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Commentary

THE year which has just ended has been a prosperous one for the electronic industry, as indeed, it has for most other industries and, while there have perhaps been no epoch making scientific discoveries or announcements, it has been one of solid achievement and has seen the culmination of several important long-term projects. Among the latter the two outstanding achievements have been the completion and opening of the Trans-Atlantic Telephone Cable and of Calder Hall. While the latter is of considerable importance and scientific interest, the telephone cable is of more immediate interest to the electronic engineer. It is, in fact, a great tribute to the courage of those who planned it and a fine example of international co-operation. Whether or not the confidence of its sponsors will be justified, time alone will tell for, to be a success the cable and its submerged repeaters must have a life of many years. To those who have designed it, it must be a somewhat disquieting thought that by the time the cable has proved itself, probably most of the materials and techniques used in its manufacture will have become, to say the least, obsolescent.

Another feature of the past year has been the steady progress of the electronic control of machine tools. Most of the major companies have carried out considerable research and development work on this project and there are now in existence, in this country, several different methods of control, some of them already being beyond the development stage and actually in use on the toolroom floor. Most of the effort so far has been concentrated on the control of precision boring and milling machines and it would seem that for some time to come this type of control will be most profitably applied to the production of press-tools, dies and to complex components which are required in relatively small numbers, as for example aircraft equipment and the integrated type of waveguide component. For the large scale mass production type of work the more conventional mechanical methods are, in general, superior and are likely to remain so for some time.

The general purpose computer field has also seen considerable activity. Several new computers have made their appearance while one or two other firms have announced their intention of entering this market. Some firms have set up comparatively large production departments to manufacture these machines. The modern computer is a well engineered product and is one of the finest examples of electronic engineering; the most notable features being the reduction of bulk and ease of accessibility compared with its prototype predecessors. Presumably the companies which are spending large amounts both of capital and manpower on the research, development and production of these machines are well satisfied that there is an adequate market for this type of computer. Certainly the potential of these machines is very great and their integration into the industrial and commercial worlds will be of ultimate benefit to the country, yet it does seem that the technique of producing these machines is perhaps outstripping the research into their economic application and that intensive research into the best use of the computer in industry and commerce would pay handsome dividends.

On the entertainment side it has been a year of consolidation rather than of spectacular progress. Independent television has ceased to be a matter for argument and is now an accepted part of the country's life. The plans for country wide coverage of both the ITA and the BBC's v.h.f., f.m. broadcasting are progressing steadily. The most notable occurrence in the BBC's television plans was the move from Alexandra Palace to Crystal Palace, where they now have one of the world's finest television transmitting stations. On the progress of colour television there is little that can be said. Both the BBC and industry have been, and still are, conducting experiments to determine the most suitable standards and system to adopt for a public service, but there is, as yet, no agreement on this matter. In any case it is unlikely that colour television will be introduced until the country's economic position becomes more stabilized, which may well be several years hence.

The attendances at the various exhibitions have shown that interest in all branches of the industry continues to increase and it has, all in all, been a very successful and satisfying year. It is to be hoped that the events which have clouded the last few months of 1956 will be satisfactorily resolved and that the coming year will bear witness to further progress in the peaceful uses of science and the application of electronic engineering to an increase in our productivity with the ensuing betterment of our economic position and the raising of our standards of living.

The Junction Transistor as a Computing Element

(Part 1)

By E. Wolfendale*, B.Sc., A.M.I.E.E., L. P. Morgan*, B.A., and W. L. Stephenson*, B.Sc.

This paper is intended to demonstrate the possibilities and limitations of junction transistors when used in the various types of circuit which make up a computer.

With this aim in mind the paper starts with a description of the small signal and transient characteristics of the transistor. It then shows how these characteristics are used in the design of the basic circuits, and ends with a description of the performance of a number of examples of computer elements made up with junction transistors.

A LARGE number of articles have been published describing particular junction transistor computing circuits, but there has been no comprehensive article showing how the junction transistor characteristics may be generally applied to the design of such circuits. This paper is written to fill what is felt by the authors to be a need for a general paper describing the application of junction transistors to computing circuits and their limitations in those circuits.

The paper first describes the transistor characteristics which are used for the design of computing circuits. It then shows how these characteristics are used for the design of individual circuits and how they limit the performance of these circuits and finally gives some examples of the practical use of the circuits.

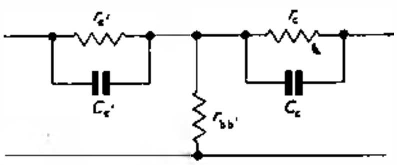

Fig. 1. The equivalent circuit of the cut-off state

The Junction Transistor Characteristics

The performance of a junction transistor in computing circuits is defined by three states, the cut-off state, the bottomed state and the transient state between cut-off and the bottomed state.

The Cut-off State

The transistor is cut off by biasing the emitter diode in the reverse direction. The equivalent circuit for this state is thus two diodes biased in the reverse direction as shown in Fig. 1. r_e' and r_c are the reverse resistances of the emitter and collector diodes respectively, and C_e' and C_c are the reverse capacitances.

The circuit is shown in the grounded base connexion. As r_c and r_e' are very much greater than $r_{bb'}$ the input impedance is very high and approximately equal to r_e' shunted by C_e'. A simple re-arrangement of the circuit shows that in the grounded emitter and grounded collector configurations the input impedance is approximately the parallel combination of r_e' and r_c shunted by the sum of C_e' and C_c.

The equivalent circuit does not show the reverse saturation currents of the two diodes. These have to be taken into account as they determine the actual cut-off current of the circuit. Ebers and Moll[1] have derived expressions for the currents flowing in the cut-off state

$$I_c = \frac{I_{eo}(1-\alpha_N)}{1-\alpha_N\alpha_I} \quad \ldots\ldots\ldots\ldots\ldots\ldots (1)$$

$$I_e = \frac{I_{co}(1-\alpha_I)}{1-\alpha_N\alpha_I} \quad \ldots\ldots\ldots\ldots\ldots\ldots (2)$$

where I_{eo} is the saturation current of the emitter diode with zero collector current,

I_{co} is the saturation current of the collector diode with zero emitter current,

α_N is the normal current gain of the transistor,

α_I is the current gain with the emitter and collector potentials reversed.

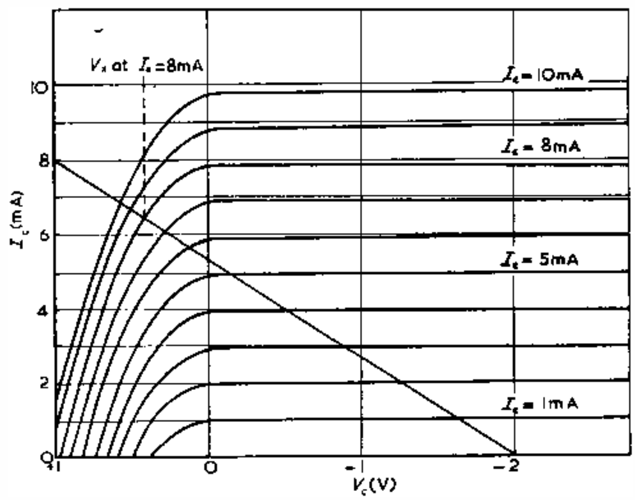

Fig. 2. The grounded base static characteristics and the bottomed voltage

These expressions may be used for an accurate analysis of the currents in the cut-off state, but for a large number of applications it is sufficient to approximate $I_e = 0$ and $I_c = I_{co}$.

Whichever way the cut-off currents are derived it should be remembered that they are an exponential function of temperature. As an approximate guide the cut-off current doubles for every 9°C rise of temperature.

The Bottomed State

The bottomed collector voltage may be obtained by drawing a load line on the static characteristics as shown in Fig. 2. The knee of the characteristics is not sharply defined and for some design work it is necessary to use a more accurate method.

* Mullard Research Laboratories

Under normal operating conditions the holes diffusing across the base are rapidly swept across the collector junction by the field across the depletion layer. In the bottomed state the field across the depletion layer is reduced by the voltage drop across the external load resistance, but holes continue to flow across the collector junction to supply the current in the load.

In order to obtain the bottomed state many more holes are emitted by the emitter than are required to supply the load current, and the additional holes crossing the collector junction cause the junction to be biased in the forward direction. The additional holes then recross the junction into the base.

The collector current is given by the expression

$$I_c = \alpha I_e - I_{cd} \quad \text{.................} \quad (3)$$

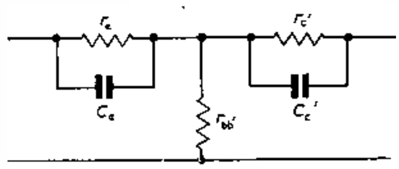

Fig. 3. The equivalent circuit of the bottomed state

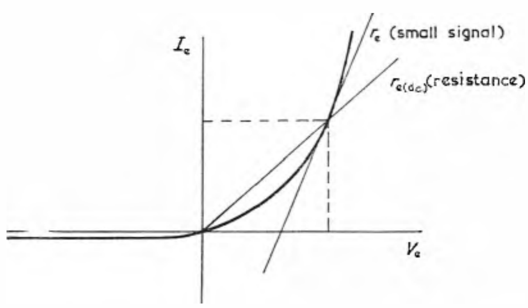

Fig. 4. Resistance characteristics of the emitter diode

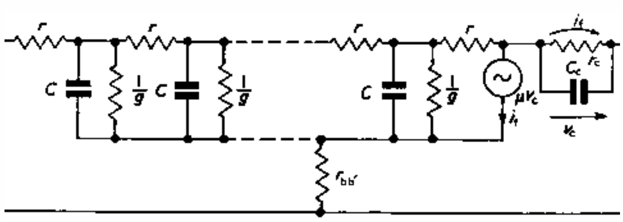

Fig. 5. The junction transistor equivalent circuit

where I_{cd} is the forward current across the collector junction. I_e is known, α is known and I_c is known, and the bottomed state can be accurately defined from I_{cd} and the current voltage characteristic of the collector junction.

The equivalent circuit of the bottomed state is shown in Fig. 3.

r_c' and C_c' are the forward resistance and capacitance of the collector diode.

For d.c. analysis, where C_c' and C_e may be neglected, the values of r_e and r_c' are not the small signal values, but the d.c. or chord resistances of the diodes as shown in Fig. 4.

The Transient State

Transient analysis is required for both linear pulse amplifiers and non-linear switching circuits. There are two approaches to this analysis, one from the transistor equivalent circuit and the other from an examination of the minority carrier storage generally known in pnp transistors.

The Transistor Equivalent Circuit

A good approximation to the transistor equivalent circuit is shown in Fig. 5; where $r_{bb'}$ is the ohmic base resistance; μv_c is the Early feedback generator[3]; r, c and g make up the emitter transmission line; and r_c and c_c are the collector resistance and capacitance.

The ohmic base resistance $r_{bb'}$ may be regarded as a constant except when the transistor is used for large pulse currents. Then the value of $r_{bb'}$ may be reduced by modulation of the base conductivity by the additional carriers in the base.

The Early feedback generator increases the input impedance of the transistor at low frequencies. Early has shown that at low frequencies its action may be represented by an additional base resistance in the equivalent circuit. This has led to some confusion as in fact there is no additional base resistance in the transistor, but a feedback caused by base width modulation due to the collector voltage as described by Early. The value of μ is inversely proportional to the square root of the collector voltage.

The emitter transmission line represents the diffusion of carriers across the base and determines the phase and amplitude of the current gain of the transistor. Over a wide

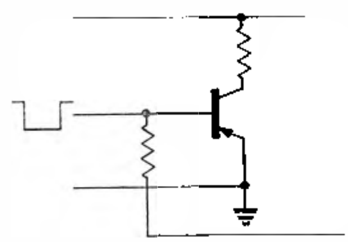

Fig. 6. The circuit used to show hole storage effects

frequency range this line may be approximated to a parallel combination of resistance and capacitance. These are generally known as the emitter resistance and capacitance. The emitter resistance is inversely proportional to the emitter current, and the emitter capacitance is directly proportional to the emitter current.

The collector resistance and capacitance represent the collector diode which is biased in the reverse direction. Almost the whole of the external voltage applied to the transistor appears across this junction and the resulting field sweeps the carriers from the base to the collector. This action is represented by the current generator across the collector resistance. The collector resistance is inversely proportional to the collector current, and the collector capacitance is inversely proportional to the square root of the collector voltage.

Analysis of this equivalent circuit is possible for small signal pulse amplifiers, but usually the hole storage approach is simpler. However, a knowledge of this circuit is invaluable when working with transistors.

Hole Storage Effects

As all transistors have a finite base width all transistors must show hole storage effects due to the time taken for holes to cross the base from emitter to collector. This transit time may be quite long in low frequency transistors.

Hole storage effects may be considered in three parts, the first occurs when the transistor is taken rapidly from the cut off to the conducting state, the second occurs when the transistor is taken rapidly from the normal conducting state to the cut off state, and the third occurs when the transistor is taken rapidly from the bottomed to the cut off state. All these conditions are best explained by reference to an example.

In Fig. 6 a transistor with a very low base resistance is connected in a grounded emitter circuit. The transistor is cut off by a small positive bias on the base. A negative going rectangular pulse is applied to the base from a low impedance source and the currents in the transistor are examined for the normal conducting state and for the bottomed state.

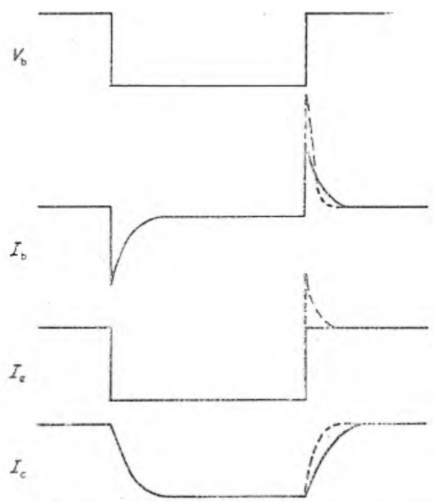

Fig. 7. Waveforms for the normal conducting state

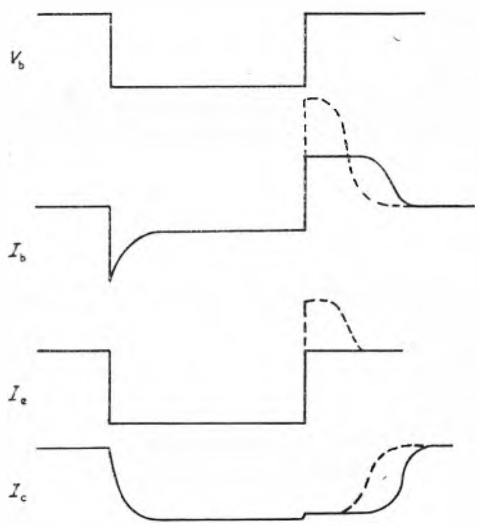

Fig. 8. Waveforms for the bottomed state

The first set of waveforms in Fig. 7 apply where the transistor is taken from cut off into normal conduction.

When the base is taken negative, base emitter current flows in the emitter diode and initially the base current equals the emitter current. Holes emitted by the emitter now diffuse across the base and collector current starts to flow. As the collector current increases so the base current falls till the steady state d.c. conditions are reached. At this point the base, emitter and collector currents are related by the d.c. gain of the transistor.

The base is now taken positive thus cutting off the emitter diode. The emitter current falls to zero, but as there are still holes diffusing across the base the collector current continues to flow and the base current now equals the collector current. The remaining holes are now collected by the collector and the collector and base currents fall together to the cut off current.

The waveforms in Fig. 8 demonstrate the second case where the transistor is taken from cut off into the bottomed state.

When the base is taken negative the conditions in the transistor are the same as in the first case, except that the final value of base current is greater than that required to maintain the collector current at its bottomed value.

The additional base current means that there are considerably more holes present in the base than those required to maintain the flow of collector current.

When the emitter diode is cut off and the emitter current ceases to flow, the additional holes in the base are collected by the collector thus maintaining the collector current at its bottomed value. When all the additional holes have been collected the collector current falls in a similar manner to the first case.

The delay before the fall of collector current may be quite long in low frequency transistors which are heavily bottomed. The maximum delay is of course limited by the hole lifetime in the base material.

If a large positive bias is applied to the base when the transistor is being cut off, the emitter diode will act as a collector. Holes present in the base near the emitter junction are collected by the emitter and instead of the emitter current falling to zero, a considerable reverse current flows in the emitter diode.

The result of the positive bias is therefore to increase the base current at switch-off and to reduce the delay in the fall of collector current. This is shown by the dotted lines in Figs. 7 and 8.

If the emitter is driven from a low impedance source the hole storage effects in the transistor are identical with those outlined for the grounded emitter connexion.

If the transistor is driven from a high impedance source the effect on the waveforms in the grounded base connexion is very small, but in the grounded emitter connexion the times of rise and fall of the collector current are increased considerably. The slower response times are entirely due to the limited base current which can be taken at the instant of switching on or off.

At the instant of switching off the emitter diode cannot cut off till the collector current has fallen to the peak value of base current available from the source. This causes a further delay in the fall of collector current.

The internal base resistance of the transistor is in series with the source impedance and may be regarded as a part of that source impedance. With the normal low frequency transistors this base resistance may be of the order of 500Ω and will delay the fall of collector current as described in the last paragraph.

Basic Circuits

An Eccles-Jordan Type Bistable Circuit

The basic circuit shown in Fig. 9 is in appearance an exact equivalent of the valve circuit but in fact various characteristics of the transistor modify both its performance and design.

The most noticeable effect of the transistor is that it provides a limitation on the switching time; this is primarily due to the hole storage phenomena discussed previously. In this circuit the feedback is to the bases of the transistors, and to obtain a fast switch-on, or-off, it is necessary to provide an initial pulse of current. This is equivalent to saying that the input impedance of the transistor is very capacitive. Therefore relatively large cross-coupling capacitors are necessary, and this results in an increase in the switch-off time. There is an optimum value of capacitance which gives equal switch-on and switch-off times.

The transistor is normally bottomed in one of its stable states in order to fix the d.c. level at the collector. This causes an increase in switch-off time and also a decrease in trigger sensitivity. The transistor can be prevented from bottoming by connecting a diode from each collector to a slightly negative potential.

The design of the potentiometer chain is complicated by the transistor being a current operated device, and allowance must be made for the base current. Also when the transistor is cut off there is a base current of I_{co} which is dependent on temperature, and care must be taken to ensure that an increase in this current with temperature will not cause the circuit to lose one of its stable states.

The usual trigger points of this circuit are the bases of the transistors and these points are low impedance.

The necessary relations for the design of the basic circuit, not using catching diodes, will be developed in three

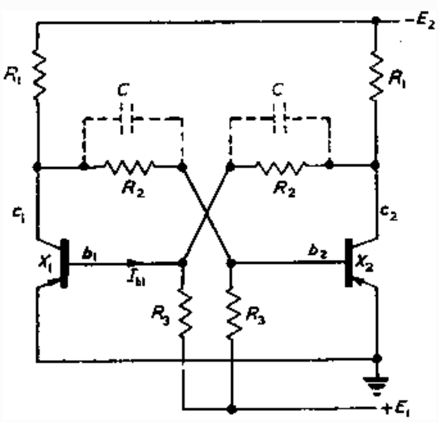

Fig. 9. Basic Eccles-Jordan bistable circuit

sections. Firstly a condition for regeneration will be derived. Then the design of the potentiometer chain to fix the stable states will be given, and the effects of different component values on trigger sensitivity will be shown.

An exact analysis of the loop gain over the whole frequency range is difficult due to the complex nature of α. However, it can be shown that at low frequencies, when all the parameters are real, the circuit is regenerative if $R_2 < \alpha' R_1$. A simple derivation of this is as follows. Consider a current I_{b1} to be flowing into the base of transistor 1

$$\therefore I_{c1} = \alpha' I_{b1}$$

Assuming $R_2 \gg R_1$, $R_2 \gg r_{in}$ and $R_3 \gg r_{in}$ which is usually so in practice.

Then $V_{c1} = \alpha' I_{b1} R_1$

$$\therefore I_{b2} \simeq \frac{\alpha' I_{b1} R_1}{R_2}$$

$\therefore$ gain from b_1 to $b_2 \simeq (\alpha' R_1 / R_2)$

$\therefore$ for the circuit to be regenerative $R_2 < \alpha' R_1$.

When the circuit is in a stable state one transistor is bottomed and the other cut off.

The potentiometer chain must be designed so that when one transistor is bottomed, the base of the other transistor is held positive for all possible values of I_{co}, and that when cut off, the base of the other is held sufficiently negative to bottom the transistor.

Fig. 10(a) shows the voltages and currents when transistor 1 is cut off.

The potential at b_1 with respect to earth is therefore given by

$$-V_{cex} - \frac{R_2}{R_2 - R_3}(E_1 + V_{cex}) - \frac{R_2 R_3}{R_2 + R_3} I_{co}$$

For the transistor to be cut off this must be greater than or equal to $V_{bey} + I_{co} r_{bb}'$, where V_{bey} is the reverse bias needed to cut off the emitter diode.

$$\therefore -V_{cex} + \frac{R_2}{R_2 + R_3}(E_1 + V_{cex}) - \frac{R_2 R_3}{R_2 + R_3} I_{co} \geqslant V_{bey} + I_{co} r_{bb}'$$

$$\frac{R_2}{R_2 + R_3}\left\{E_1 + V_{cex} + R_3 I_{co}\right\} \geqslant V_{bey} + V_{cex} + I_{co} r_{bb}'$$

$\therefore$ for transistor to be cut off

$$R_2 / R_3 \geqslant \frac{V_{bey} + V_{cex} + I_{co} r_{bb}'}{E_1 - V_{bey} - I_{co} r_{bb}' - I_{co} R_3}$$

This equation gives two results. It gives the ratio R_2/R_3 necessary to ensure that the transistor is cut off, and also it gives a maximum value of R_3 for a particular value of I_{co}, since if $I_{co} R_3 > E_1 - V_{bey} - I_{co} r_{bb}'$, R_2/R_3 is negative indicating that it is then impossible to cut the transistor off.

Fig. 10(b) shows the voltages and currents when transistor 1 is on. The potential at b_1 with respect to earth is

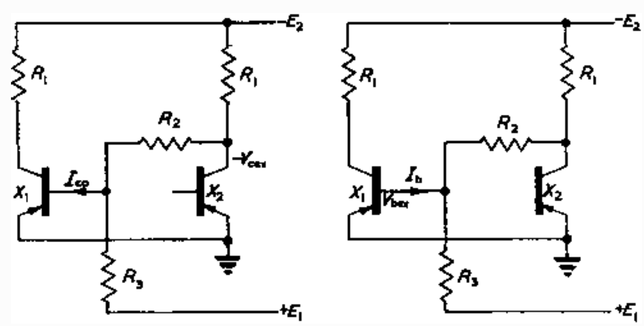

Fig. 10. Stable state design
(a) Transistor 1 cut off; (b) transistor 1 bottomed.

given by

$$-E_2 + \frac{R_1 + R_2}{R_1 + R_2 + R_3}(E_1 + E_2) - I_b\left\{\frac{(R_1 + R_2)(R_3)}{R_1 + R_2 + R_3}\right\}$$

neglecting the effect of I_{co} in transistor 2. For transistor 1 to be bottomed, this potential must be more negative than or equal to the maximum base emitter voltage $-V_{bex}$ required for I_b.

$$\therefore -E_2 + \frac{R_1 + R_2}{R_1 + R_2 + R_3}(E_1 + E_2) + I_b\left\{\frac{(R_1 + R_2)(R_3)}{R_1 + R_2 + R_3}\right\} \leqslant -V_{bex}.$$

$$\frac{R_1 + R_2}{R_1 + R_2 + R_3}\left\{E_1 + E_2 + I_b R_3\right\} \leqslant E_2 - V_{bex}$$

$\therefore$ for transistor 1 to be bottomed

$$\frac{R_1 - R_2}{R_3} \leqslant \frac{E_2 - V_{bex}}{E_1 + V_{bex} + I_b R_3}$$

where $I_b \simeq \dfrac{E_2 - V_{cex}}{\alpha' R_1}$

So that for the transistor to maintain both its stable states the following conditions must apply

$$I_{co} R_3 < E_1$$

$$R_2 / R_3 > \frac{V_{bey} - I_{co} r_{bb}' + V_{cex}}{E_1 - V_{bey} - I_{co} r_{bb}' - I_{co} R_3}$$

$$\frac{R_3}{R_1 + R_2} < \frac{E_2 - V_{bex}}{E_1 + V_{bex} + I_b R_3}$$

The last equation also ensures that the condition for regeneration, $R_2 < \alpha' R_1$, is satisfied.

If the circuit is to be insensitive to temperature change the term $I_{co}R_3$ must be small compared with E_1, a useful criterion being $I_{co}R_3 < (1/5)E_1$.

This therefore means that the current through the potentiometer chain is approximately $5I_{co}$, and is drawn through R_1 and this current in turn must be small compared with the collector current when the transistor is on, a factor of 5 again being practical $\therefore I_c > 25\, I_{co}$ for reliable operation of the circuit.

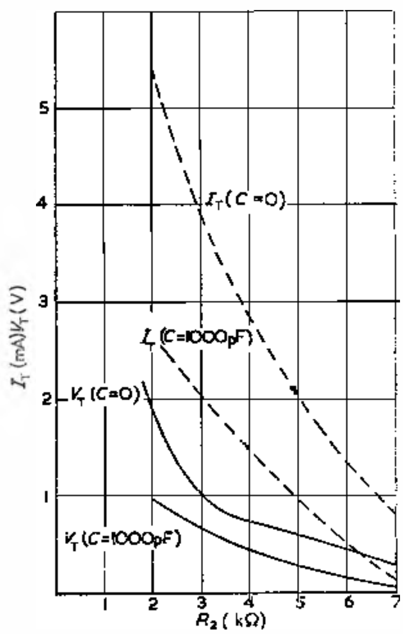

Fig. 11. Trigger sensitivity

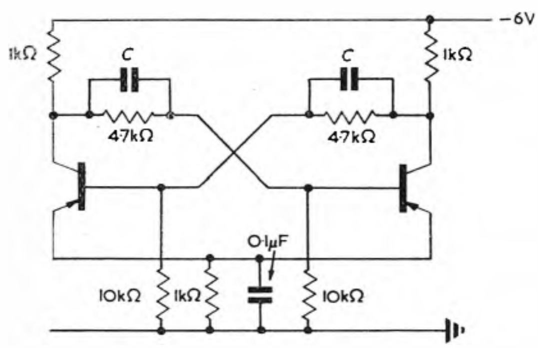

Fig. 12. Practical Eccles-Jordan bistable circuit

The trigger sensitivity of this circuit is very dependent on the degree of stability of the stable states. In the design of the stable states this can be controlled by R_2. Fig. 11 shows the variation of trigger sensitivity with R_2 for the typical circuit, (Fig. 12) using positive trigger on the base. These results also show the marked improvement in trigger sensitivity as the cross coupling capacitors are increased. Experiment shows that though the best trigger sensitivity is obtained with R_2 as large as possible there is an optimum value of R_2 midway between the limits imposed by the stable state design. With this value of R_2 the circuit is least likely to be affected by component tolerances and also has the best general performance.

Fig. 12 shows a typical circuit, and its performance with both low and high frequency transistors is shown in Table 1.

TABLE 1

	L.F. (OC71) α cut-off=500kc/s	H.F. (OC45) α cut-off=5Mc/s
C Cross coupling ..	1000pF	100pF
Switching time ..	5μsec	approximately 0·6μsec
Trigger pulse amplitude	1V	1V

A Monostable Circuit

The Eccles-Jordan bistable circuit can be converted to a monostable circuit if the d.c. coupling between one collector and base is replaced by capacitive coupling, Fig. 13.

The basic principle of operation of this circuit is again similar to that of the valve circuit.

The most noticeable effect of the transistor is to reduce the accuracy of the timing. The duration of the quasi-stable state is not only controlled by the discharge of C_2 through R_5, but also by the base current which flows when the transistor is cut-off. This current is very dependent on temperature and will therefore cause the timing to vary with temperature. This effect can be reduced by making

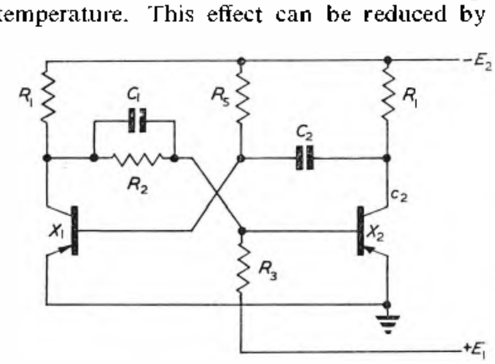

Fig. 13. Basic monostable circuit

the discharging current through R_5 large compared with I_{co}.

Also due to the fall off in gain at high frequencies the coupling and timing capacitor C_2 must be relatively large, with the result that pulses of shorter duration than $10/f_{co}$ are not easily obtained.

The values of all the components except C_2 and R_5 can be fixed by the design procedure for the bistable circuit. R_5 must pass sufficient current to ensure that transistor 1 is bottomed, but not too heavily bottomed, otherwise the circuit needs excessive trigger.

R_5 is given by

$$E_2/R_5 > (I_{cx}/\alpha'_{(min)}) \simeq (E_2/R_1\alpha'_{(min)})$$
$$\therefore R_5 < R_1\alpha_{(min)}'$$

If R_1 is fixed, the duration of the quasi-stable state can therefore only be controlled by the coupling capacitor C_2.

The duration of the quasi-stable period can be shown to be given by:

$$t = C_2R_5 \ln \left\{\frac{2E_2 + I_{co}(R_5 - R_1) - V_{bex1} - V_{cex2}}{E_2 + I_{co}R_5}\right\}$$

where V_{bex1} is the base emitter voltage of transistor 1 when conducting and V_{cex2} is the bottoming voltage of transistor 2.

The terms involving I_{co}, V_{bex1} and V_{cex2} are small and can be neglected if an approximate expression for t is required and this gives

$$t = 0{\cdot}96\ C_2R_5$$

If an accurate value of t is required it is essential to use the full expression.

Also it can be seen that t is dependent on I_{co}, which in

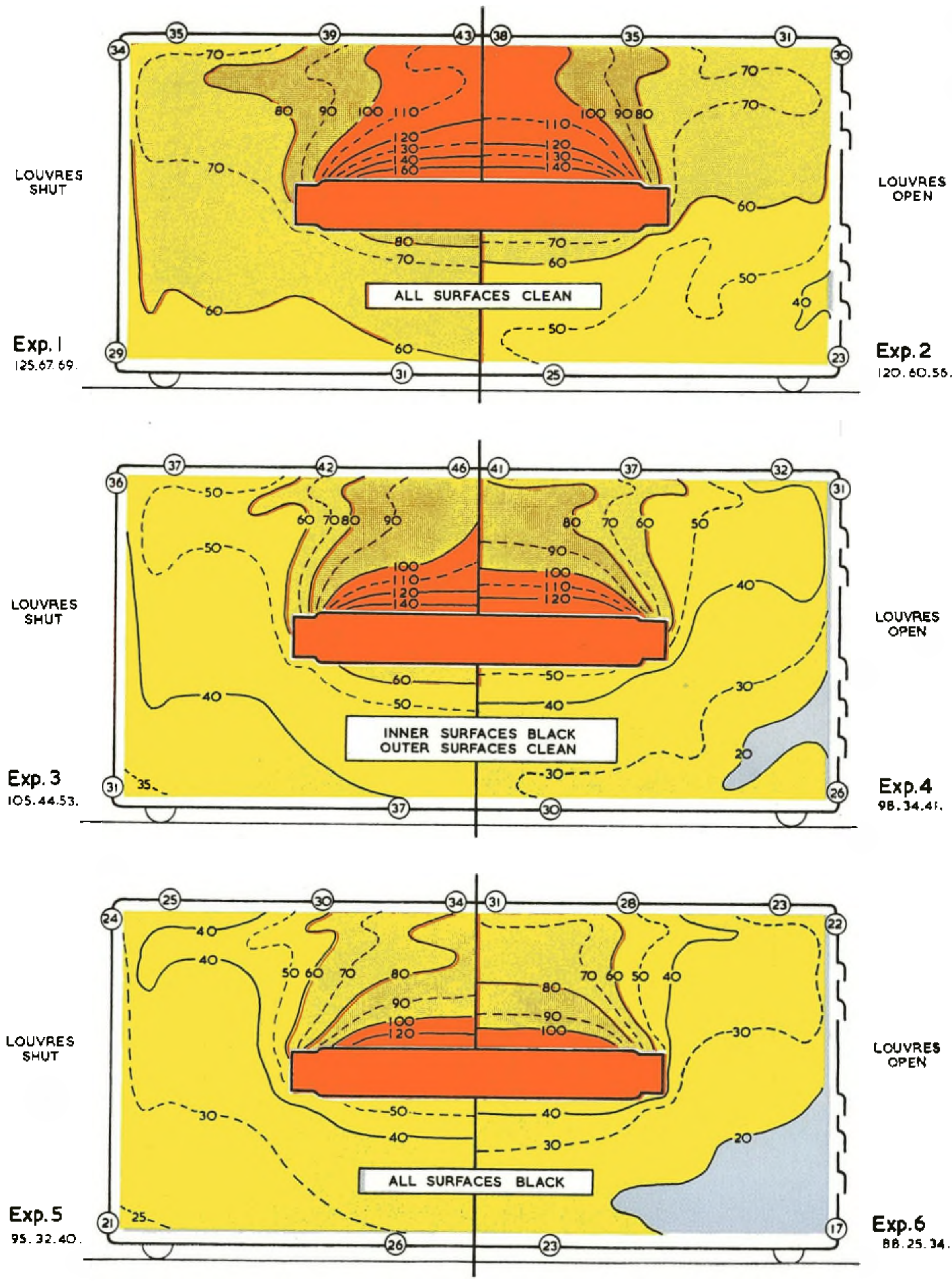

Plate 1. Radiation versus ventilation

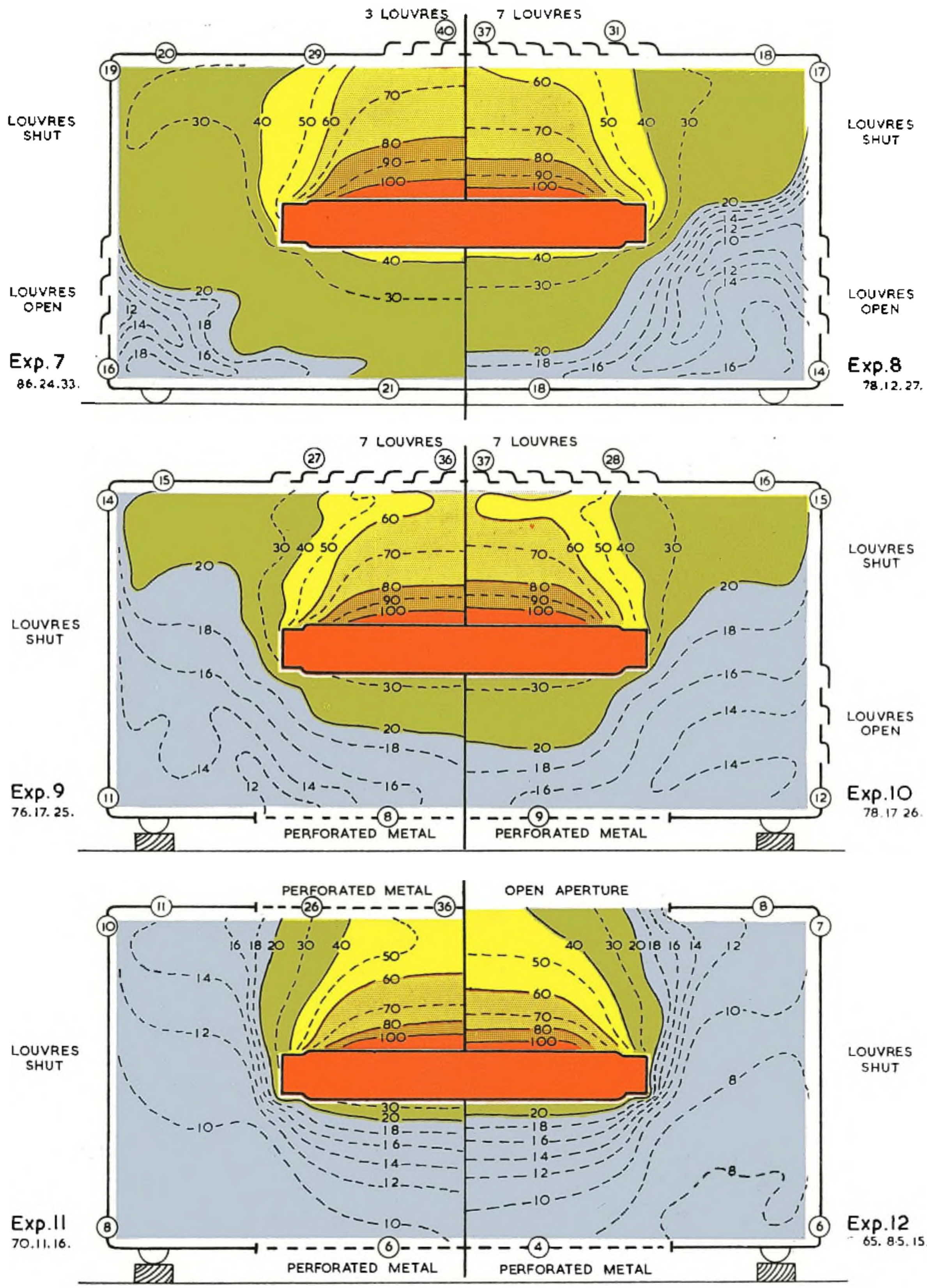

Plate 2. Improved ventilation

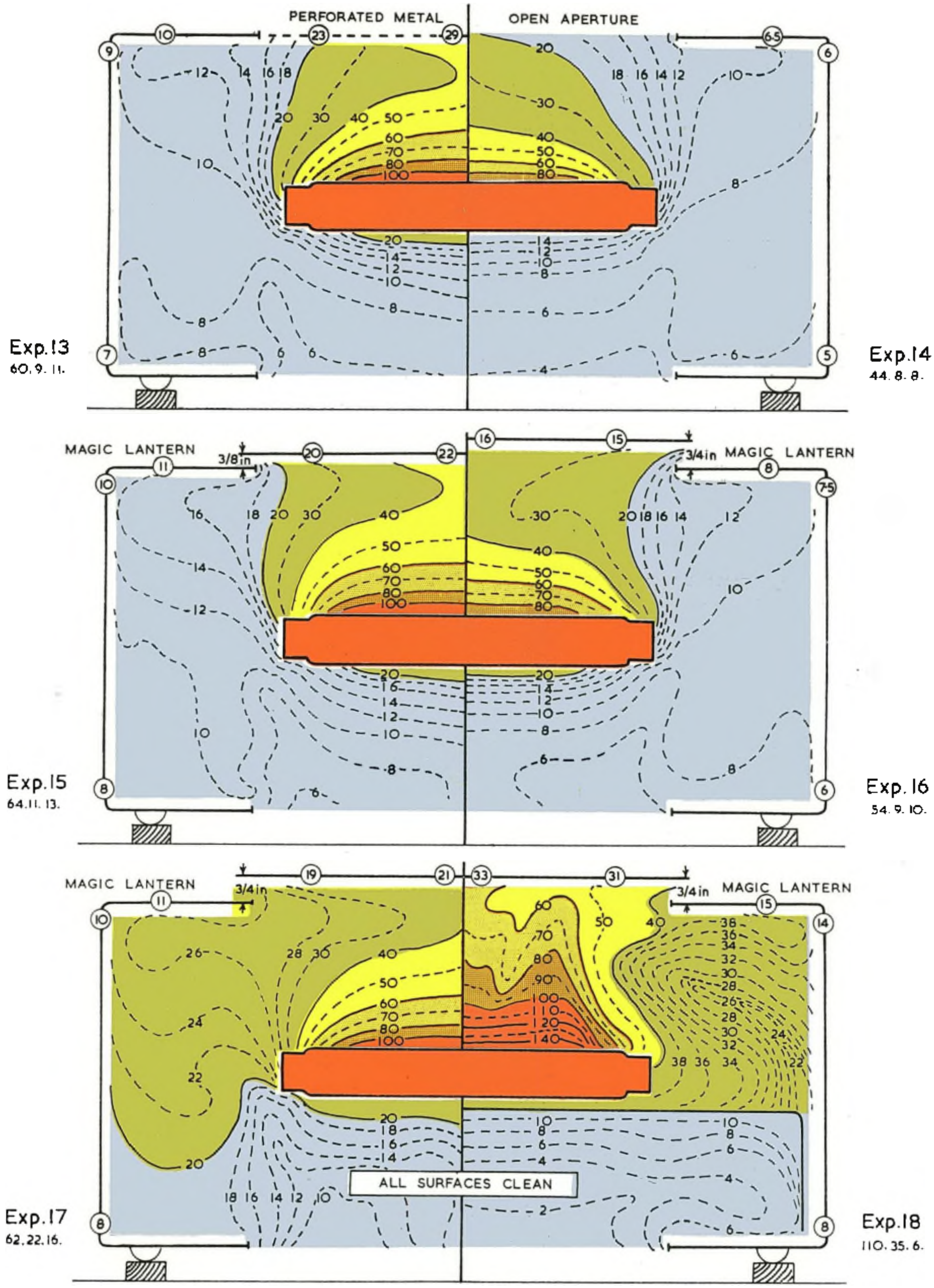

Plate 3. Improved ventilation (continued)

