feature of the book of great significance, in view of the increasing amount of work which is now being carried out on the solid state, is the excellent series of chapters on dielectric properties, Van der Waals forces, chemical bonds and the forces acting in solids.

The discussion of spectra is not limited to those of hydrogen-like atoms, but deals also with molecular spectra and the excitation of crystal lattices and of conductors.

The book ends with a most lively and stimulating discussion of the constitution and properties of nuclei, the state of matter and energy-production in stars and of various theories of the origin of elements.

In so widely-ranging a book there are bound to be one or two statements which are still very debatable. Such are the suggestion of the linear accelerator as the best source of particles with energies of 10^9eV or more, or that meteorites probably originate from broken-up planets. A reference to the "equations of state" in the index would be helpful to those who are used to the gas equations, Van der Waals' equation, etc., but have forgotten the more general term.

Altogether the book can be most heartily recommended, and forms an outstandingly successful introductory book to the Structure of Matter Series.

J. H. FREMLIN.

An Introduction to the Theory and Design of Electric Wave Filters

By F. Scowen. 188 pp. 71 diagrams. 1 table. (Second Edition revised). Published by Chapman & Hall. Price 18s.

IN this new revised edition the author has added a chapter which provides the first account in an English text book of the insertion-loss methods of design developed in America and Germany.

The "reference filter" design technique is described in detail and a particular filter meeting certain specified performance requirements is designed as an illustrative example. No account is given of the theoretical foundation or the insertion loss method of design, but as the book is intended primarily for the practical filter designer this is not a serious criticism. It is a pity, however, that the author has not dealt with the elliptic function design process, which, while necessarily more restricted than the "reference filter" method, can be used in a surprisingly large number of practical cases and results in a considerable saving of design time and labour.

Reading again over that part of the book which appeared in the first edition one is bound to regret the omission of those excellent filter design tables which first appeared in Shea's book; many authors since Shea have used them and the present one would be advised to follow their example.

The publishers, on the inside of the front dust cover refer to "Darling-

REVIEWS

ton's invention—loss method of filter design"; this should be "insertion-loss" of course.

On page 151 in the section dealing with the measurement and adjustment of inductors, the author states that an overall adjustment and bridge accuracy of ± 0.7 per cent can be obtained if care is taken in screening and earthing; surely this is a very pessimistic figure. Most manufacturers of filter coils work regularly to a specification tolerance of ± 0.5 per cent and would expect to measure inductance to a tolerance less than ± 0.1 per cent.

Apart from these points, however, the author is to be congratulated on presenting a book which is up to date and is probably the best of its kind available to the specialist filter designer.

L. I. FARREN.

"Electron Tubes" Vols. I and II.

Published by R.C.A. Reviews, U.S.A. 1949 475 pp. and 453 pp. Price $2.50 each.

THESE two books, which form part of the R.C.A. Technical Book Series, collect together an extensive number of papers which have been published since 1935 by the research workers of the Radio Corporation of America. Vol. 1 covers the period 1935-1941 and contains fifteen main papers; the second volume contains reprints of twenty-one main papers, published in the period 1942-1948. In addition many papers are summarized, and bibliographies of articles by R.C.A. workers are appended.

These volumes are not intended to be handbooks of electron tube data but deal in the main with what are regarded as more fundamental problems. The contents are divided into four sections, headed, general, transmitting, receiving and special. The general section comprises papers on space charge effects and electron ballistics, while the special section is devoted mainly to photo-electric devices, excluding, however, television pick-up tubes.

Most readers of this journal will be familiar with the original articles, many of which have appeared in the "Proceedings of the Institute of Radio Engineers" or in the "R.C.A. Review," and it is probably superfluous to remark that they will greatly appreciate their presentation in book form; especially since the authors are so often able to present the results of their abstruse discoveries and considerations in a particularly lucid way.

