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The Deeper the Better?

We have recently received a list of abstracts of technical reports from the Telsa National Corporation which now represents the Czechoslovak electronic industry, and one abstract in particular deserves more than a passing glance.

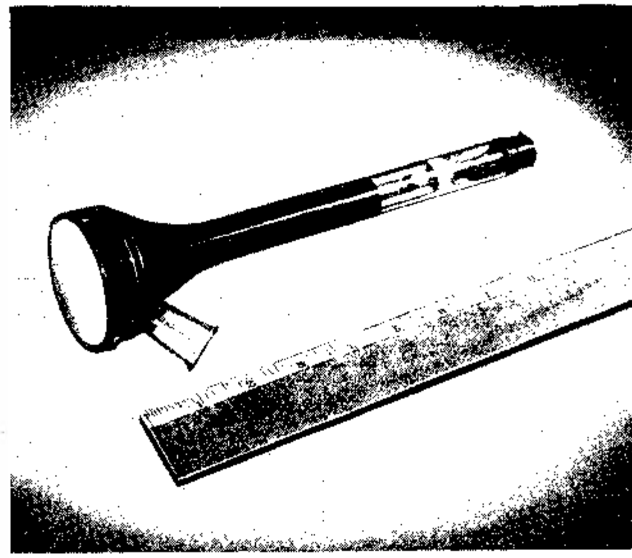
It is entitled "Underground Loudspeakers," by Joseph Merhaut, and the author claims that waterproof loudspeakers of a new type developed by him, buried at a uniform depth below ground and spaced at regular intervals are ideally suited for installation at a sports stadium, where no obstruction should be offered either to the free movement of performers or to the view of the spectators.

Perhaps the author's more important claim is the elimination of the inevitable time lag associated with the usual public address system with a number of spaced loudspeakers.

Now this problem of time delay is a very old one which can be solved to some extent by the grouping of loudspeakers into a single unit, but it was all very forcibly brought to our attention again at the recent National Air Races at Elmdon, Birmingham, over the August week-end. Here in the public enclosures were installed at regular intervals a number of loudspeakers all pointing the same way, so that a spectator at the point where the visual interest was the greatest heard *all* the loudspeakers. The positioning of the speakers was possibly done in this way to prevent the commentary being overheard outside the aerodrome, but the result was, of course, that unless one stood directly in front of the nearest loudspeaker, intelligibility was practically non-existent and Charles Gardner's breezy commentary was entirely lost.

There may have been difficulties of which we are not aware in installing a suitable public address system at Elmdon, and it is possible that danger to low flying aircraft precluded the use of a grouped system of loudspeakers, but it is probably true to say that the listener at home had a clearer idea of some of the proceedings! It is not surprising therefore that our curiosity was aroused by the Czechoslovak article and we shall await the fulfilment of the author's claim with interest.

A PROJECTION SYSTEM FOR DOMESTIC TELEVISION RECEIVERS



THE very real demand for larger television pictures for home viewing and for larger groups, such as clubs—to say nothing of cinema size pictures for public exhibition—sets a problem which has both technical and economic aspects.

Two technical solutions are available—direct viewing using picture tubes of large screen diameter, and projection viewing using small tubes of very high screen brilliance.

For domestic mass-produced receivers a 15 in. × 12 in. picture at present approaches the upper limit for direct viewing on economic grounds, owing to the high cost of the glass envelope and also the size and cost of the cabinet.

There remains projection viewing as an alternative. Projection television is not a new idea, but much research had to be undertaken before apparatus in a form suitable for mass production was evolved.

Mullard are now making, however, available to set manufacturers, samples of a projection television system giving adequate picture resolution up to 15 in. × 12 in., and suitable for use by them in conjunction with most existing television receiver chassis.

The equipment comprises three units:—

1. A $2\frac{1}{2}$ in. diameter screen picture tube.
2. An optical unit embodying a tube mounting, focusing coil and deflector coil, and a folded version of the Schmidt optical system.
3. A compact E.H.T. supply unit.

The Picture Tube

From the purely economic standpoint, the dimensions of the picture tube should be as small as possible consistent with adequate screen brightness and good picture resolution, for then the optical enlargement system will be correspondingly small and comparatively inexpensive to manufacture. Tube size, however, is governed by spot size, and this, in turn, depends upon the value of the beam current, which should be low, and the anode voltage and current density which should be high for minimum spot size. Limitations to these values are set by such practical considerations as tube life, safety, and manufacturing difficulties.

The MW6-2 picture tube has a $2\frac{1}{2}$ in. diameter screen and is designed to operate at an anode voltage of 25KV. Operated at the recommended beam current of 100 μA, the spot size is 58 μ (0.0023 in.). The tube is quite conservatively rated and can, if necessary, pass peak currents up to 500 μA. The spot size is then somewhat increased, with consequent reduction in picture definition, but as this current corresponds to peak white, slight loss of definition can be tolerated.

The tube is designed for electromagnetic deflexion and focusing—a decision taken mainly on the score of simplicity in gun construction but which also has circuit and performance advantages.

The design of the electrode assembly is, of course, governed mainly by consideration of beam size. Requirements are that the beam must be of such dimensions as to produce a spot of the required diameter, yet not so concentrated that mutual repulsion of the electrons produces blurring of the spot. Further, the beam at its widest diameter must be small enough to keep the electrons well clear of the tube walls and the anode when fully deflected; and finally, the I_a/V_g curve must be sufficiently steep to allow the tube to be driven by a normal video output valve.

It has been possible to meet these requirements with a triode construction.

A sectional view of the tube is given in Fig. 1. An important feature is the spark trap—a ring-shaped electrode situated between the grid and the anode and connected to earth via one of the base contacts. Any discharge which might occur due, for example, to the release of a small amount of gas as the result of unintentional overload, will take place between the anode and the spark trap, thus avoiding damage to the cathode.

The photo above shows the external appearance of the MW6-2 picture tube. The external surface of the neck and cone is

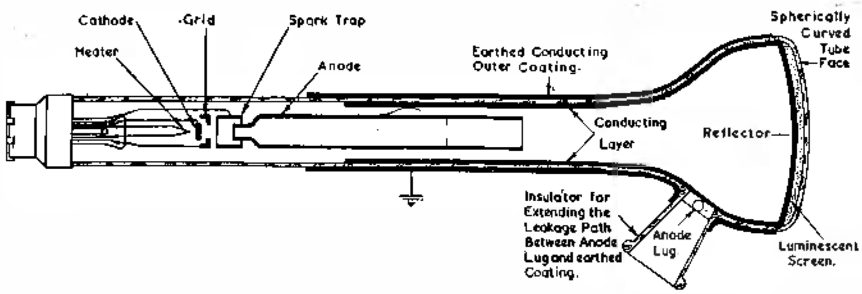


Fig. 1. Sectional view of the MW6-2 tube

coated with a graphite preparation, and must be earthed. This coating, with the glass envelope and the internal metallising of the tube, forms a capacitance of approximately 300 pF which, with a 1 MΩ resistor in the 25 KV lead, serves as the final smoothing for the E.H.T. supply.

The most conspicuous external feature of the tube is the glass shield surrounding the anode terminal. This shield obviates risk of flashover or leakage from the E.H.T. connexion to the deflector coils or to the graphite coating.

As previously stated, the fluorescent screen is backed by a metallic coating. The metal used is aluminium, which has the necessary high optical reflectivity and also a low atomic weight to permit easy penetration by the electron beam. The thickness of the layer is in the order of 0.4 micron. The metal is evaporated within the tube by conventional methods, and is deposited as a thin film on the rear surface of the screen and on the neck of the tube, to which has been previously applied a thin layer of nitro-cellulose to provide an even surface.

The metal-backed screen has many advantages. In the first place, the metal coating reflects outwards a large proportion of the emitted light which would otherwise be directed to the rear of the tube. The increase in the output of forward-going light due to this feature is, under correct operating conditions, as much as 75 per cent to 80 per cent. The metal backing also eliminates the internal reflexions and screen irradiation which would otherwise result from the light directed towards the rear of the tube. Picture contrast is therefore considerably enhanced.

A third advantage of the metallic coating is a substantial improvement in secondary emission from the screen, thus avoiding the building up of a negative charge on the screen and improving the stability. Finally, the metallic film serves as an effective ion trap as the negative ions, which would otherwise ultimately be responsible for ruining the screen, are not able to pass through the film.

The tube face is made from a special type of glass which is free from the discolouration so often occurring in tubes after a long period of operation at anode voltages above some 15 KV. This discolouration, which ranges in colour between purple, brown and black according to the type of glass, appears over the area on which the raster is formed, and is mainly due to the generation of soft X-rays resulting from the bombardment of the screen by the high-velocity electron beam.

The glass face of the tube is formed by pressing to the required convex radius to suit the restricted depth of focus of the Schmidt optical system, and this radius is controlled to a very high degree of accuracy.

The Optical System

The principle of the Schmidt mirror system for projection television is already well documented, and it will suffice here to remind the reader that it consists essentially of a concave spherical mirror and a "corrector plate." The spherical mirror concentrates the rays into a convergent beam, and the corrector plate is, in effect, a lens of special contour which compensates for spherical aberration and thus ensures that all the rays reflected from the spherical mirror come to focus at the same point, whence they again diverge to give the enlarged image.

The version of the Schmidt mirror system here employed is shown in the diagram Fig. 2, and the complete optical unit in Fig. 3. The light emitted from the face of the picture tube is directed to the concave mirror, whence it is returned as a convergent beam to the plane mirror fixed at an angle of 45° to the direction of the beam. The light is reflected vertically by this mirror, and passes through the corrector plate, after which it can be projected to any desired type of viewing screen either direct, or, at some slight sacrifice of optical efficiency, further deflected by additional mirrors to a screen in any desired position.

It will be observed that the picture tube projects through the centre of the first plane mirror. The loss of light due to the hole in the mirror and to interception by the face of the tube is practically the same for the corners as for the centre of the picture, and there is, of course, no interference from the coils and other components as these are situated behind the mirror.

The arrangement of mirrors whereby the light path is "folded up" permits adequate picture magnification within a very compact space, and has an important bearing upon the overall dimensions, and therefore the cost, of the cabinet.