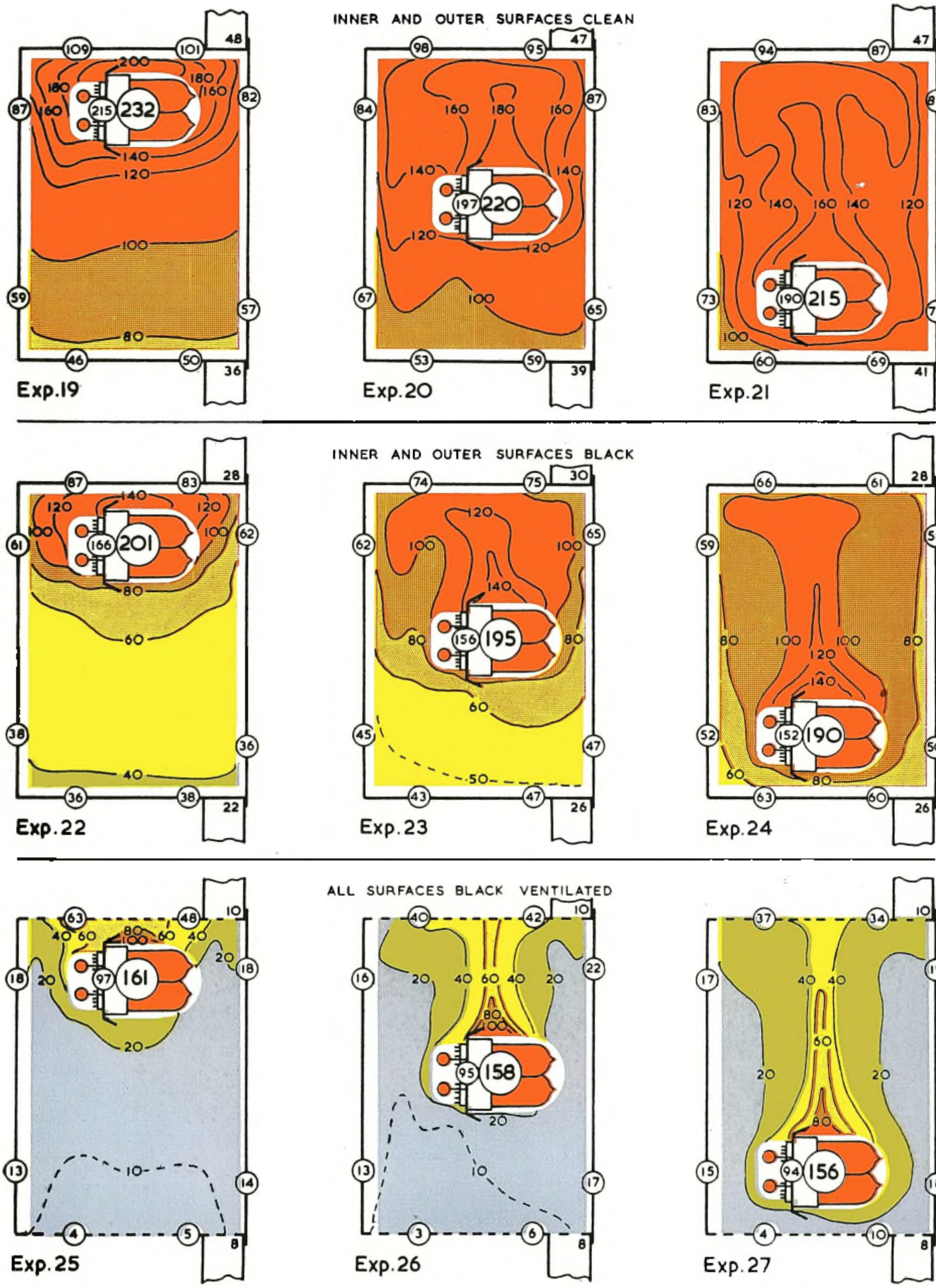

Plate 4. Exploratory measurements

turn is dependent on temperature, and if the circuit is to be operated at varying temperatures, it is essential to make the terms $I_{co}R_3$, and I_{co} $(R_3 - R_1)$ small compared with E_2.

ASTABLE ECCLES-JORDAN CIRCUIT OR MULTIVIBRATOR

If the d.c. couplings in the Eccles-Jordan circuit are removed and replaced by capacitive coupling an astable circuit results. As this circuit is a development from the bistable and monostable circuit all the effects of the transistor characteristics have been discussed. One point that should be mentioned again is that it is essential to ensure that the transistors are not too heavily bottomed during their on period, otherwise the circuit will be very slow switching and this will cause inaccurate timing.

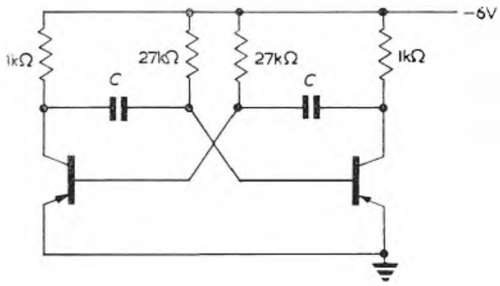

Fig. 14. Practical multivibrator circuit

Fig. 14 shows a typical multivibrator circuit and its performance with low and high frequency transistors is shown in Table 2

TABLE 2

	L.F. (OC71) α cut-off=500kc/s	H.F. (OC45) α cut-off=5Mc/s
Switch on time ..	1μsec	0·15μsec
Minimum C ..	2000pF	200pF
Maximum repetition rate	15kc/s	150kc/s

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1. EBERS, J. J., MOLL, J. L. Large Signal Behaviour of Junction Transistors. *Proc. Inst. Radio Engrs. 42*, 1761 (1954).
2. EARLY, J. M. Effects of Space Charge Layer Widening in Junction Transistors. *Proc. Inst. Radio Engrs. 40*, 1401 (1952).

(*To be continued*)

Non-Linear Modulation *

The application of negative feedback to modulators has been well-known practice for very many years and invariably the object has been an improved degree of linearity in modulation. Negative feedback may also be used when it is required to modulate in accordance with some arbitrary non-linear law, and a scheme which does this and allows for rapid and convenient change of the law of modulation is outlined below.

The scheme entails a cathode-ray tube with deflexion in, say, the X-direction by the carrier-wave signal to be modulated, modulation signals deflecting the beam in the Y-direction. A mask with an edge representing the required law in the two co-ordinates is placed in front of the tube so as partially to hide the face of the tube from a photocell. The photocell is excited by the portion of the trace not obliterated by the mask and its output is smoothed and amplified and used to control the gain of an amplifier through which the X-deflexion signals of carrier frequency are fed before they are applied to the tube. In the absence of the control the gain of the amplifier tends to be high and the function of the control is to reduce the gain so that a very small portion only of the trace is visible beyond the mask at any time. The carrier signals thus become modulated according to the required law. Obviously the law is very simply changed by substituting a different mask. The accuracy of operation is only limited by the sensitivity of the feedback loop, and there are clearly uses in the field of electronic analogue computers.

V.H.F. Car Radio Aerial *

The recently-introduced v.h.f. broadcasting and the wish to adapt car radio receivers to these transmissions have brought about a small problem in aerial and aerial feeder design which is of some interest.

Obviously to pick-up the f.m. channels an aerial of the resonant dipole type is desirable and to avoid the use of a second aerial for normal a.m. reception the resonant aerial system must somehow be pressed into service to give this reception as well. Schemes of this type have, of course, long been known and at first sight the step to take is to use the f.m. feeder as the a.m. aerial. This, however, encounters difficulties from the large capacitance of the feeder which loads up the input circuit of the receiver with too much stray capacitance.

A solution that has been put forward to overcome this problem arranges in the first place that the f.m. aerial is stood off from its feeder by an isolated transformer, and secondly that the feed to the a.m. input of the set is taken from a centre-tap on the transformer primary winding, that is to say the winding connected between the limbs of the dipole. The capacitance of the f.m. feeder is then only effective through the relatively small capacitance between the windings of the transformer.

* *Communications from Electric and Musical Industries Ltd.*

Telecommunications Equipment for the Azores

Marconi's Wireless Telegraph Co. Ltd. have been awarded a contract by the Portuguese Postal and Telegraph Authorities for the provision of radio-telephone communication equipment. The order is for four duplicated v.h.f. multi-channel terminal units, type HM.100, and ancillary equipment.

This will be used to establish direct telephonic communication between the islands of San Miguel and Terceira, and between Terceira and Faial, in the Azores group. Initially, the links will carry seven speech channels and six voice-frequency telegraph channels; standby equipment will come automatically into operation in the event of breakdown of any transmitter, receiver or power supply.

The Marconi type HM.100 terminal unit is designed for high-grade multi-channel radio communication, being capable of carrying up to 48 telephone channels, any of which may be sub-divided to provide a number of telegraph channels. It will operate continuously and unattended for long periods of service.

Previous contracts awarded to Marconi's by the Portuguese Postal and Telegraph Department include a radio link between Santa Maria and San Miguel in the same island group; the present order represents a considerable extension of the telephone network in the area.

A Versatile Rectangular Pulse Generator

By G. O. Crowther*, B.Sc., A.C.G.I., L. H. Light*, B.Sc., A.M.I.E.E., A.Inst.P.
and C. F. Hill*, B.Sc.

The instrument described generates rectangular positive or negative pulses of 1μsec to 12msec duration, at repetition frequencies of 1c/s to 150kc/s and amplitude 0 to 100V. The controls are substantially independent. An electronic delay circuit enables a triggering pulse to precede the main pulse by a short variable period; this facility is useful when the instrument is used in conjunction with an oscilloscope.

Its wide range of adjustment makes the generator very useful either as an integral part of an installation or for testing valves and circuits.

THE development of the generator was undertaken with the following design considerations in mind:

(1) Widest possible range of variation of the pulse parameters.

(2) Stability of calibration and independence of calibration on valve characteristics.

(3) Minimum interaction between controls.

(4) Ease of construction and maintenance.

(5) Minimum number of individually calibrated ranges.

(6) Good waveform.

(7) Low output impedance.

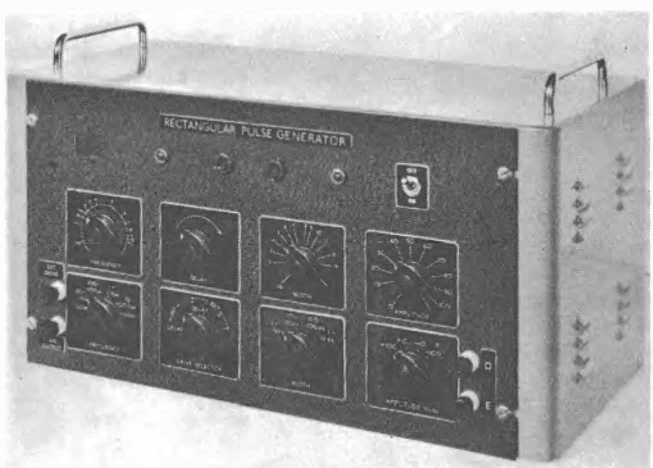


Fig. 1. View of the complete instrument

The complete instrument is shown in Fig. 1, while Fig. 2 illustrates the main elements of the equipment.

The pulse frequency can either be determined by an external pulse source of positive or negative pulses or by an internal thyratron relaxation oscillator. Pulses from either source are fed to a variable delay circuit which generates an output trigger pulse for triggering an oscilloscope followed after the delay period by a pulse for triggering the output pulse-forming circuit. The resulting rectangular pulse from the pulse width generator is passed to the output stages (one for negative-going pulses and one for positive-going outputs). H.T. is supplied from a conventional electronic stabilizer.

Some features of the pulse generator, such as external triggering or delay, will not be required in particular applications. The design allows these circuits to be omitted without modifications to the remainder of the pulse generator.

The various parts of the circuit (Fig. 3) are considered in detail below.

Pulse Frequency Generator

V_2 is a thyratron, type EN31, connected as a free-running time-base. The grid is taken to a fixed potential determined by the resistive potentiometer R_{11}, R_{12}, R_{13}, care being taken to keep the capacitive load on the grid at a minimum.

The output from the time-base is taken across the discharge current limiting resistor R_7. This output consists of a sharp positive-going pulse, whenever the thyratron strikes, whose amplitude is sufficient to switch on the following isolating valve V_3.

The repetition frequency of these pulses is controlled by the values of $C_{2\text{-}7}$, giving a coarse control, and by the continuous control R_{10}. The values of $C_{3\text{-}7}$ are chosen to give frequency ranges of 10 to 120c/s, 100c/s to 1·2kc/s, 1 to 12kc/s, 10 to 100kc/s and 20 to approximately 150kc/s. The highest frequencies given by the last range are limited by valve performance, and therefore this last range is not calibrated. An uncalibrated low frequency range of approximately 1 to 10c/s is obtained with a high value capacitor C_2. No trouble should be experienced with an electrolytic capacitor in this position providing it is not put to excessive continuous use.

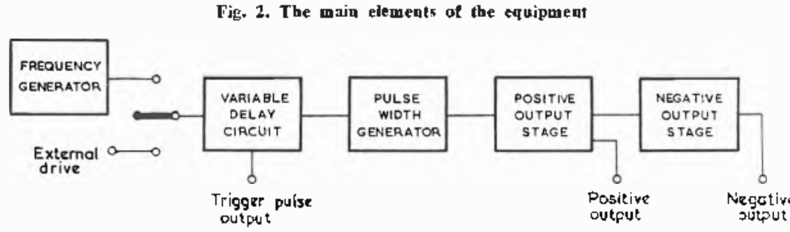
It should be noted that the centre-tap of the heater winding is taken to a potential of +150V by means of the resistive potentiometer R_{14} R_{15} so that the heater to cathode voltage of V_2 is not excessive.

The Input Trigger Circuit

The circuit consists of V_1 (EF80) connected as a simple phase-splitter. The output is taken from either the anode or cathode, dependent on the polarity of the input trigger pulse, the required output being selected by switch S_2.

Positions 1, 2 and 3 of the selector switch S_2 connect either one of the two outputs of V_1 or

Fig. 2. The main elements of the equipment



* *The Mullard Radio Valve Co. Ltd.*

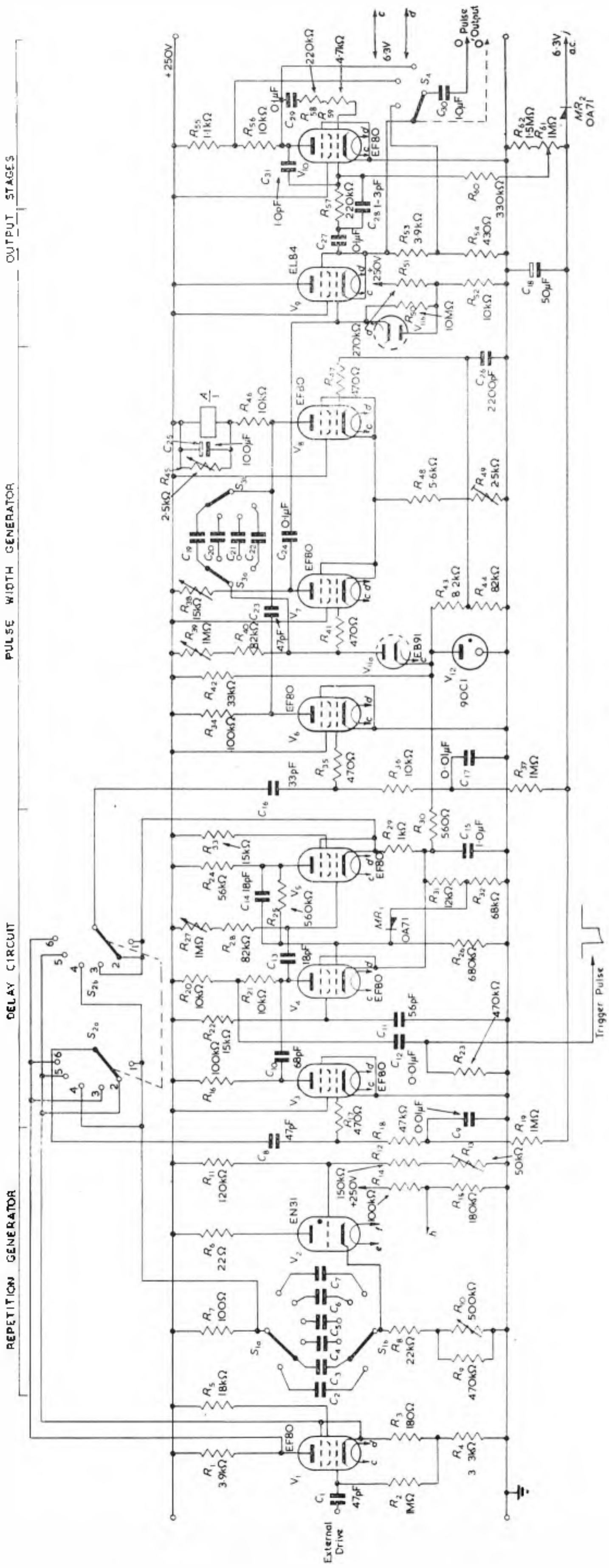


Fig. 3. The complete circuit

the output of the thyratron frequency generator to the delay circuit input amplifier valve V_3 (EF80). In positions 4, 5 and 6, S_2 selects one of the three above outputs and connects it directly to the input amplifier (V_6) of the pulse width generator, thus by-passing the delay circuit. The delay circuit, however, is still triggered, so that the oscilloscope trigger pulse is available at the same instant as the output pulse.

Delay Circuit

The delay circuit consists of V_4 and V_5 (EF80) connected as a cross-coupled monostable multivibrator in which V_5 is normally conducting and V_4 non-conducting. The circuit is triggered to its quasi-stable state by a negative pulse at the anode of V_4 fed from the anode of V_3. The oscilloscope trigger pulse is a negative pulse taken from a tapping point in the anode load of V_4, while the delayed positive pulse for triggering the pulse width generator is formed by differentiating the negative pulse which appears at the cathode of V_5 by means of $R_{36}C_{16}$. The duration of the quasi-stable state, and therefore the delay period, is determined by R_{27}, R_{28} and C_{13}. The magnitude of the delay can be varied continuously by the potentiometer R_{27} from $0{\cdot}7\mu$sec to 7μsec approximately. The delay control is not calibrated.

The function of the germanium diode MR_1 (OA71) at the grid of V_4 is to ensure that the grid-cathode potential of V_4 is constant and equal to the minimum value necessary to ensure cut-off (independent of the V_5 anode voltage). This in conjunction with the speed-up capacitor C_{14} gives a faster and more sensitive switching action.

The current consumption of the circuit in the quiescent state is comparatively small (8mA). However, due to the small capacitor C_{11} in the screen grid circuit of V_4 the peak current which flows during the changeover is large, ensuring a sharp pulse. The time-constant of $R_{22}C_{11}$ is adjusted so that the capacitor is recharged substantially to h.t. potential during the period in which V_4 is cut off. Thus, momentarily, when the valve is switched on it has full h.t. on its screen grid, allowing very high anode currents to flow which decay to a smaller value as the capacitor is discharged by screen grid current.

The negative trigger pulse is taken from a tapping in the anode load of V_4 rather than direct from the anode to eliminate the effect of external loading on the performance of the circuit. The function of R_{30} and C_{15} is to ensure that sharp changes in current do not occur in the stabilizer tube V_{12} (90C1).

Pulse Width Generator

This timing circuit consists of V_7 and V_8 (EF80) connected as a cathode coupled univibrator, biased so that V_7 is conducting and V_8 cut off. A negative pulse applied to the anode of V_3 causes the circuit to 'flip' to its quasi-stable state in which V_8 is conducting and V_7 is cut off.

Coarse control of the pulse width is obtained by selecting one of the capacitors C_{19} to C_{22} by switch S_3 and a fine control is given by R_{39}. The values of C_{19} to C_{22} are chosen to give ranges 1μsec to approximately 12μsec; 10μsec to approximately 120μsec; 100μsec to approximately 1 200μsec and 1msec to approximately 12msec. Amplitude control is obtained by means of the variable resistor R_{38}.

The cathode-coupled univibrator circuit was chosen since both the pulse width and the amplitude are substantially independent of valve characteristics. The amplitude for any given setting of R_{38} is determined by the anode current of V_7 which is defined by the grid potential of V_7 with respect to earth and the common cathode load R_{45} and R_{49}. The grid potential of V_7 is held constant by the stabilizer tube V_{12} (90C1). The pulse width is dependent on the voltage fall across R_{46} when V_8 conducts in addition to R_{39}, R_{40}, and C_{19} to C_{22}. The anode current of V_8 is defined by taking its grid to the fixed potential given by the potentiometer R_{43} and R_{44}. In the steady state this potential is sufficient to keep V_8 cut off.

The rise and fall times of the output pulse are limited by the stray capacitance at the anode of V_7 and the setting of the amplitude control R_{38}. It is found that for pulse amplitudes of 33V or less, the rise time is sufficiently small to provide a good pulse shape down to 1μsec; at 60V, pulse widths of 2μsec are possible, while at 100V deterioration occurs below 3μsec.

The catching diode V_{11a} (EB91) prevents any overswing of the control grid of V_7 and also provides for a faster recovery time. The accuracy of calibration deteriorates for a duty ratio greater than 50 per cent. A relay is placed in the anode of V_8 and its shunt resistor R_{45} adjusted such that when the duty ratio of the pulse output exceeds 50 per cent the relay operates and switches on an indicating lamp. It should be noted that the relay does not pull out until the mark-to-space ratio is reduced to well below 1 : 1.

The variable resistor R_{49} is a preset control allowing for circuit tolerances encountered in the calibration of the pulse amplitude.

The positive rectangular output is taken via C_{24} to the output stage.

Output Stages

A positive output is taken from the cathode load of the cathode-follower V_9 (EL84). The diode V_{11b} (EB91) provides d.c. restoration for the input of V_9. The tapped cathode resistor ($R_{53}+R_{54}$) provides for an attenuated output of one tenth of the full output.

A negative output is obtained from the anode-follower V_{10} (EF80) the output from V_9 being fed via C_{27} and R_{57} to V_{10}. The bias of V_{10} is adjusted to give a standing voltage of 170V on the anode. This allows an upward anode swing of 50V which is necessary with a pulse output of 100V and a mark space ratio of 1 : 1. The tapped anode load (R_{55} + R_{56}) provides for an attenuated output of one tenth of the full negative output.

The positive and negative outputs from V_9 and V_{10} are taken via the switch S_4 and capacitor C_{30} to the output terminal 0.

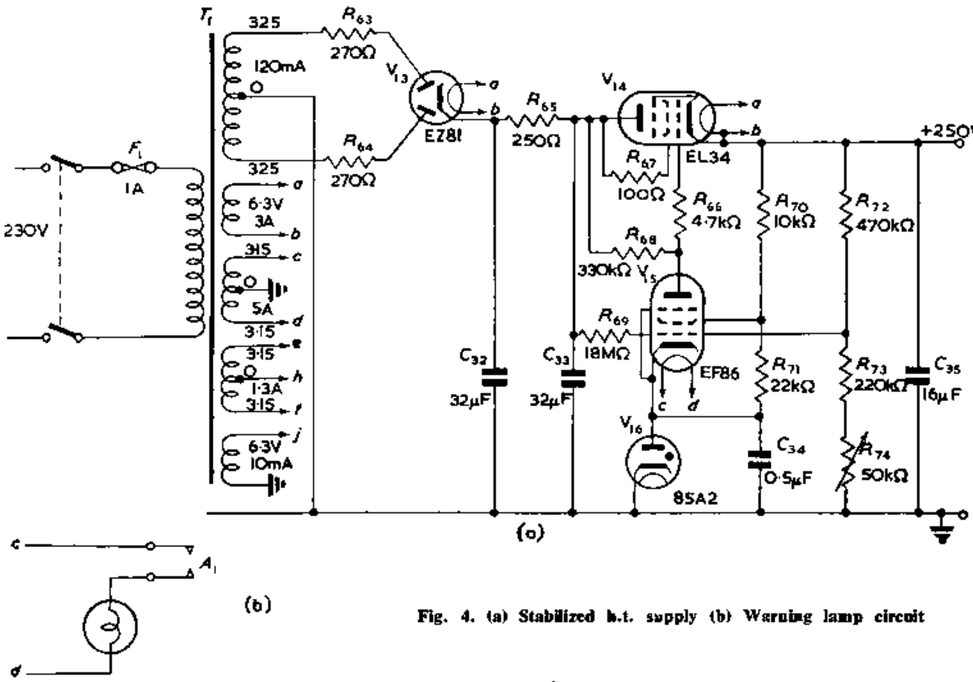


Fig. 4. (a) Stabilized h.t. supply (b) Warning lamp circuit

Power Supply

The pulse generator is designed to work from a 250V positive stabilized supply. The stabilized h.t. supply employs a conventional two valve plus voltage reference tube circuit (see Fig. 4).

Calibration and Setting Up

The h.t. supply must first be set to +250V by adjustment of the preset resistor R_{74} in the power supply.

The grid of V_2 is set to 150V by adjustment of the preset control R_{13}. This control can be used to correct for valve differences, and also for strike voltage drifts in individual valves. The fine frequency control of the pulse output is by means of the carbon potentiometer R_{10}. Owing to variation of the law followed by individual potentiometers a standard frequency scale is not possible; each pulse generator should be individually calibrated.

It is suggested that a convenient method of frequency calibration is against a calibrated sine wave by way of

Lissajous figures on an oscilloscope, see Fig. 5(a). The calibrated sine wave is fed on to the Y-plates through R_2 and on to the X-plates via the phase-shifting network R_1C_1. In the absence of any pulse input an ellipse is obtained on the oscilloscope. The pulse is applied via R_3 to the Y-plates. When the sine wave and pulse frequencies are equal, a Lissajous figure of the form shown in Fig. 5(b) is obtained. The 10c/s to 120c/s, 100c/s to 1·2kc/s and 1kc/s to 12kc/s ranges can all be calibrated on the same scale, using the following procedure. R_9 is first chosen

such that $\left(\dfrac{R_{10(max)}R_9}{R_{10(max)}+R_9}+R_8\right)\dfrac{1}{R_8}=12$

and then with R_{10} at its maximum value C_3, C_4, C_5 are chosen to give frequencies of 10c/s, 100c/s, and 1kc/s respec-

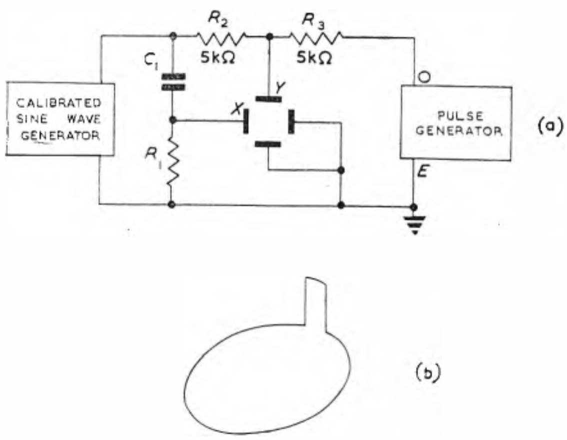


Fig. 5. (a) Frequency calibration against calibrate sine wave (b) Form of Lissajous figure obtained with equal sine wave and pulse frequencies

tively, which correspond to the bottom end of their respective ranges. It should then be found that the upper ends of the ranges give frequencies of 120c/s, 1·2kc/s, and 12kc/s, giving an overlap on each range. The 10kc/s to 100kc/s range requires a separate scale, C_6 being chosen to fix the lower end of the scale at 10kc/s. The other two ranges are not calibrated; C_2 is chosen to give an approximate range 1 to 10c/s and C_7 is made half the value of C_6 to give a range 20kc/s to approximately 150kc/s.

In the original model, nominal values of the capacitors to give the required ranges are as follows:

$C_2 = 16\mu F$ electrolytic

$C_3 = 1{\cdot}4\mu F$

$C_4 = 0{\cdot}127\mu F$

$C_5 = 0{\cdot}0118\mu F$

$C_6 = 900pF$

$C_7 = 560pF$

Amplitude Calibration

This is carried out before the pulse width calibration.

The amplitude control R_{38}, being a wirewound potentiometer, allows the effective rotation of the manual control of R_{38} to be calibrated linearly over the range 0 to 100V. Hence a standard linear 0 to 100V scale can be used for the amplitude calibration. It is still necessary to line up the pulse output with the amplitude scale. This is carried out by turning the amplitude control R_{38} to its maximum setting, and then adjusting the preset control R_{49} to give a 100V positive output.

The setting of R_{49} must not be subsequently disturbed, as this affects the pulse width calibration by controlling the anode current through V_8.

Pulse Width Calibration

Here again, owing to variation in the law followed by carbon potentiometer control R_{39}, each pulse generator should be calibrated with its own individual pulse width scale.

A suggested method of calibration is the metering of the anode current through V_8. The average current passed by V_8 is determined by the current i passed by V_8 during the quasi-stable state of the 'flip-flop' mechanism, the duration of the pulse m and the repetition frequency f.

$$I_{av} = imf$$

i is independent of the frequency and pulse width control setting, being affected, however, by the preset control R_{49} which is fixed in the previous amplitude calibration. The value of i is determined by removing V_7 and measuring the anode current of V_8; it will be approximately 10mA. The current meter should be placed at the top end of the anode load of V_8 thus introducing no active stray capacitance into the action of the pulse width generator. V_7 is replaced and the average anode current I_{av} of V_8 metered under normal pulsing conditions. The frequency f having been calibrated previously, the pulse width m is given by

$$m = I_{av}/(if)$$

The calibrating procedure is then as follows: R_{39} is set to its minimum value and C_{19}, C_{20}, C_{21} and C_{22} are chosen to give the pulse widths set out in the tables below. The recommended frequency for calibration is also given

C_{19}	1μsec	10kc/s
C_{20}	10μsec	1kc/s
C_{21}	100μsec	100c/s
C_{22}	1msec	10c/s

It should be found that one scale suffices for all ranges.

In the original model nominal values for the various capacitors were

$C_{19} = 33pF$

$C_{21} = 3\,300pF$

$C_{20} = 330pF$

$C_{22} = 0{\cdot}033\mu F$

Invertor Stage Over-swing Correction

The negative pulse output contains over-swings, these being usually associated with anode-follower circuits of this type. The amplitude of these over-swings varies with the capacitance loading on the output. With an average load of 70pF (input capacitance of an oscilloscope, say,) the over-swing on the leading edge of the pulse may be eliminated by choosing a suitable value of the capacitor C_{31} (approximately 1pF). There may exist, still, an over-swing on the trailing edge, but this is usually of no consequence. Capacitor C_{28} is adjusted to give optimum rise and fall times.

Construction and General Features

Warning Circuit

The shunt resistor R_{45} across the relay, is adjusted so that when the mark-to-space ratio exceeds 1 : 1 the relay contact makes and completes a warning lamp circuit (Fig. 4(b)).