When so much of interest is included, such as the classic papers by Salzberg and Haeff on "Space Charge in the Grid-Anode Region of Vacuum Tubes," or the series by Thompson, North and Harris on current fluctuations, it is perhaps too much to wish that the complete papers by Ferris and North could also have been published along with these, in Vol. 1. The reviewer proposes to criticise one article, by O. H. Schade on "Beam Power Tubes." The theory given is not capable of explaining the principle facts; which is the reason why it was already abandoned in 1934 by the English inventors who carried out "some development work on the replacement of the suppressor grid by space charge."

In conclusion one would recommend that the invaluable record of R.C.A. activities and achievements presented in these two volumes should be supplemented and compared with similar compilations on the Electronic Art now being published in Europe.

S. RODDA.

Principles and Practice of Radar

By H. E. Penrose. 692 pp. 478 figures. George Newnes, Ltd., 1949. Price 42s.

RADAR, based on the fundamental principles of radio, owes much of its development to the war work of the television engineers and to physicists engaged in research on microwaves.

Twenty-six chapters and four appendices cover the wide field of radar. Sections of the book have been devoted to detailed consideration of pulsed and pulse forming circuits at high and low levels, timing circuits, gate circuits and applications of microwave technique to transmitters, receivers, feeder and aerial systems.

In the microwave field the resonant magnetron and its associated units have received special attention in a long chapter; while the many problems arising from the employment of ultra-high frequencies in transmitting and receiving circuits have received suitable consideration.

The book aims to demonstrate the fundamental principles of applying radar to various requirements, rather than describing a single one of the wide variety of possible combinations in detail.

The four appendices deal respectively with transmission line theory, waveguides, cavity resonators, and radar test equipment. A full index is included, and the figures are clearly produced.

BOOK REVIEWS (Continued)

The Quarterly Journal of Mechanics and Applied Mathematics

Volume II, Part 4 (December 1949) 480 pp. Oxford University Press. Price 12s. 6d.

TWO papers in this issue may be of special interest to readers. The first is by Dr. Love on the exact solution for the electrostatic field between two circular disks. It is, however, very mathematical and is concerned with the validity of earlier work. It arose from a discussion on the oil-drop method of measuring the electronic charge, as used by Dr. Hopper of Melbourne.

The other is by Dr. Booth, on the numerical solution of non-linear simultaneous equations, by the method of "steepest descents." Such equations are only too familiar, and are usually avoided if possible. This is the second of what will probably develop into a series of papers from Birkbeck College, London, due to the necessity of devising suitable numerical techniques for the electronic computer which is being constructed.

G. J. KYNCH.

Questions and Answers on Electronics

By E. Molloy. 128 pp. 65 figures. George Newnes, Ltd., 1949. Price 5s.

THIS small book presents in the form of questions and answers basic facts on electronic theory and the principles involved in the construction of various types of electronic equipment and instruments.

The wide range of subjects covered includes: radar, electron microscopes, high frequency heating, cathode ray tubes, television and atomic structure. As the book is small the subjects are only briefly dealt with, but the information is well presented and should appeal to those people with only an elementary knowledge of electronics.

1949/50 F.B.I. Register

22nd Edition. 807 pp. Published for the Federation of British Industries by Kelly's Directories, Ltd., and Iliffe & Sons Ltd. Price 42s.

THE 1949/50 F.B.I. Register lists nearly 6,000 firms and their products, and includes much information of interest to overseas buyers. Especially useful is the Products and Services Section in which the manufacturers are listed under more than 5,000 headings facilitating the identification of supply sources. There are reference facilities in French and Spanish, while the classification has been improved and greatly extended.

Fourier Methods

By Phillip Franklin, Ph.D. 290 pp. 77 figures. 1st Edition. McGraw Hill, Ltd., 1949. Price 34s.

THIS useful book is about the solution of partial differential equations and gives full details of Fourier series, Laplace transforms and the other techniques now used to solve them. It is therefore quite valuable. The methods are illustrated by examples, mainly from circuit theory, which are examined in considerable detail, and at the end of each section there are a large number of extra examples for the reader to tackle.

Whether a book on a special subject should be self-contained and include all the advanced mathematics required, depends a great deal on the reader as well as the subject. The reader, or the library, who buys one book of this type is often inclined to theory and has several more. If that happens, the inclusion of general sections merely means, for an increase in cost, a number of overlapping accounts which tend to be scrappy.