The chief problem in designing the optical unit was the manufacture of the corrector plate to the desired degree of accuracy at economic cost. Production in glass by conventional grinding technique is a lengthy and costly process, and even moulded plastic lenses are expensive to produce in moderate quantities. The technique ultimately adopted is moulding in gelatine on to a glass plate. A negative mould of exaggerated contour is used, with a gelatine solution of controlled con-

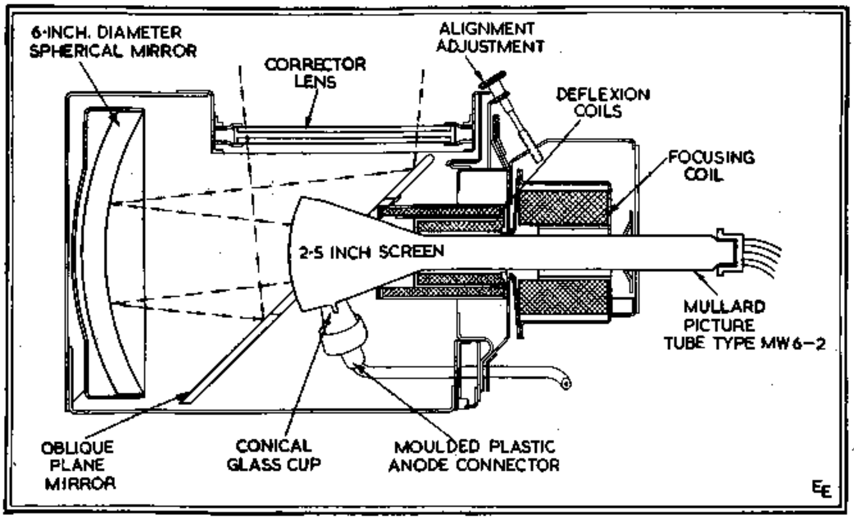
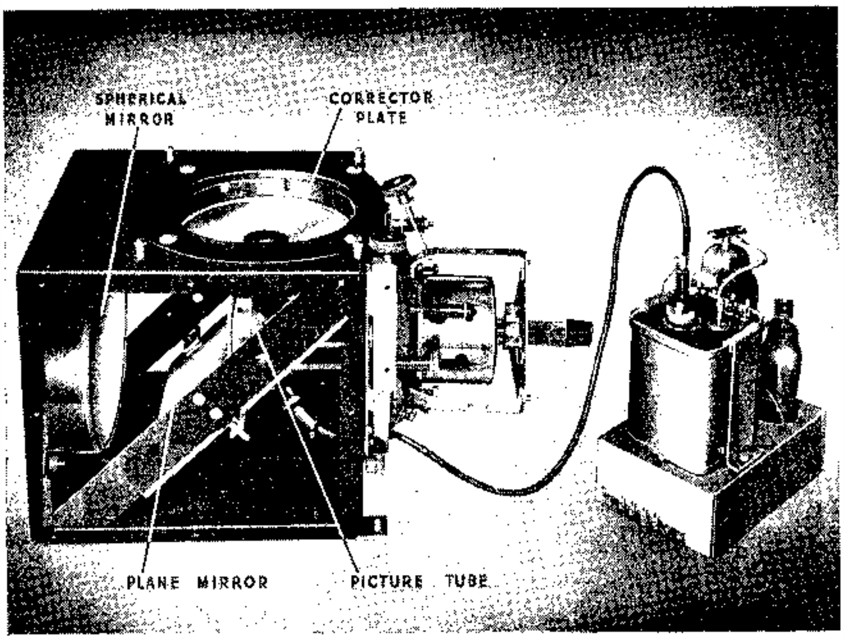


Fig. 2 *(above)*. **Diagram illustrating the Schmidt optical system**

Fig. 3 *(below)*. **Complete projection system with E.H.T. supply unit**



centration. Because the gelatine adheres firmly to the glass plate, shrinkage of the gelatine during drying is only in the direction of its thickness, and by controlling the concentration of the solution, correct contour of the film after drying is assured.

The sharpness of the image produced by this optical unit with a gelatine corrector plate is extremely good, and is, in fact, more than adequate for reproduction of a 600-line picture. The durability of the gelatine plate is comparable with that of a photographic negative.

The E.H.T. Supply Unit

Production of a 25 KV supply direct from the normal 50 c/s mains is uneconomic by reason of the large physical dimensions and weight of such a unit and its high cost particularly the cost of insulation. In addition, the unit would inherently be capable of a much greater output than the few hundred microamperes required for the cathode ray tube, so that operation at supply frequencies would involve considerable danger to life.

Of the available methods of increasing the frequency of the input voltage, that known as the "pulsed oscillator" system has been adopted for the Mullard projection television system.

A very compact construction is achieved by the use of Ferroxcube for the transformer core, and of miniature rectifier valves Type EY51.

Fundamentally the unit comprises a blocking oscillator, a driver valve capable of withstanding high peak anode voltages, and a cascade voltage tripler rectifier circuit.

The circuit diagram is reproduced in Fig. 4 and the complete unit is shown to the right in Fig. 3. The first valve, Type EBC33, is a double diode triode, the triode portion of which is connected as a blocking oscillator. The voltage on the capacitor of the R.C. network rises exponentially until the short circuit pulse from the EBC33 discharges it rapidly and the cycle repeats. The resulting voltage, which has an approximately saw-tooth waveform, is applied to the grid of the driver valve, Type EL38.

This valve is biased well below cut-off so that anode current flows during only a part of the saw-tooth input cycle. On cessation of anode current the energy stored in the inductive component of the anode circuit causes the circuit to "ring" at a frequency determined by the inductance of the anode circuit tuned by all the stray capacitance. In the equipment under review this frequency is of the order of 25 Kc/s.

The peak value of the oscillatory voltage can be derived from the energy equation.

$$\tfrac{1}{2}L_a i_a^2 = \tfrac{1}{2}C_p V_o^2 \quad \ldots\ldots\ldots\ldots \quad (1)$$

Where:

L_a = inductance in anode circuit.

i_a = instantaneous value of anode current at the moment of interruption.

C_p = total stray capacitance including that of the valve.

V_o = peak value of output voltage.

From the above equation

$$V = i_a\sqrt{\frac{L_a}{C_p}} \quad \ldots\ldots\ldots\ldots \quad (2)$$

and substituting practical values (i_a = 100 mA; L_a = 0.5 H and C_p = 50 pF) the peak voltage is:

$$V_o = 0.1\sqrt{\frac{0.5}{50 \times 10^{-12}}} = 10{,}000 \text{ V}$$

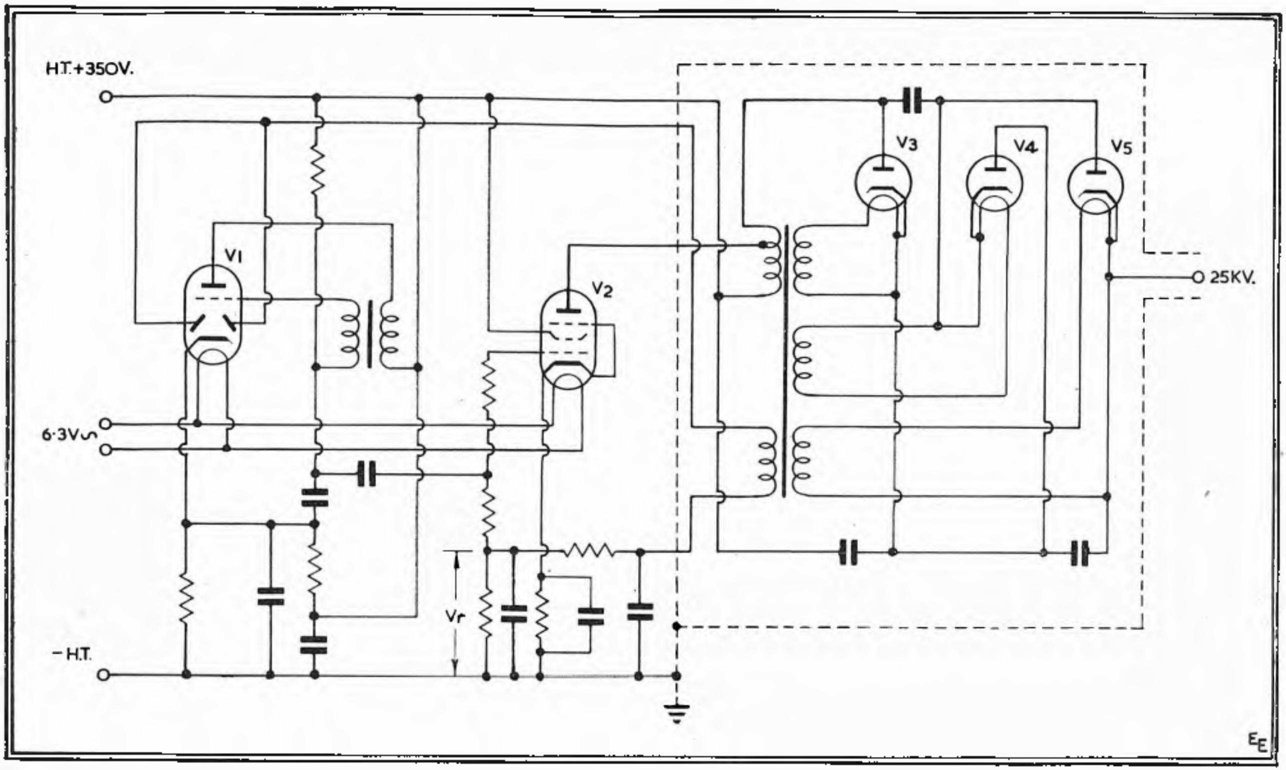


Fig. 4. Circuit diagram of the E.H.T. supply unit

The interruption frequency is given by the formula:

$$L_a \frac{\partial i_a}{\partial t} = \frac{L_a i_a}{\alpha T} = V_a \quad \text{.............. (3)}$$

Whence

$$f_i = \frac{\alpha V_a}{L_a i_a} \quad \text{........................ (4)}$$

Where

$f_i = \dfrac{1}{T}$ = interruption frequency

α = anode current pulse ratio

V_a = permissible anode voltage drop.

Substituting practical values (α = 0.2; L_a = 0.5 H; i_a = 100mA and V_a = 250 V), the interruption frequency is:

$$f_i = \frac{250 \times 0.2}{0.5 \times 0.1} = 1{,}000 \text{ c/s.}$$

This is the value adopted for the repetition frequency in the unit under review.

The peaks of the damped oscillatory wave trains produced by the driver stage are available for rectification and smoothing to give the D.C. E.H.T. supply.

The maximum voltage, as calculated from formula (2) above is, however, somewhat limited as it is impossible to reduce C_p to much below 50 pF, while i_a is restricted by the valve used to about 120 mA and there are practical limitations to the physical dimensions of the inductor L_a, particularly where a very compact power unit is desired.

The circuit is therefore designed to generate a voltage in the order of 8 to 9 KV which is applied to a three-stage voltage tripler rectifying circuit of conventional design using three EY51 half wave rectifiers and producing a rectified output at 25 KV.

It is essential that the E.H.T. power unit for the projection television system has good voltage regulation. Normally, good regulation demands a supply unit of low internal resistance, but it can be shown that, in a unit of the type described, the internal resistance is inversely proportional to the input power, and since the output power required is small, low internal resistance could be achieved only at very considerable sacrifice of overall efficiency.

The problem of voltage regulation has, however, been solved in a simple manner by automatic regulation of the biasing voltage applied to the grid of the driver valve.

An additional winding is provided on the "ringing" transformer, and the voltage induced in this winding is rectified by the diode section of the EBC33 and is then fed via a filter network as grid bias to the driver valve.

As a result of this feature the regulation of the E.H.T. unit is very good for outputs up to about 175 µA after which it falls off rapidly. The regulation curve is reproduced in Fig. 5. This falling-off on overload is very advantageous as it protects the supply unit from damage in the event of short circuits, and also serves as an additional emergency safety precaution.

The three units described form a complete system which, as previously stated, can be adapted to most commercial television receiver chassis without major modification. One precaution, however, should be taken. If either of the time base units should cease to function, the

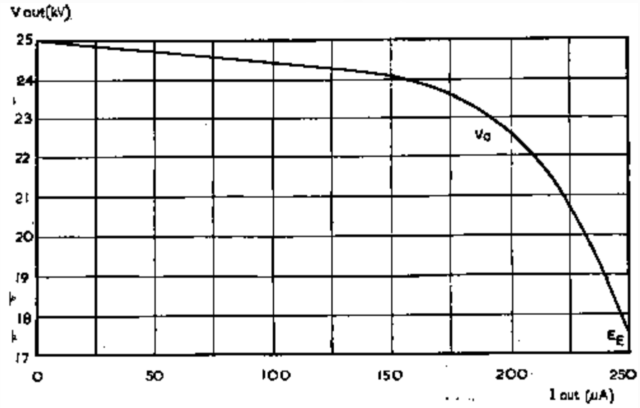


Fig. 5. Regulation curve of the E.H.T. supply unit

high-velocity electron stream will destroy the screen surface of the picture tube along a line which will be horizontal or vertical according to which time base is inoperative.