Supply Socket for Auxiliary Shaping Circuits

A socket is provided on the front panel to supply simple valve or diode circuits which may be used in conjunction with the equipment. The maximum load which may be placed on the 250V line is 15mA while the heater windings provide for a load of 0·5A at 6·3V a.c.

Layout

The unit is portable, mounted on a 19in by 10½in front panel, and housed in a case, measuring 10½in by 10½in by 21½in.

Point-to-point wiring should be carried out wherever possible, and lead lengths kept short to diminish stray capacitance. Particular attention should be paid to the wiring of the switch S_3 and amplitude control R_{38}.

Performance

The pulse generator described has the following performance.

Parameter Ranges

Pulse Repetition Frequency

1c/s to 150kc/s (approx.) in the following ranges:

(a) 1c/s to 12c/s (approx.) uncalibrated

(b) 10c/s to 120c/s
(c) 100c/s to 1·2kc/s } on common scale
(d) 1kc/s to 12kc/s

(e) 10kc/s to 100kc/s — on separate scale
(f) 20kc/s to 150kc/s (approx.) — uncalibrated

Pulse Width

1μsec to 12msec in the following ranges:

(a) 1μsec to 12μsec
(b) 10μsec to 120μsec
(c) 100μsec to 1 200μsec
(d) 1msec to 12msec

A common scale is provided for all ranges.

The pulse width may be up to 50 per cent of the repetition interval.

For pulse widths less than 3μsec, satisfactory pulse shapes can only be obtained at amplitudes below the maximum setting.

Pulse Height

0 to 100V positive or negative in two ranges.

(a) 0 to 100V positive
(b) 0 to 10V positive
(c) 0 to 100V negative
(d) 0 to 10V negative

Stability

Pulse width and amplitude have been made largely independent of valve changes by the use of stabilized voltages and negative current feedback. The long term stability of these two parameters is better than some 3 per cent. The stability of the repetition frequency up to some 30kc/s should be of the same magnitude, worsening to perhaps 10 per cent at 100 kc/s.

Interdependence of Controls

The repetition frequency is independent of all other settings.

The pulse width is not entirely independent of the ratio of pulse width to repetition interval (the duty ratio). The pulse width may decrease from its calibrated value by up to 5 per cent as the duty cycle is increased from a small value to 50 per cent. (The amount of decrease depends on the setting of the 'Pulse Width—Fine' control, and is in general much less than 5 per cent).

The pulse amplitude is independent of other adjustments on the positive output; on negative output the constancy of amplitude is better than 2 per cent as the duty ratio is varied up to 60 per cent and then deteriorates rapidly.

The degree of interdependence of the adjustments is thus seen to be slight, provided that the duty ratio does not exceed 50 per cent (square wave operation). The indicator light fitted gives warning when this duty ratio is reached.

Waveform

The performance of the pulse generator when its output is lightly loaded is as follows:

Rise and fall times: These depend on the setting of the 'Amplitude-Fine' control. They are 0·3μsec at 30V and 1μsec at full output. Decay of pulse amplitude during its duration (sag): This is apparent only at the greatest pulse widths and is less than 2 per cent.

Effect of load: Stray capacitances >100pF across the output will increase the fall time on the positive output and the rise time on the negative output. Some over-shoot may also appear on the trailing edge of the negative output.

The decay on wide pulses will be increased if the load resistance is low owing to the limited size of the output capacitor (1μF). The amount of decay due to the output time-constant is given by

$$\text{Fractional decay} = \frac{\text{Pulse width}}{\text{Output time-constant } (=\text{load} \times 1\mu\text{F})}$$

Output Impedance

The output impedance in the +100, +10, −100 and −10 positions of the output switch is approximately 100, 400, 300 and 1 000Ω respectively

Possible Modifications

Conversion for Internal Triggering Only

When the facility of triggering the pulse from an external source is not required, V_1 with its associated circuit can be omitted. S_2 then can be a two-position switch giving delay/no delay.

Conversion When Delay is not Required

The three valves V_4, V_5 and V_6 with their associated circuits can be omitted when the delay facility is not required. C_{23} is then taken direct to the anode of V_3.

Pulse Width Control

The pulse generator as described is open to the criticism that the percentage reading accuracy at low settings of the 'Pulse Width-Fine' control is low, because a linear potentiometer is used to cover a wide (12:1) range of widths. A semi-log law potentiometer would largely eliminate this difficulty. An alternative solution is to provide the 'Pulse Width-Coarse' control with a greater number of steps, so that the 'Pulse Width-Fine' control need only cover a smaller range.

Direct Coupled Positive Output

In some applications it is an advantage to have access to the pulse output directly and not via the output capacitor. As shown dotted in the circuit diagram, an extra socket may be provided, connected to the positive output (cathode of V_9) but care is necessary in its use as external short-circuits will cause damage to the pulse generator. (A 60mA fuse in the connexion will give a substantial measure of protection.) It is not desirable to allow direct connexion to be made to the output of the invertor stage, for no great advantage is obtained and the danger of causing damage is much greater.

Acknowledgments

The authors wish to express their thanks to The Mullard Radio Valve Co. Ltd, for their permission to publish this article and to Mr. G. F. Jeynes, for checking the circuits.

Heat Control in Electronic Equipment

(Part 1)

By E. N. Shaw*, A.M.I.E.E.

A study is made of natural methods of cooling compact equipment with a view to improving thermal stability and reducing component failure-rate due to overheating. Rejecting forced ventilation as an increase in bulk and complexity that defeats the object of miniaturization, a number of experiments were made to determine the mechanism of heat loss from basic units of simple design.

The resulting appreciation of the relative effectiveness of loss by conduction, convection and radiation made possible the optimum use of each in a system of construction in which heat flow is controlled into desired channels. Units constructed on this principle are self-cooling and may be mounted in assemblies of similar units with the minimum of heat transfer between them.

THE advent of automation will not only present the designer of electronic equipment with a new field in which to exercise his ingenuity, but will demand a standard of reliability higher than hitherto considered adequate. When a manufacturing process is entirely dependent upon the correct functioning of an 'electronic brain', the failure of a single component will completely disrupt production for the time necessary to effect a replacement, make routine tests and regain full control. Moreover, higher orders of accuracy and stability will be required in equipments called upon to gauge the quality of output from the machines they control if the inadvertent manufacture of worthless sub-standard articles is to be avoided.

Judging by current practice, the desired reliability will not be attained unless very much more thought is given to protecting components from the effects of temperature rise, particularly in miniature forms of construction that will doubtless be essential in equipments of the complexity envisaged. Statistical evidence on this point is hard to obtain through sheer lack of experience over considerable periods of time, although figures quoted[1] for a large analogue computer in the United States suggest that one third of the total time is devoted to maintenance, giving a utilization factor of only 66 per cent. While this can be tolerated in a device that solves in a few moments problems that would otherwise take months of laborious calculation, a comparable instrument controlling a continuous manufacturing process would require to be a great deal more reliable if it is not to defeat the very purpose of its introduction.

With a range of more reliable valves now becoming available, an equipment designer need have less anxiety on that score and may be expected to devote more of his time to becoming better acquainted with the shortcomings of components under thermal stress, with a view to increasing the reliability of his products by building-in improved methods of cooling. In an assembly of mixed components the effects of heating can range from slight changes in value due to temperature coefficient, through quite marked deterioration in characteristics, to catastrophic failure, if the maximum permitted temperature is exceeded. A brief survey of the magnitude of some of the effects encountered shows how desirable it is to attain a minimum operating temperature within equipments, miniature or otherwise.

Temperature-Sensitivity of Components

All components are temperature-sensitive to some extent, either as regards their accuracy, stability or life-expectancy. Typical examples are examined below :—

(1) High-stability (Grade 1) cracked carbon resistors have a temperature coefficient of the order of 0·04 per cent/°C. This means that a resistor selected to a tolerance of 1 per cent at room temperature is forced out of tolerance by being placed in an ambient only 25°C above room temperature and can, without any dissipation at all in the resistor itself, attain a departure of 5·2 per cent from nominal when raised in temperature by 130°C to a maximum of 150°C, the point on the de-rating curve at which the permitted dissipation is zero. Fully loaded at its rated ambient of 70°C, the change in value might not be quite so marked, but no amount of de-rating could reduce the departure to less than the 2 per cent due to the 50°C increase above 20°C room ambient.

(2) The insulation resistance of paper capacitors is halved for each 10°C increase in working temperature. To guard against a catastrophic failure of the insulation it is necessary to reduce the voltage rating at 100°C to some 60 per cent of that at 70°C. If the voltages within an equipment are irreducible, the designer has the choice of sacrificing miniaturization by using bulkier capacitors with thicker insulation, or reducing the ambient to a more tolerable level. Failure to do either will materially reduce the working life of a capacitor.

(3) Semiconductor devices are highly temperature-sensitive. Firm figures are not available for transistors or crystal diodes, but selenium rectifiers are rated at 55°C maximum, above which rapidly increasing losses decrease both rectifier efficiency and working life. On the other hand, if an ambient temperature not exceeding 35°C can be guaranteed, current ratings may be doubled, or the same current supplied by a rectifier of smaller size—an excellent example of miniaturization by temperature control.

The figures above not only indicate maximum temperature limits, but also draw attention to the different temperature levels that must be considered as being critical for various components. It is this disparity between safe working temperatures that is the most acute problem facing equipment designers, particularly when components are so tightly packed in miniature assemblies that it is difficult to make intelligent use of the temperature gradients that can be found in less congested containers.

Is Miniaturization Worthwhile ?

The answer to this question will remain in the negative for as long as the rapid advances in electronic circuit design are allowed to outstrip the technique of assembling components into reliable equipments. Components differing in temperature-sensitivity cannot be packed into the same environment without risk of early failure of the more sensitive elements unless some form of segregation is adopted. This need for segregation is hotly contested by the advo-

* *Radar Research Establishment, Malvern.*

cates of miniaturization-at-any-price, who cover their ineptitude by clamouring for components capable of withstanding higher and higher temperatures. While such demands are reasonable for military applications where extreme forms of miniaturization are of paramount importance, the cost of developing and manufacturing high-temperature components may well prohibit their use in equipments operating under less stringent domestic conditions.

Even if the economic factor is ignored, it is not likely that a complete range of components capable of reliable operation at, say, 200°C will suddenly become available. There will always be some items that will wilt in so torrid a climate unless special steps are taken to keep them cool.

The accepted method of dealing with such exceptions is, to put it mildly, somewhat illogical. A collection of components, each capable of almost indefinite life at room temperature, is bundled into the smallest possible con-

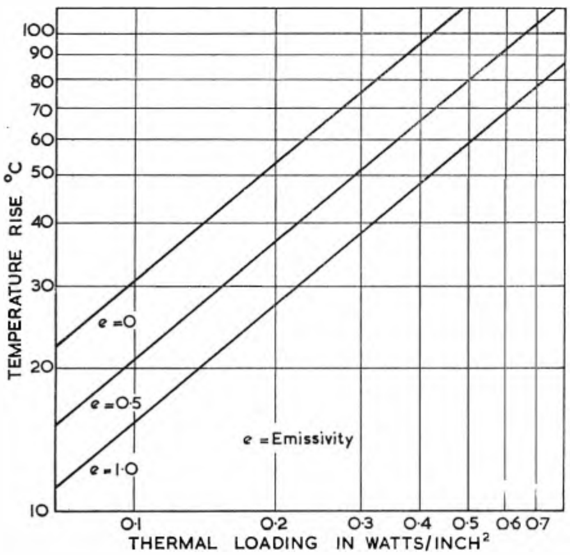


Fig. 1. Skin temperature as a function of thermal loading for various surface emissivities

tainer, within which the temperature immediately rises to a level that cannot be tolerated by the more delicate components inextricably buried in this wantonly created oven. The impossibility of giving them individual protection is overcome by the "brute force" method of blowing air through the whole equipment to reduce the internal ambient to a level at which the accuracy, stability and reliability of the most highly vulnerable components will not be jeopardized. To achieve the required temperature reduction the blower must not only supply an adequate volume of air, but must also overcome the considerable impedance to air flow offered by a series of compact assemblies, the ducting between them, and the numerous baffles necessary to maintain some semblance of temperature uniformity. Power requirements for the blower may, under these conditions, approach those of the equipment itself, with the added complexity of ventilating plant making maintenance more difficult and nullifying the saving in space that must be presumed to be the object of miniaturization.

Always excluding military equipments, which must function under conditions that defy a logical approach, the absurdity of allowing the power dissipated in an equipment to heat it to the point where it needs further power to cool it becomes apparent, and is quite inexcusable under the conditions existing in the laboratory or on the factory floor. This does not mean that miniaturization is not worthwhile for such applications: on the contrary, it will be shown that a mastery of the natural methods of heat dissipation enables heat flow to be controlled so effectively that forced ventilation can be dispensed with without sacrificing miniaturization.

Fundamentals of Heat Dissipation

Cooling equipment by skill instead of brute force requires the ability to control heat flow by a more complete understanding of the fundamentals of heat transfer and

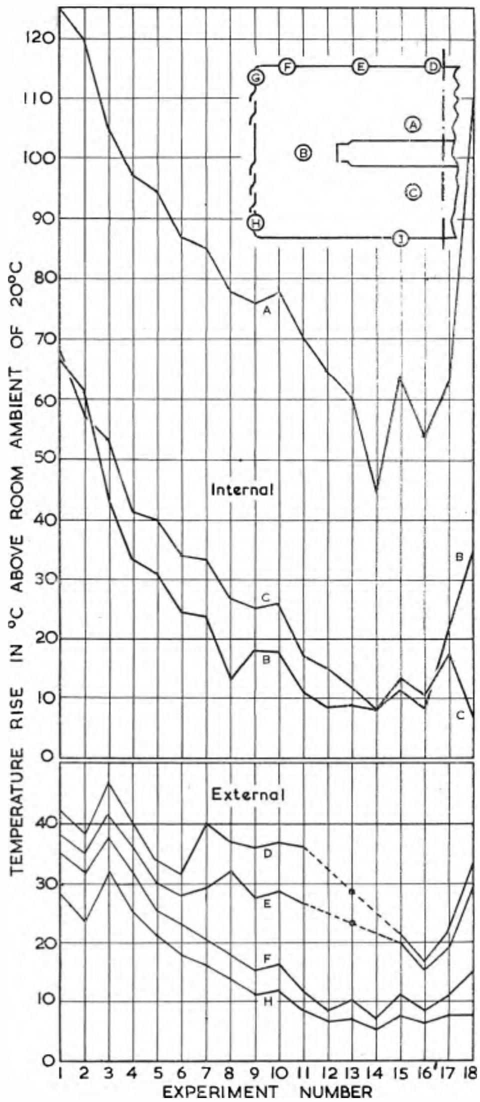


Fig. 2. Temperature reductions due to improved radiation and ventilation

dissipation. Any textbook on heat will inform the reader that heat transfer by conduction depends upon temperature difference, that heat loss by convection is a function of temperature to the power 5/4, and that radiation loss is determined by the difference between the fourth powers of the absolute temperatures. Given a set of constants and the thermal properties of the materials involved, it is easy to estimate the heat loss under idealized conditions due to any one of the three means of propagation in the absence of the other two.

Idealized conditions do not, unfortunately, exist in complex electronic equipments and the need for simultaneous consideration of three closely interdependent temperature relationships makes it as difficult to calculate the total loss as it is to predict the takings at a country fair, in which the losses on the swings are made up by gains on the roundabouts, or both eclipsed on wet days by profits from the covered sideshows.

Empirical relationships between container skin temperatures and thermal loading per unit area, as in Fig. 1, are of little value in determining temperature levels within a container because these figures represent arbitrary points on the temperature gradient between room ambient and the heat source in the very much hotter interior. A glance at Fig. 2 shows that internal temperatures not only vary with position in the container, but can be slightly lower and very considerably higher than external skin temperatures. The general trend is for internal and external temperatures to follow one another, but conditions can arise (as in Exp. 3) when reduced internal temperatures are accompanied by increased external temperatures.

Although the empirical relationships in Fig. 1 fail to provide a short cut to the solution of thermal problems in equipments, they are valuable when considering the implications of miniaturization and its effect on the operating temperature of bodies in thermal equilibrium. As an example, 100W in a container of surface area 500in² gives a loading of 0·2W/in² and requires (from Fig. 1) the container temperature to rise 37°C above room ambient to maintain thermal equilibrium between energy supplied and that dissipated by convection and by radiation from a surface of emissivity 0·5. Greater emissivity increases the dissipation rate and drops equilibrium temperature below 30°C, while poorer emissivity decreases the dissipation rate and raises equilibrium temperature to the region of 50°C above room ambient. Changing to miniature components within the container will have no effect on skin temperature if the input remains at 100W, but taking advantage of their smaller bulk by reducing the container surface to, say, 250in² will double the loading and force the skin temperature rise up to 67°C for emissivity 0·5, with a range of some 50° to 90°C for different emissivities. Temperature levels in the interior are unpredictable, except that they will follow, and at every point exceed, those on the exterior.

The same laws apply to components. A miniature valve must, by virtue of reduced envelope surface area, attain an equilibrium temperature higher than that of its conventional counterpart of equal dissipation. Restricting its rate of heat loss by placing it in a container will increase its equilibrium temperature, with a further increase if the container itself is miniaturized. As temperature is both the driving force behind all three means of heat transfer and a limiting factor in the operation of a wide range of components, the cumulative effect of more compact assemblies comprising hotter components in a smaller container can be devastating.

The implication is inescapable. If miniature assemblies are to give reliable service, every heat-dissipating element will have to be regarded as a potential thermal menace and appropriate steps taken to cool it in such a way that heat transfer to temperature-sensitive components is minimized.

Measuring Internal Temperatures

Temperature levels within containers being of so much greater interest to equipment designers than skin temperatures, it was decided to determine the former experimentally in order to study the mechanism of heat transfer within a container in relation to its skin temperature, and to investigate the effectiveness of different methods of cooling.

Thermo-couples were used for all temperature measurements, both external and internal. In anticipation of more detailed exploration at a later stage 30 thermo-couples were prepared at the same time to ensure uniformity, and connected to the switching system shown in Fig. 3. As all three means of heat transfer depend upon temperature difference, the thermo-couples were made double-ended and all temperatures recorded as a rise above room ambient which was maintained at 20°C. By breaking the copper conductor only, this double-ended arrangement also avoided spurious temperature effects from dissimilar metals in switching circuits. A mirror galvanometer was made direct-reading on three ranges by pre-set resistors in series to swamp any small changes in circuit resistance due to switch contacts or the accidental stretching of thermo-couple leads. The 30 thermo-couples agreed to better than 0·25°C.

A welded aluminium instrument case of standard pattern 19in long, 9in high and 8¼in deep with a surface area of 804in² was used in the first series of experiments.

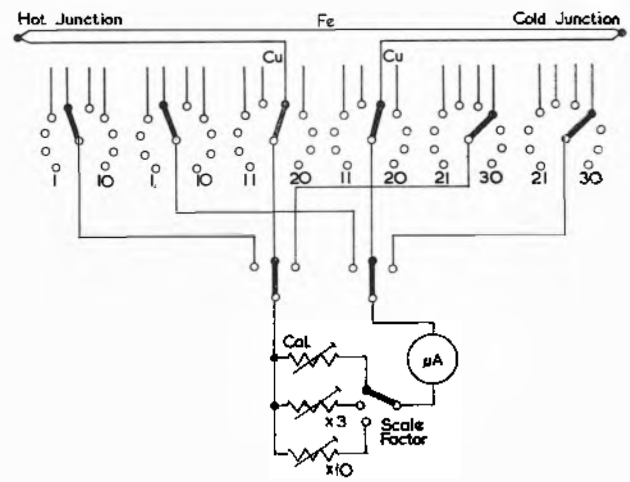


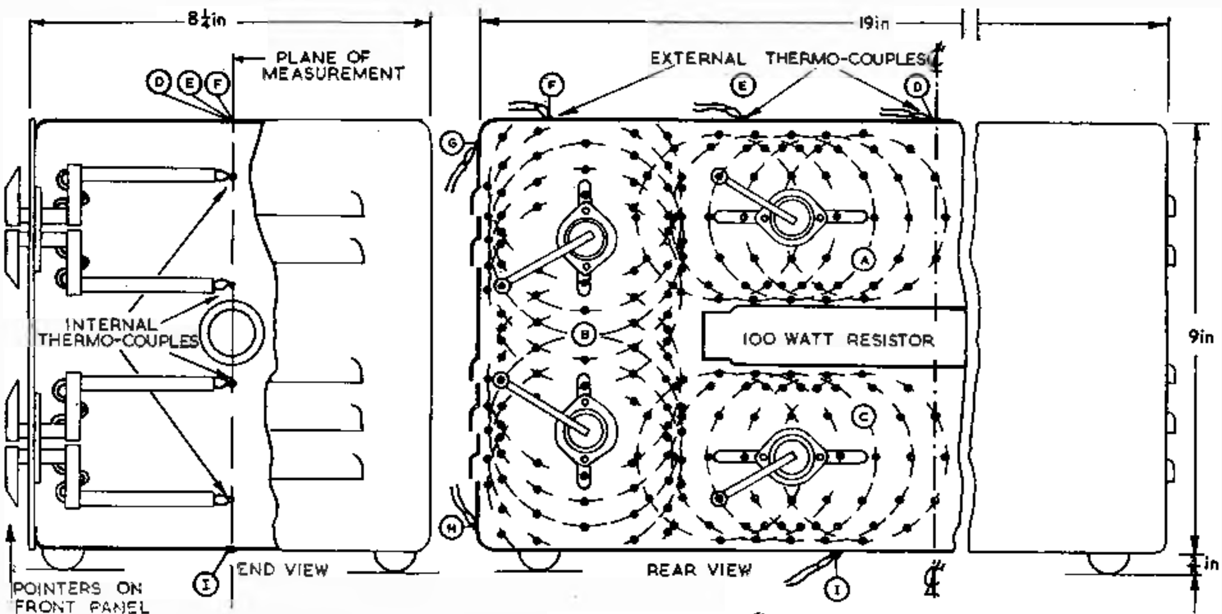
Fig. 3. Switching arrangement for 30 thermo-couples

The heat source was a fully-loaded 100-watt vitreous resistor centrally mounted to give 0·125W/in² surface loading and a predicted skin temperature rise (from Fig. 1) between 19° and 37°C, depending upon the surface emissivity.

Ten thermo-couples were employed, with six at fixed points on the exterior and four made mobile within the container by the mechanism shown in Fig. 4. Rotated by controls on the front panel, each thermo-couple had twelve accurately indexed positions round a circle whose centre could be located at any of five predetermined points by moving the mechanism along horizontal or vertical masked slots in the front panel, giving $4 \times 12 \times 5 = 240$ points of measurement in the form of a mosaic, Fig. 4, suitable for the plotting of isothermal patterns in the plane of measurement. As the mosaic was reproducible from experiment to experiment, three points A, B and C were selected as being representative and the temperatures at these points plotted in the form of a histogram, Fig. 2, in which they could be compared with skin temperatures in the same plane of measurement. Out of consideration for the feelings of the manufacturer whose product was operated at full load in a very confined space, the temperature rise of the 100W resistor was not measured, although it might be mentioned that it withstood its ordeal with commendable fortitude.

Presentation of Results

The strictly symmetrical layout permitted both measurement and presentation to be restricted to half-sections, with a dividing line between them to mark a change in conditions. Thus, in the six experiments recorded on Plate 1, the half-sections on the left had all the louvres shut and those on the right had all the louvres open. Half-sections with common surface finishes were grouped together, the top pair having all surfaces clean, the two below sharing an internal coat of black paint, and the lowest pair being black on all surfaces. Ringed figures on the container skin record the temperature rise at those



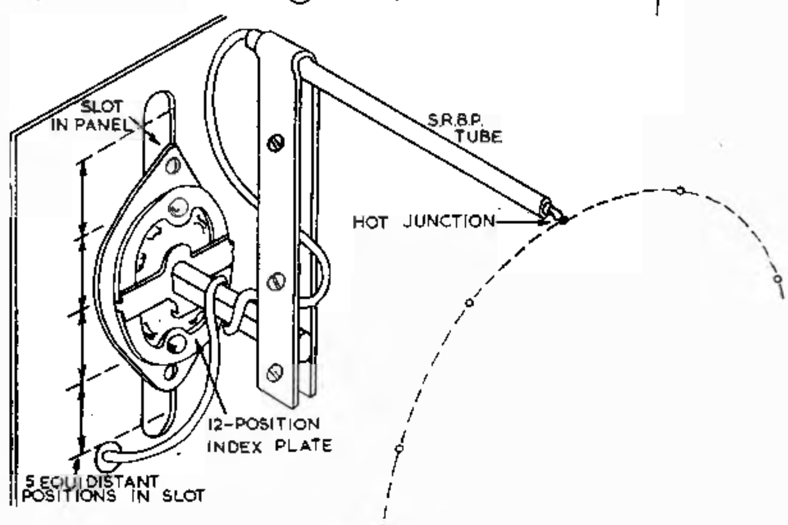


Fig. 4. Instrument case containing a 100W load, with means of measuring both external and internal temperature rise

(a) (above) Disposition of fixed (external) and mobile (internal) thermo-couples in a plane of measurement passing through the centre of the instrument case

(b) (right) Mobile thermo-couple mechanism

points, while isotherms drawn for every 10° of temperature rise show the internal distribution. Colour bands 20° wide define progressively warmer areas up to a maximum of 100°C, above which red indicates a 'very hot' area. By noting the dwindling of the 'hot' areas and the growth of those 'cool', a subjective impression is obtained of improvements made in successive experiments, with the temperatures at the selected points A, B and C extracted and placed (in that order) below each experiment number for the purpose of more objective comparison.

Radiation Versus Ventilation

There appears to be a tendency, when considering the relative merits of the three means of cooling, to disregard the contribution made by radiation unless the surface in question reaches a temperature at which it advertises its dissipation by glowing a cherry-red. The first six experiments were therefore devoted to demonstrating the considerable drop in internal temperatures that is attainable by improving the emissivity of container surfaces a mere 30° or 40°C above room ambient and to showing, by contrast, the relative inadequacy of the conventional louvres provided in the sides of the container for cooling by ventilation.

The instrument case was obtained in a clean (i.e. unpainted) condition, with the welded joints carefully smoothed to prepare external surfaces for a high-gloss finish calculated to overcome the sales-resistance of potential customers. Fitted with a clean front panel, and with ventilation inhibited by shutting all the louvres with adhesive tape, the 100W load produced a temperature rise of 125°, 67° and 69°C at the selected points A, B and C respectively in Exp. 1. Restoring ventilation by opening all the louvres in Exp. 2 reduced these temperatures to 120°, 60° and 56°C. The temperature drop due to ventilation was therefore :—

	Exp. 1		Exp. 2	
A	125°	–	120°	= 5°
B	67°	–	60°	= 7°
C	69°	–	56°	= 13°

25° ÷ 3 = 8·3° mean ventilation drop in clean container.

Omitting the intermediate stage, making the same comparison between Exp. 5 and Exp. 6 yields :—

	Exp. 5		Exp. 6	
A	95°	–	88°	= 7°
B	32°	–	25°	= 7°
C	40°	–	34°	= 6°

20° ÷ 3 = 6·7° mean ventilation drop in black container.

Reductions due to increased radiation loss from surfaces of higher emissivity are given by comparing Exp. 1 with Exp. 5, thus :—

	A	B	C
Exp. 1	125°	67°	69°
Exp. 5	95°	32°	40°

30° + 35° + 29° = 94° ÷ 3 = 31·3° mean radiation drop in unventilated container.

And again, comparing Exp. 2 with Exp. 6:—

	A	B	C
Exp. 2	120°	60°	56°
Exp. 6	88°	25°	34°

32° + 35° + 22° = 89° ÷ 3 = 29·7° mean radiation drop in ventilated container.

From these comparisons it is seen that, irrespective of finish, introducing ventilation reduced internal temperatures by some 8°, while improvements in surface emissivity produced a drop of about 30° in both unventilated and ventilated containers. The sum of the two should result in a total drop of 38° between the clean and unventilated container of Exp. 1 and the high-emissivity ventilated container of Exp. 6. This supposition is supported by a subjective impression that isothermal levels had changed by about 40° and confirmed by the more objective method of comparison:—

	A	B	C
Exp. 1	125°	67°	69°
Exp. 6	88°	25°	34°

37° + 42° + 35° = 114° ÷ 3 = 38° total mean drop.

It will doubtless surprise instrument case makers to learn that louvres provided in the pious hope that they would accomplish all that was necessary in the way of cooling made only 8° difference, whereas sacrificing æsthetic values by substituting a more utilitarian finish dropped internal temperatures by 30°, relegating the louvres to the realms of little more than decorative appendages. Emerging from this is the sober thought that there must exist many thousands of glamorous instrument cases of dazzling beauty concealing blistering components, including (in those claiming a higher order of precision) compensating devices to combat a temperature rise that could have been avoided by a more complete understanding of the factors contributing to efficient cooling.

Heat Transfer Within Containers

Having demonstrated that deliberate suppression of ventilation had very little effect, the mechanism of heat loss from a totally enclosed container could be considered in detail.

Dealing first with external surfaces, heat loss by conduction was eliminated by the rubber feet provided and loss by radiation minimized by the initially clean surfaces in Exp. 1, leaving convection as the sole effective means of dissipation. Convection loss depends upon surface temperature to the power 5/4, is nearly independent of finish, but is twice as effective from upward-looking surfaces as from those downward-looking, with vertical surfaces intermediate between the two. To obtain optimum convection loss it is clearly desirable to subtract heat from the less effective surfaces and to add it to those capable of a greater rate of dissipation.

This function was performed by the internal closed convection systems clearly visible in Exps. 1, 3 and 5. Heated air in the immediate vicinity of the heat source expanded and, becoming less dense, rushed upwards in an attempt to establish free convection. Deflected from its purpose by the horizontal container top to which it gave up most of its heat, the air stream searched for a means of escape by moving towards a cooler surface until, having traversed the top, it slithered frustrated down the vertical side to a level at which its loss of heat and gain in density allowed it to repeat the cycle. By this means a practically constant temperature difference of some 14° was maintained between the hottest and coldest points on the container throughout this series of experiments despite the fact that the high conductivity of the metal container tended to equalize the temperature rise, as did the radiation reaching the internal surfaces from the heat source.

Had these experiments been conducted in a vacuum, a symmetrical radiation pattern would have appeared around the source, with an equal heating of the equidistant surfaces above and below it. As heat transfer by radiation takes place irrespective of the intervening medium, the existence of this pattern had to be taken for granted however heavily it was masked by the much more obvious convection stream. With conduction prevented by an insulated heat source, internal convection aided by radiation transferred energy to the container skin at a rate that produced an equilibrium temperature averaging 35·1°C., agreeing with that predicted from Fig. 1 for convection loss alone.

The contribution to heat transfer by radiation within the container could not have been very great owing to the reluctance of the clean internal surfaces to absorb radiated energy, which was reflected into the interior and gave rise to large temperature gradients between points on the skin and the nearest isotherms, ranging from some 30° at low levels to more than 60° at top centre. This reluctance to absorb radiated heat was overcome by painting all internal surfaces matt black to increase their absorptivity, with the results shown in Exp. 3. Reduced temperature gradients indicated that energy previously reflected was now being absorbed, lowering internal isothermal levels by approximately 20°. The greater absorption rate had to be equalled by a higher rate of dissipation from external surfaces by convection alone, a demand that could be met only by an increase in equilibrium temperature, which actually rose by 3° to an average of 38·1°C, while objective comparisons gave an internal drop of 19·7°. This seeming anomaly would doubtless confuse those pundits who, by placing their hands on the container, would judge Exp. 3 to be the hotter, whereas the internal temperatures were some 20° lower than in Exp. 1.

Continuing the process of increasing heat transfer by radiation, all external surfaces were painted matt black to increase their emissivity in order that radiation loss could aid dissipation by convection. Objective comparisons gave a further drop of 11·6° in internal temperatures to indicate a higher rate of heat transfer through the container in Exp. 5. As no change had been made in the absorptivity of internal surfaces, the greater rate of transfer could have been achieved only by increasing the temperature difference between heat source and container, which actually became cooler by 11·4°. This reduction in equilibrium temperature was the result of vastly increased radiation loss from external surfaces of improved emissivity, offset by a smaller reduction in convection loss due to the selfsame reduction in surface temperature: an example of the 'swings-and-roundabouts' relationship that makes calculation of thermal losses so intricate and justifies a non-mathematical approach based on experiments and logical deduction.

The overall effect of increased radiation transfer by improved absorptivity and emissivity of container surfaces was to reduce internal temperatures by the total of 31·6° previously noted, with the average skin temperature rise falling by 8·4° from 35·1°C in Exp. 1 to 26·7°C in Exp. 5. Because emissivities of unity and zero are not attainable in practice, these reductions are less than suggested by Fig. 1, while recorded temperatures may be considered as the maximum attainable because the plane of measurement passed through the hottest portion of the container.

Ineffective Ventilation

Compared with the figures above, the intended method of cooling by ventilation contributed an average of only 8·2° internally and 4·3° externally in Exps. 2, 4 and 6 with surface finishes as in the left-hand half-sections, but with louvres open to admit a stream of cold air at lower levels and extract heated air at upper levels. Had this theory worked, a change in isothermal patterns would have given a subjective impression of the desired effect, but the patterns remained substantially similar to those in the totally enclosed containers except in the neighbourhood of the side louvres remote from the heat source. Some improvement might have been made by placing the outlet louvres as high as possible on the container side with their apertures uppermost to reduce the frustration of the internal convection stream by providing an easy escape path for heated air, with the inlet louvres lowered to allow cold air to enter at the very lowest level.

From the tortured shapes of the 50° and 30° isotherms in Exp. 2 and Exp. 4 respectively, it can be deduced that the incoming cold air, instead of assisting the internal convection stream in its clockwise movement, disrupted the pattern by trying to escape via the outlet louvres without doing useful work, with more than a suspicion that some of this 'lazy' air took the shortest path up the side of the container in a stream so thin that it could not be detected by the means of measurement employed.

The disappointing temperature reductions obtained by this form of ventilation were undoubtedly secondary effects due to cooling of the container by an air stream that could not attain its primary target, the heat source. The fault lay not in the louvres, but in their positioning—in this regrettably representative instrument case—as far as possible from the heat source.

Improving Ventilation

In marked contrast to totally enclosed containers in which the skin is first encouraged to absorb internal heat and then to dissipate it from external surfaces, the object of ventilation should be to reduce all forms of heat transfer to the container and to cool the heat source more directly by an air stream entering and leaving the container through properly positioned low-impedance vents. The following experiments show how modifications to the container attained this object with more substantial temperature reductions.

As frustration of the internal convection current appeared to be the main cause of heat transfer to the container, the logical point of attack was the container top. An aperture 11in by 3½in was therefore cut in the upper surface directly above the heat source and covered with a blackened plate in which were punched a total of 14 louvres, 7 per half-section.

The two 5½in outlet louvres in the side of the case were sealed with adhesive tape, as were all but three (per half-section) of the 3½in louvres on the top, the idea being to stimulate a 'transfer' of outlet vents of comparable impedance. The result, Exp. 7 on Plate 2, was subtle rather than spectacular. While internal temperatures dropped by a mean of only 1·3°, container surfaces at the upper corner grew cooler by 3° as an indication that less heat reached them by means of a circulating convection current and more was escaping through the top vents, which was the desired effect.

Reducing the escape path impedance by opening all seven louvres in Exp. 8 removed almost all evidence of frustration, the upper corners growing cooler by a further 2° and internal temperatures falling by 8° at point A and 6° at C, with a disproportionately large drop of 12° at point B due to 'lazy' air taking the diagonal path between inlet and outlet vents.

Sealing the side inlets and cutting an aperture 11in by 3½in in the base immediately below the heat source removed this effect without, however, showing any improvement in cooling. The impedance of the black perforated metal sheet (of the meat safe type) covering the aperture was suspected until it was realized that the controlling factor was the height of the container base above the bench. The ¼in provided by the rubber feet was extended to ⅞in by standing them on wooden blocks with the results shown in Exp. 9.

The side inlets were re-opened in Exp. 10 to see if additional cold air would be of benefit. Repeated measurements showed that, having established 'straight through' ventilation, the admission of cold air at a point where it could do no useful work reduced cooling efficiency because its comparatively high density added dead weight to the warmer air to which it attached itself, retarding the upward flow and raising both internal and external temperatures, in this experiment, by 1°. The effect was small, but the principle so clearly indicated that the side inlets were promptly sealed up and not used again.