This book claims to be self-contained and the author has included the extra material very skilfully. The mathematics required is mostly covered by a simple post-H.S.C. course, and the rest, by a careful choice of examples, is included in the text so as to form part of the development of the subject.

It is a great pity that so few books of this type are published in this country. There are many people who write well, who see the need and have the teaching experience, but so few who can find the time.

G. J. KYNCH.

Fundamentals of Vacuum Tubes

By Austin V. Eastman. 644 pp. Third Edition. McGraw Hill Book Co., Inc. Price 47s. 6d.

THE treatment in this text book is designed to rest midway between the purely descriptive, and the purely mathematical. Of the two parts, the first covers the theory and the second some of the commoner applications of the vacuum tube. No attempt is made to deal with any of the modern ultra high frequency tubes.

A fourth year National Certificate student, or an engineering undergraduate will have little difficulty in using this book. Most chapters conclude with a group of questions, to which a separate answer book is planned. The general presentation is lavish, numerous illustrations (some of little value) being included. The price is such that the ordinary student will not be able to afford this volume: but there are several books of equal merit on the market at a more modest price.

K. G. LOCKYER.

Radio Operator's Q. and A. Manual

By Milton Kaufman. 608 pp. 193 figures. John F. Rider Publisher Inc., 1949. Price 6 dollars.

THIS book, by an instructor in radio operating at America's R.C.A. Institutes, gives a list of questions and answers representative of the Federal Communications Commission's examinations for obtaining commercial licences in radiotelegraphy, radiotelephony, and amateur radio operation. The answers are supplemented by a discussion on the answer given, and illustrations are incorporated to make difficult technicalities clear.

Five useful appendices are included, which are based upon information from the Government Study Guide and F.C.C. releases. Of particular interest are the ones on Small Vessel Direction Finders and Automatic Alarms.

For a quick review of essential theory or as a refresher this volume would prove useful.

"Wireless World" Diary 1951

80 reference pages, with the usual diary pages of a week to an opening. Iliffe & Sons, Ltd. October, 1950. In Morocco leather, price 5s. 6d., in rexine, price 3s. 8d.

THE reference pages of this diary contain information mainly of a technical nature and data on design, maintenance and the use of radio equipment. A useful summary of the existing regulations affecting the wireless user, including the new Wireless Telegraphy Act, is included.

Other interesting features include a large selection of formulae, abacs for easy graphical estimation of such things as coil windings and circuit constants, lists of unit abbreviations, definitions and classifications, and a number of circuit diagrams. The valve base tables give connexions for more than 500 valves in convenient form.

NOTES FROM THE INDUSTRY

R.S.G.B. Amateur Radio Exhibition

The Fourth Annual Amateur Radio Exhibition organized by the Radio Society of Great Britain is to be held at the Royal Hotel, Woburn Place, London, W.C.1, from Wednesday, November 22 to Saturday, November 25.

The exhibition will be opened at 2.30 p.m. on November 22 by Mr. Hugh Pocock, M.I.E.E.

Admission will be by catalogue obtainable at the door (price 1s.) or on application from the Radio Society of Great Britain, New Ruskin House, Little Russell Street, London, W.C.1 (price 1s. 3d. post free).

The exhibition will be open at 11 a.m. on November 23, 24 and 25 and will close at 9 p.m. each evening.

Industry and Technical Colleges

As an acknowledgment of the part played by technical colleges in supplying trained and experienced technologists, Messrs. Acheson Colloids, Ltd., have made a contribution to equipping a chemical research laboratory at the Plymouth and Devonport Technical College. The company's works have been established at Plymouth for 40 years.

N.P.L. Annual Report for 1949

Details of the work of the National Physical Laboratory, are given in the Annual Report for the year 1949. The Report has been published for the D.S.I.R. by H.M.S.O. price 1s. 9d. (45 cents U.S.A.), by post 1s. 11d. It is divided into reports from each of the divisions of the Laboratory.