There are many circuit devices which could be used as a safeguard and the choice will depend upon personal preference. A single example, therefore, will suffice. From a few extra turns on both line and frame output transformers equal voltages are fed, after rectification, to the grid of a valve via a balanced network. Any change in the output of this network due to either or both time bases becoming inoperative results in a change of anode current in this valve, and thus a change in the anode voltage. The anode of the valve is connected to the grid of the cathode ray tube, and the circuit is so designed that the variation of voltage due to time base failure is sufficient to bias the tube to cut-off.

Samples of the Mullard projection television system are now available to setmakers. They reduce the problems of design and manufacture of a projection television receiver to the familiar ones associated with the direct viewing system, since the receiver circuit requirements are substantially the same, and the special problems of high voltage and optics are taken care of in the Mullard projection television system.

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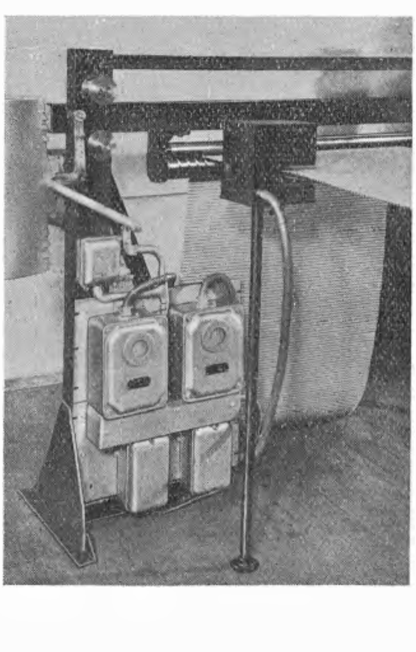
Lateral Control of Strip Materials by Photocells

WHEN strip material such as paper, fabric plastic, etc., has to pass through a machine during a manufacturing process, there is often a tendency for the web to move from side to side, with the result that the roll of material does not layer properly and the overhanging edges become damaged. Lateral wandering also makes it difficult to run the material over rollers for printing.

The difficulty, however, has been overcome by the Electrical Equipment Co. (Leicester) Limited, of 106, London Road, Leicester, with the aid of photo-cell equipment supplied by The General Electric Co., Ltd.



The equipment consists of a slotted lamp and cell unit (see photo) with the lamp in the top half and two photo-cells in the bottom, the cells being masked so that light can fall on one, both or neither, according to the lateral position of the web passing through the control head.

When the web is too far in one direction, both cells are covered, when it is correctly positioned one cell is covered and if it moves too far towards the opposite side neither cell is covered.

These three conditions are used to control, via relays, two solenoids which move an idler roller sideways under the web and gradually edge it, if necessary, in the desired direction. In the "neutral" position, when only one call is uncovered, neither solenoid is energised and the idler roller revolves with the web.

Accuracy of control to within a few thousandths of an inch can be obtained in operation and correction starts immediately there is any deviation from normal position. The maximum standard correction is 2 in. of lateral movement, but for special cases, up to 12 in. can be arranged.

The solenoids used for correction are D.C. operated and have a constant rating, with a linear pull throughout the stroke. Speed of operation is controlled by an adjustable hydraulic damper of the self-filling reservoir type.

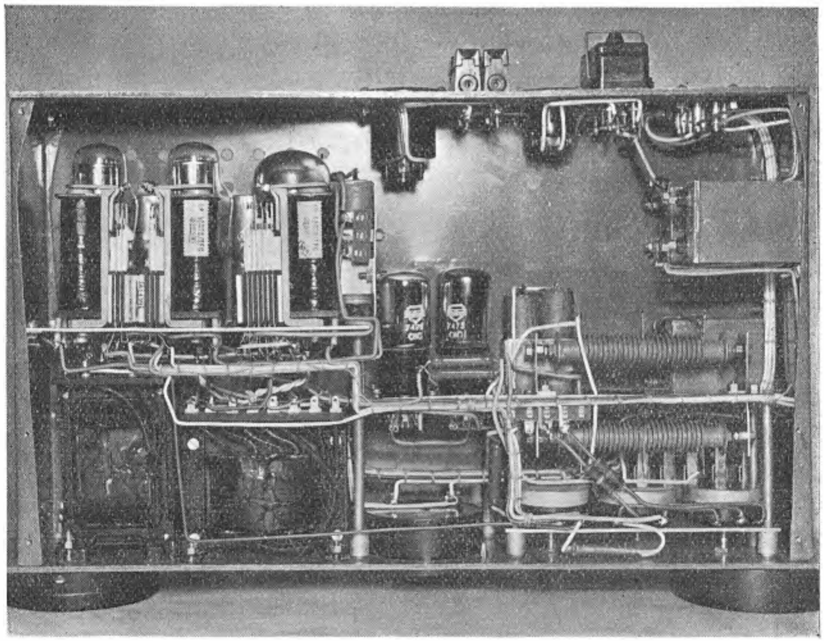
A further adaptation of this principle is guiding the web from a coloured line. When printing paper, etc., it is often desirable to work from a printed line, as opposed to the edge of the web.

This is controlled by the varying degree of light reflexion on the photo-cells, *i.e.*, when light is reflected from white paper on one cell and from a coloured line on the other a similar position is arrived at as in the first case.

In conjunction with this adaptation the equipment is so designed, that, instead of the roller moving the fabric from a fixed light head, the light head moves in a lateral direction on free bearing and follows the datum line. The equipment is designed according to the application. In the case of trimming, cutters are joined to the light head and trim the edge of the web at any desired position; secondly, a guider can be fixed to the head to ensure that plastic material in a semi-liquid state is guided back from the edge of the web to obviate spillage on to the roller, and wastage.

AN ELECTRO-MECHANICALLY STABILISED MAINS SUPPLY UNIT

By A. E. Maine*



Internal view of the Voltage Sensitive Unit. The thyratrons and relays are mounted on the raised chassis to the left and the bridge neons are approximately in the centre with associated controls on the panel below

RECENTLY in this laboratory the necessity arose for the provision of some form of stabilised A.C. mains supply unit in order to meet the power requirements of a group of instruments whose operation is adversely affected by fluctuations of the mains voltage.

The characteristics of a suitable regulator were dictated not only by the large mains voltage variations experienced in the laboratory, but also by special features of the instruments requiring the regulated supply, as well as other factors. The main requirements affecting design procedure are listed below:

(i) Output voltage to be regulated to within 1 per cent for loads up to a maximum of 1.5 KVA.

(ii) To cater for voltage changes extending from −15 per cent to +5 per cent.

(iii) To effect the necessary correction in as short a space of time as possible.

(iv) To introduce no additional distortion of the waveform.

(v) To be robust in every way and capable of satisfactory operation in situations where high levels of vibration are present.

Consideration of the above led to the adoption of an electro-mechanical system, chiefly on account of the fourth requirement.

* Vibration Dept., De Havilland Propellers.

At this stage the basic arrangement of the regulator was laid down in broad outline, and—in common with the generally accepted form—was to comprise a voltage sensitive unit with output relays controlling a bi-directional motor geared to the brush-adjusting mechanism of a smoothly variable ratio transformer. The action is such that departures from the set voltage datum are detected by the voltage sensitive device and control the motor in a way that readjusts the transformer tap until the voltage is reset to the datum, and the system is once more balanced. Here it may be added that, as the power involved is comparatively small, it is permissible for the variable ratio transformer to supply the load directly; whereas, in the case of a higher powered system, the familiar series booster transformer would be employed for reasons of economy.

The fifth requirement listed precluded the use of all fragile devices and components such as sensitive telegraph type relays, galvanometer movements, etc.; hence, an electronic form of voltage sensitive unit was developed. This employs only standard radio type components and the mechanical layout is such that the whole unit is contained within a ventilated casing mounted on a standard relay rack panel 19 in. wide and $3\frac{1}{2}$ in. deep; an internal view of the unit is shown in the photograph above. The unit extends behind the front panel for 10 in. in order to line up with other instruments mounted in the same rack, no attempt having been made to reduce size to a minimum.

The circuit diagram of the complete regulator is shown in Fig. 1. Considering the voltage sensitive device to the right of the diagram, this may be broadly divided into two parts—a neon valve bridge and a voltage sensing circuit employing gas triodes.

The employment of bridge circuits containing non-linear elements for the detection of voltage changes is by no means new, and a bridge containing neon valves as the non-linear elements was selected for the purpose as this offered the possibility of a large unbalance signal for a comparatively small change of input—providing at the same time the requisite stability for this type of apparatus. Also, the associated circuit is made quite simple as no signal amplifier and attendant complication is necessary.

In order to provide the required sensitivity, the bridge, as shown, contains two neon valves in series for each non-linear arm, the resistance values being so chosen that the neons work under conditions of maximum stability at the regulating voltage decided upon,

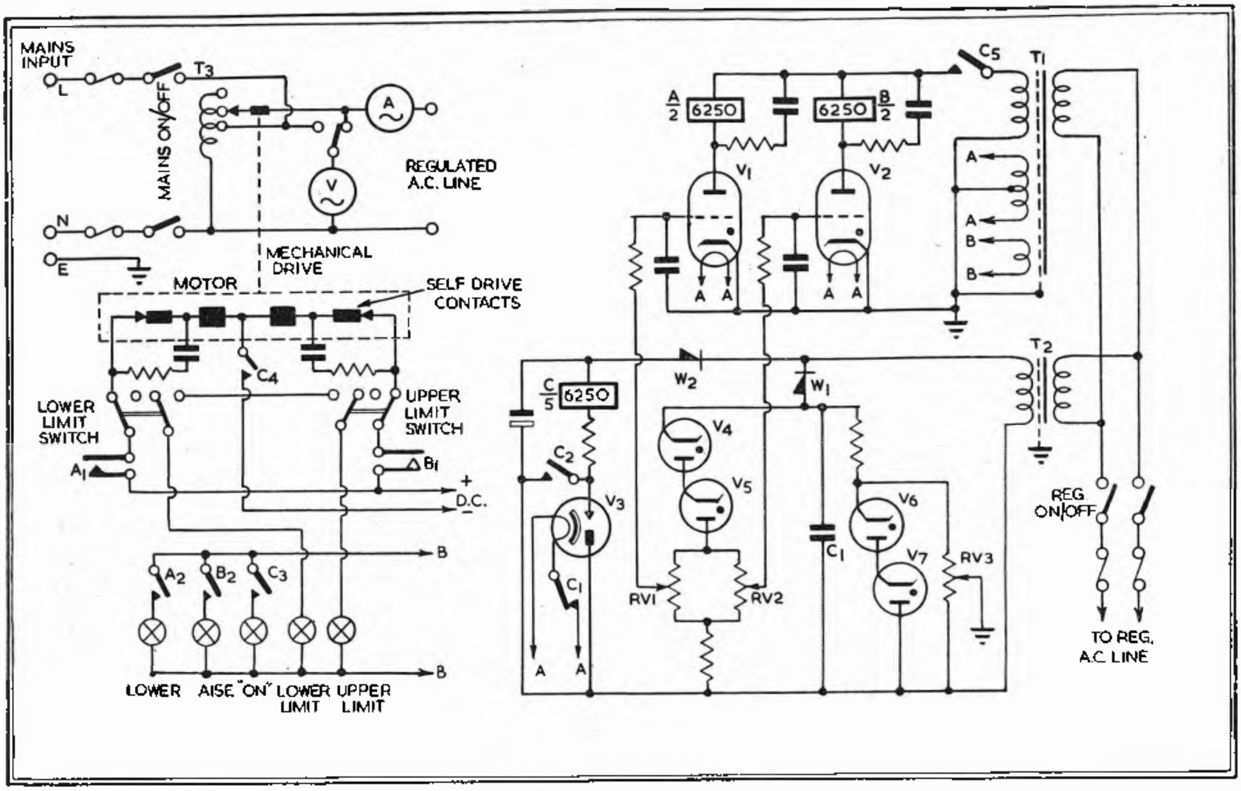


Fig. 1. Circuit diagram of the complete Regulator

with the bridge almost symmetrical.