It was noted, in all experiments with top louvres, that quite a high temperature was attained by the central louvre in the hottest portion of the air stream, but that the corners of the case were isolated from the centre by the louvre apertures interrupting a conductive path along the plane of measurement, although they conducted freely to the front and rear of the container. To obtain a measure of isolation in both planes the louvres were replaced by a black perforated sheet in Exp. 11, obtaining a record drop of 4° at the upper corner, with internal temperatures some 7° cooler than in Exp. 9 and a smooth isothermal pattern suggesting the minimum of impedance to ventilation.

Presented with identical aperture covers, the opportunity was taken of comparing their impedances in inlet and outlet positions by removing the upper one in Exp. 12 and the lower one in Exp. 13 on Plate 3, from which it appeared that greater impedance was offered to the entry of cold air of higher density than to the exit of hot air of lower density. No valid reason could, in fact, be found for obstructing the inlet aperture, which was left open for the remaining experiments. Manufacturers of instruments intended for operation in the tropics may interpret the term 'open' quite liberally by providing an expanded metal grille to prevent the entry of scorpions and other local fauna.

Opening both apertures in Exp. 14 gave a glimpse of the ultimate in cooling, an arrangement made more practical in Exp. 15 by placing a 12in by 4in black metal cover plate $\frac{3}{8}$in above the aperture in imitation of the Victorian magic lantern. As in Exp. 9, ventilation was impeded by a surface too close to the aperture, but increasing this distance to $\frac{3}{4}$in in Exp. 16 attained internal and external temperatures lower than those with the perforated metal top of Exp. 13 and only slightly higher than in the completely open Exp. 14.

Two details not made obvious in the diagram contributed to the satisfactory results in Exp. 16. The first was that $\frac{1}{4}$in spacers of insulating material prevented re-heating of the cool container by conduction from the hotter cover plate, and the second was that frustration heating of the cover plate was not as extensive as it appears from measurements along the 6in half-length because a 2in half-width escape path was available along 22in of the aperture circumference perpendicular to the plane of measurement. This is mentioned because some manufacturers favour a cover plate of 'top hat' section with apertures at the ends, an illogical arrangement offering the longest possible frustration path with maximum heat transfer to a cover plate integral with an instrument case intended to be kept cool by ventilation.

Results of Improved Ventilation

Considering Exp. 16 to be representative of good practice, the step-by-step temperature reductions in the histogram (Fig. 2) may be more vividly presented by direct comparison of the separate, and joint, results of enhanced radiation and improved ventilation. With the effect of an all-black container already evaluated, the additional temperature reductions due to improved ventilation are given by comparing Exp. 16 with Exp. 6, thus:—

	INTERNAL		EXTERNAL	
	A	B	C	F
Exp. 6	88°	25°	34°	23°
Exp. 16	54°	9°	10°	8°
	34°	16°	24°	15°

Additional temperature reductions due to improved ventilation.

The combined results of enhanced radiation and improved ventilation are given by comparing Exp. 16 with Exp. 2, thus:—

	INTERNAL		EXTERNAL	
	A	B	C	F
Exp. 2	120°	60°	56°	31°
Exp. 16	54°	9°	10°	8°
	66°	51°	46°	23°

Total temperature reductions due to improved radiation and ventilation.

Stripping the high-emissivity finish off both internal and external surfaces of Exp. 16 without any other modifications raised temperatures at all points in Exp. 17, but enabled an estimate to be made of temperature reductions due to improved ventilation in the original clean condition by comparing Exp. 17 with Exp. 2, thus:—

	INTERNAL		EXTERNAL	
	A	B	C	F
Exp. 2	120°	60°	56°	31°
Exp. 17	62°	22°	16°	11°
	58°	38°	40°	20°

Temperature reductions due to improved ventilation alone.

Appraisal of Results

As the wide variation between temperature reductions at various points would have made average figures misleading, final comparisons were made on a percentage basis with the low-emissivity surfaces of the original container and its decorative louvres in Exp. 2 representing the least effective cooling by either means.

TEMPERATURE REDUCTIONS
(Per cent)

	INTERNAL		EXTERNAL		
Exp.	A	B	C	F	Method.
2 - 6	27	58	39	26	Improved radiation.
2-17	48	63	71	65	Improved ventilation.
2-16	55	85	82	74	Combination of both.

While the means of measurement demanded an absence of internal structures, the exclusion of conduction did not change the total dissipation because it could only have transferred heat from the source to points remote from the plane of measurement and blunted the sharp contrast it was desired to make between the mechanism of heat loss by radiation and ventilation.

The very substantial reductions obtained by combining these two methods of cooling stress the importance of each when efforts are made to give components maximum protection from thermal effects. Ventilation is naturally superior because aiding the internally-generated convection current is a more direct method of removing heat than forcing it into the container skin and dissipating it from external surfaces, although this process must continue due to the temperature gradient still existing between heat source and surrounding container surfaces. Even under the nearly ideal ventilating conditions of Exp. 16 the reduction of radiation loss by stripping off the high-emissivity finish resulted in the higher temperatures recorded at all points in Exp. 17 because a greater load was placed on the ventilating system—a reminder that the 'swings and roundabouts' principle should never be neglected.

Sacrificing utility to glamour being a compromise often practised in industry, the designer may be forgiven for having been forced into an inferior solution providing he does not allow his efforts to be ruined by the crowning folly pictured in Exp. 18. The container was a clean version of that in Exp. 16 with emphasis upon cooling by 'straight through' ventilation. Large apertures were provided directly above and below the heat source with frustration reduced to a minimum by adequately spacing the insulated cover plate and by supporting the base a reasonable distance above the bench. Nevertheless, introducing a conventional 'inverted tray' chassis denied cold air access to the heat source by forcing it towards the sides and out through the upper aperture in a clearly-defined stream of 'lazy' air that cooled external surfaces without making much impression on internal temperatures, which shot up to levels encountered in the original container before modification. Increased radiation loss would have undoubtedly eased the situation, although it is obvious that the better solution is to design the container and its interior as an entity to be cooled by the combination of every possible means of heat transfer and dissipation.

Finally, it should be pointed out that the cooling methods outlined above, while applicable to single instruments standing on a bench, are practically useless in an assembly of similar instruments mounted one above another on a rack, a condition that makes greater—but not insuperable—demands upon the ingenuity of the designer.

Rack-Mounting Units

Having made the point that internal arrangements can thwart an otherwise effective form of cooling, it seemed futile to develop ideal instrument case layouts if these did not also meet the very much more exacting requirements of rack-mounting units which, once mastered, would suit both forms of construction. Attention was therefore transferred to a more detailed study of heat loss from rack-mounted units in an endeavour to evolve a system of cooling, aided rather than inhibited by internal layout, that would be as effective on the bench as in a rack full of similar units, the exacting requirement in the latter case being that cooling should take place with maximum efficiency without heat transfer from, or to, other units in the rack.

Starting with exploratory measurements, this second series of experiments was devised to show the shortcomings of conventional arrangements (including an interim solution) before passing on to the final phase in which logic and experience were combined in a systematic control of heat flow to attain the desired results.

The basic assembly comprised a 19in by 8¾in front panel with a 6in deep, 8½in high and 16½in long dust cover forming an enclosure with a surface area of 600in². The plane of measurement having to include internal structures, all measurements were made across the centre of the container with mobile thermo-couples exploring the 8¼in by 6in cross-section and fixed thermo-couples measuring the temperature rise at various points on the container skin and panel.

Temperature-measuring facilities were kept very flexible to allow both mobile and fixed thermo-couples to be added during the progress of experiments until all 30 were, in fact, deployed.

Changes in internal structure did not permit the extraction of temperatures at selected points for purposes of objective comparison, although the coloured isothermal patterns enabled subjective comparisons to be made, with heat source and rack temperatures as additional criteria of the improvements made from experiment to experiment.

Heat Source

As the proportions of the rack-mounting unit resembled those of the instrument case, the heat source was made up in a long and narrow form similar to the 100W vitreous resistor. Comparable thermal output was attained by using ten miniature pentodes with a maximum dissipation of 12·5W each, mounted at 2in centres in two parallel rows ¾in apart, the rows being staggered to reduce convection heating of the upper valves by those below when the valves were disposed horizontally. To add realism to the measurements each valve was provided with a 3W vitreous resistor mounted below the 2in wide metal deck, whose up-turned edges gave rigidity and provided a means of fixing to structural members introduced at a later stage. This arrangement, shown in Fig. 5, resulted in a compact heat source without undue cramping, the distance between centres of adjacent B9A valve bases being 1¼in.

Triode-connected as simple cathode-followers in accordance with the circuit diagram Fig. 6, the dissipation of the valves (and their loads) could be controlled over a wide range by adjustment of grid and anode potentials, although it was found convenient to eliminate one variable by running the valves at a constant current of 45mA to ensure that each load resistor dissipated its full rated wattage. Under these conditions it was found possible to vary the valve dissipation from a maximum of 12·5W to a minimum of some 6W each, below which the reduced anode potential would not sustain a current of 45mA. For any combination of anode and grid potentials that

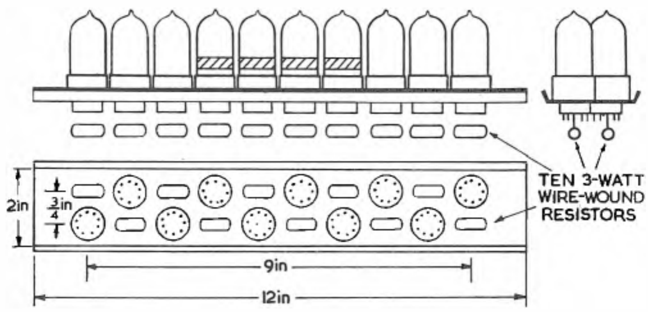


Fig. 5. Dimensions and layout of valve deck used as a heat source. Shaded areas indicate temperature measuring windings on four of the ten 12·5W pentodes employed

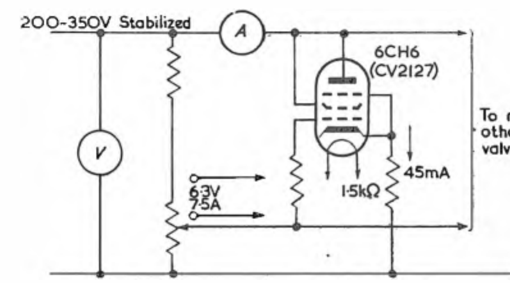


Fig. 6. Circuit of valve deck and power supply

gave a total current of 450mA, the anode dissipation was taken as the product of the indicated current (I) and the indicated voltage (V) minus the 67·5V dropped across the cathode loads.

The total dissipation thus comprised a controllable 60 to 125W from the anodes, plus a constant contribution of 47·25W from the heaters and 30·375W from the load resistors, giving a range of some 140 to 200W from all sources. Preliminary tests showed this range to be more than adequate for the effects it was desired to demonstrate and all experiments were actually made at a constant dissipation of 10 watts/valve, equivalent to a total dissipation of 178W from the heat source.

With greater dissipation in a smaller space the thermal loading rose to 0·3W/in², more than double that in previous experiments with the instrument case.

Valve Temperatures

However convenient for other purposes, thermo-couples are by no means ideal for measuring valve temperatures. It is difficult to ensure good thermal contact with the glass, and the temperature gradient along the envelope can be so considerable that a single measurement made at a random point on one of a group of valves is likely to prove misleading, especially if changes are made in the mounting position, convection stream or re-radiation from structural elements. It was thought that more reliable results would be obtained from a single measurement if it indicated the average temperature rise over appreciable areas on several valves. This was achieved by winding some 20 to 25 turns of p.t.f.e.-insulated 38-gauge copper wire round four valve envelopes just above the valveholder skirts. The four windings were connected in series and the average temperature rise calculated from their change in resistance.

As other measurements were to be made at the centre of the container, these windings were placed (Fig. 5) on the four valves in the middle of the deck with the remaining three valves on either side guarding against end-effect. No difficulty attended the measurement of deck temperature rise, a thermo-couple being soldered to a tag at its geometric centre.

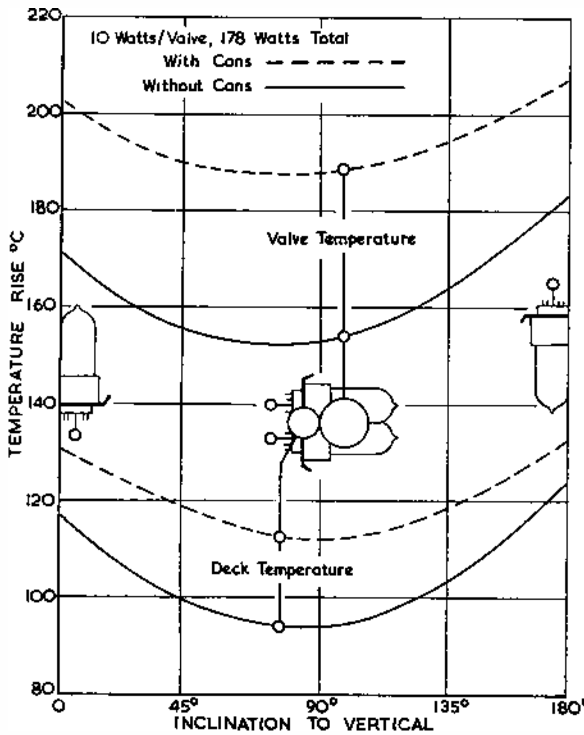


Fig. 7. Temperature-orientation curves for heat source with, and without, screening cans on the valves

Before building the heat source into a rack-mounting unit some preliminary measurements were made in free air with, and without, screening cans. Fig. 7 shows the effect of different mounting positions upon valve and deck temperatures, from which it is clear that, with no conduction and low radiation, optimum convection cooling takes place with the valves horizontal. The effectiveness of convection in this optimum position is emphasized by the temperature rise recorded when cans were placed on the valves. Despite the addition of comparatively large black surfaces of high emissivity, which must have increased the radiation loss, the net effect of shielding the valves from direct contact with free air was to raise their temperatures by some 30° and to increase conductive flow into the deck, which grew hotter by some 20°. At positions other than optimum the effect of the cans is not so marked, indicating that the deck itself impedes the free flow of cooling air when it is in a horizontal position. Having established that convection cooling is optimum with the valves horizontal, curves were prepared (Fig. 8) of valve and deck temperatures with the valves in this position.

Apart from stressing the importance of avoiding horizontal surfaces when relying upon convection cooling, these preliminary measurements in free air gave some indication of the temperatures to be aimed at in the process of cooling the heat source after it had been placed in its container. It could not be expected that ventilation would prove as effective as convection in free air, although making the deck an integral part of the structure would remove a great deal of heat by conduction. Bearing in mind this compensating action, it was estimated that it should be possible to attain valve temperatures equal to those in free air if the deck could be made considerably colder.

With an approximate target set for minimum temperatures, all that remained was to decide from what upper level to commence experiments designed to achieve this target. As published data for the valves used limits the

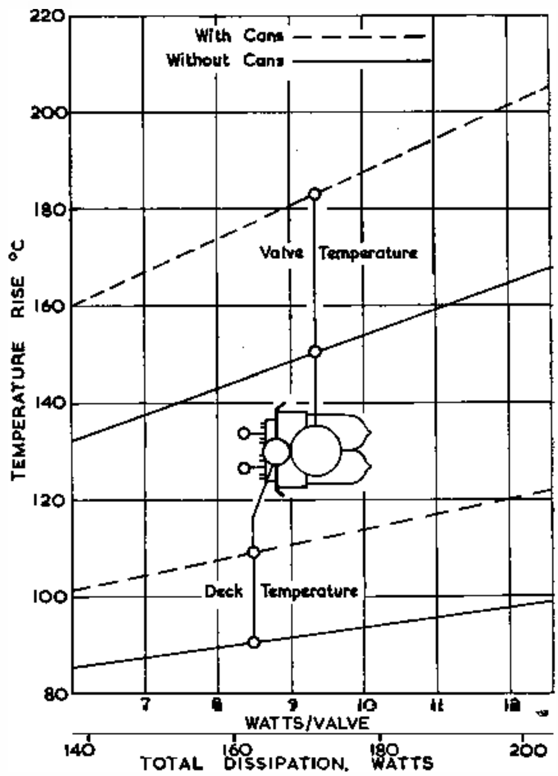


Fig. 8. Temperature/dissipation curves for valves in horizontal position with, and without, screening cans

maximum glass temperature to 250°C, this figure was adopted as the upper limit. No doubt was felt about being able to reach 250°C within an unventilated container, although it was not anticipated that before this temperature could be achieved—and maintained for long enough to enable measurements to be made—the heat source would twice melt and have to be re-built!

How Hot Can It Get—And Still Keep Working ?

The evils of excessive temperature rise in equipments were touched upon in general terms in the introduction to this work. As an indication of the faults that can occur it might be profitable to record the practical difficulties encountered in maintaining the reliability of a test-piece deliberately operated at a high temperature.

While p.t.f.e.-insulated wire was used for the temperature-measuring windings in anticipation of 250°C on the valve envelopes, it was assumed (quite erroneously) that no further precautions need be taken, ordinary phenolic valveholders being used and wired with p.v.c.-covered wire. Fitted just below the top of the dust cover of a clean and un-ventilated rack-mounting unit, without a conducting path to any of the surrounding surfaces, the heat source warmed up rapidly as its dissipated power was progressively increased. The valve temperatures had barely reached 200°C when reduced current readings indicated that all ten valves were not taking their fair share of the load.

Inspection revealed a sorry state of affairs. The p.v.c. covering, dripping from the cooler extremities of the deck, had vanished completely from the hottest central part, re-appearing as a mist-like deposit on all interior surfaces. The phenolic valveholders were warped and blistered, accounting for the poor contact with the tarnished valve pins, and some silver-plated stand-off tags were also completed blackened by the sulphur released from the rubber anti-vibration mountings used as non-conducting supports for the deck.

The remedy seemed obvious. p.t.f.e. valveholders were substituted and the wiring insulated with fibre-glass sleeving which, after it had boiled off its varnish solvents and turned black, gave no further trouble. Measurements of temperature rise in free air were repeated to cover differences in conductivity between phenolic and p.t.f.e. bases and the heat source again mounted within the container on ceramic non-conducting supports.

On this occasion all went well up to about 220°C, above which point the valve currents (now separately metered for each valve) became intermittent and sensitive to vibration, although these odd effects disappeared with the inrush of cold air immediately the container was opened. This trouble was traced to the melting of soldered joints at the hottest part of the valve deck. Re-wired anew with each joint crimped in addition to being soldered, the heat source was brought up to the desired temperature and held there despite an occasional drip of molten solder.

Bizarre as these conditions might be, it was reassuring to find there was little chance of exceeding maximum valve temperatures without ample warning from very obvious secondary effects. On the other hand, it was apparent that catastrophic damage to components could be caused by the heat emanating from valves working within their temperature limits. The crux of the matter is that the disparity between working temperatures of valves and components is so great that the only effective means of co-existence within common containers is the separation of the useful functions of valves as electronic devices from their nuisance value as thermal menaces.

That this is a feasible proposition will be shown as the experiments proceed.

Exploratory Measurements

Plate 4 shows a number of exploratory measurements made with the heat source in three positions within containers with different finishes, both totally enclosed and ventilated. Each diagram represents the temperature distribution at the centre section of a unit mounted on a rack with its front panel on the right.

Exp. 19 was the one that gave so much trouble before reliable operating conditions were established. The dissipation from the heat source to raise valve temperatures by 230°C to the permitted maximum of 250°C was found to total 178W, of which 100W (10W/valve) were contributed by anode dissipation. This was adopted as standard dissipation for all subsequent experiments in this series, one of the tasks being to reduce the valve temperature rise from the 232°C of Exp. 19 to the target figure of 154°C for the same dissipation in free air (Fig. 7) with a simultaneous reduction of deck temperature rise from 215°C to a great deal less than 94°C.

A study of the extreme conditions recorded in Exp. 19 does not support the popular assumption that the heat source should be as high as possible within the container. While it is true that the isothermal pattern shows a steep gradient between maximum and minimum temperatures, the pattern is almost completely stratified, or stagnant, with the heat source 'stewing in its own juice' because little or no heat is removed from it by internal convection. This can be very serious when the heat source includes heat-generating components that contribute to, and have to function in, higher ambients than are considered to be safe for such components.

This consideration focused attention on the ten load resistors, rated at 3W dissipation in an ambient of 70°C, sweltering in an ambient exceeding 180°C.

Immediate relief was obtained in Exp. 20 by lowering the source to a position mid way between container top and bottom. Stagnant conditions were disrupted by the stirring action of the convection current, reducing valve and deck temperatures by 12° and 18° respectively and dropping the ambient in the immediate vicinity of the resistors by about 60°. Slight asymmetry was created by moving the heat source ½in from the vertical centre line towards the front panel (this will later be used as the standard position) with the resulting anti-clockwise convection flow clearly seen in the diagram.

Further downward movement of the heat source failed to produce any significant improvement in the immediate vicinity of the load resistors, although the valve and deck temperatures in Exp. 21 dropped by a further 5° and 7° respectively. Returned to its original central position, the heat source gave rise to symmetrical convection currents with the characteristic mushroom-like frustration where the rising column of hot air reached the top of the container.

Bearing in mind the effectiveness of the high-emissivity finish applied to the instrument case, these measurements were repeated with the container painted matt black on inner and outer surfaces (Exps. 22 to 24). Comparing each of these diagrams with the one immediately above it shows that temperatures dropped by an average of 28° for the valves and 45° for the deck. The isothermal patterns remained similar in form, but experienced a mass drop of at least 40° at all points. This improvement is 10° better than that obtained with the instrument case because the lower surface of the latter was unable to radiate to the polished surface of the bench, whereas no obstacle to free radiation existed for a single rack-mounting unit. In addition, rubber feet insulated the instrument case from the bench, while the rack-mounting unit was bolted to a fairly massive rack able to act as a heat sink.

The temperature rise of this rack was measured at the two front panel fixing bolts in every experiment and recorded at the upper and lower right-hand corners of each diagram for use as one of the parameters for estimating the degree of heat control achieved.

Summarizing these results, it can be seen that improving the emissivity of the unventilated container and moving the heat source within it resulted in overall temperature reductions (Exp. 19 to Exp. 24) of 42° for the valves, 63° for the deck and about 100° for the ambient in the immediate vicinity of the load resistors. Except for the latter, these improvements were due mostly to wholesale removal of heat by enhanced radiation, with movement of the heat source relieving local conditions by changing the distribution within the container without appreciable effect upon the total heat loss. This is clearly brought out by averaging the skin temperatures for each of the six experiments in numerical order, yielding 73°, 76°, 78°—55°, 57°, 57°, together with 42°, 43°, 44°—25°, 28°, 27° for the average rack temperatures, the small changes due to movement of the source being swamped by the much larger differences between low and high-emissivity surfaces, amounting to some 20° and 16° in skin and rack temperatures respectively.

References to Fig. 1 suggests that a loading of 0·3W/in² should produce a rise in skin temperature of 77° for a surface with zero emissivity and 38° for one with perfect emissivity. It is seen that the average skin temperatures obtained were within the bracket, with a tendency towards the upper extreme because the measurements were made at the hottest centre section of the container. Apart from this, manipulation of temperature gradients by movement of the heat source produced local skin temperatures both above and below the figures predicted by the graph, implying that the designer can not only reduce hot spots by movement of the heat source, but can also control the heat flow into, and through, selected portions of the container skin. This is particularly applicable to pressurized equipments, in which the usual internal blower used to equalize the temperature distribution could be dispensed with, by so grouping the heat-dissipating components that they produce an internal circulating current depositing heat on a selected portion of the container, which could then be blown externally with maximum effect.

The remaining three diagrams on Plate 4 show thermal patterns within all-black containers with upper and lower surfaces of perforated metal permitting 'straight through' ventilation. As valve and deck temperatures decreased with downward movement of the heat source the perforated metal must have had some frustrating effect, although the final temperatures recorded in Exp. 27 differ so little from those obtained in free air that the effect could not have been very great, showing, once again, with what efficiency ventilation removes unwanted heat. Deliberate use was made of the term 'unwanted' to stress the grave disadvantage of 'straight through' ventilation when such a unit is mounted with others in a rack. Not only will it transfer heat with the greatest efficiency to the unit above, but it will also accept 'unwanted' heat from the unit below. It is therefore necessary to regard 'straight through' ventilation with a certain amount of suspicion.

Another point worthy of notice in Exps. 25 to 27 is that, although ventilation removed very nearly as much heat as was lost in free air, sufficient was transferred to raise the rack temperature by an average of 9° in all three experiments. This could have been due to the small amount of frustration present, but its constancy with movement of the source suggested that it was caused by absorption of radiation from the valves by the blackened surface of the front panel. This is supported by the fact that the blackened surface at the rear, shielded by the deck, was cooler than the panel with its heat sink exposed to direct radiation from the valves. Thus, instead of travelling to infinity, radiation was trapped and made to give up its energy to the wanton warming of a heat sink that lost efficiency thereby—a point that must not be neglected in an approach to the final solution.

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(To be continued)

Holme Moss F.M. Transmitting Station

Holme Moss transmitting station, which was brought into regular operation on December 10, 1956, is the first of the new high-power v.h.f. stations to be opened in its permanent form with a three-programme service. It is built on the same site as the BBC's Holme Moss television station, adjoining the Holmfirth-Woodhead road some 18 miles south of Huddersfield.

The new station is expected to provide satisfactory reception in Yorkshire, with the exception of the northern and extreme eastern parts of the North Riding; Lancashire as far north as Morecambe Bay; Lincolnshire with the exception of the extreme eastern and southern parts; Cheshire; Derbyshire; Nottinghamshire; north Leicestershire; north Shropshire; most of Staffordshire; the north-eastern part of Anglesey; Flintshire and most of Denbighshire. There will be some locations within this area, particularly in valleys and behind hills, where reception is difficult or even unsatisfactory, as with television reception.

The transmissions will be horizontally polarized and will be on the following frequencies: North of England Home Service, 93·7Mc/s, Light Programme, 89·3Mc/s; Third Programme, 91·5Mc/s. The effective radiated power on each programme service will be 120kW.

The six 10kW v.h.f. transmitters, two for each programme service were manufactured by Marconi's Wireless Telegraph Co. Ltd. The transmitters use the 'FMQ' frequency modulation system, which consists essentially of a quartz crystal oscillator connected through a quarter-wave network to a balanced modulator, the susceptance of which is varied by the modulating signal, and this in turn varies the frequency generated by the crystal oscillator. The output of the crystal oscillator is multiplied by three stages of frequency doubling and one tripling stage, to produce the required carrier frequency.

There are five stages of carrier frequency amplification, two being of the conventional push-pull type, while three are single-ended earthed-grid stages, provided with coaxial line tuning elements. The final stage uses two BR191B valves operating in parallel. All valves are air-cooled and employ a.c. for filament heating.

The Holme Moss station is equipped with monitoring equipment which can give a continuous indication of the centre frequency of any of the three transmissions. It also includes high-precision low-frequency oscillators for accurate measurements of frequency deviation, noise and distortion.

The outputs of one Home Service transmitter, one Light Programme transmitter and one Third Programme transmitter are combined and connected to one half of the slot aerial. Similarly the outputs of the other three transmitters are combined and fed to the other half of the aerial via a separate feeder. Thus all three programmes are radiated by both halves of the aerial, thereby ensuring continuity of service in the event of failure of either half of the feeder/aerial system. This arrangement of transmitters also enables continuity of service to be maintained, at reduced power, should either transmitter in any or all of the three pairs of transmitters cease to operate.

Automatic equipment ensures that the outputs of the two transmitters on any one programme are correctly phased. As the v.h.f. equipment is designed to run virtually unattended the transmitters are switched on and off by time-switches and automatic monitoring equipment is provided. The associated fault-indicator panel and alarm system have been designed to call attention to any faults which develop during transmission, and also to give warning if the equipment is not in a suitable condition for automatic operation.

The v.h.f. aerial system is carried on the same mast as the television aerials. The mast is 750ft high and has a total weight of 140 tons. The maximum downward thrust on the base pedestal under the most severe conditions is 350 tons. The base is located by a two-inch diameter steel ball in a socket, which forms a pivot to allow angular movement of the mast in high winds. Up to the 610ft level the cross-section is triangular, each face being 9ft wide. Between 610ft and 710ft is located the cylindrical v.h.f. aerial. This consists of a hollow galvanised steel cylinder 6ft 6in in diameter and 100ft long, in which there are 32 slots arranged in eight tiers of four, giving a net power gain factor of 6:1. Thus the 20kW output from each pair of transmitters becomes 120kW effective radiated power.

Specially designed notch filters are included between the transmitter outputs and the combining units to reduce the possibility of spurious radiation caused by inter-modulation between the Home, Light and Third programme transmitter outputs.

Above the cylindrical v.h.f. aerial is a short square-section topmast which supports the television aerial.

Power for the television and v.h.f. equipment is supplied by the Yorkshire Electricity Board, at 11kV, three-phase, 50c/s over duplicate feeders which terminate on switchgear on the sub-station. This switchgear is so arranged that, in the event of a failure of supply over one feeder, it would be automatically disconnected and the second feeder switched in.

High Value Resistors and their Measurement

By G. France*

Very high value resistors must be measured many times during the production process. Various methods of measuring such resistors are considered from this point of view. An instrument is described which enables resistors of up to $10^{14}\Omega$ in value to be measured at 10V to an accuracy of about 1 per cent.

RESISTORS of value up to $10^{13}\Omega$ have become available in recent years, and find ready application when it is necessary to measure very small currents[1,2]. An application of particular importance is to be found in the expanding field of nuclear physics, it being often necessary to measure currents as small as 10^{-15}A, such currents usually originating in ionization chambers. If a current of this order of magnitude is made to flow through a high value resistor, the resistance of which is known and which is sufficiently stable, then the current may be measured in terms of the voltage drop produced across the resistor. It is, therefore, necessary that resistors having a range of values extending from 10^{8} to $10^{13}\Omega$ or higher be available and the properties of these components should be known to a high order of accuracy.

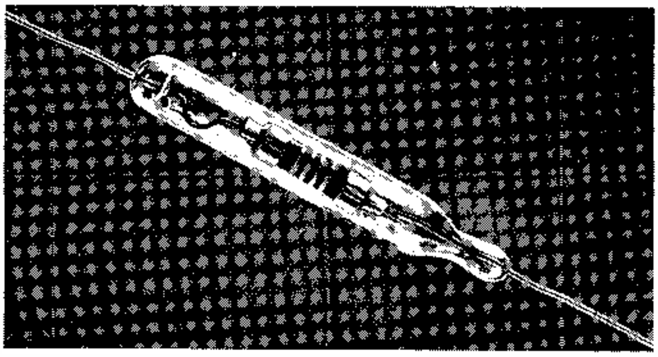

Fig. 1. A typical glass-enveloped very high value resistor

The resistors usually comprise high resistivity carbon composition films deposited upon superior grade porcelain rods, the films being cut in a helical manner in order to increase the length/width ratio of the conducting layers and thus increase the maximum resistance value which can be made from films of a given resistivity. After being fitted with caps, the components are sealed into siliconed glass envelopes, to protect them from the atmosphere, and are then given a lengthy stabilizing treatment. The electrical properties of the components are carefully recorded at various stages of manufacture and resistors of doubtful performance are eliminated, the remainder being a generally reliable product. The finished product is illustrated in Fig. 1.

The important properties of these resistors are:

Stability

The stability of the components is dependent upon the resistance value and operating temperature. Fig. 2 shows the relation between stability and value at ambient temperature. It will be seen that an annual drift of about 1 per cent occurs in resistors of low value, while the higher values may change 2 or 3 per cent per annum. This process is accelerated at higher temperatures and at 100°C the rate of change may be as high as 5 per cent per month.

* Welwyn Electrical Laboratories Ltd.

Temperature Coefficient

The temperature coefficient varies considerably between individual mixes, but it does not normally exceed $-0{\cdot}15$ per cent/°C.

Voltage Coefficient

This property varies markedly with resistance value; Fig. 3 illustrates this variation. For low values it is of

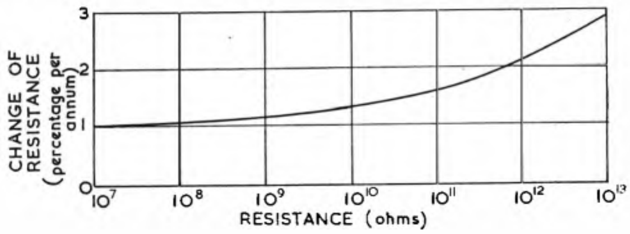

Fig. 2. Percentage change of resistance per annum under normal operating conditions

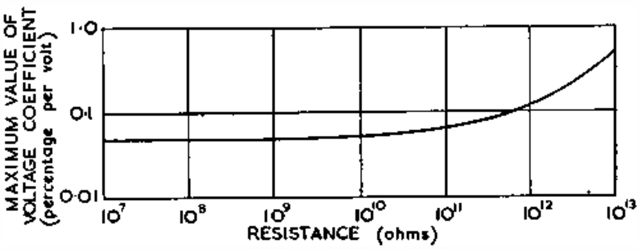

Fig. 3. Maximum value of voltage coefficient plotted against resistance value

the order of $-0{\cdot}05$ per cent/V; for higher values it might be as much as $-0{\cdot}25$ per cent/V.

Noise

The noise produced by these resistors, in excess of the normal thermal agitation noise, is of the order of 1mV/V. Its period is extremely long, being equivalent to a frequency of about 1c/s, this effect being due to the extremely narrow bandwidth over which it can be observed.

Measurements

The repeated measurements necessary at all stages of manufacture of this component call for the use of methods of measurement which require neither considerable skill on the part of the operators, nor an unduly long time for each measurement. Some consideration of various methods is necessary before deciding on the most suitable circuit for an instrument capable of measuring up to at least $10^{13}\Omega$ to an accuracy of about ± 1 per cent under production conditions.

The techniques which have been used for the measurement of very high value resistors may be conveniently subdivided into four main groups for further discussion:

(1) Time-constant methods, using capacitors.

(2) Constant current methods.

(3) Constant voltage methods.

(4) Bridge methods.

Time-Constant Methods

If a resistor of unknown value, R_x (Fig. 4) is connected in series with a capacitor C, and the capacitor is charged to a voltage E, by closing switch S, then the voltage V, across C at any instant t after closing S is given by

$$V = E\,(1 - e^{-t/CR_x}) \quad \ldots\ldots\ldots\ldots\ (1)$$

The time T, taken for C to charge to 63·2 per cent of its limiting value E, is given by

$$T = CR_x$$

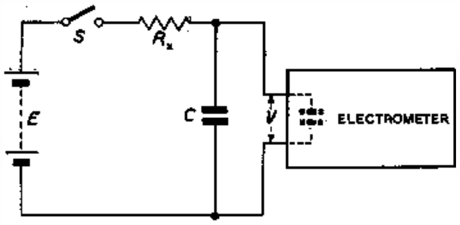

Fig. 4. Capacitor, rate of charge method

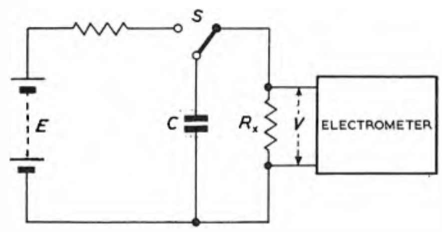

Fig. 5. Capacitor, rate of discharge method

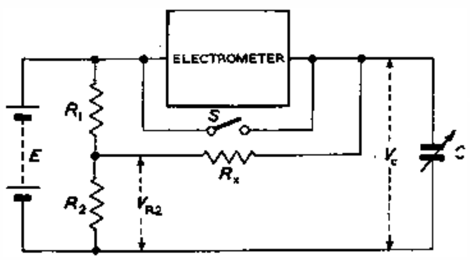

Fig. 6. Variable capacitor method

where T is in seconds

R_x is in mega-megohms

C is in picofarads.