Identification Scheme for Television Aerials

A uniform scheme of catalogue suffix numbers for television aerials to indicate clearly the channel for which an aerial is suitable has been agreed upon by the manufacturers.

The channel number will also be marked clearly on aerial packages. Some manufacturers will use a colour code as tabulated below.

The scheme has been initiated by the Aerials Panel of the Radio and Electronic Component Manufacturers' Federation in the hope that it will simplify storekeeping and records, not only for manufacturers but for wholesalers and retailers.

Channel number	Location	Frequencies Mc/s Vision	Sound	Catalogue Suffix	Optional Colour Code
1	London	45.00	41.50	/1	Yellow
2	N. England	51.75	48.25	/2	L. Blue
3		56.75	53.25	/3	Red
4	Midlands	61.75	58.25	/4	Green
5		66.25	63.25	/5	D. Blue

Brochure of Instruments and Accessories for Radio-Isotope Applications

The Scientific Instrument Manufacturers' Association, in collaboration with the A.E.R.E. Harwell, has produced a brochure on the instruments and accessories for radio-isotope applications.

It is intended to summarize the equipment available from British sources for isotope techniques in medicine, research and industry.

Copies are available free on application to the S.I.M.A. offices, 17 Princes Gate, London, S.W.7.

Tender for a Television Installation in Australia

The United Kingdom Trade Commissioner at Melbourne has reported that the Postmaster-General's Department are calling for tenders for the supply, delivery and certain technical services of the television installation at Sydney.

Copies of the tender documents (No.C.6423) can be obtained from the Chief Supply Officer. Supply Branch, Room 406. Australia House, Strand, London, W.C.2.

The closing date for the receipt of tenders is November 21, and they should be addressed to the Deputy Director, Posts and Telegraphs, Melbourne, C.1.

An Important Centenary

We regret that an error over which we had no control occurred in the Leader for the October issue of the journal. In the fifth paragraph *The Times* mentioned should of course be that of August 31, 1850, not 1950.

Accurate Triggering of Multi-vibrators

We regret that an error occurred in the last paragraph of this article on page 377 of the September issue. The seventh line should read ". . . of the pair. The second pulse then fires the multi-".

PUBLICATIONS RECEIVED

Tufnol Technical Information Catalogue

This catalogue consists of technical data of interest to design-engineers and other technicians concerned with Tufnol materials. *Tufnol Ltd., Perry Barr, Birmingham 22B.*

Taylor Electrical Measuring Instruments

Price List of Electrical Measuring Instruments

Two booklets giving complete details of the well known Taylor meters. *Taylor Electrical Instruments, Ltd., Montrose Avenue, Slough, Bucks.*

Electronic Control, Information Sheets

These leaflets describe how electronic control is carried out by *Sargrove Electronics, Ltd., of Effingham, Surrey.*

Marconi Television

This booklet describes a new range of the firm's equipment for television. *Marconi Wireless Telegraph Co., Ltd., Marconi House, Chelmsford, Essex.*

R.C.A. Catalogue of Products for Export

The catalogue contains sections on equipment for telecommunication, aviation, broadcasting, motion picture, scientific and industrial services. *R.C.A. International Division, 745 Fifth Avenue, New York 22, N.Y., U.S.A.*

Press Bulletin of B.I. Callender's

This bulletin describes the company's activities and development during the past year. *British Insulated Callender's Cables, Ltd., 21 Bloomsbury Street, London, W.C.1.*

T.C.C. Capacitors

The capacitors listed in this booklet are representative of the post-war T.C.C. range. *The Telegraph Condenser Co., Ltd., North Acton, London, W.3.*

R.C.A. Phototubes and Cathode Ray Tubes.

This booklet describes the current range of R.C.A. phototubes and cathode ray tubes, obtainable from *R.C.A. Photophone, Ltd., 36 Woodstock Grove, London, W.12,* price 1s. 6d.

The Brush Group Report and Accounts.