The supply to the bridge is obtained from a half-wave metal rectifier (W_1) of adequate thermal rating and this is joined in series with the secondary of a step-up transformer whose primary is connected across the regulated side of the mains. The smoothing capacitor C_1 is made much smaller in value than that conventionally used, in order to remove an otherwise intolerable time delay. As may be expected, the supply to the bridge contains a pronounced ripple at the mains frequency, but as this is arranged to be in phase with the A.C. supply to the gas triode it is of no consequence as far as operation is concerned.

The action of the bridge, when supplied with pure D.C., is easily visualised in view of the fact that the neons may be considered to be constant voltage devices over their working range. It follows that, if the resistive arms are properly dimensioned, a condition of balance can be obtained for a suitable value of applied voltage; if this is subsequently changed, the full magnitude of this change is developed across the resistive arms and so appears across the "signal" corners of the bridge.

In the case of the regulator, however, the supply to the bridge, as stated, contains a pronounced ripple and this, by the action described, appears across the corners of the bridge. It is necessary, then, to consider the bridge as balanced when the bridge signal contains no D.C. component.

It is required that the signals from the bridge be capable of operating a pair of gas triodes in such a manner that correct "sensing" results; this is arranged by taking two signal feeds from the left-hand side of the bridge (RV_1, RV_2) with respect to a common point (RV_3) at the right-hand side of the bridge and joining them to the appropriate terminals on the input to the sensing circuit.

To clarify the operation of this part of the circuit, consider the case when the bridge is correctly balanced, as described, and all three bridge controls are set to their "upper" limit of travel in the diagram. The resultant signal will be at mains frequency containing no D.C. component and both thyratron grids will be driven alternately positive and negative. If now the slider of RV_3 is lowered, the common cathodes of the thyratrons will receive a positive D.C. voltage which acts as negative bias on the grids. It is clear than that a point can be reached when only the positive crests are capable of initiating conduction within the gas triodes. If now RV_1 and RV_2 are both set to their mid-travel point, a further downward movement of RV_3 is required to re-establish the conditions just mentioned. Finally, if RV_1 is turned fully down, the negative bias on V_1 is decreased and so conduction commences earlier in the positive half-cycle, and if RV_2 is turned fully up V_2 is cut off altogether. As far as the operation of the regulator is concerned, these conditions represent balance at the particular regulating voltage employed.

Now each gas triode has a relay connected in series with its anode and the contacts of these relays control the corrective movements of the brush adjusting mechanism of the transformer T_3. The contacts of the relay $A/2$ are "break" sets, while the contacts of the relay $B/2$ are "make" sets. Referring to the balance conditions mentioned in the last paragraph, it will be apparent that at balance, as V_1 is conducting and V_2 cut-off, both sets of relay contacts will be open and the motor will not run. If now the mains voltage rises above the pre-determined value as set in by RV_3, both gas triodes are cut-off as the D.C. component drives both grids more negative; hence, the break contacts close and so the motor runs, re-adjusting the line voltage in the manner already described. Conversely, if the mains voltage falls, the negative bias on both grids decreases, *i.e.*, the grids are driven more positive, hence both thyratrons conduct and the "make" contacts close the circuit to the reversing side of the motor, once more correcting the line voltage.

To sum up: with both thyratrons "on" the motor raises the voltage, and with both thyratrons "off" the motor lowers the voltage; one valve conducting and the other cut off is the condition of balance, with the motor stationary.

Before proceeding further, it is to be noted that the thyratrons are not used purely as marginal devices owing to their instability in the region of critical bias. With the circuit values chosen, it is possible to set the standing bias at balance to the "critical" +20 per cent on one valve and to the "critical" −20 per cent on the other, and yet maintain a sensitivity equivalent to 1 per cent of mains input change. Changes of grid conditions outside these limits are not normally encountered. Maximum stability may be expected by employing only argon filled gas triodes, feeding their heater transformers from the regulated side of the mains, paying due attention to mounting and ventilation, and ensuring that the peak anode current is kept well below the maximum rating of the valve.

So far no mention has been made of the type of drive motor employed; the characteristics of this unit are, of course, of prime importance to the operation of the regulator as a whole. In this particular case, the motor was specially designed for another apparatus, but has been found ideally suitable for this application as well. The motor employs the ratchet principle and bears many similarities to the Post Office type of uniselector mechanism and to other similar types of stepping relay. Essentially, it comprises two electro-magnetic ratchet units mounted back to back in a single case with their respective drives coupled to a common shaft by means of an epicyclic type of differential gear. Thus, by energising one or other ratchet unit, a forward and reverse drive is possible. The motor, which by means of self-drive contacts provides its own current pulses, is D.C. operated and hence requires a rectifier in circuit with the mains supply; no smoothing is necessary.

The available torque from the motor is constant under all conditions of load up to the stalling point and is independent of supply voltage over the working range of the motor. However, the most attractive features of the motor for this and similar applications are that the effects of inertia are eliminated —the motor starting, stopping and reversing immediately upon operation of the associated relays—and also that the drive shaft is rigidly locked in position at all times when the motor is not running.

Employing this drive unit in the regulator, recovery speeds of 600 volts per minute have been reached with complete stability and still higher speeds are available. The effective suppression of transient changes is apparent when it is considered that a 10 volts fluctuation can be corrected in one second.

Referring once more to the circuit diagram, there are several practical points of some interest. All the relays used are of the P.O. 3000 type fitted with 6250Ω coils and, in the case of the control relays, suppression components are connected in parallel with the operating coils to prevent back E.M.F.'s from interfering with the operation of the thyratrons.

The safety devices and monitoring circuit incorporated into the regulator are somewhat elaborate for this class of apparatus but, at the same time, it is considered that their inclusion lends to reliability and assists in servicing. The safety devices consist of fuses in all supply leads, limit interlock switches on the brush adjusting mechanism of the control transformer, a delay switch and associated relay ($C/5$) for protecting the thyratron cathodes on first switching on and, finally, a slipping clutch in the gear train to the motor.

With respect to the monitoring circuits, an A.C. voltmeter is fitted to the front panel with provision for reading either input or output volts and, in addition, there are five coloured lamps. Two of the lamps indicate the normal functioning of the regulator, being controlled by auxiliary contacts on the motor relays; the third lamp indicates the expiration of the initial delay period, while the remaining two indicate the engagement of the transformer tap arm with either of its mechanical stops.

The variable ratio transformer, the motor and gear unit and other auxiliaries, are mounted together as a common unit, connexion being made to the Voltage Sensitive Unit by way of a multicore cable. The motor unit is easily detachable and provision is made for hand control to be assumed in the unlikely event of breakdown.

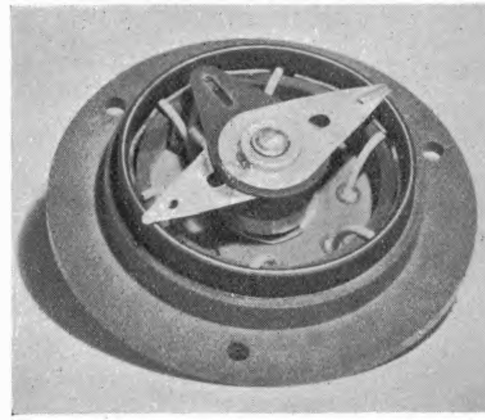
The initial setting of the regulator is facilitated by arranging all the bridge controls on the front panel in the form of screwdriver-operated pre-sets. Once set, RV_1 and RV_2 are locked in position and RV_3 is left unclamped to enable the precise value of regulating voltage to be set as required. Having adjusted the transformer primary taps on T_1 and T_2 and set RV_3 for a particular input voltage, this may subsequently be changed merely by adjusting RV_3. If the change exceeds 5 per cent, the transformer taps must be adjusted as well.

With respect to general performance, the regulator described has been in service almost continuously for some months and it has only been necessary to make a slight adjustment to RV_3 on two or three occasions. In each case the discrepancies were small and are attributed to the ageing of the type 7475 neons used in the bridge. During this running period mentioned, 48-hour recordings of mains input versus mains output have been made from time to time and in no case were errors found in excess of 1 per cent. The regulator is installed in a situation where considerable vibration is present, but this has not interfered with the operation in any way.

In conclusion the author wishes to thank Mr. R. N. Hadwin, Chief Vibration Engineer of Messrs. De Havilland Propellers, Ltd., who initiated the work, and also Messrs. De Havilland Propellers both for giving permission for the article to be published and for providing the photographic illustration.

Resistive Phase Shifters

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



RECENTLY, many new applications for phase comparison have appeared in telecommunications, particularly in radio navigation and remote position control.

A remote-position-control system using phase comparison is less affected by changes in the transmission system than one using amplitude comparison. Frequency comparison involves greater component stability in the receiver to obtain a similar accuracy. An accuracy of 1° in 360° is adequate for many purposes. Neglecting the effect of transmission, this figure represents the practical limit of precision for a simple servo system, part of which will be mentioned briefly. A resistive phase-shifter supplied from an electronic single to six-phase conversion unit of the type to be described was demonstrated at the exhibition of the Physical Society in April last year.

There are three important classes of phase-shifter. The most commonly used instrument has been the magnetic goniometer. This phase-shifter has been developed for various frequencies, and it has been reduced in weight and size to meet airborne equipment requirements. A typical magnetic field phase changer weighs 8¾ oz. and has an accuracy of phase change of $\pm\frac{1}{2}°$. Higher accuracy is obtainable, but it is believed the limit is about $\pm 10'$. It is inductive and careful design is necessary to avoid phase errors in coupling the stator to an amplifier. Amplitude variations may then result in considerable errors of phase.[1] The capacitance goniometer has been used for some special applications involving high frequencies, but it appears to be of limited use for low frequencies because of the difficulty in making the coupling impedances sufficiently small. The third class is the resistive phase-controller. The attractions are compactness, light-weight and easily attained accuracy of the order of $\pm\frac{1}{2}°$ in its mechanical construction. The photograph, above, illustrates an experimental model weighing 1.9 oz. It is housed in a drawn aluminium case, in which a long bearing for the control shaft is fitted. The surface, which supports the potentiometer, and the bearing, are machined after assembly to avoid eccentricity. The total track resistance is about 10,000 ohms. The most important details in construction are concentric assembly of the track, the bearing of the wiper arms, and the driving shaft. The track must have a regular winding and accurate angular positioning of the winding. The limit of accuracy is determined by the pitch of the winding. If precision manufacture is required, it appears desirable to avoid using specially shaped tracks and to make this simple phase shifter from a continuous resistance of toroidal form. Fig. 1 gives alternative supply arrangements. There may be one or more wiper-arms positioned to give outputs of various phase differences.