Thus if the time taken for the capacitor to charge to 63·2 per cent of E was measured the value of R_x could be found from

$$R_x = T/C$$

It will be seen from Fig. 4 that the input capacitance of the electrometer is effectively in parallel with the capacitor C and hence it is not easy to arrange for the total capacitance to be made much less than 10pF. It is necessary, for accurate measurements, that the total capacitance be known to at least the same order of accuracy as it is desired to find R_x. The time involved in the measurement of a resistor of $10^{13}\Omega$ in value, assuming a minimum capacitance of 10pF and that the capacitance is to charge to 63·2 per cent of the applied voltage E, would be about 100sec.

The accuracy of this measurement is dependent upon the accuracy with which the total capacitance is known, the accuracy of measuring the time interval T and the voltage V to which the capacitance charges during this time. Additional errors could be introduced by any non-linearity in the resistor R_x and the capacitor C as the voltage across each is not constant throughout the measurement period. The self capacitance of R_x (usually nearly 1pF) must not be ignored[3] and, at extremely high values of test resistor the input resistance of the electrometer would be significant; for 1 per cent accuracy it must be maintained at a value at least 100 times greater than that of the test resistor.

The variation of this method whereby the capacitor is first charged to a voltage E and then the discharge through R_x is timed, is shown in Fig. 5. In this case the voltage across C at any instant t after the start of the discharge is given by

$$V = E\,(e^{-t/CR_x}) \quad \ldots\ldots\ldots\ldots\ldots\ (2)$$

This variation offers no advantage over the method just described.

A modification to the capacitor discharge method which overcomes many of the objections to the original is shown in Fig. 6.

Capacitor C is adjusted to maximum value and charged by closing switch S; if S is now opened, the voltage across C may be kept constant, as C discharges through R_x, by reducing the value of C continuously and at the appropriate rate.

The current I from C is given by

$$I = V_c\,(dC/dT)$$

Now if the electrometer input resistance is very high compared with R_x, the current I may also be written as

$$I = \frac{V_c - V_{R2}}{R_x}$$

So

$$\frac{V_c - V_{R2}}{R_x} = V_c\,(dC/dT)$$

and

$$R_x = \frac{V_c - V_{R2}}{V_c\,(dC/dT)} = \frac{V_{R1}\Delta T}{V_c\Delta C} \quad \ldots\ldots\ldots\ (3)$$

provided that the electrometer is maintained in the null condition throughout the time of measurement. The electrometer provides a means of continuously monitoring the difference between the voltage across C and the potential to which C was charged initially. The ratio $R_1/(R_1 + R_2)$ and the supply voltage V determine the current through R_x and the rate of discharge of C which, for any specified accuracy, determines the time taken by the measurement.

An important source of error in this method lies in variations in the rate of discharge of C, these variations resulting from failure of the operator to keep the electrometer strictly at null indication throughout the discharge time. Scott[4,5] describes a motor driven capacitor system which is capable of accuracies of 0·1 per cent when measuring resistance of $10^{12}\Omega$ at voltages of 1·5 to 180V and there is no reason to suppose that accuracies of 1 per cent and better for values of $10^{15}\Omega$ are not possible with automatic or manual drive of the capacitor, provided a null indicating electrometer of sufficient sensitivity is employed.

The time for each measurement may be quite long and a range of calibrated precision variable capacitors of high quality is required if one is to measure a wide range of resistance. For these reasons this method is considered to be more suitable for use in the laboratory than in a production department.

Constant Current Methods

Possibly one of the most obvious methods which might be employed in the measurement of high value resistors is to make use of the Ohm's Law relationship and deduce the resistance value from the voltage drop produced by the passage of a known, constant current through the resistor; in other words, measure the resistor by a method similar to that in which most of the resistors are to be ultimately employed. A constant current source is not difficult to provide when the resistance to be measured is not greater than about $10^7\Omega$. Well-known devices which have been used to give a constant current output include saturated diodes and pentodes operating in the flat region of their anode current-anode voltage characteristics. A radioactive source, enclosed in an ionization chamber, can give practically constant currents of about 1×10^{-9}A with a half-life of hundreds of years[6], but the simplest constant current device of all is merely a resistor of many times the value it is intended to measure, in series with the test resistor across a constant voltage supply (Fig. 7). If the standard resistor is high enough in value compared with any test resistor which it may be required to measure, the current will be almost independent of the test resistor value, being determined by the test voltage E and the standard resistor R_s, so that

$$R_x \simeq R_s(V/E) \qquad (4)$$

When $R_s \gg R_x$ and where V is the voltage across R_x.

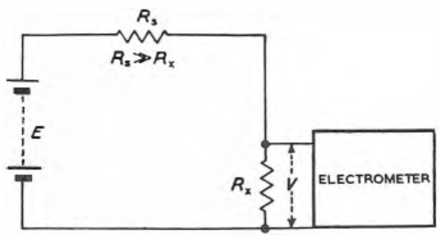

Fig. 7. Resistor as a constant current device

If the range of values of the resistors to be measured extends upwards from 10^7 to about $10^{13}\Omega$, the provision of a constant current source is by no means as easy. For instance, if a resistor is to be employed as a constant current device it must be at least 100 times greater in value than the greatest value of test resistance, otherwise the current in the measuring circuit will vary by more than 1 per cent with changes in the test resistor value. This means that a source resistance of $10^{15}\Omega$ is required when measuring test resistors of $10^{13}\Omega$ to 1 per cent; furthermore, the standard resistor of $10^{15}\Omega$ would be required to be accurate and stable to 1 per cent. At the present time no resistor of suitable stability and accuracy is available and so this method cannot be applied to the highest values of resistance. In such high values of resistance the stability of any type will be doubtful and it seems wrong in principle to measure a component in terms of a similar component of inferior quality.

Similar considerations indicate that the saturated diode or pentode constant current source would not satisfy the stability requirements when used with resistors much in excess of $10^7\Omega$.

It is, of course, possible to use a standard resistor comparable with the test resistor in value and to compute the value of R_x from

$$R_x = \frac{VR_s}{E - V}$$

but as this expression contains V as a reciprocal it is not convenient from a calibration point of view and any method involving calculation is not desirable in a production instrument.

Fig. 8 gives the circuit of a method of measuring very high value resistors which utilizes the displacement current of a capacitor as a constant current[7,8]. In this circuit R_1 is a potentiometer across a source of constant potential E_1; if the slider of R_1 is arranged to move so that V, the voltage between slider and ground, rises linearly with time, the displacement current through C, which is given by

$$I = C\,(dV/dt)$$

is constant.

This current, on passing through R_x produces a voltage drop across R_x which can be balanced by the voltage drop across the lower part of the second potentiometer R_2, thus producing a null in the electrometer. The potentiometer R_1, which may be constructed with a high degree of linearity, may be driven by a synchronous motor to provide a constant rate of change of voltage V applied to C.

R_x is given by

$$R_x = \frac{E_{R2}\Delta T}{C\Delta V} \qquad (5)$$

ΔT must be long enough to make any error in time measurement negligible and, in addition, time must be allowed for the stray capacitances of the circuit to charge in order that transients do not affect the constancy of the current produced during the observed period. Crawford[9] describes an electronic circuit which provides a constant current by displacement.

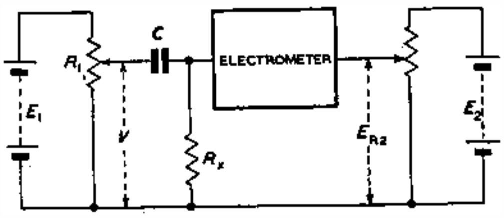

Fig. 8. Constant current method using displacement current of capacitor

A disadvantage of all constant current methods of measuring very high value resistors is that, in general, the input resistance of the indicating device, which is usually connected across the test resistor, must be very high indeed if it is not to make a significant contribution to the errors; at least 100 times as great as R_x if its shunt effect on R_x is to be less than 1 per cent. A circuit which has a source resistance about 100 times greater than the test resistor and has an indicating device of similar input resistance will have an input time-constant which is principally determined by the value of test resistor and the electrometer input capacitance. Considering a test resistor of $10^{13}\Omega$ in value and an electrometer of 10pF input capacitance, the time required for the voltage across the electrometer to approach 1 per cent of final value would be of the order of 400sec. This time is excessive for the applications being considered.

A further disadvantage of all constant current methods of measurement is that the voltage coefficient of the test resistor can introduce significant errors, as the voltage across the resistor being measured is dependent upon its value.

Constant Voltage Methods

Another method suggested by Ohm's Law is to apply a constant voltage to the test resistor and to measure the current which flows as a result. It has previously been stated that very small currents can be measured by allowing the current to pass through a standard resistor and measuring the voltage drop produced across the standard with an electrometer (Fig. 9).

The value of R_x can be seen to be

$$R_x = R_s \frac{(E - V)}{V} \quad \ldots\ldots\ldots\ldots\ldots (6)$$

or, if $R_x \gg R_s$

$$R_x \simeq R_s (E/V) \quad \ldots\ldots\ldots\ldots\ldots (7)$$

where R_s is the value of the standard resistor

E is the applied voltage

V is the voltage drop across the standard.

In order that the error in equation (7) shall not exceed 1 per cent, R_s must not exceed $R_x/100$ for any value of R_x likely to be encountered. This limitation severely restricts the magnitude of the voltage V, which is to be measured across the standard resistor, when the test voltage E is only a few volts. Taking the case of R_x being a maximum of $10^{13}\Omega$ and a minimum of $10^{12}\Omega$ on a given range, then R_s must not exceed $10^{10}\Omega$ if the error on some part of the range is not to exceed 1 per cent; when a test voltage of 10V is applied to the circuit, the voltage across R_x is only 10mV, when R_x is a maximum on that range. To measure 10mV to 1 per cent accuracy the electrometer used must be capable of discriminating against changes of 100μV in the 10mV signal. In addition, if 1 per cent is to be maintained, the input resistance of the electrometer must be better than $10^{12}\Omega$ in order that it does not shunt R_s by an amount which would introduce an error of 1 per cent.

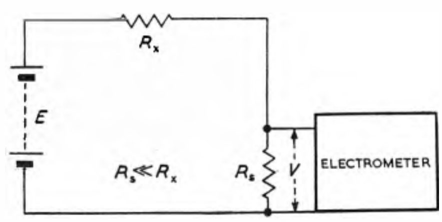

Fig. 9. Constant voltage method

It is evident that there are two primary advantages to be gained by using this constant voltage method over the constant current methods previously discussed; firstly, the standard resistor is not required to be as high in value, in fact, for comparable accuracy and stability, the standard resistor required to measure $10^{13}\Omega$ to 1 per cent is $10^{10}\Omega$ in the constant voltage case and $10^{15}\Omega$ in the constant current case, a value, as we have already seen, which is not practical. It is possible to obtain stable $10^{10}\Omega$ resistors without undue difficulty.

The second evident advantage of the constant voltage method is that the input resistance required of the electrometer is very much lower than is the case with the constant current circuits. Furthermore the constant voltage method eliminates the possibility of errors due to the voltage coefficient of the test resistor.

A disadvantage of the method is that it requires an electrometer which is much more sensitive than is the case with the constant current methods. The application of modern techniques to d.c. amplifiers, especially of the vibrating capacitor type, have shown that these sensitivity requirements can be met.

Bridge Methods

The Wheatstone bridge circuit is generally difficult to apply to the measurement of multi-megohm resistors, because

(a) Either a very high bridge ratio must be used and the sensitivity is consequently very low, or if a low bridge ratio is used the balance must be obtained by varying one ratio arm of the bridge with respect to the other, as there is clearly a limit to the maximum value of decade resistance which it is convenient to use.

(b) Either the supply to the bridge or the detector input circuit must be of the balanced type.

(c) The detector input resistance must be very high, in order that reasonable sensitivity be obtained and as the input capacitance of the detector cannot be negligible, the input circuit will have an appreciable time-constant. This time-constant will have the effect of making the act of balancing the bridge tedious because all variations in the detector input signal, produced by the balancing procedure, are subject to the time-constant of the detector input circuit.

A circuit* has been developed which eliminates the need for either a balanced supply or detector input circuit and which does not have a long time-constant to changes in signal level as the bridge is brought to balance. It combines the advantages of the constant voltage method (i.e., elimination of the errors due to voltage coefficient of the test resistor and the use of standard resistors which are considerably lower in value than the test component) with the ease of operation of a conventional bridge instrument.

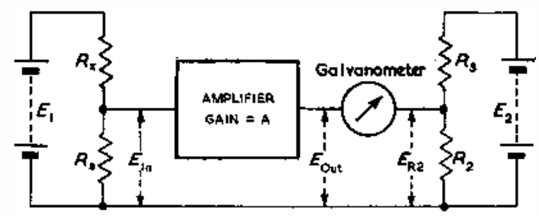

Fig. 10. Modified Wheatstone bridge method

If a d.c. amplifier, of very high input resistance (compared with the standard resistor R_s) and low output resistance, is interposed between the high resistance arms and the galvanometer and ratio arms of a Wheatstone bridge, as in Fig. 10, then

$$E_{out} = A\frac{E_1 R_s}{R_s + R_x} \text{ and } E_{R2} = \frac{E_2 R_2}{R_2 + R_3}$$

where A is the gain of the amplifier.

Now at balance

$$E_{out} = E_{R2}$$

So

$$\frac{E_2 R_2}{R_2 + R_3} = \frac{A E_1 R_s}{R_s + R_x}$$

and

$$R_x = R_s \; [(AE_1/E_2)\;(1 + (R_3/R_2)) - 1] \quad \ldots\ldots (8)$$

If it is postulated that $(AE_1/E_2) = 1$

then

$$R_x = R_s \,.\, (R_3/R_2) \quad \ldots\ldots\ldots\ldots\ldots (9)$$

which is the familiar Wheatstone bridge equation. $A(E_1/E_2)$ may be made equal to unity by making $A = (E_2/E_1)$ for any value of E_2/E_1.

The stability of the circuit is, of course, dependent upon the stability of the gain A and the stability of the ratio of the supply voltage, E_2/E_1. By the use of considerable negative feedback in the amplifier it is possible to achieve a sufficient stability and linearity of the amplifier to meet practical requirements.

* *British Patent Application* /56.

The sensitivity of this arrangement may be expressed as,

$$i_g = E_2 \frac{\left[\dfrac{R_2}{R_2+R_3} - \dfrac{R_2}{R_2+R_3(1+\delta)}\right]}{R_o + R_g + \dfrac{R_2R_3(1+\delta)}{R_2+R_3(1+\delta)}} \qquad \ldots\ldots\ldots\ldots (10)$$

where i_g is the current in the galvanometer

δ is a fractional unbalance of R_3

R_o is the output resistance of the amplifier

R_g is the galvanometer resistance.

If in Fig. 10, R_s is very small compared with R_x, the voltage applied to R_x during measurement will be practically constant. This condition, which is similar to that of the constant voltage case, discussed earlier, ensures that the error due to the voltage coefficient of the test resistor is negligibly small. The use of a standard resistor which is small compared with R_x makes it a much simpler matter to provide stable standards.

In practice standard resistors of up to $10^{11}\Omega$ in value are available with satisfactory stability. Thus with $R_s = 10^{11}\Omega$ and $R_x \geqslant 100R_s$ the value of R_x which may be measured on the highest range is not less than $10^{13}\Omega$. If each range is to cover one decade of resistance, the maximum value which can be measured on this range is $10^{14}\Omega$.

The relation that

$$A = E_2/E_1$$

suggests an easy way of providing the facility of different test voltages with a constant bridge sensitivity. If E_2 is fixed at as high a value as is convenient, in order to obtain high sensitivity, according to equation (10), then it is only necessary to adjust the gain A as the test voltage is changed and the sensitivity is unaltered. This facility enables the voltage coefficient of R_x to be determined accurately.

In order that measurements of different resistors may be accurately compared, all measurements, except those involved in the determination of voltage coefficient and other special requirements, are carried out at a test voltage of 10V. Using the bridge circuit just described it is evident that the amplifier input signal will be between 10 and 100mA on any one range of the instrument. The input resistance required is not less than 100 times the maximum value of the standard resistor which is used, i.e. not less than $10^{13}\Omega$. The drift of zero level which is characteristic of all d.c. amplifiers[10] must be less than $100\mu V$ over a period of an hour or more in order that the drift signal shall not introduce an error of more than 1 per cent into the measurement. Operation of the

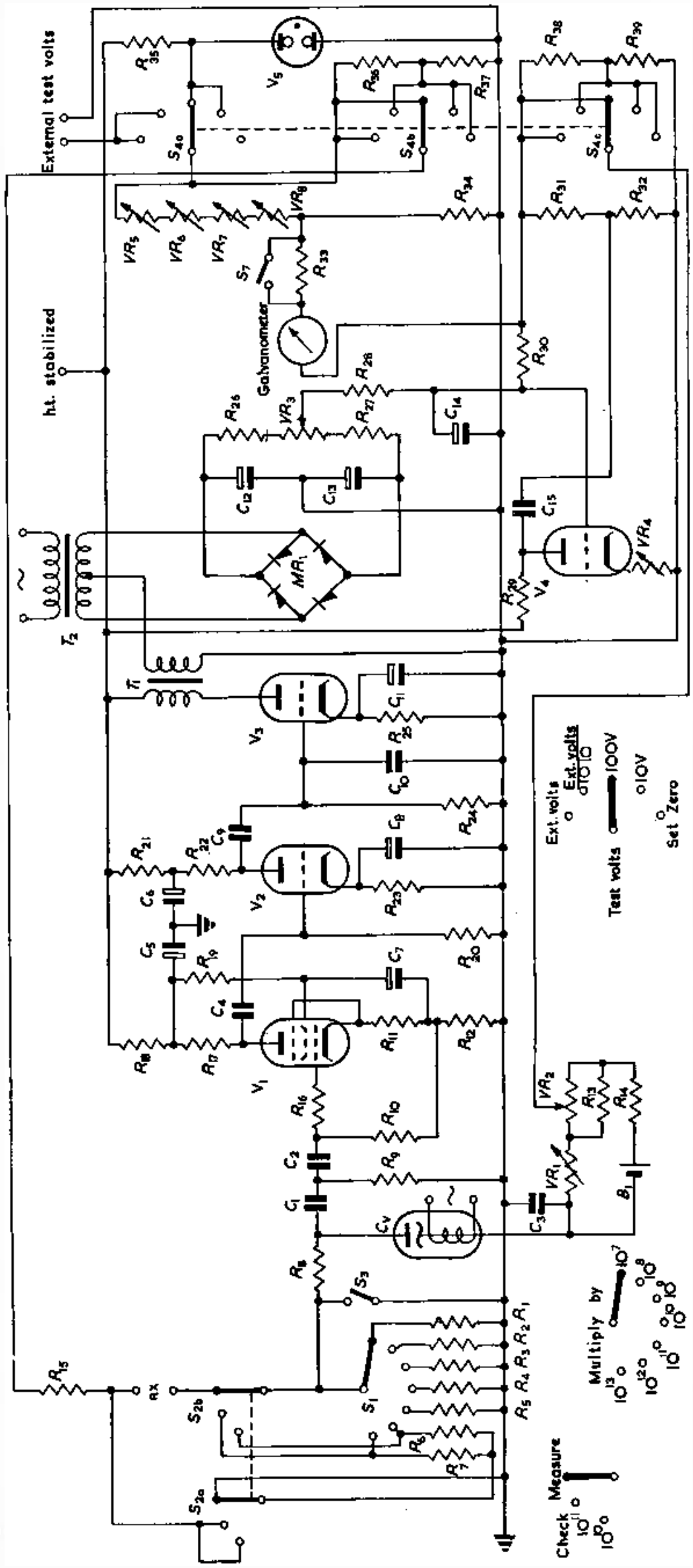

Fig. 11. Circuit diagram of bridge to measure 10^{8} to $10^{14}\Omega$

instrument by unskilled personnel makes it advisable to avoid the necessity for repeated adjustments.

The requirements of high input resistance and low zero drift suggest the use of a capacitor modulator type of d.c. amplifier. The drifts which occur in this type of amplifier are almost entirely due to variations in the contact potential of the vibrating capacitor. The design and construction of a capacitor for minimum drift has been discussed in detail by Palevsky[11], Chance[12] and others[13]. For minimum contact potential variations, the capacitor plates should be ground and optically polished, then both plates coated with pure gold simultaneously, by evaporation; the plates and supports must be scrupulously cleaned before assembly and the whole mechanism sealed in a container filled with an inert gas such as argon. To keep the leakage resistance high it is necessary that the terminal seal used to carry the connexion to the high impedance plate through the case be made of a material with very high volume and surface resistivity, such as quartz or high grade porcelain. Using a capacitor wherein all these precautions have been taken, and by the application of considerable negative feedback to the amplifier, it is not difficult to arrange for an input resistance of the order of $10^{13}\Omega$ and a drift of less than $20\mu V$ over short periods, in an amplifier of this type.

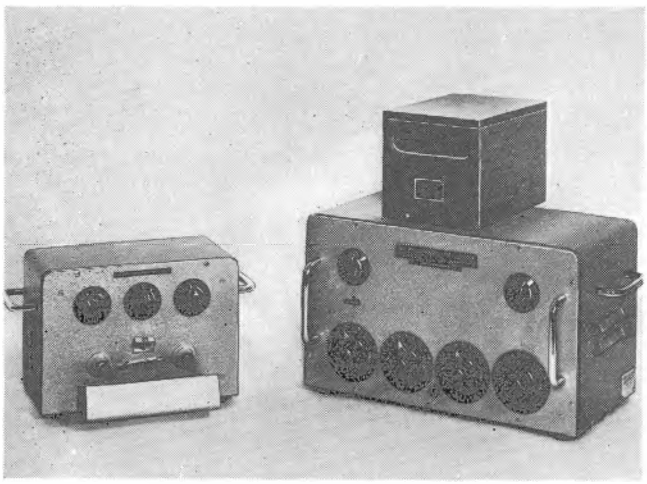

Fig. 12. The complete measuring instrument

Fig. 11 is a circuit diagram of an instrument designed to measure resistors from 10^7 to $10^{14}\Omega$ in value.

The d.c. amplifier comprises the vibrating capacitor C_v, the a.c. coupled amplifier $V_1V_2V_3$ and the phase sensitive detector, MR_1, which provides a d.c. output, a portion of which is fed back to the vibrating capacitor as overall negative feedback, via switch S_{4c}, resistors R_{38} and R_{39} and the network associated with the battery B_1. The hum cancelling circuit of V_4 improves the filtering of the phase sensitive detector output and allows greater loop gain to be used without introducing instability. R_{38} and R_{39} are adjusted to give stabilized gains of the d.c. amplifier of 1 and 10, dependent upon the setting of S_{4c}. Ganged with S_{4c} are S_{4a} and S_{4b}; S_{4b} switches the test voltage so that 100V is applied to R_x when the d.c. gain is unity and 10V is applied when the d.c. gain is 10; S_{4a} disconnects the internal 100V supply from the decade resistors VR_5 to VR_8 and ratio arm R_{36} and allows the application of other test voltages from an external source. Provision is made, on S_4, for the external test voltage to be divided by 10 before application to the test resistor, thus reducing the range of external supplies needed to test over a range of voltages. The battery B_1 and its associated network enable the residual contact potential of the vibrating capacitor to be balanced out and forms a convenient 'set zero' control.

The vibrating capacitor, together with V_1, the jig and the standard resistors are assembled in a sub-unit in order to reduce the effects of stray a.c. fields on the high impedance input circuit. The heater supply of the first three valves is derived from a d.c. source with the same object.

The test voltage is applied to R_x via R_{15} which is too low in value to affect the accuracy of measurement significantly but large enough to prevent a high current if the jib is accidentally short-circuited.

S_1 selects the appropriate standards R_1 to R_7 which range from 10^5 to $10^{11}\Omega$. All standards up to $10^9\Omega$ are high stability cracked carbon resistors; R_5, the $10^9\Omega$ resistor being made up of ten cracked carbon $10^8\Omega$ resistors in a series assembly. The 10^{10} and $10^{11}\Omega$ resistors are specially selected glass sealed resistors. In order that a facility might be provided for regular checking of R_6 and R_7 the switches S_{2a} and S_{2b} are arranged so that either R_6 or R_7 may be connected into the R_x position and hence measured against the appropriate cracked carbon standards.

S_3 earths the input to the vibrating capacitor amplifier while the jig is opened to insert or remove a resistor, thus minimizing the danger of stray charges being acquired by the input circuit.

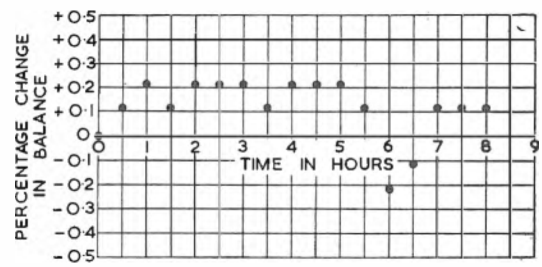

Fig. 13. Effect of zero drift expressed in terms of equivalent percentage out of balance

All insulation in the input circuit is polystyrene, similarly C_1 and C_2 are polystyrene film capacitors.

A series resistor R_3 is provided in the galvanometer circuit, which reduces the sensitivity when the bridge is far from balance and thus reduces the danger of damaging the galvanometer. S_7 enables this resistor to be shorted out for final balancing of the bridge.

Performance

The instrument, the circuit of which is shown in Fig. 11 and the appearance in Fig. 12, has an electrometer input resistance of about $2 \times 10^{13}\Omega$. The sensitivity is such that, under the worst conditions (it will be noted that equation (10) shows that the sensitivity is an inverse function of the decade resistance value) it is possible to detect an out-of-balance of 0·1 per cent. Residual noise, apparent as a jitter of the galvanometer spot, is equivalent to a variation around the balance point of about 0·25 per cent.

Zero drift on the $\times 10^{13}$ range is not greater than 0·3 per cent in 30min and on the $\times 10^{10}\Omega$ range is less than 0·1 per cent in a similar period. Fig. 13 is a plot of zero drift over a period of several hours.

Stabilizing the d.c. gain by overall negative feedback allows wide variation in the performance of the individual components without affecting the d.c. gain, in the least favourable condition, by a significant amount. The performance remains substantially unaffected by normal variations in the mains supply voltage.

Errors due to the temperature coefficient of the standard resistors have been practically eliminated by operating the apparatus in a temperature-controlled room which was already available.

Several of these instruments have been in use for some months and they have enabled high value resistors to be measured more rapidly and with consistently better accuracy than was previously possible under production conditions.

Acknowledgments

The writer wishes to thank the directors of Welwyn Electrical Laboratories Ltd, for permission to publish this article. Thanks are also due to Electronic Instruments Ltd, for the assistance rendered by making available one of their 'Vibron' vibrating capacitors.

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A Desk-Sized Electronic Computer

The Burroughs E.101 is a new digital computer which will mainly be used for engineering and scientific applications. The computer is arranged within a self-contained cabinet that is about the size of an office desk.

The following sections give a description of the various parts of the computer.

Input Keyboard

The input keyboard is centrally located before the operator. A number to be entered into the computer is entered on the keyboard by depressing the appropriate decimal keys for up to eleven decimal digits. Then one of the motor bars is depressed and the number is entered into the computer. At the same time the number is printed on the output page printer to give a record of the number. The four control keys for entering information into the computer control the page printing format through a mechanical program panel. A minus key inserts the number into the computer as a negative number, otherwise the number is entered as positive.

Output Page Printer

The output page printer system is integral with the keyboard, but will operate independently of it and consists of the gang printing mechanisms, the 18in carriage, and the mechanical program panel to control the format of the printing. The carriage can be fitted with form guides to permit front feed insertion of special forms if required. Instructions in the mechanical program panel cause the carriage automatically to close, to print and to open for removal of the form. When an instruction is received by the computer to print out a number, the instruction calls for one of four output possibilities, corresponding to the four input control keys or motor bars. These, in conjunction with the program panel, select the operations of tabbing, carriage returning, and line feeding.

The mechanical program panel controls printing format for bookkeeping and accounting applications and carriage opening and closing features.

Program Pinboard and Stepping Switches

The program pinboard is located to the right of the operator on top of the cabinet and at its side are located the stepping switches that advance the program step by step. Problems are programmed by plugging step by step operations in columns on the pinboards, a column for each operation, and when a pin is inserted into a hole, connexions are automatically made to carry out the instruction.

There are eight separate pinboards, each having its own stepping switch to select the sequence, and each containing up to sixteen program steps. The pinboards are removable so that a second program may be pinned up while one is being run. The program may also be scheduled on card templates, the cards being marked as a guide to indicate pin locations. The program can skip from one step to any other on the same pinboard, skip to another pinboard and begin at any of its sixteen steps, or repeat an instructional routine a controllable number of times.

The program is visible at all times, as indicated by the pins on the pinboard, and an illuminated panel at the side of the pinboards designates the pinboard and the program step that is currently in use.

The pinboards are divided into three principal sections for programming each step. The first section in general gives the operation to be carried out, i.e., arithmetic, read from keyboard or drum, write on drum or print, shift number right or left, or step the stepping switch. The second and third sections give either the tens and units digit of the drum location of the number to be operated on, or the pinboard location for the next instruction.

Manual Control Board

This is located to the operator's right. It contains the power control switches, several special signal lights, and all the controls necessary to operate the computer manually. The control panel is used in the initial set-up and start of a problem and also permits the operator to supplement or over-ride the automatic program contained in the pinboard. The judgment of the operator can thus be used to modify the course of the computation at any point in the problem.

Dial switches on the control panel duplicate all the computer instructions and permit manual changing of programmed drum locations. Other switches start and stop the machine and select the automatic or manual modes of operation.

Magnetic Drum

The magnetic drum is located in the cabinet beneath the pinboard section and contains magnetic recordings on the surface of the drum. Storage is provided for 100 or 220 numbers, the digits of each number being stored as from 0 to 9 pulses and the sign stored as no pulses for a plus sign and 9 pulses for a minus sign. Pulses at the rate of 75 000/sec are used as the basis for machine control and for such purposes as counting from 0 to 9, to present each digit being processed by the machine.

Electronic Circuits

These are located in the back part of the cabinet that swings open for access. The circuits perform the arithmetic operations, the control and counting operations and the output operations for printing, which are carried out by flip-flops, diode arrays, choppers, amplifiers, and thyratron pulsers, the circuit elements being grouped on printed circuit cards.

There are 51 cards to make up the circuits and these plug into place on the receptacle side of the chassis. In the performance of addition the digits are successively added together to give one number as the sum. Subtraction is performed by means of complementing and adding. Multiplication involves the over and over addition of the multiplicand with appropriate shifts, while division involves over and over subtraction with shifts.

Two arithmetic registers are provided in the machine. One of these forms a part of the accumulator which is used for addition and subtraction. The other stores the multiplicand and divisor for multiplication and division respectively.

Heater Voltage Compensation for D.C. Amplifiers

By J. B. Earnshaw*, B.Sc.

The concept of a fictitious voltage source to represent heater voltage fluctuations is examined critically and useful experimental results are quoted. This is followed by a detailed analysis of three compensation circuits whose advantages and disadvantages are discussed. Finally, experimental results and suggestions for future developments are given.

THE principal requirement in direct coupled amplification is the elimination of drift mainly due to varying supply voltage and changing cathode temperature. As long as valve characteristics vary with voltage and temperature, the quest for a stable, drift-free, d.c. amplifier will never be completely solved.

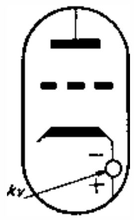


Fig. 1. Equivalent generator representing heater voltage variations

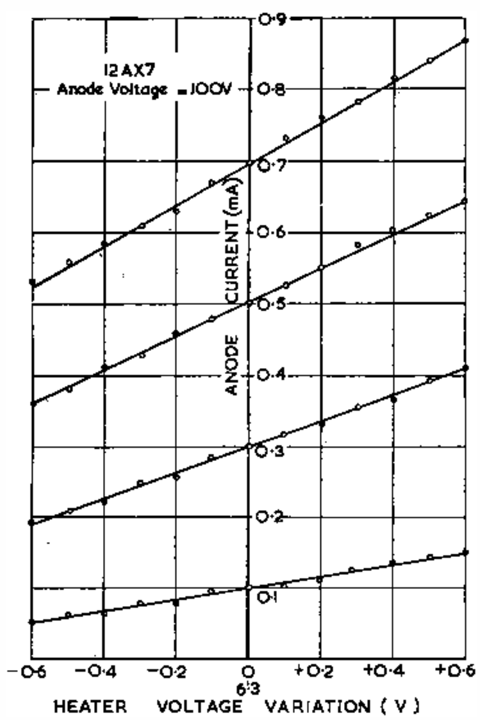


Fig. 2. Variation of i_a with v_h for different bias values

However, the problem has always been an interesting one; this fact can be judged by the extensive literature associated with the subject. Recent articles by Artzt[1], Verhagen[2], Yarwood and Le Croissette[3] give a comprehensive and up-to-date survey.

Essentially, in a perfect d.c. amplifier the valve currents depend solely upon the signal voltages, so that changes in h.t. and heater voltage (due to variations of the supply voltage of mains operated equipment) are completely compensated.

As almost any degree of stability of h.t. supplies can now be obtained[1,5], this article is concerned with alternative circuits for heater voltage (cathode temperature) compensation.

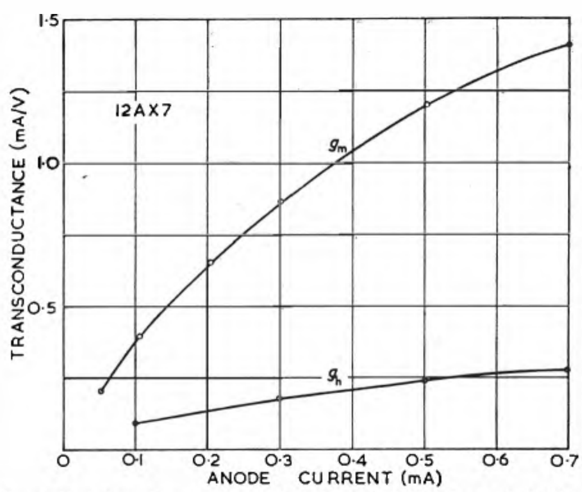


Fig. 3. Variation of g_m and g_h with i_a for a 12AX7 valve with 100V h.t.

Fictitious Voltage Source

Changes in the cathode temperature of a thermionic valve caused mainly by heater voltage fluctuations, alter the electron emission from the cathode by variation of both the work function of the cathode surface and the initial velocities of the emitted electrons. The effect is equivalent to a shift in the grid-bias with cathode temperature, and can be represented by a voltage generator in the grid-cathode circuit of the valve, located inside the valve. (See Fig. 1).

Experimental verification of this concept was obtained in the following manner: The current variations caused by a ±10 per cent change in heater voltage, for a 12AX7 triode, are shown in Fig. 2. It can be seen that at different standing currents the relationship between anode current and heater voltage is linear within the range investigated. The slope of the curves gives a value of g_h, where g_h is the heater-anode transconductance[5]. This is a function of i_a, the anode current for the nominal heater voltage. The grid-anode transconductance, g_m, was also measured and Fig. 3 shows the relationship between both g_h and g_m, and i_a.

The ratio of the transconductances is

$$(g_h/g_m) = (\partial i_a/\partial v_h)/(\partial i_a/\partial v_g) = (\partial v_g/\partial v_h) = k \text{ (say)} \quad .. \ (1)$$

* *Aukland University College, New Zealand.*

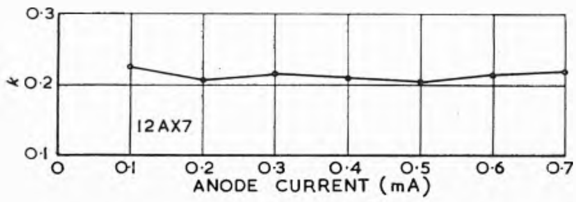


Fig. 4. Curve representing $k = \frac{g_h}{g_m} \simeq$ constant

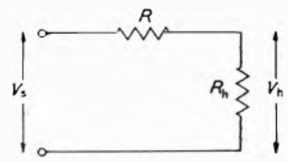


Fig. 5. Circuit representing a fixed resistance R in series with the valve filament R_h

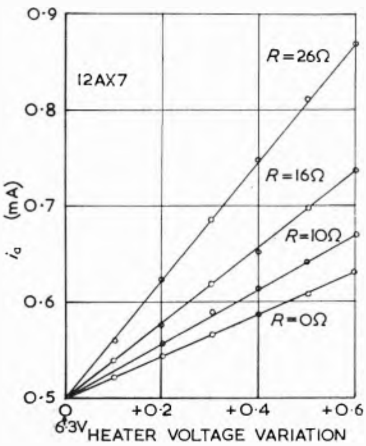


Fig. 6(a). Variation of i_a with v_h for different series resistances

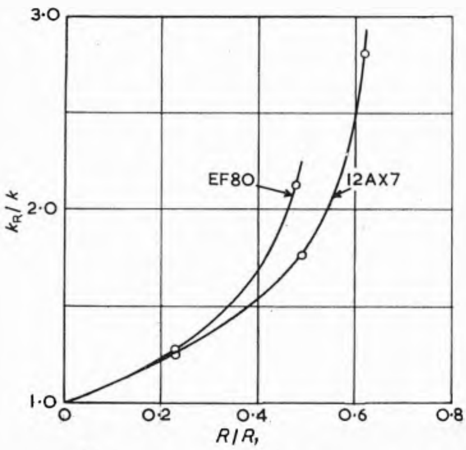


Fig. 6(b). $\frac{k_R}{k}$ as a function of $\frac{R}{R_h}$ for two specimen valves

where the suffixes h and g apply to the heater and grid circuits respectively.