This is a report of the progress of the Brush Electrical Engineering Co., Ltd., and its subsidiary companies during 1949, and includes the company's statement of accounts. *Brush Electrical Engineering Co., Ltd., Duke's Court, 32 Duke Street, St. James's S.W.1.*

MEETINGS THIS MONTH

INSTITUTION OF ELECTRICAL ENGINEERS

All meetings, unless otherwise stated, are held at the Institution of Electrical Engineers, Savoy Place, London, W.C.2, at 5.30 p.m.

Date: November 2.
Report: The Education and Training of Electrical Technicians.
Presented by: Sir A. P. M. Fleming, C.B.E., D.Eng.

Measurements Section

Date: November 14.
Lecture: The Determination of Time and Frequency.
By: H. M. Smith, B.Sc.
(Joint Meeting with the Radio Section.)

Date: November 28. Time: 3.30 p.m. and 5.30 p.m.
Symposium on: Radiation Monitoring Apparatus.
(Joint Meeting with Radio Section.)

Radio Section

Date: November 8.
Lecture: Low-Frequency Radio-Wave Propagation by the Ionosphere, with particular reference to Long-Distance Navigation.
By: C. Williams, B.Sc.

Date: November 14.
For full particulars see Measurements Section.

Date: November 20.
Lecture: The Nervous System as a Communication Network.
By: J. A. V. Bates, M.A., M.B., B.Chir.

Date: November 28.
For full particulars see Measurements Section.

District Meetings

Date: November 3. Time: 7.30 p.m.
Held at: The New Inn, Sandling Road, Maidstone.
Lecture: The Measurement of Light and Colour.
By: G. T. Winch.

Date: November 8. Time: 7 p.m.
Held at: Electricity Showrooms, Southern Electricity Board, 37, George Street, Oxford.
Lecture: Radio.
By: W. H. Lee.

Cambridge Radio Group

Date: November 14. Time: 8.15 p.m.
Held at: Cavendish Laboratory, Cambridge.
Address by the Chairman of the Radio Section.
By: C. F. Booth, O.B.E.

Date: November 27.
Lecture: Frequency Modulated Broadcasting.
By: Group Captain W. P. Wilson, C.B.E., M.Sc.(Eng.).
(Joint Meeting with Cambridge University Wireless Society.)

Mersey and North Wales Centre

Date: November 13. Time: 6.30 p.m.
Held at: The Liverpool Royal Institution, Colquitt Street, Liverpool.
Lecture: Crystal Diodes.
By: R. W. Douglas, B.Sc., and E. G. James, Ph.D.
Lecture: Crystal Triodes.
By: T. R. Scott, D.F.C., B.Sc.

North-Western Radio Group

Date: November 29. Time: 6.30 p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Lecture: Crystal Diodes.
By: R. W. Douglas, B.Sc., and E. G. James, Ph.D.
Lecture: Crystal Triodes.
By: T. R. Scott, D.F.C., B.Sc.

Scottish Centre

Date: November 8. Time: 7 p.m.
Held at: The Heriot Watt College, Edinburgh.
Lecture: A Review of Some Television Pick-up Tubes.
By: J. D. McGee, M.Sc., Ph.D.

South Midland Radio Group

Date: November 27. Time: 6 p.m.
Held at: James Watt Memorial Institute, Gt. Charles Street, Birmingham.
Lecture: Analogies.
By: Professor M. G. Say, Ph.D., M.Sc.

BRITISH INSTITUTION OF RADIO ENGINEERS

Date: November 20. Time: 6.30 p.m.
Held at: London School of Hygiene and Tropical Medicine, Gower Street, W.C.1.
Lecture: Ultrasonic Generators for High Power.
By: B. E. Noltingk, Ph.D.

Merseyside Section

Date: November 2. Time: 6.30 p.m.
Held at: Electricity Board, Service Centre, Whitechapel, Liverpool.
Lecture: Propagation of Metric Waves Beyond Optical Range.
By: D. W. Heightman.

North-Eastern Section

Date: November 8. Time: 6 p.m.
Held at: Neville Hall, Westage Road, Newcastle-on-Tyne.
Lecture: Radio Navigational Aids.
By: J. C. Martin, M.A.