If the track is not specially shaped, there will be a non-linear relationship between the output phase and the angle of rotation of the wiper-arms. Fig. 2 gives this theoretical relationship for potentiometers with a maximum of six tapping points. With six tapping points the discrepancy, 1°7′, is quite small. In the case of a remote position control system employing similar phase potentiometers for control and reset position identification, these errors have been eliminated by inserting a fixed phase shift of 90° in the reference channel. This corrects the quadrature condition that exists between the reference and control channel phases when the phase comparison discriminator is balanced. Thus both transmitter and receiver wipers are similarly disposed on the tracks.

Fig. 3 shows the amplitude variation as the wiper arm is rotated. This causes changes in sensitivity, and may sometimes be responsible for hunting in the servo. In some applications it could be removed by a form of A.G.C.

There is no difference between the floating three-phase and six-phase potentiometers so far as ultimate accuracy of phase, and amplitude variations are concerned, but if several potentiometers are to be fed from the same multiphase supply the floating three-phase potentiometer cannot be used. The efficiency of the six-phase potentiometer is higher than the floating three-phase potentiometer. The input power is 2.5 db. less for a given voltage at the slider.

Typical Applications

An experimental servo system, using phase comparison between two channels at a frequency of 500 c/s., has been built. The errors involved have been assessed as follows:

(1) Potentiometer errors are within $\pm 1°$. Improvements in design are now being made, which should reduce the error to within $\pm\frac{1}{4}°$.

(2) Total peak voltage of harmonics produced by the amplifiers and transmission system is 1 per

* Research Laboratories of The General Electric Company, Limited, Wembley, England.

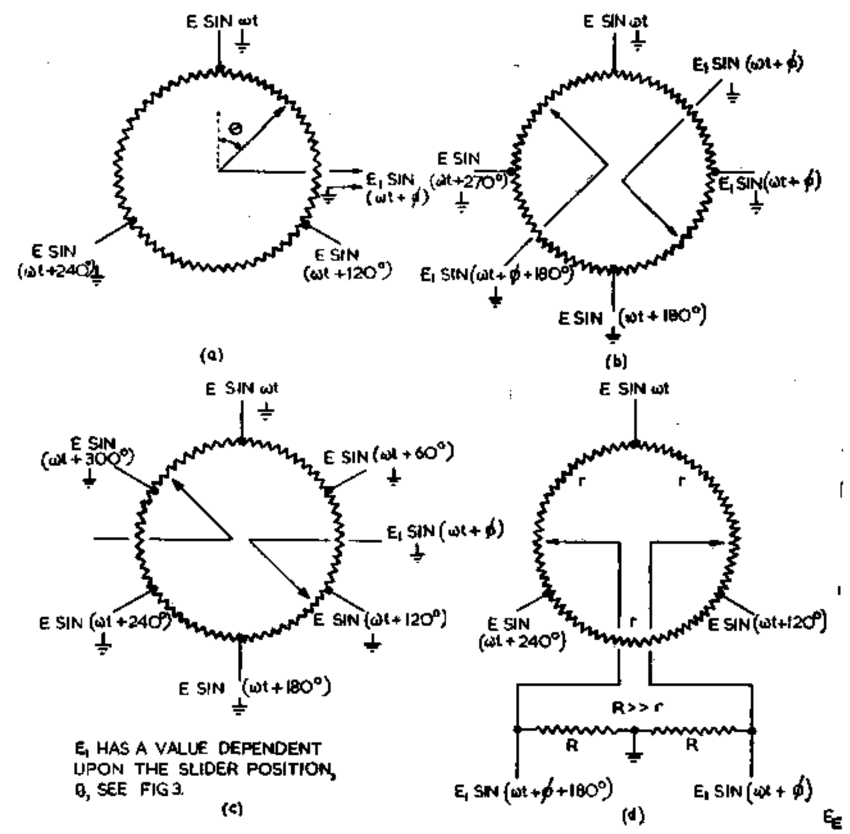


Fig. 1. Arrangements of the resistive potentiometers. The tracks are wound for linear resistance change, and the tapping points are accurately spaced

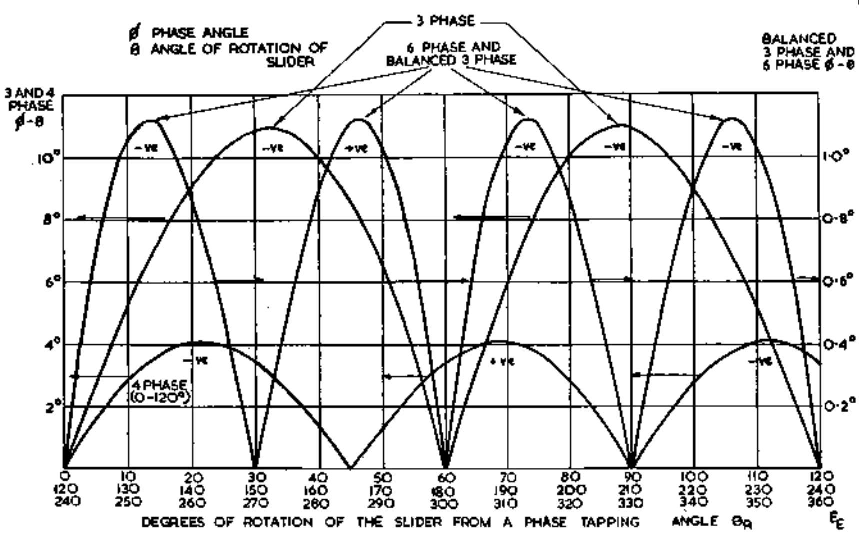


Fig. 2. Phase error

cent of the fundamental. Under the worst conditions this could introduce an error of approximately $\pm\frac{1}{2}°$. The error is introduced at the phase discriminator, and may be reduced by a wave filter.

(3) Errors in phase between the receiver and transmitter phase-conversion units are within $\pm\frac{1}{2}°$.

Phase 1	0°.	Phase 2	0°.
Phase 3	$-\frac{1}{2}°$.	Phase 4	0°.
Phase 5	$+\frac{1}{2}°$.	Phase 6	0°.

No special adjustments have been made to obtain this accuracy.

(4) Sensitivity of the phase discriminator in resetting the position of the control motor $\pm\frac{1}{4}°$.

In addition to these errors must be added phase shift due to circuit drift. Under normal conditions the total error was found to be within $\pm 2\frac{1}{2}°$. It is felt that improvements could be made which would bring the errors to within $\pm 1°$, neglecting transmission errors, but without elaborate circuits.

There are other possible applications of the resistive-phase-shifter which may make use of it's being sensibly independent of frequency over a wide range. Three of these are cited:

(1) If a transmission circuit and a phase-shift unit are fed from the same frequency source, the outputs may be compared for phase, and the phase-shifter adjusted to give the same phase output as the circuit under test. A cathode-ray oscilloscope with the vertical deflector plates connected to the output of the transmission circuit, and the horizontal deflector plates connected to the output of the phase shifter, makes a very satisfactory indicator. With a 5-in. diameter screen and a fine trace, the readability is about $\pm\frac{1}{2}°$.

(2) It is common practice to obtain a train of pulses with a fixed time interval between them, from a sinusoidal waveform, by limiting and differentiating. The onset of these pulses may be shifted in time by shifting the phase of the sine wave. This technique appears in multiplex telephony, radar range finding and associated test equipment.

(3) If A.C. is applied to the anode of a gas triode the angle of flow of the anode current may be controlled by the relative amplitude and phase of the applied grid voltage. The range of control of the angle of

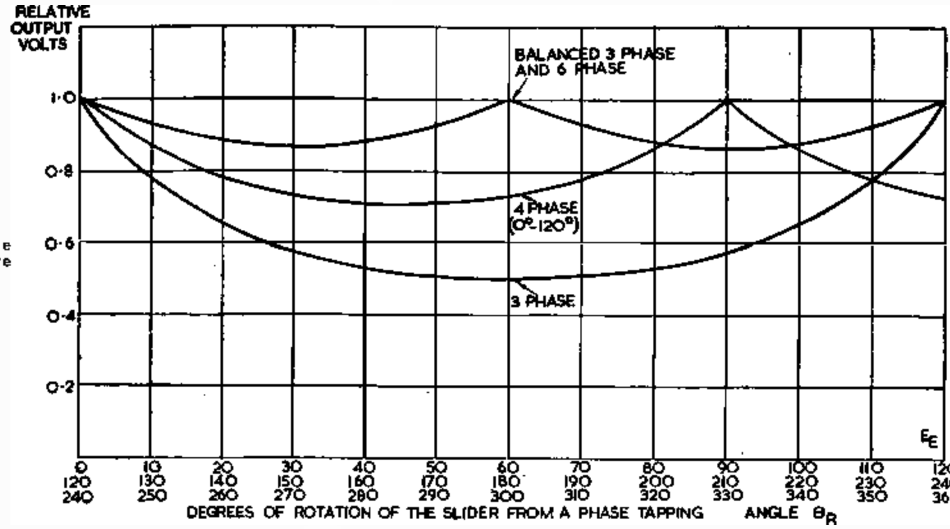


Fig. 3. Amplitude variations of the voltage at the slider of various resistive phase shifters

anode current flow is 0° to 180° for a similar range of phase at critical grid voltage. This method is due to Toulon[2], and is used to control the temperature of an oven, the speed of a D.C. motor and the output of grid controlled rectifiers. If a three-phase supply is available, the resistive phase-shifter should be satisfactory for controlling the grid voltage.

When proportionality between phase and the angle of rotation of the wiper-arms is important the following[3] is suggested as a simple method of approaching this requirement with a minimum of precision mechanical engineering.

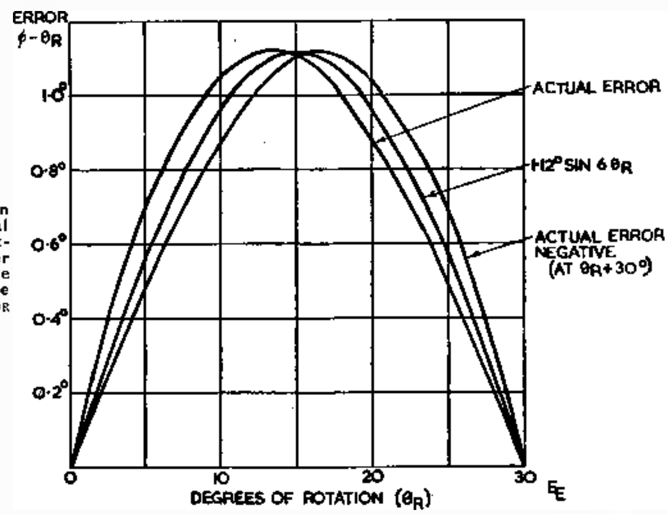


Fig. 4. Comparison between the actual error of the six-phase potentiometer and the approximate expression for the error 1.12 Sin $6\theta_R$

Mechanical Correction of Error

Fig. 4 compares the negative and positive phase errors with a sinusoid of $1.12°.\sin 6.\theta_r$. It will be seen that they diverge by less than 8′.