The ratio k is essentially independent of i_a, as shown by Fig. 4 where the experimental variation amounts to less than 0·5 per cent of the mean value. Hence k is a true constant and the equivalent generator concept is valid under all conditions of operation. Consequently, the equivalent generator voltage, v_G, for a small heater voltage change v is

$$v_G = k \,.\, v \ldots \text{(see Fig. 1, for polarity)} \quad \ldots. \ (2)$$

Modification of k

The value of k can be modified by connecting a resistance in series with the valve heater.

Fig. 5 shows a circuit with resistance R in series with R_h, the filament resistance of the valve. R_h is non-linear being a function of the voltage v_h across it. If v_s is the supply voltage across R and R_h, then

$$v_h = \frac{R_h(v_h)}{R + R_h(v_h)} \,.\, v_s \ \ldots\ldots\ldots\ldots\ (3)$$

differentiating both sides with respect to v_h.

$$1 = \frac{R_h}{R + R_h} \,.\, (\partial v_s/\partial v_h) + \frac{v_s R}{(R + R_h)^2}(\partial R_h/\partial v_h) \ \ldots.\ (4)$$

rearranging the terms and substituting for v_h/v_s from equation (3),

$$\frac{(\partial v_h/v_h)}{(\partial v_s/v_s)} = \frac{1}{1 - [v_s R/(R + R_h)^2] \,.\, (\partial R_h/\partial v_h)} \ \ldots.\ (5)$$

Now since $(\partial R_h/\partial v_h) > 0$ the denominator of equation (5) is always < 1.

Therefore

$$(\partial v_h/v_h) > (\partial v_s/v_s) \ \ldots\ldots\ldots\ldots\ (6)$$

This means that the fractional change in voltage across the filament is greater than the fractional change of the supply. So that it is only possible to increase k by this method.

Experimental results of the effect of connecting resistances in series with the valve filament are shown in Figs. 6(a) and 6(b). k_R is the modified value of k and R_h is the nominal heater resistance. Worthy of note is the fact that R can also represent the internal impedance of the filament supply.

It is a well known feature of valve design that μ, the amplification factor, can be controlled within fine limits since it depends mainly upon geometrical considerations. On the other hand, g_m, i_a, and the dynamic anode resistance r, may vary by large amounts from valve to valve due to variation in cathode emission[6]. Since k depends upon these latter quantities, it follows that k may vary considerably and must be measured for any particular valve.

Compensation

For input signals of 0·1 to 1V the cathode drift is of little consequence, but when desired input signals fall in the range 1 to 10mV, compensation methods must be employed. The amount of compensation that can be obtained is directly dependent upon the accuracy with which the compensating voltage approaches the disturbing voltage.

A popular cathode compensation circuit is described by Miller[7], whose results are shown in Fig. 7.

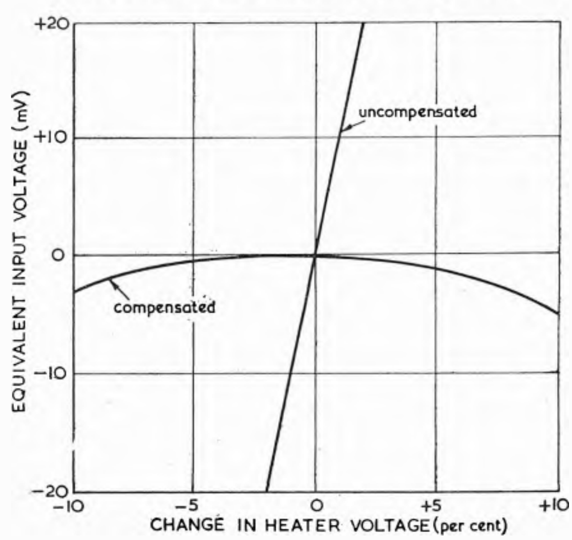
Further analysis of Miller's circuit seems profitable as it points to other methods for compensation having further advantages.

Compensation Analysis

Case 1

Miller's circuit is shown in Fig. 8. In this, for a given change in heater voltage v, let the equivalent generator

Fig. 7. Compensation obtainable with Miller's[7] circuit



voltage for the control valve, V_2, be k_2v and k_1v for the amplifier V_1.

With perfect compensation the current in V_1 will remain unchanged, so that the potential of point A (Fig. 8) must rise by k_1v so that v_k remains constant.

The current change* in V_2 is

$$i_2 = \frac{(1+\mu_2)\,.\,k_2v}{r_2 + R_1(1+\mu_2) + R_2}$$

$$\simeq k_2v\,.\,\frac{g_2}{1+g_2R_1} \quad \ldots\ldots\ldots\ldots\ldots\ (7)$$

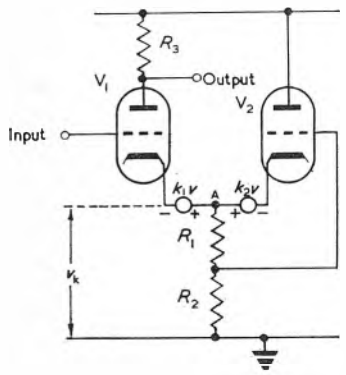


Fig. 8. Miller's circuit for heater voltage compensation

and since $v_k = 0$, then

$$k_1v = (R_1 + R_2)\,.\,i_2$$

$$= \frac{(R_1 + R_2)}{1 + g_2R_1}\,.\,g_2\,.\,k_2\,.\,v$$

and if $b = (k_1/k_2)$, then

$$b = \frac{R_1 + R_2}{1 + g_2R_1}\,.\,g_2 \quad \ldots\ldots\ldots\ldots\ldots\ (8)$$

which simplifies to

$$R_2 = (b/g_2) + (b-1)\,R_1 \quad \ldots\ldots\ (9)$$

If the valves are identical and $k_1 = k_2$, then

$$b = 1 \text{ and } R_2 = (1/g_2) \quad \ldots\ldots\ldots\ldots\ (10)$$

Under these conditions R_1 can be varied to control the current through V_2 and hence the value of g_2 until equation (10) is satisfied.

Case 2

When $b \neq 1$, the situation becomes somewhat more involved, for as R_1 varies so must R_2.

It can be arranged that $b < 1$ by connecting the valve with the smaller variation as the amplifier. From equation (9), with $b < 1$

$$R_2 = (b/g_2) - (1-b)R_1$$

so that with $R_2 = 0$ (Fig. 9), compensation can be obtained when

$$R_1 = \frac{b}{1-b}\,.\,(1/g_2) \quad \ldots\ldots\ldots\ (11)$$

In practice, the grid of V_2 can be returned to a variable bias which can be obtained from the negative supply—a common feature of most d.c. amplifiers. The value of b can also be reduced, if desired, by the addition of a small resistance in series with the filament of V_2.

Case 3

Where the addition of a battery operated bias to V_2 is not considered inconvenient the circuit of Fig. 10 is very useful.

$$\text{Change of current in } V_2 = \frac{1+\mu_2}{r}\,.\,k_2v$$

$$\simeq k_2v\,.\,g_2$$

and for compensation $k_1v = k_2v\,.\,g_2R_1$

therefore $\quad R_1 = (b/g_2) \quad \ldots\ldots\ldots\ldots\ (12)$

This equation can be satisfied for $b \geqslant 1$ or < 1 by variation of R_1 or g_2. The addition of a bias battery, operated under 'shelf life' conditions can hardly be considered a heavy price to pay for satisfactory performance with arbitrary valve selection.

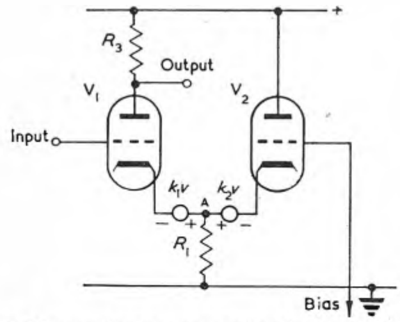


Fig. 9. Heater voltage compensation circuit when $k_1 < k_2$

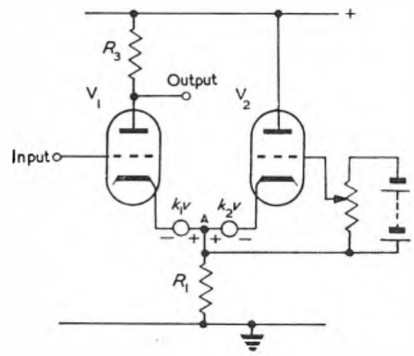


Fig. 10. Heater voltage compensation circuit when $k_1 \leqslant k_2$ or $k_1 \geqslant k_2$

Compensation Factor

The compensation factor, γ, is defined by

$$\gamma = \frac{(\partial i_a/\partial v_s)\text{ uncompensated}}{(\partial i_a/\partial v_s)\text{ compensated}} \quad \ldots\ldots\ldots\ldots\ (13)$$

where v_s is the supply voltage. Consequently $\gamma > 1$. The actual value will vary with the magnitude of δv_s and should be stated for a specific change, e.g. $\pm$ 10 per cent variation.

Figure of Merit

The improvement obtained in compensated circuits can be best expressed by a figure of merit, ϕ, defined by

$$\phi = (\text{gain}/\gamma) \quad \ldots\ldots\ldots\ldots\ldots\ (14)$$

for a given compensation factor. The circuit with the highest figure of merit will be most satisfactory. Since the input stage has a predominant effect, variations in subsequent stages are of lesser importance, unless $\phi < 1$, in which case the following stage should be compensated also.

Gain Analysis

The simplest gain analysis for Miller's circuit, Fig. 8, is to consider V_1 as an amplifier with an unbypassed cathode load of $(R_1 + R_2)/(1 + g_2R_1)$ (see Fig. 11—also Appendix equation 31).

* *See Appendix, equation (25).*

Feedback is provided by the cathode load such that

$$\beta = \frac{R_1 + R_2}{1 + g_2R_1} \Big/ R_3 \quad \ldots\ldots\ldots\ldots\ldots \quad (15)$$

Now as the circuit is adjusted for compensation, the relationships in equation (8) are valid, hence equation (15) simplifies to

$$\beta = (b/g_2R_3) \quad \ldots\ldots\ldots\ldots\ldots \quad (16)$$

The gain without feedback,

$$A = g_1 \cdot \frac{R_3r_1}{R_3 + r_1} \text{ as } R_3 \gg \frac{R_1 + R_2}{1 + g_2R_1}$$

$$= g_0R_3 \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (17)$$

where $g_0 = g_1/[1 + (R_3/r_1)]$

The actual gain of the circuit $A_1' = (A/1 + A\beta)$ and with substitutions from equations (16) and (17),

$$A_1' = \frac{g_0R_3}{[1 + b \cdot (g_0/g_2)]} \quad \ldots\ldots\ldots\ldots \quad (18)$$

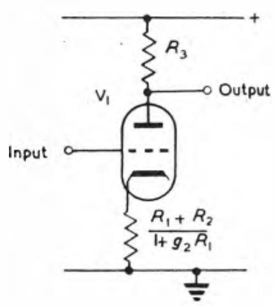


Fig. 11. Common equivalent circuit for gain analysis

Where pentodes are used to obtain higher gain, $g_0 \simeq g_1$. Cases 2 and 3 (see Figs. 9 and 10), can be analysed in a similar manner to the above. For Case 2, the cathode load is $R_1/(1 + g_2R_1)$ (see Appendix equation 32), making

$$\beta = \frac{R_1}{1 + g_2R_1} \Big/ R_3 \quad \ldots\ldots\ldots\ldots\ldots \quad (19)$$

and for Case 3, the cathode load is R_1, making

$$\beta = R_1/R_3 \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (20)$$

Substituting the respective R_1 values from equations (11) and (12),

$\beta = (b/g_2R_3)$ in both cases (cf. equation (16))

Hence in all three cases

$$A_1' = A_2' = A_3' = \frac{g_0R_3}{[1 + b \cdot (g_0/g_2)]} \quad \ldots\ldots \quad (21)$$

Summary

Case 1.

$$b \geqslant 1;\ R_2 = (b/g_2) + (b-1)R_1;\ A_1' = \frac{g_0R_3}{1 + b \cdot (g_0/g_2)}$$

Case 2.

$$b < 1;\ R_2 = 0;\ R_1 = \frac{b}{1-b} \cdot (1/g_2);\ A_2' = \frac{g_0R_3}{1 + b \cdot (g_0/g_2)}$$

Case 3.

$$b \geqslant 1 \text{ or } < 1;\ R_2 = 0;\ R_1 = (b/g_2);\ A_3' = \frac{g_0R_3}{1 + b \cdot (g_0/g_2)}$$

Although the gain appears to be the same in all cases, closer inspection will show that the various values of b used with each type of circuit will cause slight differences. Circuits employing $b < 1$, especially when pentodes are

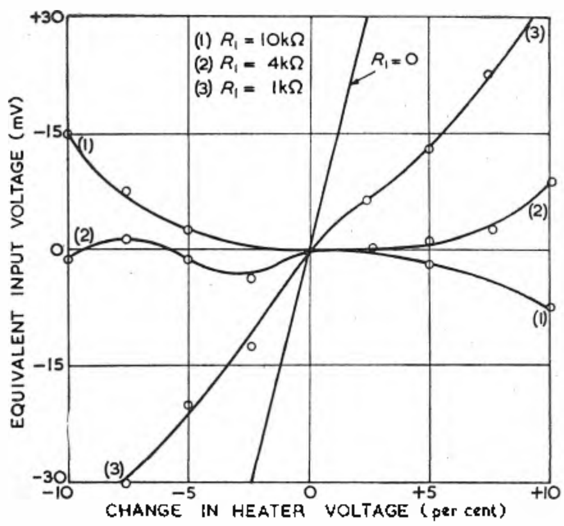


Fig. 12(a). Case 2. 12AX7 double triode with k_2 modified by connecting a 10Ω resistance in series with the filament

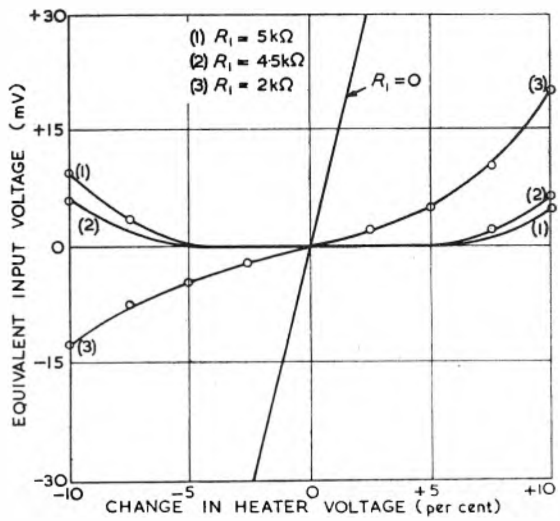
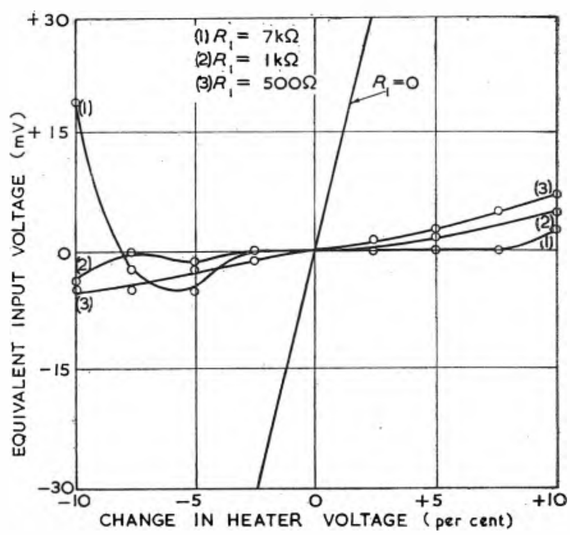


Fig. 12(b). Case 2. 12AX7 double triode with k_2 modified by connecting a 15Ω resistance in series with the filament

Fig. 12(c). Case 2. 12AX7 triode amplifier with EF80 (triode connected) as control valve



used, give the smallest reductions in gain. Improvements can also be obtained by using high μ-control valves.

Experimental Results

The graphs of Fig. 12 show the compensation obtainable with the circuit of Fig. 9. In (a), the amplifier was a 12AX7 triode passing 0·1mA, with a 12AX7 triode passing 0·5mA as control valve. A 10Ω resistor in series with the heater of the control tube increased its k value; in (b), conditions were as above except for a 15Ω series resistor; in (c) the control valve was an EF80, triode connected, passing 5mA, with a k value known to be higher than that of the 12AX7. The results of Fig. 13 were obtained using the circuit of Fig. 10, with an unmodified double triode 12AX7, one section acting as amplifier (passing 0·1mA and the other section as control valve passing 0·5mA.

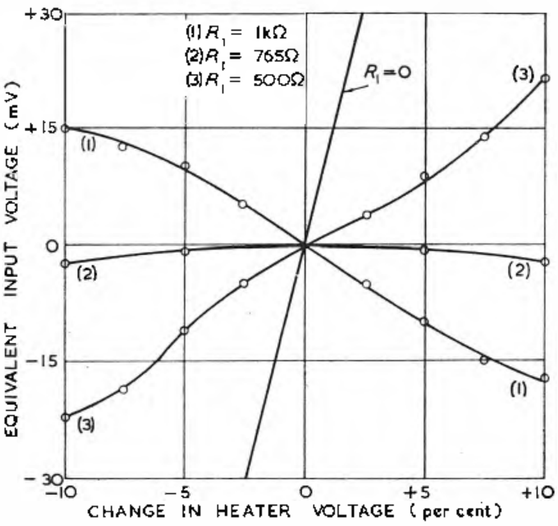


Fig. 13. Experimental results with case 3 using unmodified 12AX7 double triode

Conclusions

The circuits investigated give approximately the same amount of stabilization. The preference of any one over the others is dictated by the design considerations already discussed. Modification of k by the addition of resistors, is satisfactory for slow supply voltage variations only, as sudden changes cause unbalance due to the different time-constants in the amplifier and control valve heater circuits.

The above considerations have special significance in the design of cascode circuits as high gain d.c. amplifiers[4]. This aspect will be investigated in a further article with a different emphasis.

Acknowledgments

The author wishes to acknowledge the assistance and helpful criticism given to him by Professors P. W. Burbidge and K. Kreielscheimer during the preparation of this article.

Appendix

1. The change in current through V_2 when a voltage v is generated in series with R_1 is i, as shown in Fig. 14. As a result the grid-cathode voltage is increased by v_g where

$$v_g = v - iR_1 \quad \ldots \quad (22)$$

and the anode-cathode voltage is increased by v_a where

$$v_a = v - i(R_1 + R_2) \quad \ldots \quad (23)$$

From the accepted triode relation

$$i = gv_g + (v_a/r) \quad \ldots \quad (24)$$

and substituting from equations (22) and 23),

$$i = g(v - iR_1) + \frac{v - i(R_1 + R_2)}{r}$$

which reduces to

$$i \,.\, [r + R_1(1 + \mu) + R_2] = (1 + \mu) \,.\, v \quad \ldots \quad (25)$$

2. The effective impedance between cathode, A, and earth, for the circuit shown in Fig. 15, is the same as the parallel output impedance[8] at the cathode.

Considering the circuit as a modification of the cathode-follower, the gain from grid-to-cathode

$$A = \frac{\mu \,.\, (R_1 + R_2)}{r + R_1 + R_2} \quad \ldots \quad (26)$$

and when $(R_1 + R_2) \ll r$,

$$A = g(R_1 + R_2) \quad \ldots \quad (27)$$

The fraction of the cathode signal feedback to the grid is β where

$$\beta = \frac{R_1}{R_1 + R_2} \quad \ldots \quad (28)$$

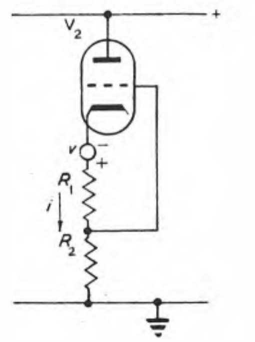


Fig. 14. Equivalent circuit showing change in current i with voltage v generated in series with R_1

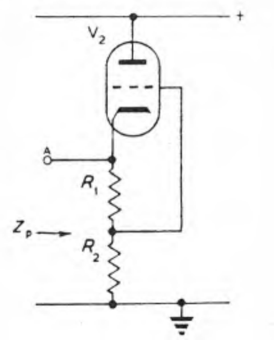


Fig. 15. Equivalent circuit comparing the effective impedance between cathode and earth and the parallel output impedance at the cathode

Thus the actual gain

$$A' = \frac{A}{1 + A\beta} = \frac{g(R_1 + R_2)}{1 + gR_1} \quad \ldots \quad (29)$$

Now the parallel output impedance, Z_p, of any simple amplifier, with or without voltage feedback is given by[8]

$$Z_p = (A'/g) \quad \ldots \quad (30)$$

so that

$$Z_p = \frac{R_1 + R_2}{1 + gR_1} \quad \ldots \quad (31)$$

Consequently the effective impedance between A and earth is given by Z_p in equation (31).

When $R_2 = 0$ equation (31) reduces to

$$Z_p = \frac{R_1}{1 + gR_1} \quad \ldots \quad (32)$$

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Surface Finish and Attenuation of Aluminium Waveguides

By J. Allison*, B.Sc., Ph.D., and F. A. Benson*, Ph.D., A.M.I.E.E., M.I.R.E.

The attenuation produced by a length of air-filled, rectangular, drawn aluminium waveguide has been measured at a frequency of 9 300Mc/s and the result is compared with figures previously obtained for drawn copper and brass and electroplated waveguides. It is shown that, as with copper and brass, chemical polishing or electro-polishing may be employed to improve the surface finish of drawn aluminium waveguides. Some information is also presented on the surface texture of cast and sprayed aluminium waveguide components.

REPORTS of large discrepancies between theoretical and experimental performances of certain conductors at microwave frequencies have been published by Vivian[1], Morgan[2], Maxwell[3] and others[4-10] and it has recently been concluded[11-13], from a series of measurements on drawn copper and brass and electroplated waveguides, that such discrepancies are due solely to surface roughness. The performances of drawn copper and brass waveguides have also been compared with electroplated, electro-polished, chemically-polished and electro-formed surfaces[13]. It has been shown that the surface finish of precision-drawn waveguide as manufactured at present is quite adequate for most applications, but it may be improved, if desired, by careful chemical or electrolytic polishing or electroplating in bright baths under closely-controlled conditions.

Measurements have now been carried out on a length of commercial drawn aluminium waveguide (internal dimensions 1in by ½in), to determine the losses in the metal walls and to compare the performance with copper, brass and electroplated guides. Samples of the drawn waveguide have also been chemically polished and electrolytically polished to determine the improvements which may be expected in the surface finish by such treatments. In addition, the drawn waveguide surface has been compared with the surface texture of cast and sprayed aluminium waveguide components. Although there is a great deal of published data on metal spraying there seems to be a scarcity of information about its possible application to the manufacture of waveguide components.

Calculation of Attenuation in Waveguides

A general formula for calculating the attenuation produced by losses in the wall metal of a waveguide has been derived by Kuhn[14] assuming ideally smooth internal surfaces. The expression can be applied to any wave mode in either rectangular or circular waveguides. It has been shown by the authors[13] that the formulae derived by Kuhn for the attenuation of either an H_{nm} wave or an E_{nm} wave in a rectangular guide may be modified to include the effects of surface roughness by introducing three additional factors K_{T1}, K_{T2} and K_p.

For the special case of the H_{01} mode, which is the one normally used, the modified formula is:

$$\alpha' = [(c/\sigma) \, . \, (\mu_1/\mu) \, (\epsilon/\mu)^{\frac{1}{2}}]^{\frac{1}{2}} \, . \, \frac{\lambda_g}{b \, . \, (\lambda_e)^{3/2}} \, . \, [(K_{T2} + (b/2a) \, . \, K_{T1}) \, \lambda_e^2/\lambda_{cr}^2 + K_p b(1 - (\lambda_e^2/\lambda_{cr}^2))/2a] \qquad \text{.............. (1)}$$

where α' = attenuation produced by wall metal taking account of surface roughness,

a = short internal dimension of waveguide,

b = long internal dimension of waveguide,

c = velocity of light,

σ = conductivity of waveguide wall metal,

μ_1 = permeability of waveguide metal,

λ_g = guide wavelength,

λ_e = wavelength in unbounded dielectric,

λ_{cr} = critical guide wavelength,

ϵ = permittivity of dielectric (normally air)

μ = permeability of dielectric (normally air).

Thus, the ratio of the calculated attenuation which takes account of surface roughness (α') to that which assumes a perfectly flat surface (α) becomes, for the H_{01} wave:

$$\alpha'/\alpha = [(K_{T2} + (bK_{T1}/2a)) \, (\lambda_e^2/\lambda_{cr}^2) + bK_p(1 - (\lambda_e^2/\lambda_{cr}^2))/2a] \div [(\lambda_e^2/\lambda_{cr}^2) + (b/2a)] \qquad \text{............ (2)}$$

Experimental Procedure

The attenuation produced in the drawn aluminium waveguide has been measured by the standing-wave ratio method previously described[11], at a frequency of 9 300Mc/s. The method consists of examining the standing-wave pattern, produced by the length of short-circuited waveguide, in the immediate vicinity of a voltage minimum. In order to use only fairly short lengths of guide the grid of the klystron oscillator was modulated with a square-wave voltage, at a frequency of 1 250c/s, and the output from the crystal in the standing-wave indicator was amplified by a variable-gain selective amplifier.

To compare the theoretical and experimental performances of a waveguide surface from attenuation measurements and calculations it is important to have an accurate figure for the d.c. resistivity if the comparison is to be of much value. The d.c. resistivity of the material used has been determined by measuring the resistance of a known length of guide. Such a measurement presents the difficulty of having to estimate the area of cross-section of a fairly long length of guide[11]. When the d.c. resistivity is known, the ratio of the measured attenuation to the calculated one, assuming a perfectly flat surface, can be found. If the values of the surface-roughness factors were known, the ratio α'/α could also be calculated from equation (2) for comparison. A method has been evolved by the authors[13] for obtaining the precise values of the surface-roughness factors so that the difference between actual and theoretical attenuation values can be predicted by calculation, without the necessity of making careful measurements on long lengths of waveguide. Briefly, the method consists of plating a sample with a protective metal, mounting a section in a plastic mould, polishing and finally examining the surface with a microscope. Because of the difficulties of plating aluminium the method could not be adopted in the present case and so it has not been possible here to confirm that equation (2) agrees with the measurement of α'. Chemical polishing of some of the drawn waveguide was carried out

* *University of Sheffield*

by immersing a sample for five minutes in the following solution[15]:

Phosphoric acid 80·5 per cent by volume.
Nitric acid 3·5 ,, ,, ,, ,,
Water 16 ,, ,, ,, ,,

After immersion, the sample was quickly washed and dried.

A sample of the drawn waveguide was also electropolished with an anodized finish. The contents of the polishing solution and the conditions under which it was used are as follows[16]:

Anhydrous sodium carbonate 15 per cent by weight
Trisodium phosphate 5 ,, ,, ,, ,,
Water 80 ,, ,, ,, ,,
Cathode—mild steel
Voltage—10V
Time—30min
Temperature—78 to 82°C.

The anodizing solution consisted of 25g of sodium bisulphate in 100g of water and was used with a stainless-steel cathode and a voltage of 12V. The anodizing operation was carried out for 20 to 30min at room temperature.

The electro-polishing procedure was as follows:

(1) Etch 20sec before switching on current
(2) Polish as above
(3) Wash rapidly in cold water
(4) Anodize as above
(5) Immerse in distilled water at 80 to 85°C for 10min
(6) Immerse in 10 per cent sulphuric acid for several minutes at room temperature to remove the white film.
(7) Rinse in cold water and dry.

To determine the improvements in the interior finish of the drawn waveguide due to the polishing procedures it was necessary to evaluate the roughness of the various metal surfaces with a "Talysurf" stylus type of instrument because the superior method due to the authors could not readily be used on aluminium as mentioned already. The principle of the "Talysurf" instrument is that a pick-up unit having a sharply pointed stylus is traversed across the surface by means of a motorized driving unit. The up and down movements of the stylus control a variable inductance by means of which the movements of the stylus are converted into corresponding changes in an electric current. This current is amplified and is then used to control:

(1) A rectilinear graph recorder providing a representation of the profile of a cross-section of the surface irregularities, and
(2) A scale-and-pointer instrument (called the average meter) which shows the centre-line average (c.l.a.) index of all irregularities having a predetermined maximum crest spacing (three ranges are provided, 0·1in, 0·03in and 0·01in). This quantity is the accepted index of surface texture in Great Britain.

Although the "Talysurf" gives an indication of both the centre-line average value of the surface roughness and the peak roughness, it does not provide a figure for the ratio of the actual length of a surface to its ideal length, owing to unequal horizontal and vertical magnifications. Care needs to be taken with centre-line average measurements because they give no indication of the shape of the profile and profiles of very different shape may have the same average height. The graphs from the recorder and centre-line average readings together do allow, however, certain comparisons of various surfaces to be made, but values of K_{T1}, K_{T2} and K_p for use in equation (2) cannot be found. The surface finishes of a cast aluminium waveguide (internal dimensions 3in by 1in) and a waveguide produced by spraying aluminium on a mild-steel mandrel have also been observed with the "Talysurf" equipment.

Results

The drawn waveguide tested had a d.c. resistivity of 2·82 × $10^{-6}\Omega$-cm, and the measured value of the attenuation at a frequency of 9 300Mc/s was 0·1185dB/m. The ratio of the measured attenuation to the calculated one, assuming a perfectly flat surface is therefore 1·115. This result should be compared with the figures for normal and precision drawn brass and copper waveguides and electroplated tubes presented in Table 1.

Tracings of typical surface-roughness recordings for the various waveguides, as obtained from the "Talysurf" instrument are shown in Fig. 1 and the mean of a number of centre-line average readings are listed in Table 2.

TABLE 1
Attenuation Ratios and Surface-Roughness Factors for Various Types of Waveguide
Frequency taken as 10 000Mc/s. Internal dimensions of guides 1in × ½in

TYPE OF WAVEGUIDE	K_{T_1}	K_{T_2}	K_p	α'/α
Normal-drawn brass*	—	—	—	1·060–1·074
Normal-drawn copper*	—	—	—	1·102–1·120
Precision-drawn copper	1·020	1·015	1·006	1·012
Precision-drawn copper*†	—	—	—	1·034
Precision-drawn brass	1·002	1·001	1·001	1·001
Precision-drawn brass*†	—	—	—	1·014
Acid-copper electroplate	1·625 (max.)	1·625 (max.)	—	—
Bright-copper electroplate	1·002	1·002	1·002	1·001
Silver - plate, from cyanide bath* ..	—	—	—	1·580
Bright-silver electroplate	1·012	1·004	1·006	1·007
Drawn aluminium* ..	—	—	—	1·115

* *These values have been determined experimentally from attenuation measurements.*

† *The measured values of attenuation cannot be compared with the estimated attenuation for similar types of guide because the roughness measurements were carried out on guides of slightly better quality.*

Conclusions

It is evident that drawn aluminium waveguide, although it has the advantage of being light in weight, has a high ratio of measured to calculated attenuation. The value of 1·115 for this ratio is much larger than the figures for precision-drawn copper and brass and normal-drawn brass guides and falls within the range of values 1·102-1·120 for normal-drawn copper tubes. That brass guides are much smoother than the aluminium and corresponding copper ones is probably due to the softer nature of the aluminium and copper and their consequent greater liability to scratching or to the higher friction between copper and aluminium and the drawing dies. Some early work by Vivian[1] showed that silver-plated aluminium guides produced large attenuations. This was probably largely due to the rather poor surface finish of the drawn aluminium tube as demonstrated here, although it should be remembered that electro-deposited surfaces can be much rougher than the original surfaces[12]. The theoretical value of attenuation produced by an aluminium waveguide is high compared with the figures for copper and silver because of the relatively large value of d.c. resistivity.

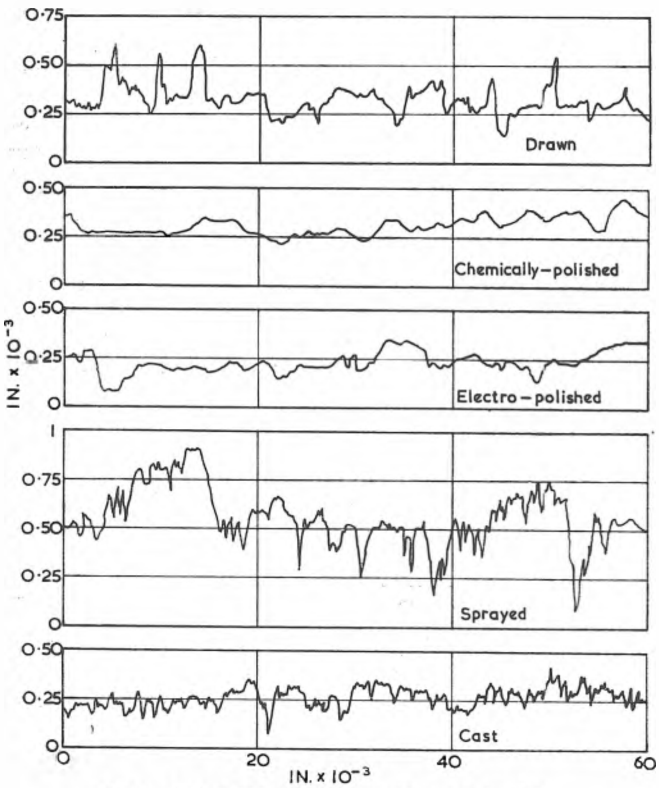

Fig. 1. Surface roughness of aluminium waveguides

TABLE 2

Centre-Line-Average Readings for Various Waveguide Surfaces as obtained from a "Talysurf" Instrument

Vertical magnification ×2000. *Cut-off wavelength* 0·03*in*

TYPE OF WAVEGUIDE	MEAN OF CENTRE-LINE AVERAGE VALUES (μin)
Drawn aluminium	57
Sprayed aluminium	50
Cast aluminium	25
Electropolished drawn aluminium	35
Chemically-polished drawn aluminium	20

From a general initial inspection of the various surfaces it was evident that both chemical and electrolytic polishing improve the surface texture of the drawn guide. Whereas electrolytic polishing removes the micro-profile only, the chemical polishing tends to obliterate draw lines too. The improvements are also apparent from the centre-line average readings and from the graphical results. It should be noted that industrial polishing techniques have been used throughout. It is seen from Fig. 1 that the cast-guide surface is fairly flat but is microscopically rough. The surface also contained a few large blow holes.