West Midlands Section

Date: November 22. Time: 7 p.m.
Discussion: The Quality of Electrical Reproduction and its effect on Receiver Design.

North-Western Section

Date: November 2. Time: 6.45 p.m.
Held at: College of Technology, Manchester.
Lecture: Communication Cables.
By: W. R. Cooper.

SOCIETY OF INSTRUMENT TECHNOLOGY

Midland Section

Date: November 17. Time: 7 p.m.
Held at: Arden Hotel, New Street, Birmingham.
Lecture: The Fundamentals of Electrical Measurement.
By: L. Hartshorn, D.Sc., A.R.C.S.

INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Date: November 6. Time: 5 p.m.
Held at: Institution of Electrical Engineers, Savoy Place, W.C.2.
Lecture: The Anglo-Continental Telephone Service.
By: J. Rhodes, B.Sc., A.M.I.E.E.

BRITISH SOUND RECORDING ASSOCIATION

Date: November 24. Time: 7 p.m.
Held at: Royal Society of Arts, John Adam Street, London, W.C.2.
Lecture: Some Features in Amplifier Design for Recording Equipment.
By: H. D. McD. Ellis, M.A., M.I.E.E.

TELEVISION SOCIETY

Date: November 2.
Held at: C.E.A., 164, Shaftesbury Ave., W.C.2.
Lecture: Music for Television.
By: Eric Robinson and Michael Mills.

Date: November 24.
Held at: C.E.A., 164, Shaftesbury Ave., W.C.2.
Lecture: Scan Linearization by Negative Feedback.
By: A. W. Keen.

Engineering Group

Date: November 10.
Held at: C.E.A., 164, Shaftesbury Ave., W.C.2.
Lecture: Design of a Combined Radio-TV Receiver.
By: A. B. Bamford.

SECRETARIES OF ASSOCIATIONS

INSTITUTION OF ELECTRICAL ENGINEERS

The Secretary, Institution of Electrical Engineers, Savoy Place, W.C.2.

Cambridge Radio Group

G. E. Middleton, M.A., University Engineering Laboratory, Cambridge.

North-Eastern Radio and Measurements Group

G. A. Kysh (Asst. Sec.), Carliol House, Newcastle-on-Tyne, 1.

North-Western Radio Group

A. L. Green (Asst. Sec.), 244, Brantingham Road, Chorlton-cum-Hardy, Manchester, 21.

South Midland Radio Group

W. H. Brent, B.Sc., Regional Director's Office, Midlands Region (G.P.O.), Civic House, Great Charles Street, Birmingham, 3.

BRITISH INSTITUTION OF RADIO ENGINEERS

The General Secretary, 9, Bedford Square, London, W.C.1.

West Midlands Section

R. A. Lampitt, A.M.Brit.I.R.E., 20, Northfield Grove, Merry Hill, Wolverhampton.

INC. RADIO SOCIETY OF GT. BRITAIN

General Secretary, New Ruskin House, Little Russell Street, W.C.1.

BRITISH SOUND RECORDING ASSOCIATION

Richard W. Lowden, "Wayford," Napoleon Avenue, Farnborough, Hants.

TELEVISION SOCIETY

Lecture Secretary: T. M. Lance, 35, Albermarle Road, Beckenham, Kent.

Engineering Group

G. T. Clack, 10, Tantallon Road, Balham, London, S.W.12.

RADAR ASSOCIATION

The Secretary, 83, Portland Place, London, W.1.

INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

W. H. Fox, A.M.I.E.E., Engineer-in-Chief's Office (T. P. Branch), Alder House, E.C.1.

INSTITUTION OF ELECTRONICS

Lecture Sec.: W. Summer, 31, Beech Road, Bournville, Birmingham, 30.

North-West Branch

L. F. Berry, 105, Birch Avenue, Chadderton, Lancs.

SOCIETY OF RELAY ENGINEERS

T. H. Hall, M.Brit.I.R.E., 23, Dalkeith Place, Kettering, Northants.

SOCIETY OF INSTRUMENT TECHNOLOGY

L. B. Lambert, 55, Tudor Gardens, London, W.3.