A mechanism for displacing the wiper-arms about the position of the driving shaft is shown diagrammatically in Fig. 5 (a) and (b). The displacement is also sinusoidal with rotation of the shaft. For one revolution of the driving shaft "C" the gear wheel "B" is rotated six times, due to its meshing with the fixed gear teeth "A". On the other side of the small gear wheel "B" is an eccentrically disposed pin. As the eccentricity is small, it may be cut on a continuation of the shaft. This eccentric pin "E" engages in a slot which is fixed to the wiper-arm assembly. The bearings and shaft of the gear wheel "B" must be accurately made, but the gear teeth do not require a high standard of machining.

Single to Multiphase Conversion

Fig. 6 illustrates one successful method of obtaining a six-phase supply from a single phase voltage.[4] Any number of output phases can be obtained by inter-connecting a suitable number of secondary windings. The two stages provide an equal amount of power to the phase shifter. The first stage provides the in-phase current for a double-Scott-connected transformer. The second stage provides the quadrature current. Fig. 7 shows the basic circuit for obtaining the quadrature input voltage for the second stage. This method of obtaining the quadrature voltage allows adjustment for phase to be made without affecting amplitude. If the capacitor C is made variable the circuit is adaptable over a range of frequencies. The effective gain of the cathode follower of the second stage is corrected to unity by reduction in the turns ratio between the primary and secondary of transformer $T2$ compared with the ratio of transformer $T1$. A preset control for adjusting the secondary voltages of $T2$ is a desirable refinement. It consists of a 500 ohm potentiometer connected across 5 per cent of the primary at the cathode end of the winding.

Fig. 8 illustrates a method of indicating when the phase shift in the "quadrature network" is correct. The connexions of the secondaries of the transformers from the star point (earth) to phases 2 and

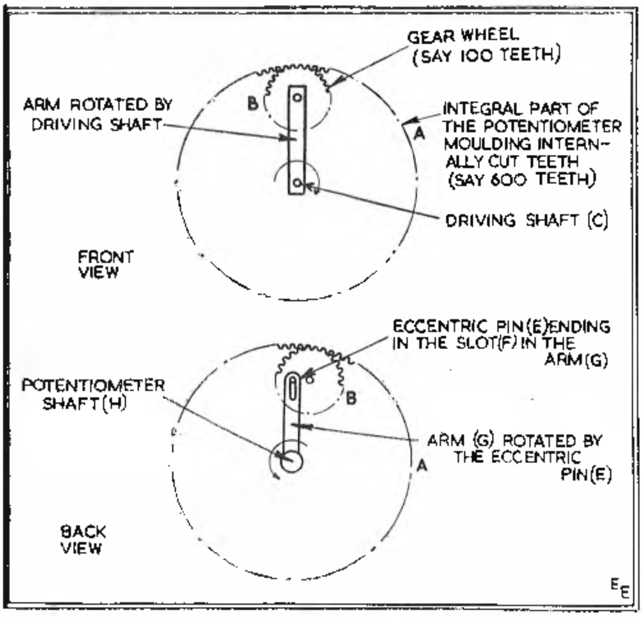


Fig. 5a. Mechanical method of correction to obtain phase angle to shaft rotation proportionality

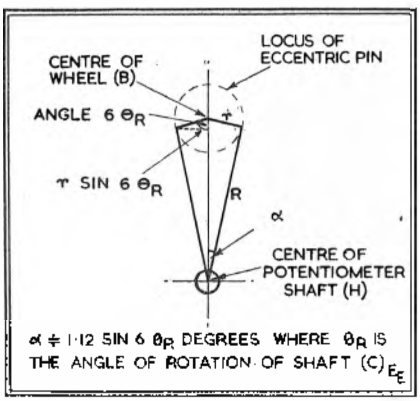


Fig. 5b. The displacement between driving and driven shafts in Fig. 5a. The actual correction is slightly better than the approximation

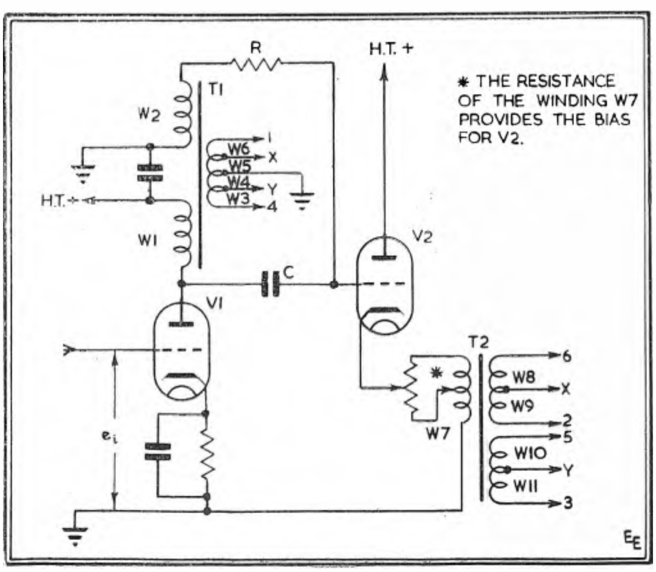


Fig. 6. A method of obtaining a six-phase supply from a single-phase voltage

and 6 may be likened to a phase discriminator. The peak voltages of these two phases will only be equal when a correct quadrature condition is obtained. Then the galvanometer will indicate zero current.

Unless considerable care is taken in the design of the transformers, total errors in this circuit greater than 10° may be experienced. It has been found that the errors may be reduced to about ½° if the primary is sectionalised and the secondaries wound between the sections. The primary should have a reactance of about four times the load resistance referred to the primary. Windings W_4 and W_5 must be wound to have a resistance half that of W_3 and W_6. The secondary windings should be disposed to have identical couplings with the primary.

The total load resistance of the phase-shifter referred to one secondary winding of transformer T_1 is $R/72$, where R is the loop resistance of the track. The windings W_3, W_4, W_5 and W_6 have an equal number of turns and W_1 is equal to W_2. The ratio of W_1 to W_3 is chosen to correctly load the valve.

$\frac{W_1}{W_3} = \sqrt{\frac{72R_0}{R}}$ where R_0 is the optimum load for the valve. The primary W_7 of the transformer T_2, has the same number of turns as W_1 less a number of turns to allow for the grid to cathode A.C. voltage drop of V_2

$$W_8 = W_9 = W_{10} = W_{11} = \sqrt{3}\, W_3,$$

in turns ratio.

At the servo transmitter the phase shift between the input voltage to V_1 and the secondaries of transformer T_1 is not important. The reference channel phase is obtained by a potentiometer connected across phases 2 and 3. At the receiver, and in other applications where this phase shift must be kept constant, voltage feedback from a secondary of transformer T_1 to the cathode of an additional amplifier stage preceeding V_1 has been found satisfactory.

For remote position control, the difficulties in making the transformers have been alleviated by identical design of the receiver and transmitter phase conversion units. The measured errors were quoted earlier. It is felt that for greater

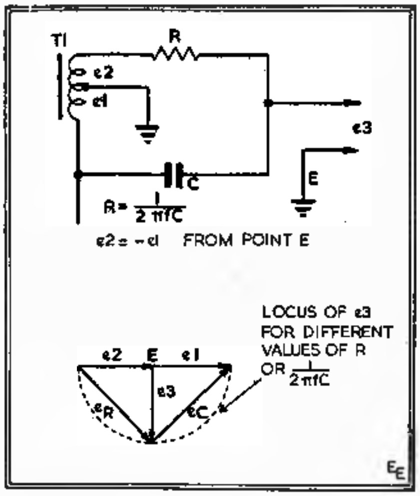


Fig. 7. The basic circuit for obtaining the quadrature voltage for the second stage

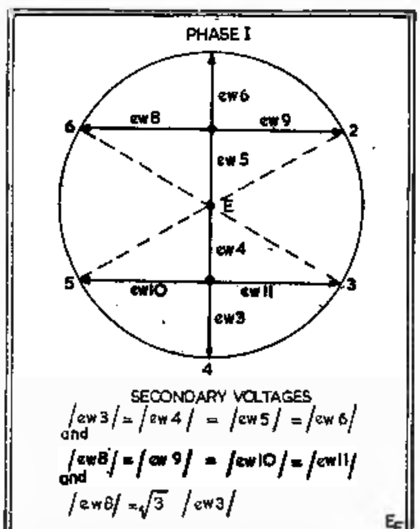
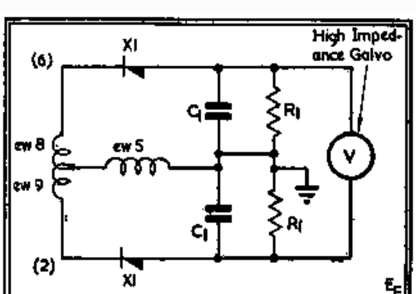


Fig. 8 (above). Voltage relationships for the secondary windings of transformers T1 and T2

Fig. 9 (below). A method of appreciating the adjustment of the quadrature network by using the transformer secondaries as a phase discriminator



precision additional resistances could be added to correct each phase.

In some cases, such as when there are large temperature variations, the temperature stability of C and R is important. A change in this time constant will cause errors in phases 2, 3, 5 and 6. In an experimental six phase convertor operating at 500 c/s., R was a little over 300 K ohms, and C was 1,000 pF. The capacitor was a high stability single plate silvered-mica component having a temperature coefficient of less than plus 40 parts in 10^6 per °C. The resistor was wire-wound and had a similar temperature coefficient. This gives the temperature coefficient of phase at the quadrature position as less than 0.01° per °C.

An alternative circuit has been published recently.[1] This has an input circuit giving a 45° advance for a first cathode follower, and a 45° lag for a second cathode follower. The cathode followers feed the primaries of the phase transformers. The effect of temperature on frequency changes is to rotate all the phases by the same amount, but an additional error is introduced by differential amplitude changes between the quadrature voltages. The total errors are about the same as for the circuit described here, but the circuit in Fig. 6 has a higher input impedance, requires a lower input level and has half the number of critical phase shift components. And no phase errors are introduced in phases 1 and 4 by changes in the receiver "quadrature network." At least one setting which has a high accuracy of positioning is often a desirable feature in remote position-control.

It is suggested that the galvanometer (Fig. 9) may be replaced by D.C. amplifiers arranged to control a motor which can automatically adjust R or C.[5] The whole circuit then becomes a network which will cause a specific phase shift independent of temperature and frequency, within its range of operation.

References

1 Ashton and Beresford: "The Omni-directional Radial Track Guide." *J.I.E.E.*, Vol. 94, Pt. IIIa, No. 16, p. 997.

2 Brit. Patent 209426, Toulon.

3 Prov. Patent 520/48, G.E.C. and Bryden.

4 Prov. Patent 18571/48, G.E.C. and Bryden.

5 Prov. Patent, G.E.C. and Bryden.

The Fallacy of the Best Two Out of Three

In scientific and engineering investigations a single measurement of an unknown quantity is seldom considered sufficient. Two or more measurements are usually made in order to establish a check on instrumental errors, operator's errors in making readings, and the reliability of the sample.

Three measurements are the minimum number that can reveal one of the measurements to be unreliable in a new experimental situation. Intuition suggests that if two of the three measurements are in close agreement while the third stands apart considerably removed from either of the others, then there may be grounds for suspecting and perhaps rejecting the third value. In terms of the difference between the two in good agreement, how different may the third measurement be before it should be suspected? Since this problem is important to all who make and interpret measurements, it is a little surprising that an answer has only recently been found.