The sprayed-aluminium surface examined is not good. The texture of any sprayed waveguide surface, however, will depend on the initial flow characteristics of the metal on the mandrel as well as on the surface finish of the mandrel itself. Thus, if the metal does not flow easily on reaching the mandrel it is quite possible to form a waveguide with a rough interior finish on a very smooth mandrel. This is assumed to be the reason for the rather poor surface finish of the aluminium sample. This limitation can be obviated by a correct choice of sprayed metal, mandrel and spraying conditions as has been shown during some preliminary investigations on sprayed zinc surfaces. A further limitation of the method is the possible decrease in conductivity of the sprayed metal, as reported by Matting and Becker[17]. Their results were not confirmed during the present investigations since the d.c. resistivity of the metal could not be ascertained because of the very short lengths of the samples examined. Another drawback of this system of manufacture may be the poor mechanical strength of the resulting component. It seems, however, that the spraying technique can be very useful for producing waveguide components with complex shapes or for connecting these to standard waveguides, e.g. a taper section joined to a length of waveguide, because generally the increase of attenuation due to the low conductivity of the sprayed metal can be tolerated. Any complex components that can be produced by electro-forming can also be made by metal spraying, with the advantages of increased speed of production and of being able to join components to standard waveguides during the process.

Acknowledgments

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WIDE BAND AMPLIFIER DESIGN

By J. Kason*, A.M.I.E.E., A.M.Brit.I.R.E.

This article describes a design of wide band amplifier and includes a discussion on alternative designs.

WIDE band amplifiers are used to-day in many branches of electronic engineering, and though their centre frequency and bandwidth may vary, the design analysis applies to all. A wide band amplifier will be considered of 30Mc/s bandwidth, covering all the channels in Band 1.

There are several uses for such an amplifier, for example:

(1) Aerial pre-amplifier on any channel in Band 1.

(2) Factory installation—5 channel signals in Band 1 are amplified simultaneously and distributed to test points.

(3) Design laboratories and service, where greater than available signal generator output is required.

(4) Repeater for Band 1 distribution.

Very great bandwidth makes these amplifiers inferior, as compared to similar single channel amplifiers with regard

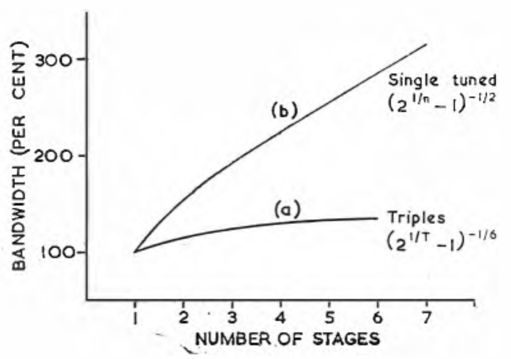

Fig. 1. Expansion of bandwidth per stage for each added stage

to noise input. Interfering signals on fringes of Band 1 may necessitate the use of an absorption circuit in the aerial down lead, as the amplifier curve is very flat. Finally, although the r.m.s. output is only limited by cross modulation, the permissible output per channel will become smaller, when more than one channel is amplified.

The following sections will deal with preliminary design considerations.

First, consider 'synchronous' cascaded, single tuned circuits, as they are cheapest and easiest to design and manufacture. The selectivity function[1] of a single tuned circuit is $F(x) = [1/(1 + jf_x)]$, where f_x is frequency deviation. Therefore for n cascaded stages the selectivity function becomes $[1/(1 + jf_x)]^n = F'(x)$ and for a deviation f_x, limited to 3dB points, the function becomes:

$$|F'(x)| = \left|\frac{1}{(1 + jf_x)^n}\right| = (1/\sqrt{2})$$

or

$$\frac{1}{\sqrt{(1 + f_x^2)}} = 2^{-1/2n}$$

$$\therefore 1 + f_x^2 = 2^{1/n}$$

$$\therefore f_x = \sqrt{(2^{1/n} - 1)}$$

Therefore

$$\text{Amplifier Bandwidth} = \frac{\text{bandwidth}}{\text{stage}} \times \sqrt{(2^{1/n} - 1)} \quad \ldots (1)$$

* *Electric and Musical Industries Ltd.*

Now $\sqrt{(2^{1/n} - 1)}$ may further be solved as follows:

Let $Y = 2^{1/n} = 2^z$, then using binomial expansion, Y becomes

$$Y = 1 + (Z/1)\ln 2 + (Z^2/2!)\ln^2 2 + \ldots$$

$$\therefore Y - 1 = (Z/1)\ln 2 + (Z^2/2!)\ln^2 2 + \ldots$$

$$= (1/n)\ln 2\,[1 + (1/2n)\ln 2 + \ldots] \simeq (1/n)\ln 2$$

$$\therefore \sqrt{(2^{1/n} - 1)} \simeq \frac{(\ln 2)^{\frac{1}{2}}}{\sqrt{n}} \simeq (0{\cdot}83/\sqrt{n}) \simeq (1/1{\cdot}2\sqrt{n}) \quad \ldots\ldots (2)$$

Therefore equation (1) can be written:

$$\text{Amplifier Bandwidth} = \frac{\text{bandwidth}}{\text{stage}} \div (1{\cdot}2/\sqrt{n})$$

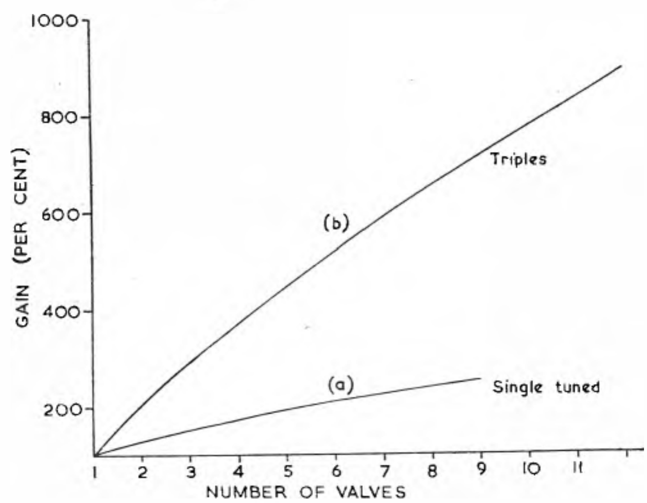

Fig. 2. Approximate increase of gain per valve for each added valve

In Fig. 1, curve (a) represents the expansion of bandwidth[2] necessary per stage for each added stage. As the gain bandwidth product is constant for any valve, it may further be seen that the gain of a stage is an inverse function of bandwidth, and will fall sharply with the addition of valve stages. (Fig. 2, curve (a)).

It is now possible to show that there is an optimum stage gain figure for a maximum bandwidth.

$$G^{1/n} \times B \simeq (G_m/2\pi C) \times \frac{1}{1{\cdot}2\sqrt{n}} \quad \ldots\ldots (3)$$

where $G^{1/n}$ = gain per stage
G_m = valve conductance
C = total input capacitance
n = number of valves

Hence let

$$A = \frac{G_m}{2\pi C \times 1{\cdot}2}$$

$$\therefore B = \frac{A}{\sqrt{n}\,G^{1/n}}$$

$$[d(B)/dn] = (1/n^2)\ln G - (1/2n)B,$$

and for a maximum

$$[d(B)/dn] = 0$$

$$\therefore B = 0, \text{ or } (1/n^2)\ln G = (1/2n)$$

$$\therefore (1/n)\ln G = 1/2$$

Hence

$$G^{1/n} = \sqrt{e} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

Hence optimum bandwidth is achieved for a gain per stage equal to $\sqrt{e} = 4{\cdot}34$dB. Thus, for a given bandwidth, using single peaked circuits, a maximum amplifier gain for a given bandwidth is obtained with a stage gain of 4·34dB. If this gain for a constant bandwidth is achieved with, say, n valves, a further addition of valves will decrease the total gain.

From the above analysis it can be seen that when very wide band amplifiers are required, the single peaked circuit could be employed for low gain requirements. Thus it can be shown that using ECC84 valves for 30Mc/s bandwidth, the maximum gain achievable is of the order of 35dB.

Let us now briefly consider overcoupled circuits. The required wide band amplifier circuits have a very low magnification factor, since $Q = (f_o/\text{bandwidth}) = (52/37) \simeq 1\frac{1}{2}$. It would therefore be extremely difficult to achieve overcoupling, as the coefficient of coupling for this must

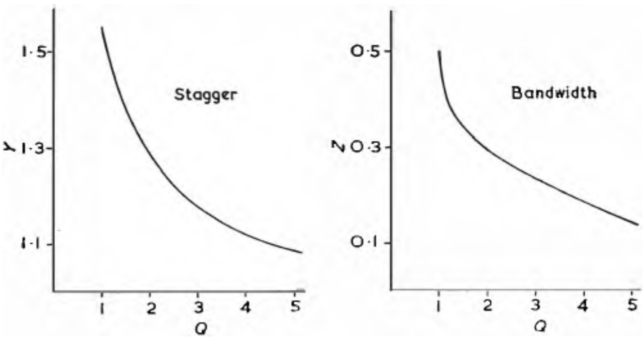

Fig. 3. Computation of triple stagger and bandwidth

be very high, the coupling index being $K\sqrt{(Q_1Q_2)}$. Circuits with equal Q's are further difficult to align and manufacture. Unequal Q double-tuned circuits give a higher gain, if loaded on one side and the tilt of the selectivity curve is compensated. The overcoupled tuned circuits were not considered further for the design, as the need for a high coupling coefficient remains, together with high production cost.

Further investigations have shown that staggered triples could be well suited for the design, as they give the required selectivity curve for the smallest number of valves. (Fig. 2, curve b).

The following sections deal with the design of a geometric triple.

The staggered circuit analysis is similar to a single peaked case. The selectivity function is $1/(1 + jx^n)$, thus for a triple the selectivity function becomes $1/(1 + jx^3)$ and for 3dB points

$$F(y) = \left(\frac{1}{\sqrt{(1 + x^6)}}\right)^T = (1/\sqrt{2}) \quad \ldots\ldots\ldots \quad (5)$$

where T is a number of triples.

Solving equation (5) similarly to equation (1),

$$x = (2^{1/T} - 1)^{1/6} \simeq \left[\frac{1}{1{\cdot}2\sqrt{n}}\right]^{1/3}$$

$$\simeq \frac{1}{1{\cdot}06\ .\ T^{1/6}} \quad \ldots\ldots\ldots\ldots\ldots \quad (6)$$

Therefore

$$\Sigma\text{Bandwidth} = B^{1/T}\ .\ (2^{1/T} - 1)^{1/6} \simeq B^{1/T} \div 1{\cdot}06\ .\ T^{1/6} \quad \ldots\ldots\ldots\ldots \quad (7)$$

when $B^{1/T}$ = bandwidth per triple.

If maximum gain required is of the order of 60dB, then gain required per triple should be approximately 20dB, and average gain per stage of the order of 7dB.

The valve chosen for the design is the ECC84 cascode, as it is equally suited as a pre-amplifier and a power output stage. The second usage of a cascode is rather unconventional, but analysing the cross-modulation effects, it may be seen that the anode curvature of a triode is more linear than that of a pentode. Therefore, as higher gain is achievable with a cascode, a greater output is possible. Furthermore, it is thought that the greatest amount of cross-modulation is contributed by the feedback from the anode to the grid. In the cascode stage the gain is contributed by the second triode, and this has its control grid grounded, and thus minimizes feedback. The bandwidth required for 3dB points for three cascaded triples ($\Sigma B = 30$Mc/s, gain = 60dB) may be evaluated from equation (7), thus

$$B^{1/T} \simeq \Sigma B \times 1{\cdot}06\ \sqrt{T^{1/6}}$$

$$\simeq 30\text{Mc/s} \times 1{\cdot}06\ \sqrt{3^{1/6}} = 38{\cdot}2\text{Mc/s}.$$

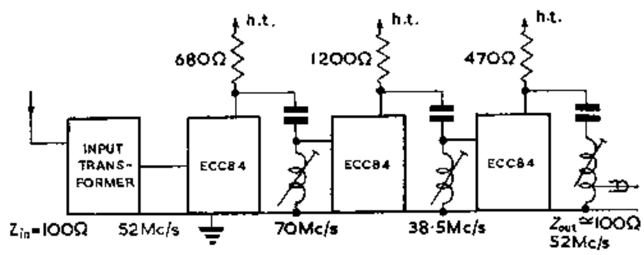

Fig. 4. Triple circuit with alignment frequencies

But the gain bandwidth product is $g_m/2\pi C$, where C is the total input capacitance. With ECC84 valves C may be of the order of 10pF. Thus the gain bandwidth product becomes 92Mc/s approximately, hence mean gain per stage,

$$G = (92/38{\cdot}2) = (2{\cdot}4/1) = 7{\cdot}6\text{dB}.$$

Hence an ECC84 valve will provide sufficient gain. The next requirement is to find the staggered three frequencies of the triple and their respective bandwidths. Now $Q = (52/38{\cdot}2) \simeq 1{\cdot}35$, and the deviation and dissipation factor can now be read from Fig. 3.

Thus the staggered frequencies become

$F_1 = 52 \times 1{\cdot}35 \simeq 70$Mc/s and $F_2 = 52/1{\cdot}35 = 38{\cdot}5$Mc/s

and the bandwidth for the staggered stages becomes

$F_o = 52$Mc/s $\quad \Delta f_o = 38{\cdot}2$Mc/s

$F_1 = 70$Mc/s $\quad \Delta f_1 = 70 \times 0{\cdot}39 \simeq 27{\cdot}2$Mc/s

$F_2 = 38{\cdot}5$Mc/s $\quad \Delta f_2 \simeq 38{\cdot}5 \times 0{\cdot}39 \simeq 15$Mc/s

The dynamic impedances can easily be calculated for each stage from $B = (1/2\pi RC)$ = bandwidth per stage (for 3dB points[3], thus $R = (1/2\pi BC)$.

Damping R for stage 1 becomes

$$(F_o) : \frac{1}{2 \times \pi \times 38{\cdot}2 \times 10^6 \times 10^{-12}} \simeq 470\Omega \quad \text{(Nearest preferred value)}$$

Damping R for stage 2 becomes

$$(F_1) : \ldots\ldots\ldots\ldots\ldots\ldots\ldots \simeq 680\Omega \quad ,,$$

and Damping R for stage 3 becomes

$$(F_2) : \ldots\ldots\ldots\ldots\ldots\ldots\ldots \simeq 1\,200\Omega \quad ,,$$

Thus a triple circuit becomes as shown in Fig. 4.

It now remains to design an input transformer to give a correct impedance match and output transformer providing a maximum power output into a 75/100Ω coaxial cable and the two selectivity curves to give us a suitable combined selectivity curve, i.e. to peak at 52Mc/s for the designed drop at the skirts.

An input transformer giving a fairly constant input impedance over the required band has been designed (**Fig. 5**).

Because of tilted response curve of the above transformer it has been found necessary to shift the resonance frequency

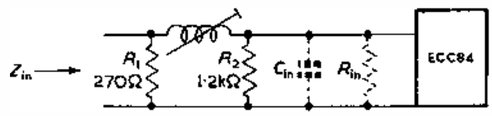

Fig. 5. Input transformer

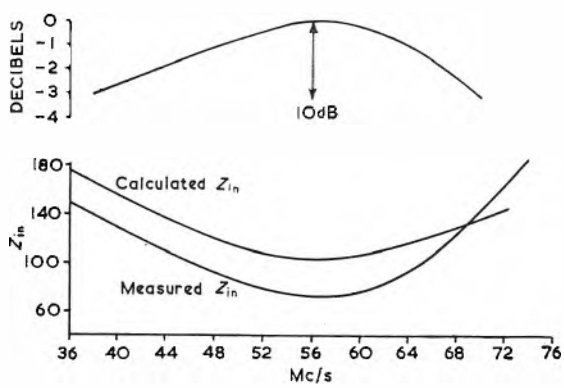

Fig. 6. Input transformer response curve (above). Variation of Z_{in} with frequency (below)

to 56Mc/s. Some theoretical impedance figures are listed below:

At 56Mc/s $\quad Z_{in} = 114\Omega$

At 36Mc/s $\quad Z_{in} = 172 - j39, \quad \therefore Z_{in} = 175\Omega : 13°$

At 72Mc/s $\quad Z_{in} = 123 + j74, \quad \therefore Z_{in} = 144\Omega : 31°$

It is worth noting that both R_1 and R_2 control the magnification factor which is of the order of $1\frac{1}{2}$, and R_1 controls the phase shift. Fig. 6 shows two impedance response curves, i.e. measured by substitution and calculated. The average impedance, measured, is more correct than the calculated one, as the strays could be only determined approximately. However, the two curves are consistent, and only show a discrepancy at the high frequency end. The rise-time of the input transformer is of the order of 270Mc/s and the voltage gain, calculated, is approximately 10dB.

The output transformer is a conventional auto-transformer, designed so as to compensate for the tilt of the

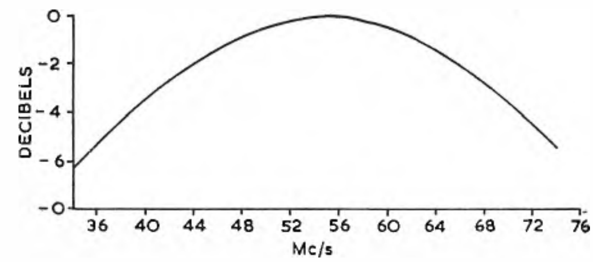

Fig. 7. Combined selectivity curve for the input and output transformers

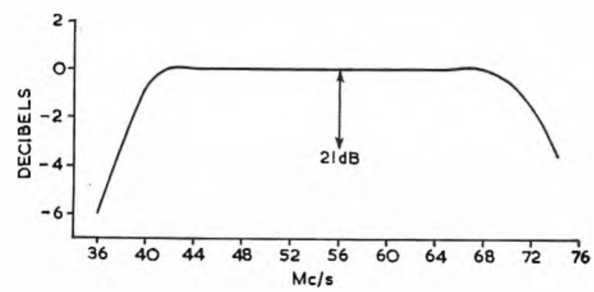

Fig. 8. Normalized response curve for the single triple

input transformer. The combined selectivity curve for the two transformers is shown in Fig. 7.

The total gain derived from a triple is of the order of 22dB for 1dB drop at 40 and 70Mc/s. Three combined triples give a gain of approximately 65dB with a drop of 3dB at 40 and 70Mc/s. The gain/frequency curve for single triple is shown in Fig. 8.

Subsequent field tests proved the design.

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A Simple Square-Law Circuit With High Frequency Response

By H. N. Coates, Ph.D., B.Sc.

By applying a signal to both control and suppressor grids of a pentode valve, it is possible to balance out changes in the valve anode current proportional to the signal fundamental and to generate a component proportional to its square, which is available without additional filtering and consequent restriction of bandwidth.

THE need often arises, particularly in analogue computing devices, for a circuit having a square-law output/input relationship and it is not always realized that the multi-electrode valve provides a simple solution where a very high degree of accuracy is not required.

Let x be the maximum rate at which the cathode of a pentode valve is capable of emitting electrons under normal conditions. This flow of electrons is controlled by the first grid, g_1, so that the cathode current of the valve can be regarded as:

$$px$$

Let the electron stream passing g_1 be modulated by a signal applied to this grid to vary p by an amount a.

Then the electrons passing g_1 can be considered as:

$$(p \pm a)\,x$$

The electrons are now divided between the second grid g_2, and the anode of the valve by the potential of the third or suppressor grid g_3.

Let the effect of g_3 be such that the anode current becomes:

$$q(p \pm a)\,x$$

and let the same modulating signal as before influence the effect of g_3 by an amount b, so that:

$$I_a = (q \pm b)\,(p \pm a)\,x$$

between the limits over which the two grids can be considered linear.

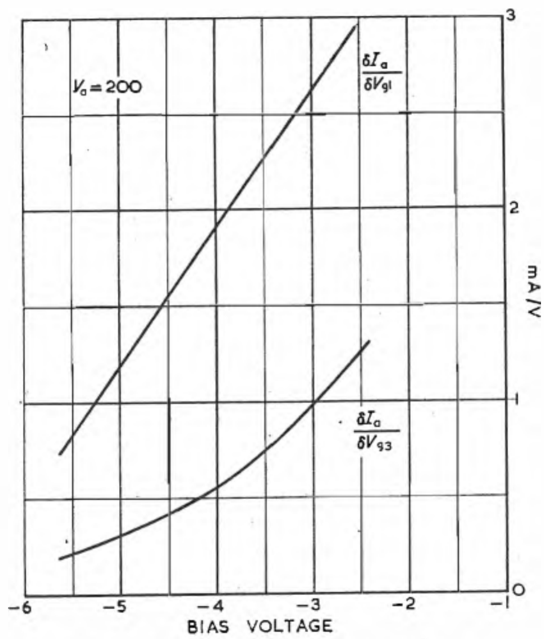


Fig. 1. Control of anode current by g_1 and g_3

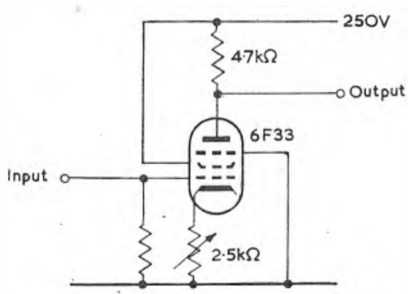


Fig. 2. Simple self-balancing circuit

If the modulating signal is applied to g_3 in the opposite sense to that in which it appears at g_1 (i.e. 180° phase shift in the sinusoidal case):

$$I_a = x(qp - bp + aq - ab).$$

In this expression, pq = a constant for fixed bias conditions, bp and aq are terms in the modulation voltage and ab is a term in the square of the modulation voltage.

Consequently, if $bp = aq$, the anode current of the valve will be proportional only to the square of the input voltage and no further filtering will be necessary.

With a conventional valve, the voltage required to produce a significant change b is certain to be large and the alternative is to keep change a small, resulting in a very small square term and poor signal-to-noise ratio. With the Mazda valve type 6F33, however, conditions are quite favourable for meeting the above requirement, and Fig. 1 shows curves of the control of the anode current by g_1 and g_3 as bias is applied to both grids simultaneously (an easier condition to meet in practice than providing a different bias voltage for each grid).

The very simple circuit of Fig. 2 can be adjusted to meet the conditions outlined above. As the bias resistor is increased, the signal voltage between cathode and g_1 is reduced and that between cathode and g_3 increased in the correct sense. At a value of cathode resistor of approximately 1kΩ, the term in the anode current proportional to the input voltage will disappear and the square term only will remain.

While a ratio better than 36dB between square term and fundamental can be achieved with an average valve in this circuit, this holds for one particular value of input level only and is of little use when a variety of input levels and waveshapes must be accommodated. By using two squaring circuits in parallel, fed 180° out of phase, this difficulty

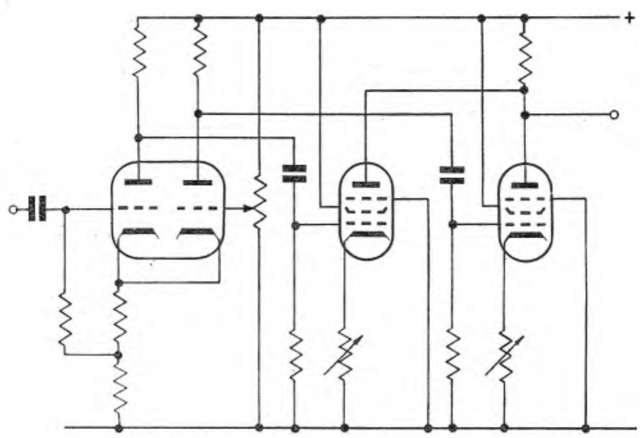


Fig. 3. Push-pull balanced square-law circuit

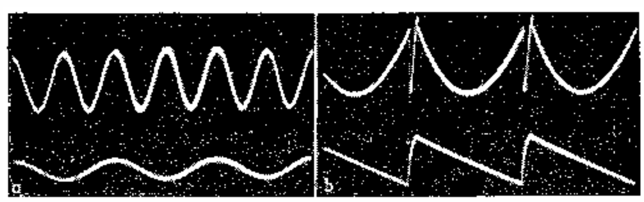


Fig. 4. Response of square-law circuit to (a) 100c/s sine wave (b) 1kc/s sawtooth

can very largely be overcome and the results obtainable with the circuit of Fig. 3 are illustrated in Figs. 4(a) and (b), the lower trace in each case being the input and the upper trace the output.

The circuit may be d.c. coupled if required and the frequency response is limited only by the valve and circuit capacitances. A good degree of stability is required of the h.t. voltage as changes in either g_2 or anode potential will upset the circuit balance. For the same reason, a large output swing is undesirable and low values of anode load are to be preferred.

Another application of the circuit is its use as a balanced modulator. If the anode load of the circuit of Fig. 2 is replaced by a tuned circuit and the resonant frequency applied to g_1, the valve can be balanced to produce no output at the anode. Modulating signals can be applied between g_3 and earth and a suppressed carrier r.f. output obtained from the tuned circuit. One valve generally suffices for this application as the r.f. input at g_1 is of constant amplitude.

LETTERS TO THE EDITOR

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

High Impedance Current Generators

DEAR SIR,—Without wishing to reflect on the desirability of high impedance current generators for certain purposes, or the ingenuity of your correspondent Mr. B. G. L. Braun (August issue), I think it is only fair to point out that the introduction of a valve of amplification factor μ into a circuit will produce an improvement as a rule of about $0{\cdot}7\mu$, if efficiently used, no matter what the configuration of the circuit.

In this particular case the efficient use of the valve is complicated by the necessity to preserve voltage levels; however, for the circuit of Fig. 2 (in Mr. Braun's letter) an improvement of 20 is claimed, whereas if the valve and neon were used conventionally, as an additional amplifying stage (Fig. A) an improvement of 40 to 50 could be expected. I contend therefore that before a valve is introduced to perform some unusual function, its use in a conventional manner to provide a straightforward increase in the gain may well be more satisfactory.

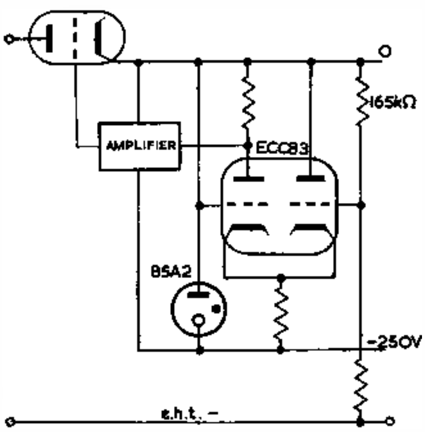


Fig. A. Conventional arrangement of valve and neon

To turn to the subject of high impedance current generators, a circuit I have used is of the type shown in Fig. B. For $R_1 = R_2 = R_3$ and the main amplifier, A, phase-reversing, analysis shows that the output impedance $Z_0 - (A\ R_4)/3$.

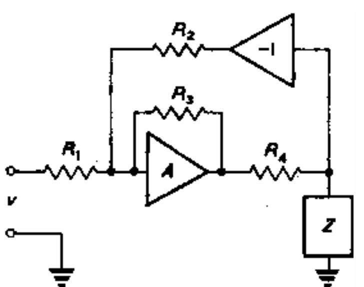


Fig. B. High impedance current generator

By increasing the positive feedback, i.e. reducing R_2, Z_0 can be made infinite, or even negative. Stability of course will depend on numerous factors including A, and the nature of the load Z.

In the particular application for which this circuit was developed, there was no necessity to earth the load and a push-pull amplifier was used, thus permitting the phase-reverser to be dispensed with, the positive feedback being obtained by cross-coupling.

The reason for the use of such a circuit was to keep the current in the coil which formed the load Z independent, as far as possible, of the frequency of the applied voltage V.

Yours faithfully,
D. C. PRESSEY.

Bagshot,
Surrey.

The correspondent replies:

DEAR SIR,—In reply to Mr. Pressey's letter, there are some points which I think should be clarified.

(1) Referring to the circuit I gave in Fig. 2 I do not see why there should be any difficulty in preserving voltage levels compared to a conventional d.c. amplifier.

(2) In this circuit there will be an apparent increase of gain as R_1 is increased until $R_1 > 10\text{M}\Omega$, taking the overall effective gain from the input—which is the negative e.h.t. line—to the output. In my circuit this gives a figure of approximately 0·4 on a 15kV supply and 0·17 on a 50kV supply; in the case of Mr. Pressey's circuit in Fig. A, using the values he has quoted, this gives 0·77 on 15kV and 0·23 on 50kV. When these circuits are connected to the amplifier there will normally be greater attenuation to the output of Mr. Pressey's circuit than mine, owing to the difference in potential levels, which would tend to make the ultimate gain comparable.

The factors which I think give my circuit some advantage are, (1) the output of low impedance which is useful if the amplifier is located away from the constant current circuit and (2) the change of output for a change of the 250V line is less than unity. I think this is important in view of the subsequent amplifier. Mr. Pressey's circuit amplifies such changes by the gain of the output stage, i.e. (approximately 50).

Yours faithfully,
B. G. L. BRAUN.

Great Malvern,
Worcestershire.

DEAR SIR,—In the light of Mr. Braun's reply to my first letter, perhaps I may be allowed to add the following.

The circuit of Fig. A is not recommended for use, as it was designed independently of the following stages. It was designed to use the −250V line as a reference potential as well as a power supply, and it incorporates a neon stabilizer as well as a double triode merely to show that a conventional arrangement of these components will give a comparable performance to Fig. 2. I had overlooked the fact that the output voltage of the circuit of Fig. 2 will be dependent more upon the neon voltage than upon the −250V line. With a sufficiently high gain overall, changes of the stabilized voltage in Fig. A will be related to changes of the −250V supply by the ratio of these voltages.

Since writing my first letter, I have realized that the introductory generalization no longer holds if positive feedback is incorporated in the circuit. This brings attendant disadvantages, notably a dependence of the circuit performance on heater voltages; nevertheless, very high performance can be obtained. Such a circuit, especially suitable for very high voltage stabilizers, has been given by Mackenzie[1].

Yours faithfully,
D. C. PRESSEY.

REFERENCE

(1) MACKENZIE, R. B., A New D.C. Electronic Voltage-stabilizing Circuit, *Proc. Inst. Elect. Engrs.*, *101*, Pt. 2, 59 (1954).

BOOK REVIEWS

Hi-Fi Loudspeakers and Enclosures

By Abraham B. Cohen. 360 pp. 183 figs. Demy 8vo. John F. Rider Publisher Inc, New York; Chapman & Hall, London. 1956. Price 37s 6d.

WITH every new phase in communication techniques, we find a corresponding increase in the relevant literature. We are now in the middle of a "Hi-Fi" era and this is a rather unusual development inasmuch as it involves an alliance between art and science. So far, no common language has been found to assign quantitative values to the expressions universally used for the individual appraisement of musical sounds.

This results in special difficulties when writing a book of the kind under review, for the subject as a whole is of as great an interest to the musician as to the engineer. Generally, it is better to avoid technicalities and to concentrate on the effects which these will have on the listener.

Mr. Cohen's book is quite non-technical in the engineering sense, and its appeal is directed to the intelligent amateur in electronics or music. The division of subject matter is neatly carried out by means of sub-titles to each section, so that it is not necessary to read any great amount of print to find a particular point of interest. While the author uses rather a lot of words to convey his meaning, this is perhaps a good thing for the musical reader. At the same time, some of the statements made are too general, technically incomplete and loosely expressed.

The author has attacked this difficult subject in a most comprehensive way. The uninitiated will derive a great deal of benefit from his painstaking explanations of the construction and functioning of all types of loudspeaker and cabinet. Certainly no previous writer has been so detailed and fundamental in dealing with the many possible combinations, although it is felt that more actual response curves for these various arrangements would have provided added guidance.

The section on horn loudspeakers is particularly interesting in view of their many advantages as mid-frequency range reproducers. Electrostatic units are not so fully treated, but then we are well in advance of our American cousins in this art.

A noteworthy feature is the inclusion of a whole chapter devoted to the practical design of loudspeaker enclosures. All American in origin, there are 18 dimensioned drawings on a reduced scale which should provide the amateur with a design for any condition which can be reasonably envisaged in the home. Again, this information would have been still more useful if response curves had been given for each design.

Mr. Cohen is to be congratulated on producing a book which will be of absorbing interest to the amateur, is extremely good value as book prices go, and embraces the effect of a far greater range of fundamentals in design than anything hitherto published on domestic loudspeaking equipment. A special word of praise should be given for the very large number of excellent drawings and photographs.

ALAN DOUGLAS

Progress in Semiconductors. Vol. 1

Edited by A. F. Gibson, R. E. Burgess, P. Aigrain. 220 pp. 70 figs. Demy 8vo. Heywood & Co. Ltd. 1956. Price 50s.

THIS book is the first volume of a new series which it is intended to publish annually. The articles deal with:— the preparation, properties and applications of silicon; the measurements which can be made using filaments of germanium; the theory of the thermoelectric effects in semiconductors; the electrical properties of phosphors; the design of high frequency transistors; and the photo-magneto-electric effect and the field effect in semiconductors. The articles cover therefore the physics of devices, some of the essential measurements and their underlying principles, and some of the fundamental aspects.

The progress of this subject is so rapid that any book dealing with it is liable to be out of date in some respects by the time it is published. With this volume the interval between writing and publication was about twelve months, which is no doubt as short as can be expected. Thus this fault, though present, is minimized. The bibliographies are extensive and up-to-date, numerous papers published in 1955 being included. Except for Johnson's account of the thermo-electric properties, the physical aspects are dealt with rather than the highly theoretical. Partly for this reason the articles have a less sophisticated air than those for example in Volumes VI and VII of Advances in Electronics and Electron Physics, or in Volumes I and II of Solid State Physics.

There has until recently been a shortage of good integrated accounts of the various branches of semiconductor physics, whether in text books or in series of review articles, but the situation is rapidly changing, and we are entering a phase of considerable competition between editors and publishers, and selection among readers. This applies especially to review series, so that comparison with other volumes such as those mentioned is inevitable. The present volume will be of considerable value to students and newcomers to the subject, to workers in related fields, and to those concerned mainly with the applications and technology of semiconductors. It will probably be of less value to those more advanced in the subject. It is perhaps not without significance that in America, where the technology of semiconductors is well ahead of that in this country, volumes of review articles are published which do cater for the reader more advanced in the fundamental and academic aspects of the subject.

D. A. WRIGHT

Variable Resistors

By G. W. A. Dummer. 176 pp. 101 figs. Demy 8vo. Sir Isaac Pitman & Sons Ltd. 1956. Price 30s.

THIS is the second volume of a proposed series of books on components, and with its ten chapters is a worthy successor to Fixed Resistors by the same author.

It combines 176 pages of the most useful data concerning variable resistors of every conceivable kinds with a strong bias towards Service requirements and specifications.

Commencing with the enumeration of types and their characteristics the book progresses through measurements, noise and wear to special and experimental variable resistors ending with a forecast of future developments based on current trends.

Very many manufacturing processes are dealt with in considerable detail together with methods of testing and fault finding which, although this is obviously a work for the design engineer, makes it suitable and easy to read for the not so technically minded in the electronic industry.

The final chapter contains an extensive and carefully thought out bibliography combined with many pages of comparison charts in which are illustrated constructional details of many types of commercial potentiometers with comprehensive operating characteristics. Also in this chapter are suggestions for a "Standard" terminology.

For the sake of completeness, certain tables given in Volume 1 are reproduced again with suitable cross references.

The chapter on special types of variable resistor is most illuminating and particularly lucid where it deals with hypocycloid systems, interpolating sine-cosine and also inductive potentiometers as well as sub-miniatures for use in telemetering and guided missile technique.

Clarity of diction and the general layout of the numerous tables and diagrams are of high standard. In fact, reading through this volume, and it is very readable except perhaps in the few places where the author discusses such subjects as track laws, an impression is created that it would be difficult indeed to find any item connected with this subject that the author might have omitted.

R. H. MAPPLEBECK

Electronic Data Processing for Business and Industry

By Richard G. Canning. 313 pp. 53 figs. Demy 8vo. John Wiley & Sons, New York. Chapman & Hall Ltd, London. 1956. Price 56s.

THIS is a book which packs a great deal of information into ten chapters. It lays out in a matter-of-fact way, what all business men, engineers, and workers have known must come about, that electronic machines are capable of taking over, not only the drudgery of clerical work, but make decisions in minor matters, and greatly influence management decisions on policy.

The impact of the book springs from the fact that the author not only explains how machines can be made to do this, but actually details the steps that must be taken to enable a company to change over from a manual clerical system to data processing.

To illustrate this point, he describes the changeover of an American aircraft component manufacturer, in which he actually participated as a "systems engineer", commissioned to investigate and plan the change.

This is, therefore, first-hand information, which anyone contemplating the use of electronic machinery must study.