An approximate solution has been obtained in the National Bureau of Standards statistical engineering laboratory through an empirical study of triads drawn at random from a large group of measurements constructed to conform to the characteristics of a normally distributed set entirely free from any gross errors. The following table shows, for certain ratios, the results of an empirical sampling study of 400 sets of three measurements compared with those predicted by the exact mathematical solution of the problem.

Ratio of large to small difference	Frequency in 400 sets	Theoretical frequency
4.0 or more ...	149	145.2
9.0 „ „ ...	76	69.4
19.0 „ „ ...	38	33.9

These results reveal that in an average of one out of every twelve sets, one of the measurements will be at least 19 times farther away from its neighbour than the difference separating the two closest. Since in every 12 sets one shows such a spacing for measurements with no gross observational errors, it appears that measurements that should be retained are often dropped from the records.—From the *Technical Bulletin*, National Bureau of Standards (U.S.A.)

Distortion of Scanning Waveforms

due to Inadequate Frequency Response

By G. G. Gouriet, A.M.I.E.E.*

IN designing mass-produced commercial television receivers, it is common practice to obtain a linear saw-tooth current waveform for the purpose of magnetic scanning by arranging that various distortions shall cancel. Since such apparatus is designed to operate under prescribed and fixed conditions this procedure is justified, but in the case of laboratory or transmitting equipment where independent adjustment of scanning amplitude and linearity may be required, it is very desirable that all circuit elements should introduce minimum distortion.

When magnetic scanning is used it is invariably necessary to use scanning current transformers in order to avoid using excessively large scanning coils, or, alternatively, valves with a very high emission.

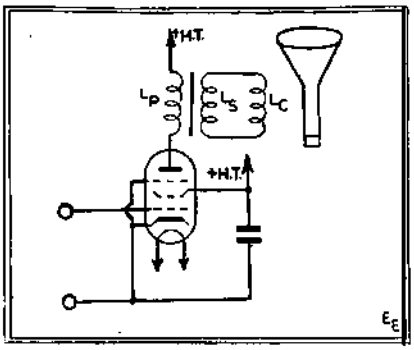
The problem of designing such transformers at line frequency is largely that of obtaining a response to high frequencies sufficient to permit of a rapid flyback. With modern transformer design this can usually be accomplished without much difficulty. At frame frequency the problem is to obtain sufficiency response at low frequencies, and here the requirement is far more stringent since the low frequency response determines the linearity of the scan.

In this article it is proposed to discuss the requirements for a linear scan in general terms, and then to deduce the distortion that will result from insufficient bass response. Finally, suitable means of compensation will be discussed.

General Requirements for Magnetic Scanning

The most fundamental requirement of a magnetic scanning system is that a linear saw-tooth current waveform should pass through an inductance without distortion. This can only be achieved in two ways: firstly, the current may be derived from a generator of infinite impedance, in which case it will be independent of the coil impedance, or alternatively a voltage waveform derived from a zero impedance source, and proportional to the first derivative of the desired saw-tooth, may be applied across the coil, in which case the current through the coil will be of the required saw-tooth form, *provided* the coil is a pure inductance. Since it is easier to produce a current source with a relatively high impedance than it is to produce a coil and generator with a total resistance that is substantially zero, the former method is normally adopted.

The natural choice of generator is a pentode valve. Fundamentally, the only requirement for linearity is that the design of the scanning coil shall be such that its impedance is low compared with the valve impedance at all the significant frequency components of the saw-tooth waveform. Failure to comply with this requirement will naturally have the greatest effect at high frequencies, since the coil reactance increases with frequency, and therefore the distortion will occur mainly during the fly-back period. Practically, for a given valve current the scanning coil must be sufficiently large to provide the ampere-turns product required for deflexion, and while the necessary compromise presents no difficulty, the physical size of such an inductance is normally considered too large to support around the neck of a cathode-ray tube, and a current transformer is therefore used in conjunction with a relatively small scanning coil.



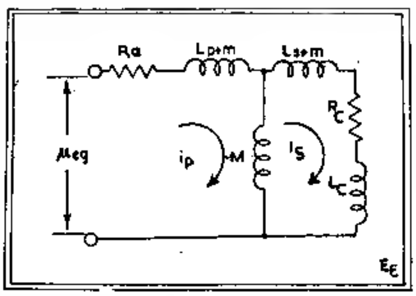


Fig. 1a (above). Practical scanning circuit
Fig 1b (below). Equivalent circuit at low frequencies

Assuming that an ideal transformer is used, the conditions are virtually unchanged. The ampere-turns product of the scanning coil, and the inductive load presented to the valve will be unaltered, but in practice we can only approach the performance of an ideal transformer by making the primary open circuit impedance high compared with the referred load. The problem is, of course, no different from that involved in the design of an audio-frequency transformer used for speech currents, except that, as will be shown later, a different quantitative interpretation must be given to the requirement, " . . . high compared with the referred load," if linearity of the saw-tooth is to be preserved.

Practical Design

In practice the output circuit of a typical scanning wave-form generator will be of the form shown in Fig. 1(a). At low frequencies the equivalent circuit will be approximately that of Fig. 1(b) where R_c is assumed to include the transformer secondary resistance.

Assuming a perfect pentode, so that

$$R_a = \infty$$

and a transformer coupling factor of unity, whence,

$$M = \sqrt{L_p L_s}$$

we obtain for the secondary current

$$i_s = e_g g_m \frac{\sqrt{L_p L_s}}{L_s + L_c} \cdot \frac{\omega(L_s + L_c)}{\sqrt{R_c^2 + \omega^2 (L_s + L_c)^2}} \Big/ \tan^{-1} \frac{R_c}{\omega(L_s + L_c)} \quad \text{(1)}$$

* Research Dept. Engineering Division, B.B.C.

The term $\frac{\sqrt{L_p L_s}}{L_s + L_e}$ is the current transformation ratio which if $L_e << L_s$ becomes $\sqrt{\frac{L_p}{L_s}}$ as is well known.

The remaining expression on the right is the familiar complex factor which expresses the gain and phase characteristic as a function of frequency. This same factor expresses the voltage developed across the inductance of the circuit shown in Fig. 2(a).

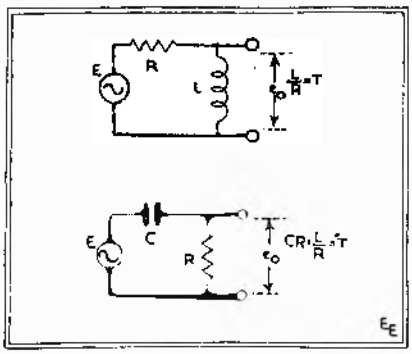


Fig 2. Showing equivalent bass-loss circuits
(a) across inductance
(b) across capacitor

More generally we may write the factor in the form

$$\frac{1}{\sqrt{1 + \frac{1}{\omega^2 T^2}}} \Big/ \tan^{-1} \frac{1}{\omega T} \quad \ldots\ldots (2)$$

where T is the time constant of the circuit, and it will apply also to the capacitative circuit of Fig. 2(b). This form of expression is useful for the purpose of comparison with the transient response, which will be dealt with in the following section.

In the case of a practical circuit tested in the laboratory, the values associated with the equivalent circuit in Fig. 1(b) were:—

$L_e = 6$ mH. $R_e = 8\Omega$
$L_s = 50$ mH. } Impedance ratio $= 675 : 1$
$L_p = 34$ H. }

Substituting the appropriate values in (2), where

$$T = \frac{L_e + L_s}{R} = 7 \times 10^{-3}$$

we obtain for the current attenuation at 50 c/s.

$$A = 20 \log_{10} \frac{1}{\sqrt{1 + \left(\frac{1}{7 \times 10^{-3} \times 6.28 \times 50}\right)^2}}$$

$$= 20 \log_{10} \frac{1}{\sqrt{1.21}} = 0.85 \text{ db.}$$

Regarding the justification of the assumption made at the beginning of this section, the value of R_s was known to be greater than 300 $K\Omega$ which, compared with the referred load of approximately 5.4 $K\Omega$, may justifiably be regarded as infinite, while the transformer coupling factor was measured and found to be 0.988.

For many audio frequency purposes, a transformer introducing an attenuation of less than 1.0 db. at the lowest frequency limit would undoubtedly be considered satisfactory. In the following section we shall see how the same circuit will affect the linearity of a 50 c/s. saw-tooth waveform.

Steady-State Response of Network to Saw-Tooth Waveform

In order to calculate how the circuit of Fig. 1(b) will affect the linearity of a current saw-tooth waveform, we shall examine the steady-state response of the circuits of Fig. 2 to an impressed saw-tooth

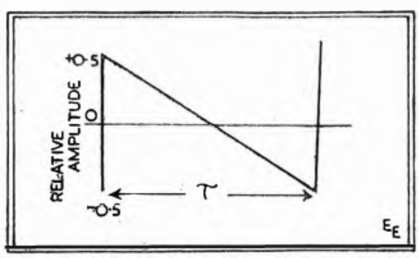
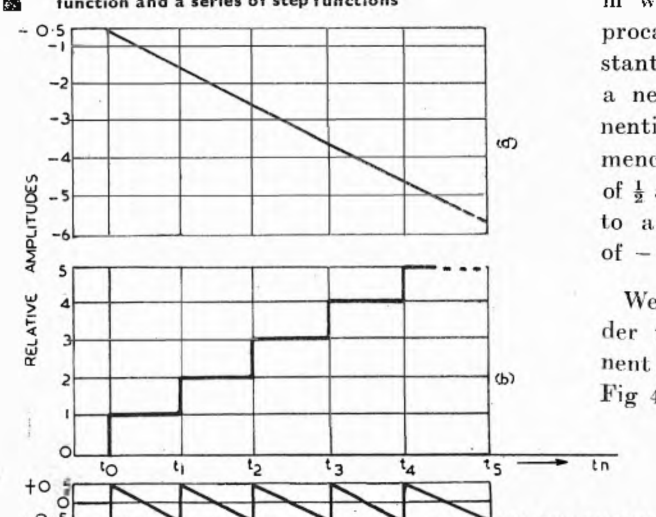


Fig. 3. One cycle of saw-tooth defined as $\left[\frac{1}{2} - t/\tau\right]_{t=0}^{t=\tau}$

Fig. 4. Showing how a saw-tooth waveform may be regarded as the sum of a continuous function and a series of step functions



voltage. The characteristics of these circuits have been shown to be identical if in the latter case voltages are considered instead of currents.

We shall consider a recurrent saw-tooth waveform, one cycle of which is shown in Fig. 3, and which may be defined by:—

$$e = \left[\frac{1}{2} - \frac{t}{\tau}\right]_{t=0}^{t=\tau} \quad \ldots\ldots\ldots\ldots\ldots (3)$$

For the purpose of analysis it will be convenient to consider the complete periodic waveform as the sum of two functions, one of which is continuous and may be written

$$e_1 = \left[\frac{1}{2} - \frac{t}{\tau}\right]_{t=0}^{\infty} \quad \ldots\ldots\ldots\ldots\ldots (4)$$

and the other a series of unit steps commencing at times t_0, t_1, t_2, etc., and which may be expressed symbolically as:

$$e_2 = \sum_{n=0}^{\infty} 1 \quad (t - t_n) \ldots\ldots\ldots\ldots\ldots (5)$$

It will be seen from Figs. 4(a), 4(b) and 4(c) that the sum of these two functions is equal to the saw-tooth waveform of Fig. 3, and by the theory of superposition we are entitled to consider the effect of each component separately.