The reviewer uses the word study, because readers should be warned that it is quite definitely not "a book at bedtime", and although it contains no mathematics, those familiar with management problems will find difficulty with the electronic and engineering concepts, and engineers difficulty with unfamiliar management problems.

The author has used a rather unfortunate formula, whereby at the end of a chapter he explains what the next is about, starts the new chapter with a synopsis of what it contains, and finishes this chapter with a résumé.

Some of the americanisms are difficult to interpret, and one phrase on page 200 took a good deal of re-reading to understand—"That is, yesterday's up-dated master records can be merged with yesterdays non-selected master records to make a complete up-dated master file at the same time that today's transactions are used to select out the master records to be up-dated today".

Nevertheless, the reviewer can thoroughly recommend this book to both management and engineers, because they may rest assured it will form a very useful guide at some time in the unspecifiable future, when all businesses will have to contemplate the installation of electronic equipment as an aid to reducing overhead costs and stream-lining their procedures.

W. WOODS-HILL

Radio Electronics

By Samuel Seeley. 487 pp. Demy 8vo. McGraw Hill Publishing Co. Ltd. 1956. Price 52s. 6d.

RADIO *Electronics* is a companion volume to the author's *Electronic Engineering*. Together they are a revision of his *Electron Tube Circuits*, but there is some common material designed to make them independent of each other. *Radio Electronics* deals mainly with the steady state analysis of lumped circuits involving valves. A reader with a moderate mathematical background need not be discouraged by the use of determinants and integration or by the tendency to express relationships as differential equations as the mathematical development, as distinct from notation, rarely passes beyond ordinary j-notation algebra and a few simple integrations. In general the explanations are clear, but at times the mathematical style complicates what could be expressed more simply and directly. A knowledge of the basic theory of valves and a.c. circuits is assumed. One unusual feature of the book is the effort made to encourage students to be logical and consistent particularly in their ideas of polarity and reference direction as applied to a.c. waveforms. A distinction is drawn, too, between true vectors and the phasors and sinors which represent the complex numbers associated with a.c. waveforms.

The range of subjects covered is extremely wide including most parts of sound transmitters and receivers and their power supplies but omitting wave propagation, television, metal rectifiers and transistors. While single stages are dealt with in detail and some space is devoted to the broad principles of communication systems, little is included about the characteristics of the receiver or transmitter as a unit and there is nothing at all about the operational requirements.

Indeed, the general outlook of the book is not practical in spite of the veneer given by numerous valve characteristics and sets of problems at the ends of chapters. The impression given is of a set of very full University lecture notes particularly as the sequence of presentation of the material is based more on a continuous development of a theme than the needs of an engineer who wishes to refresh his memory on a particular point. Towards the end of the book though in about thirty pages the author compares the effects of noise and interference on a.m. and f.m. signals and gives an introduction to Information Theory, both of which could be read profitably by many radio engineers.

Very few misprints were noticed in the book and the printing and production are at the publishers' usual high standard. The drawings and symbols follow American practice which differs in many ways from our own.

C. R. G. REED

Analysis of Bistable Multivibrator Operation

By P. A. Neeteson. 82 pp. 60 figs. Demy 8vo. Philips Technical Library. Cleaver Hume Press Ltd. 1956. Price 15s.

IN this book an analysis of the dynamic behaviour of the bistable multivibrator is given. Although the bistable multivibrator was conceived by Eccles and Jordan as early as 1919, it has only comparatively recently started to play an important role in electronic pulse apparatus.

BOOK REVIEWS (Continued)

Electrical Interference

By A. P. Hale. 118 pp., 100 figs. Demy 8vo. Heywood & Co. 1956. Price 15s.

IT is not unfair advertising to state that the author is a member of one of the pioneer firms in this field: the book in fact appears to be partly a revision of earlier books on the subject by other members of the same firm, but this is not claimed. The book deals with the practical aspect and mainly ignores theory: this should result in the maximum use to the many who are confronted with the task of suppressing any particular equipment. Only "man-made" interference is covered, so that a solution to all such interference problems is possible, although in some cases only knowledge and wide experience will be able to give an effective solution. Standard methods are quoted and in many cases will suffice, but the book alone will not always be the answer.

Initial chapters deal with causes and effects of interference and with aerial systems, the last rightly insisting that an efficient aerial is the first step to the elimination of radiated interference by increasing the signal noise level. Both sound radio and television are included in the chapter of effects of interference, with some useful illustrations of the types of effect on a television screen. Interference levels are discussed, with the apparatus necessary for such measurements. Specific figures are given, taken from B.S. 800 and other recent regulations based on the Wireless Telegraphy Act. A useful chapter deals with the location of sources of interference and mentions the availability of G.P.O. services. The chapter on avoidance of interference is very brief—six pages—and might well have been extended. It covers motors, lamps, switches and thermostats but for working values of capacitance etc, it refers to a later chapter on safety, in which the maximum allowable values of capacitance for various cases are given, extracted from B.S. 613 and C.P. 1006, these maxima being imposed by shock danger through capacitance current which is proportional to voltage, frequency and capacitance. Typical diagrams of arrangements for small motor suppression are very useful but might have included the motor with "double insulation", although this is quoted in the table.

Basic Filters, and practical Filters deal quantitatively with figures taken from the table also. The final chapter deals with screening and Faraday cages.

The reader is given some unnecessary difficulty through the employment of some obsolete terms and symbols: the terms "capacitor" and "condenser" are used almost alternately and a newcomer to the subject might well wonder what is the difference. Similarly "aerial" and "antenna" are used indiscriminately. Graphical symbols are sometimes in error, such as the chassis symbol where earth appears to be intended, and some confusion may arise through a wrong direction of flow being shown, sometimes from right to left and sometimes from bottom to top.

To summarize, this book can be recommended to those who have to deal with the practical problem of suppressing electrical equipment to reduce interference to radio and television receivers, and also to advise on aerial systems for least interference. The references to British Standards and other publications give the necessary backing for the quantitative values given, not least from the safety aspect to the user of the suppressed equipment.

E. H. W. BANNER.

Vacuum Deposition of Thin Films

By L. Holland. 541 pp. 60 figs. Demy 8vo. Chapman & Hall Ltd. 1956. Price 70s.

THE author of this book is head of the Vacuum Coating Research Laboratory of Edwards High Vacuum Ltd. While it was obviously impossible to deal in detail with all the numerous applications of thin films, the main subject-matter has been planned to cover plant design, film production, and the physical properties of thin films, all of which are of fundamental importance irrespective of the particular purpose for which the film is required.

Electronics in Industry

By George M. Chute. 431 pp. Demy 8vo. 2nd. Edition. McGraw-Hill Book Company, Inc. New York and London. 1956. Price 56s 6d.

THIS valuable addition to the available literature on industrial electronic circuits and equipment, comes from the pen of one who has had years of experience as an engineer with the General Electric Company, and in the teaching of the subject at the Detroit University. The text is written in a readable style, requires no specialized knowledge of mathematics and the book is valuable as the basis of a general course in this section of electrical engineering. The field covered is a wide one, and the first edition is well known to electrical engineers.

In this second edition, there is a more comprehensive study of resistance welding controls, and an interesting additional chapter deals with electrical control systems with special reference to electronic amplifiers and is of interest as a study of feed-back controls. Stability is the greatest problem in the design of a modern closed-loop system, and in this chapter, general ideas of servo behaviour are presented by illustrations and examples of the graphic methods used by engineers in the design of a closed-loop system. Since 1940, servo-mechanisms have attracted considerable attention and have found application in gun directors on warships, in the automatic-tracking radar, the automatic "follow up" control of precision machine tools and in the remote handling of dangerous materials. The author emphasises that in order to obtain high accuracy and fast response servos require exceedingly precise design. By this device, mechanical speeds may be compared and the resulting error signal can be changed into a corresponding electric voltage, for use with an electronic amplifier by a transducer, and there are many such familiar devices, e.g. a gramophone pick-up converts needle vibration into an electric signal, a thermocouple and a strain gauge can change a temperature signal into corresponding electric signals and an oscilloscope change an electric signal into a visible pattern. All these are examples of transducers.

This second edition is most valuable in that it embodies much first hand experience, this coupled with a wide knowledge of published information, provides a clear, logically reasoned account which will be invaluable to all those engaged in the electronic industry. It is a reasonable compromise between compactness of form and unnecessary repetition of material available elsewhere. Professor Chute has produced a volume which will be of great assistance to the student and of interest to the expanding number of people in industry concerned with complex electronic devices.

E. LEWIS

Short News Items

Cable and Wireless Ltd, in conjunction with the P & O Steam Navigation Co Ltd, recently carried out an experiment to investigate the possibilities of a radio-telephone service between London ships at sea in the Far East and similarly distant waters. Trials were carried out during which a radio-telephone call took place between a shipping office in the heart of London and a British ship at sea, over a distance of 10 000 miles. The ship was the P & O Company's new cargo liner *Salsette*. At the time of the talk she was steaming off Taku Bar, Tientsin, approximately 1 500 miles north of Hong Kong and 11 000 nautical miles from London. The radio-telephone circuit was relayed via Hong Kong and Nairobi to the London Terminal Exchange. The call was then connected by landline to the P & O Company's office in Leadenhall Street.

Pye television was used when Sir Cyril Hinshelwood, President of the Royal Society, officially opened Dido, the new heavy water reactor at Harwell. Over 200 official visitors watched the opening ceremony on Pye closed circuit television.

Pye Ltd have also recently announced an addition to their range of underwater television cameras. It is the first all-purpose underwater television camera, being capable of dealing with every practicable application and able to operate down to a depth of 3000ft. It can be hand-held by a frogman or diver, suspended from or trawled by a ship, or fitted to an underwater vehicle.

A meeting of members and prospective members of the Institution of Electronics was held in Belfast recently to consider the desirability of forming a local branch of the Institution. An organizing committee was elected until a branch could be formed and a further meeting will be held on 15 January. Further information may be obtained from the acting Honorary Secretary of the organizing committee, Mr. L. Mee, 11 Holland Park, Belfast, Northern Ireland.

The Plessey Group of Companies have established an office in Amsterdam to provide local contact and expand existing markets in Northern Europe. The office is situated at Singel 160, Amsterdam C.

Standard Telephones and Cables Ltd, have recently received an order, placed by the Portuguese Administration, Administracao Geral Dos Correios, Telegrafos e Telefones, for the provision of a comprehensive telecommunications network between Lisbon and Oporto and other localities in the presidency. This project, jointly planned by engineers of the CTT and the ST & C organization, embraces the supply and installation of equipment to provide telephone, telegraph and broadcast circuits and the provision of the necessary cable system. Manufacture of the equipment will be shared between Standard Telephones and Cables Ltd and their associates Standard Electrica Lisbon, and will proceed concurrently in Great Britain and Portugal. The British share of this contract is in excess of £1 600 000.

Dr. A. M. Uttley has been appointed superintendent of the Control Mechanisms and Electronics Division of the National Physical Laboratory, Teddington, which was formed in 1954. The Division studies the automatic control of experimental, industrial and administrative operations and the development of techniques and equipment for data processing and computation. Dr. Uttley succeeds Mr. R. H. Tizard who has taken up an appointment at the London School of Economics.

The Radio Industry Council announces that the 24th National Radio Show will be held at Earls Court, London, from 28 August to 7 September 1957.

The Decca Windfinding Radar W.F.1, is among the complex scientific and electronic apparatus carried by the *Magga Dan* which sailed recently for Antarctica. This new radar is the first of its kind and is the result of extensive research by Decca Radar Ltd, who have pioneered the application of radar to meteorological problems. During the International Geophysical Year, upper air stations all over the world are co-operating to measure the velocities of the winds from the earth's surface to 100 000ft.

Mr. D. L. Johnston, formerly Manager of the Component Division of Fortiphone Ltd, is joining Sir Walter Puckey and Mr. J. Sargrove at Automation Consultants and Associates Ltd, Barratt House, 341 Oxford Street, London, W.1. Mr. H. L. Ranson is joining Fortiphone Ltd, as Sales Manager of the Component Division.

British Communications Corporation Ltd are to equip fifty further lifeboats with their v.h.f. radio telephone equipment. This decision was taken by the Royal National Lifeboat Institution following the successful installation of this equipment in about thirty of the first fifty lifeboats.

The Midland Office of Marconi Instruments Ltd, formerly situated at 19 The Parade, Leamington Spa, has been transferred to 24 The Parade.

Regentone Products Ltd of Romford have received an order worth one million dollars from the United States for radio and audio equipment.

RCA Great Britain Ltd announce that the recently published RCA Transmitter Tube Manual T.T.4 will shortly be available in this country, price 12s. 6d. per copy, and will be obtainable from RCA Great Britain Ltd, Lincoln Way, Windmill Road, Sunbury-on-Thames, Middlesex.

20th Century Electronics Ltd announce that details have recently been released relating to their Boron-10 Separation Column. This is a new development in the nuclear energy field, a chemical separation column for stable isotopes now being operated by 20th Century Electronics. It is the result of close collaboration between the United Kingdom Atomic Energy Authority and this company.

Southern Instruments Computer Division announce the sale to the Canadian Defence Research Board, Ottawa, of one of their K.1001 film analysers. The order, worth about £11 000, was secured by their agents, Electrodesign Ltd of Montreal.

Marconi Instruments Ltd have manufactured f.m. station monitor equipment to a total value of nearly £22 000 for the BBC's new v.h.f. service. This apparatus, Type OA1082, forms part of the contract signed by Marconi's Wireless Telegraph Co Ltd with the BBC for f.m. broadcast equipment.

J. Langham Thompson Ltd have secured a Ministry of Supply contract with a value of £25 100 for development and supply of transducers for the measurement of acceleration.

Sir Austin Hudson, Bt. The death is announced of Sir Austin Hudson, Bt., which occurred on 29 November. He was Conservative Member of Parliament for Lewisham North, and was also the Chairman of Morgan Brothers (Publishers) Ltd, proprietors of ELECTRONIC ENGINEERING.

ELECTRONIC EQUIPMENT

A description, compiled from information supplied by the manufacturers, of new components, accessories and test instruments.

PAPER CAPACITORS

(Illustrated below)

The Plessey Co. Ltd, Ilford, Essex

A PAPER dielectric capacitor which has a performance comparable with the more expensive metal-cased types is now being manufactured.

The operative temperature range is −30 to −85°C and the component complies with tropical storage requirements R.C.S.11, Grade H2.

The range is from ·005 to 1μF in either standard or logarithmic ranges. Normal tolerance is ±20 per cent but tolerances of ±10 or ±5 per cent are available at slightly higher cost. Working voltages are 150, 250, 350, 500, 750 and 1 000V d.c. at 20°C; for operation at temperatures over 70°C, it is necessary to derate these figures by 25 per cent.

The shape of the unit is roughly elliptical with solid cross-section, allowing the capacitor to have a reduced volume.

The component is wound with the outer foil extended to provide an electrostatic screen and the tinned copper terminal lugs are inserted simultaneously to produce a non-inductive unit.

Impregnating and coating materials are of similar chemical composition, but with different physical constants. The finishing process, applied after impregnation, is such that each coat provides a key for, and is effectively bonded to, the next, thus producing a homogeneous covering which extends just past the soldered joint and forms a complete seal around the component. A flexible lacquer final seal completely envelops the capacitor and assists in completely excluding all moisture from the unit.

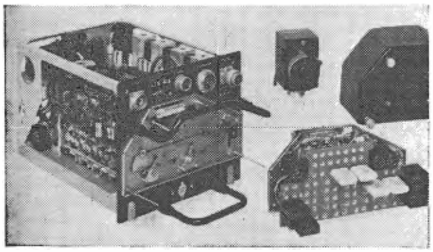
Insulation resistance is greater than 4k$M\Omega/\mu F$ at 20°C at the rated d.c. working voltage, even after prolonged periods of storage. Average d.c. breakdown voltage at normal temperatures is over five times the rated d.c. working voltage, thus ensuring an adequate safety factor at elevated temperatures.

MODIFICATION KIT FOR AIRBORNE COMMUNICATION EQUIPMENTS

(Illustrated above right)

Standard Telephones and Cables Ltd, Connaught House, Aldwych, London, W.C.2

THE STR.9X series of 10-channel v.h.f. airborne communication equipment is now standard in 24 air forces and is used by numerous airlines and charter companies.



The new modification increases the number of channels to 44 and incorporates automatic tuning of both the receiver and transmitter.

A modification kit providing all the necessary parts to modify existing 4- or 10-channel equipment is now in production.

The modification requires no changes other than the pilot's control unit, which is the same size and uses the same mounting and the same plug and socket as the existing one. This changeover can be made in a matter of minutes.

The main transmitter/receiver unit can be modified on the bench in approximately 4 hours to take the new crystal mounting panel and automatic tuning head.

44 crystals, type 10XJ, 10XAJ or 10XAU (or 10XAH with adaptor), can be plugged into the new crystal panel.

With the new modification incorporated, tuning of both the transmitter and receiver is automatic, no pretuning whatsoever being required.

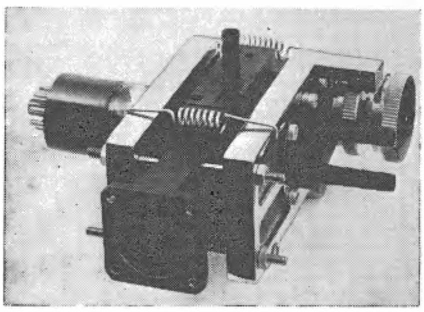
HIGH POWER X-BAND TUNABLE KLYSTRON

(Illustrated below)

Mullard Ltd, Century House, Shaftesbury Avenue, London, W.C.2

KLYSTRON type OKL1B, is a double-cavity tunable oscillator valve capable of delivering output powers up to 200W and of operating in the frequency range 8 600 to 10 000Mc/s. It is therefore useful in c.w. and Doppler radar systems, in microwave test gear, and as a driver for larger klystrons.

The valve is water cooled and is fitted with an indirectly heated dispenser cathode. Each valve is fitted with two tuning knobs with indicator dials. The frequency can be pre-set with the help of a graph supplied with each valve.



MINIATURE RELAY

(Illustrated below)

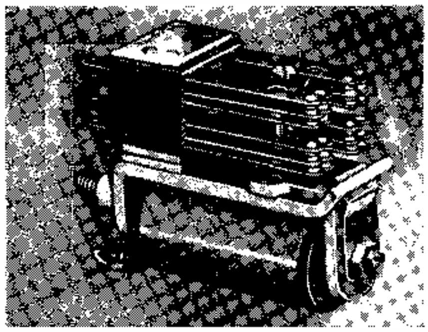
D. Robinson & Co, 58 Oaks Avenue, Worcester Park, Surrey

THIS relay is a sub-miniature version of the '3 000' type, retaining its most valuable features, yet being in scale with the modern range of miniaturized electronic components.

Even when fitted with the maximum number of contacts (four changeovers) it occupies only ½in × 1 3/16in of chassis space. and projects above by 1 1/4in.

The coil can be of any resistance value up to 5kΩ and consumes, according to contact build-up, from about 0·05 to 0·2W.

The blades are buffered and fitted with twin contacts, normally of fine silver rated at 2½A at 24V or ½A at 100V (non-inductive).



Platinum and other contacts with higher ratings can also be fitted.

An important feature is that the fall-off voltage can be set to anywhere between about 10 per cent and 60 per cent of the pull-up voltage.

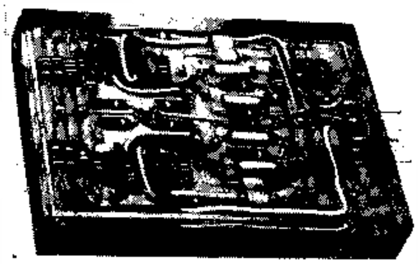
RELAY CONSTRUCTION KIT

W. H. Sanders (Electronics) Ltd, Gunnels Wood Road, Stevenage, Herts

TO meet the problem of supplying development engineers with 'one off' relays at very short notice, a kit of relay stacks, coils and associated items has been produced. Also enclosed are a limited selection of tools and comprehensive design data.

Within the very large range of coils and contact configurations, the aim has been to give as wide as possible a choice at a price well within the budget of the average laboratory.

The coils, which are complete with yokes and armatures, are of 500Ω and 1, 2, 2 + 2, 6·5, 10 and 20kΩ resistance. Contact banks are two off 1 changeover, two off 2 changeovers, two off 3 changeovers, one off 1 make-before-break and one off 2 make-before break, all in light duty silver; in heavy duty Elkonite, two off 1 make, two off 2 make and two off 2 break. Two off of each buffer block are also supplied, as are screwdriver, box spanner, feeler gauges and setting pliers.



TRANSISTOR DECADE COUNTER AND SCALER

(Illustrated above)

Venner Electronics Ltd, Kingston By-pass, New Malden, Surrey

THIS unit has been developed to provide a small and compact device having a low consumption which can be used either as a stable 'divide-by-ten' stage or, in conjunction with a meter, as a counting stage.

The unit employs transistors and germanium diodes throughout, arranged in four Eccles-Jordan stages with feedback. The device will count or divide from zero up to 30kc/s.

It only requires one power supply of 10V and total consumption is 100mW as a scaler and 160mW as a counter.

Some 86 components, including 8 diodes and 8 transistors are assembled on a double-sided tagboard, size 4in × 3in. This is then encapsulated in epoxy resin to form a block 4in × 3¼in × ¾in which in turn is fitted with an eight pin plug so that the unit is ready for direct insertion into equipment.

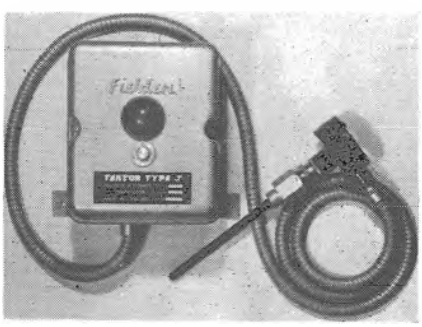
One decade may be directly connected to another, with no additional components, and thus by using four in cascade fed from a 10kc/s crystal oscillator, an output pulse every second may be obtained. Alternatively, by connecting a meter to each stage and again feeding the decades from the crystal oscillator, but having a switch to interrupt the flow of pulses from the latter, a short period timer is obtained which indicates directly on the meters to the nearest 0·1msec the time for which the switch has been closed.

LEVEL CONTROLLER

(Illustrated below)

Fielden Electronics Ltd, Manchester, 22

THE design of the 'Tektor' incorporates a simple oscillatory circuit which, when subject to a change of electrical capacitance, operates a relay, for the control of electrical contactors, visible and audible alarms, etc. This change of capacitance is brought about by the presence, or absence, of the material under measurement, either liquid or solid, and this is initiated by a metal rod inserted in the container at the point at which control is required. This method eliminates any moving parts in the container.



The 'Tektor' is offered in two standard forms, type J.3 which is housed in a die-cast aluminium case, suitable for normal industrial applications, and type J.5, housed in a weatherproof sheet metal case for more arduous industrial conditions. Flameproof models are also available.

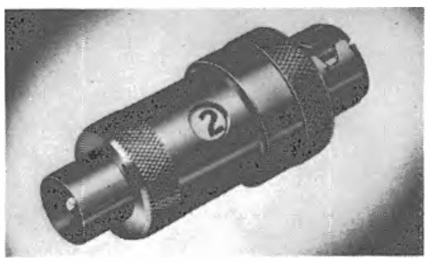
ADJUSTABLE ATTENUATOR

(Illustrated below)

Egen Electric Ltd, Charfleet Industrial Estate, Canvey Island, Essex

RECENT increases in the power of the London BBC and ITA television transmitters often require the fitting of an attenuator to receivers in areas where the signal strength of either station is too great. Type 141 attenuator in which simplicity of adjustment is one of the most important features, is little larger than a standard coaxial plug and can be easily inserted between the aerial feeder plug and receiver aerial socket.

Adjustment is by unscrewing the end cap of the attenuator and rotating the



internal moulded insulator to any one of six positions, indicated by the figures 1 to 6 which appear in a window in the outer casing, figure 1 giving the smallest and figure 6 the greatest attenuation.

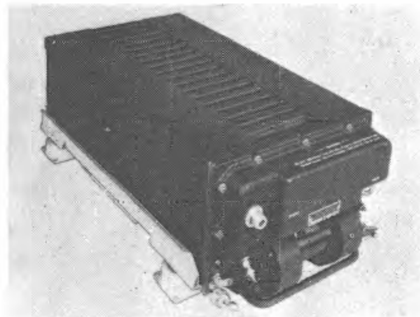
AIRCRAFT U.H.F. TRANSMITTER-RECEIVER

(Illustrated above right)

The Plessey Co. Ltd, Ilford, Essex

THE equipment known as type AN/ARC52 has been designed by the Collins Radio Company of U.S.A. and provides a.m. radiotelephone communication in the u.h.f. range 225 to 399·9Mc/s, between aircraft and ship, aircraft and shore, or between aircraft. It employs a double conversion superheterodyne circuit and the approximate overall dimensions of the transmitter-receiver unit are only 21 × 10⅛ × 7 3/16in. A pressurized case ensures operational efficiency at any altitude.

There are 1 750 frequency channels of operation available through the use of a remote manual control unit and provision has been made for selection of any one of 18 pre-set channel frequencies within the specified range. A guard channel frequency, within the range 238 to 248Mc/s and employing an entirely separate fixed tuned receiver unit, permits constant watch for emergency signals.



DIELECTRIC TEST SET

Microcell Electronics, Imperial Buildings, 56, Kingsway, London, W.C.2

THIS instrument is intended for laboratories engaged in the development of high temperature resins and laminates for microwave propagation. Direct readings can be made of permittivity and loss tangent for small samples at fixed spot frequencies; three different models are available to cover the frequency range 6 000 to 17 000Mc/s. Provision is made for dielectric measurements to be undertaken at temperatures up to 450°C.

Insertion of a dielectric sample in arm 2 of a 3dB klystron coupler causes a change in electrical lengths and some of this reflected energy passes to arm 4 which is terminated in a detector unit where output current is indicated on the panel meter. By noting the shift of the plunger necessary to restore the meter reading to a minimum, the electrical length of the sample is obtained. This enables the permittivity to be calculated.

A measure of loss tangent for the sample can be obtained from the difference in amplitude of the reflected waves due to the loss in the sample.

ELECTRONIC MEGAPHONE

(Illustrated below)

Pye Ltd, P.O. Box 49, Cambridge

THE 'Transhailer' is a self-contained power megaphone consisting of a folded horn loudspeaker, a 3W germanium transistor amplifier, a sensitive electromagnetic microphone and batteries.

The weight of the unit, with batteries is 5lb and the horn diameter is 10½in.

Average consumption on speech is 120mA giving an approximate battery life of 12 hours on a duty cycle of 20sec off, 20sec on. It is expected that the life of batteries on normal use will be up to 6 months, or sufficient for about 20 000 10sec announcements.

Range in reasonably quiet conditions is approximately 500 yards.

Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 30 January. Time: 6.30 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Lecture: a.m.-f.m. Battery-operated Receivers.
By: R. A. Lampitt and J. P. Hannifan.

South Wales Section

Date: 16 January. Time. 6.30 p.m.
Lecture: Radio Techniques in Post Office Engineering.
By: C. T. Lamping.

North Western Section

Date: 3 January. Time: 6.30 p.m.
Held at: Reynolds Hall, College of Technology, Sackville Street, Manchester 1.
Lecture: Projection Television Systems.
By: C. Somers.

West Midlands Section

Date: 9 January. Time: 7.15 p.m.
Held at: Wolverhampton and Staffordshire Technical College, Wulfruna Street, Wolverhampton.
Lecture: Design and Application of Magnetic Amplifiers.
By: R. G. Russell-Bates.

Merseyside Section

Date: 2 January. Time: 7 p.m.
Held at: The Council Room, Chamber of Commerce, 1 Old Hall Street, Liverpool, 3.
Lecture: Instrumentation for Frequency Modulation.
By: D. R. Willis and A. G. Wray.

Scottish Section

Date: 11 January. Time: 7 p.m.
Held at: The Department of Natural Philosophy, University of Edinburgh.
Lecture: Potted Components and Assemblies.
By: H. G. Manfield.
Date: 17 January. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow.
Lecture: Industrial Applications of High Speed Pen Recorders.
By: R. Kasler.

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings, unless otherwise stated, will be held at the Institution, commencing at 5.30 p.m.

Measurement and Radio Sections

Date: 8 January.
Lecture: A Theoretical and Experimental Investigation of Anisotropic-Dielectric-Loaded Linear Electron Accelerators.
By: R. B. R. Shersby-Harvie, L. B. Mullett, W. Walkinshaw, J. S. Bell and B. G. Loach.
Date: 29 January.
Discussion: The Performance of d.c. amplifiers with special reference to the use of Transistors.
Opened by: K. Kandiah and G. B. B. Chaplin.

Ordinary Meetings

Date: 10 January.
Lecture: Some Aspects of Heat-Pump Operation in Great Britain, with particular reference to the Shinfield Installation.
By: Miriam V. Griffith.
Date: 24 January. (Time not yet announced).
Symposium of papers on the Transatlantic Telephone Cable. Joint meeting, linked by the cable, with the American Institute of Electrical Engineers and the Engineering Institute of Canada.

Informal Meeting

Date: 14 January.
Discussion: The Influence of Maintenance Requirements on the Design of Industrial Electrical Equipment.
Opened by: H. C. Fox and J. H. Harris.

Radio and Telecommunications Section

Date: 23 January.
Papers on Transistor Applications.

Supply and Measurement Sections

Date: 30 January.
Lectures: Power System Protection, with particular reference to the Application of Junction Transistors to Distance Relays.
and: A Dual-Comparator Mho-Unit Distance Relay Utilizing Transistors.
By: C. Adamson and L.M. Wedepohl.

East Midland Centre

Date: 8 January. Time: 6.30 p.m.
Held at: The College of Art and Crafts, Waverley Street, Nottingham.
Lecture: Electrical Energy from the Wind.
By: E. W. Golding.

Date: 23 January. Time: 6 p.m.
Held at: The Temple, Dale Street, Liverpool.
Lecture: The Electronic Control of Machine Tools.
By: N. Milne.
Date: 24 January. (Time and place as above).
Lecture: The Pressurized Water Reactor as a Source of Heat for Steam Power Plants.
By: J. M. Kay and F. J. Hutchinson.

North-Eastern Radio and Measurements Group

Date: 21 January. Time: 6.15 p.m.
Held at: King's College, Newcastle-upon-Tyne.
Informal Evening on Electronics and Automation.
Talk on Some Industrial Applications.
By: H. A. Thomas.

North Midland Utilization Group

Date: 22 January. Time: 6.30 p.m.
Held at: Offices of the Central Electricity Authority, Yorkshire Division, 1 Whitehall Road, Leeds 1.
Lecture: Germanium and Silicon Power Rectifiers.
By: T. H. Kinman, G. A. Carrick, R. G. Hibberd and A. J. Blundell.

North-Western Measurement and Control Group

Date: 15 January. Time: 6.15 p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Lecture: The Control of Nuclear Reactors.
By: R. J. Cox.

North Scotland Sub-Centre

Date: 9 January. Time: 7.30 p.m.
Held at: The Caledonian Hotel, Aberdeen.
Lecture: Colour Television.
By: G. N. Patchett.
Date: 10 January. Time: 7 p.m.
Held at: The Electrical Engineering Department, Queens College.
Repeat of above lecture.

South-East Scotland Sub-Centre

Date: 15 January. Time: 7 p.m.
Held at: The Carlton Hotel, North Bridge, Edinburgh.
Lecture: A Transatlantic Telephone Cable.
By: M. J. Kelly, Sir Gordon Radley, G. W. Gilman and R. J. Halsey.

South-West Scotland Sub-Centre

Date: 30 January. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow.
Lecture: Choice of Insulation and Surge Protection of Overhead Transmission Lines of 33kV and above.
By: A. Morris Thomas and D. F. Oakeshot.

Rugby Sub-Centre

Date: 16 January. Time: 7.30 p.m.
Held at: The Temple Speech Room, Rugby.
Faraday Lecture on Nuclear Energy in the Service of Man.
By: T. E. Allibone.

Southern Centre

Date: 9 January. Time: 6.30 p.m.
Held at: The Brighton Technical College.
Lecture: The Crystal Palace Television Transmitting Station.
By: F. C. McLean, A. N. Thomas and R. A. Rowden.
Date: 16 January. Time: 7.30 p.m.
Held at: The R.A.E. Technical College, Farnborough.
Lecture: Time Sharing as a Basis for Electronic Telephone Switching: A Switched Highways System.
By: F. C. Nelson.
Date: 25 January. Time: 6.30 p.m.
Held at: The South Dorset Technical College, Weymouth.
Lecture: Germanium and Silicon Power Rectifiers.
By: T. H. Kinman, G. A. Carrick, R. G. Hibberd and A. J. Blundell.

Western Supply Group

Date: 21 January. Time: 6 p.m.
Held at: The Demonstration Theatre, South Western Electricity Board, Colston Avenue, Bristol.
By: D. F. Welch.

West Wales (Swansea) Sub-Centre

Date: 10 January. Time: 6 p.m.
Held at: The Conference Room, South Wales Electricity Board Showrooms, The Kingsway, Swansea.
Lecture: The Instrumentation of a 14in Experimental Rolling Mill.
By: S. S. Carlisle and G. W. Alderton.

THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Date: 7 January. Time: 5 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2.
Lecture: Auto Exchange Maintenance Procedure—Past, Present and Future.
By: S. Rudeforth, K. S. Laver and E. R. Driver.
Date: 23 January. Time: 5 p.m.
Held at: The Conference Room, 4th Floor, Waterloo Bridge House, London, S.E.1.
Lecture: The Utilization, Review and Planning of a Large Junction Network.
By: J. B. F. Williams.

THE PHYSICAL SOCIETY

Acoustics Group

Date: 24 January. Time: 5.30 p.m.
Held at: Imperial College, Imperial Institute Road, South Kensington, London, S.W.7.
Lecture: Detection and Measurement of Vibration.
By: M. L. Parsey.

THE TELEVISION SOCIETY

Date: 11 January. Time: 7 p.m.
Held at: 164 Shaftesbury Avenue, London, W.C.2.
Lecture: Automatic Gain Control Circuits in Television Receivers.
By: S. N. F. Doherty and P. L. Mothersole.
Date: 24 January.
The Fleming Memorial Lecture-Luminescence.
By: H. G. Jenkins.
(This lecture is not held at Shaftesbury Avenue, and admission is by ticket only, obtainable from any member or from the office of the Society.)

PUBLICATIONS RECEIVED

THE WIRELESS WORLD DIARY 1957, now in its 39th year of publication includes in its 80 page reference section base connexions for 600 current valves, design data, frequency allocations, addresses of radio organizations in this country and abroad and, in fact, most of the technical and general information frequently required. "Wireless World," Dorset House, Stamford Street, London, S.E.1. Price 6s. (leather) and 4s. 3d. (rexine).

AUTOMATION IN PERSPECTIVE is the subject of a booklet issued by the Department of Scientific and Industrial Research. The text opens with a brief review of the technical trends and prospects and includes a description of what a fully automatic factory may look like. The remaining pages set out the possible effects of automation on management, on labour and on the economy as a whole. Her Majesty's Stationery Office, Kingsway, London, W.C.2. Price 1s. 6d.

AN AUTOMATIC INTEGRATOR FOR DETERMINING THE MEAN SPHERICAL RESPONSE OF LOUDSPEAKERS AND MICROPHONES is No. 8 in the BBC series of Engineering Division monographs. About six of these monographs will be published each year. BBC Publications, 35 Marylebone High Street, London, W.1. Price 5s. each or annual subscription of 20s. a year, post free.

INVERSE FEEDBACK, BASICS OF PHOTOTUBES AND PHOTOCELLS and TV MANUFACTURERS' RECEIVER TROUBLE CURES VOL. 8 are three recent books published by John F. Rider Publisher, Inc., 480 Canal Street, New York 13, N.Y., applicable to American practice. Prices 90c., \$2.90 and \$1.80 respectively.

DICTIONNAIRE FRANCAIS-ANGLAIS DES TERMES RELATIFS A L'ELECTROTECHNIQUE L ELECTRONIQUE ET AUX APPLICATIONS CONNEXES by H. Piraux has recently been published by Editions Eyrolles, 61 Boulevard Saint-Germain, Paris VE. Price 1 035F.