The response of the circuit of Fig. 2(a) to the first component τ given readily by means of Duhamel's superposition theorem (see Appendix 1), from which we obtain:—

$$e_{\alpha t} = \left(\frac{1}{\alpha\tau} - \frac{1}{2}\right) e^{-\alpha t} - \frac{1}{\alpha\tau} \quad \ldots\ldots\ldots\ldots (6)$$

in which α is the reciprocal of the time constant, T. This is simply a negative-going exponential voltage commencing with a value of $\frac{1}{2}$ at t = 0 and falling to a steady-state limit of $-1/\alpha\tau$.

We shall now consider the second component with reference to Fig 4(b).

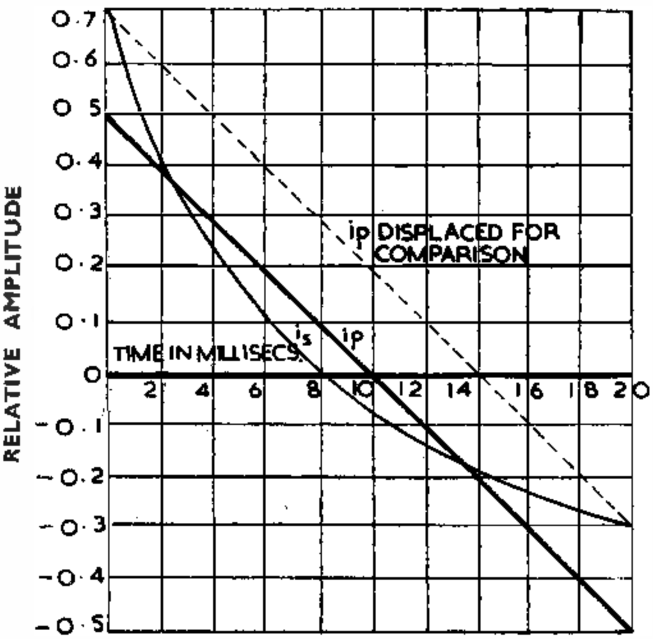


Fig. 5. Showing distortion due to bass loss of 1 db. at 50 c/s.
i_p=Primary Current i_s=Secondary Current

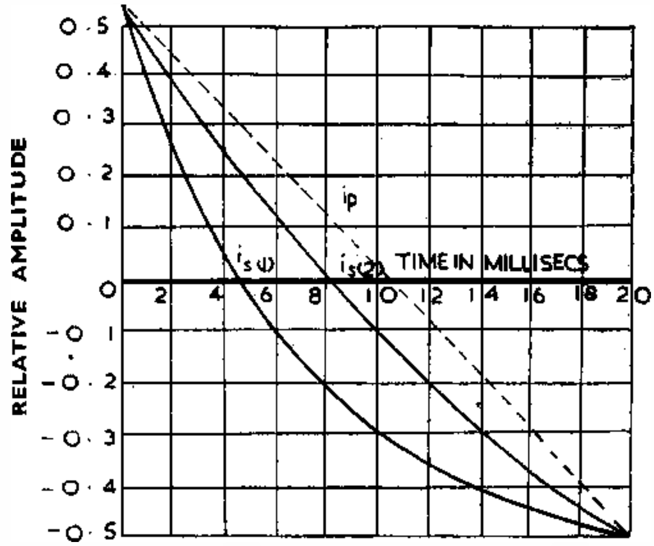


Fig. 6. Comparison between distortions resulting from bass loss of 1 db. and 0.1 db. at 50 c/s.
$i_s(1)$= bass loss of 1 db. (L/R= 7 millisecs.)
$i_s(2)$= bass loss of 0.1 db. (L/R=23 millisecs.

The first step will give rise to a response voltage.

$$e^{-at}\,1$$

in which Heaviside's unit function, 1, is included as a reminder that the voltage is zero for $t < o$. The second step will result in a voltage

$$e^{-a(t-t1)}\,1$$

where again the 1 has been included as a reminder that nothing occurs until $t > t_1$, while for the third step we have

$$e^{-a(t-t2)}\,1$$

and so on. Thus, between the limits of $t = t_n$ and $t = t_n + 1$, the total voltage at any instant will be given by

$$e_{o2} = \left[e^{-at} + e^{-a(t-t1)} + e^{-a(t-t2)} \dots\dots + e^{-a(t-tn)}\right]_{t=tn}^{t=tn+1}$$

$$= \left[e^{-at}(1 + e^{at1} + e^{at2} + \dots\dots e^{atn}\right]_{t=tn}^{t=tn+1} \quad \dots (7)$$

As successive steps occur at equal time intervals of τ we may write t_n as $n\tau$. Also changing the limits we may write

$$\left[e^{-at}\right]_{t=tn}^{t=tn+1} = \left[e^{-a(tn+t)}\right]_{t=o}^{t=\tau}$$

$$= e^{-an\tau}\left[e^{-at}\right]_o^\tau \quad \dots\dots\dots\dots\dots\dots (8)$$

whence

$$e_{o2} = \left[e^{-at}\right]_o^\tau \times e^{-an\tau}\left[1 + e^{a\tau} + e^{2a\tau} \dots\dots + e^{na\tau}\right] = \left[e^{-at}\right]_o^\tau \times$$

$$\left[1 + e^{-a\tau} + e^{-2a\tau} \dots\dots + e^{-na\tau}\right] \dots\dots (9)$$

The sum of the power series to n terms is

$$1 + e^{-a\tau}\left[\frac{1 - e^{-na\tau}}{1 - e^{-a\tau}}\right] \quad \dots\dots\dots\dots\dots (10$$

For the steady state conditions we must put $n = \infty$ and we obtain finally

$$e_{o2} = \left[e^{-at}\right]_o^\tau \times \left[\frac{1}{1 - e^{-a\tau}}\right] \dots\dots (11)$$

Adding the steady state term of (6) we obtain for the complete steady state response for one cycle

$$e_o = \left[e^{-at}\right]_o^\tau \times \left[\frac{1}{1 - e^{-a\tau}}\right] - \frac{1}{a\tau} \dots\dots (12)$$

It is interesting to note that the form of the response is identical to that resulting from unit step over a period equal to one cycle of the saw-tooth, modified by a scale factor which makes the total excursion equal to that of the input saw-tooth, in this case unity. The added constant $-1/a\tau$ is, as would be expected, of such value that the positive and negative areas of the waveform are equal, which may easily be verified by integrating (12) with respect to time.

In the example quoted in the previous section, the values were

$$a = \frac{R_c}{L_s + L_c} = \frac{8}{56 \times 10^{-3}} = 143$$

and $\tau = 1/50$ secs. $= 20 \times 10^{-3}$ secs.

In Fig. 5 one cycle of the steady state response corresponding with these values is shown plotted together with the input waveform for comparison. The degree of distortion resulting from a relatively insignificant bass loss is somewhat surprising.

In Fig. 6 an additional curve is plotted showing the distortion corresponding with a bass loss at 50 c/s. of 0.1 db. In order to achieve this result it would be necessary to increase the secondary inductance of the current transformer from 50

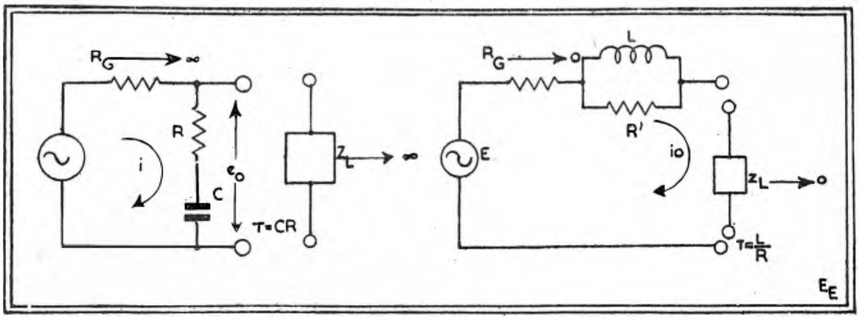


Fig. 7. Two compensating circuits in simplified form

mH to 180 mH, and the primary inductance to 120 H, and even then the departure from linearity is far greater than could be tolerated in any high grade apparatus.

Instead of increasing the size of the transformer, a far more economical procedure is to introduce some form of frequency compensation, and it is proposed to discuss suitable methods of compensation in the following section.

Methods of Compensation

In order to obtain perfect compensation for the frequency characteristic of the circuit of Fig. 2(a), it would be necessary to introduce a network into some part of the circuit which possessed precisely the inverse characteristic. Such a network would, therefore, have to provide infinite gain at zero frequency, and with increasing frequency, a characteristic falling asymptotically to a finite gain, which would remain substantially constant over the greater part of the working band. While such an ideal network is unrealisable in practice, we may approach the ideal by means of either of the two fundamenal circuits shown in Figs. 7(a) and 7(b). The circuit of Fig. 7(a) will provide a compensated voltage across an infinite impedance load, while that of Fig. 7(b) will provide a compensated current through a zero impedance load, and if

$$RC = L/R' = T$$

the two circuits will have identical characteristics. Both circuits operate on a similar principle, that is, by introducing, in the limit, infinite attenuation which is reduced to zero as the frequency is reduced to zero. The gain characteristic of either circuit for the limiting condition may be written

$$\sqrt{1 + \frac{1}{\omega^2 T^2}} \quad \big/ -\tan^{-1}\frac{1}{\omega T}$$

which, were it realisable in practice, would be the precise inverse of the characteristic expressed in (2) for which we desire to compensate. The degree of compensation we may achieve, therefore, depends entirely upon how much attenuation we are prepared to introduce.

In practice, if it is desired to compensate for a loss of, say, 1.0 db. at the lowest limit of the working frequency range, it will be found sufficient to introduce a limiting attenuation for the higher frequencies of the order of 20 db., and the characteristic over the working band will be substantially the same as the ideal characteristic.

To introduce a suitable compensating circuit into either the transformer primary or secondary circuits is not only difficult but very uneconomical, since additional current amplification would be required in order to obtain sufficient scanning current. A simpler procedure is to introduce the compensation before the distortion has occurred. When a separate current amplifying stage is used, the available input voltage supplied from the saw-tooth generator is usually large and attenuation must necessarily be introduced in some form in order to prevent overloading the current amplifying valve. In this case the compensating circuit may be included as part of the attenuator without introducing any additional loss.

Figs. 8(a) and 8(b) show the two alternative practical forms of compensation. The circuit of Fig. 8(b) has an important advantage over that of Fig. 8(a) in that over the working band of frequencies the cathode is virtually not decoupled and, therefore, provides current feedback which not only reduces the distortion introduced by the valve, but considerably increases the valve impedance. This is, of course, true only if the degree of compensation at the lowest working frequency is small compared with the reduction in gain introduced by the cathode resistance. Adjustment of linearity is achieved simply by varying the resistance, R'.

To take a practical example, suppose we wish to compensate for the distortion introduced by the circuit of Fig. 1(b), with values as previously quoted. Assuming a mutual conductance of 6 mA/V and a cathode resistance of 3.17 KΩ, the gain at high frequencies will be reduced to one-twentieth, i.e., a reduction of 26 db.

The required size of compensating inductance (Fig. 8(b) will be given by

$$\frac{L}{R'} = \frac{L_s + L_c}{R_c} = 7 \times 10^{-3}$$

whence

$$L = 7 \times 3.17 \text{ H} = 22 \text{ H approx.}$$

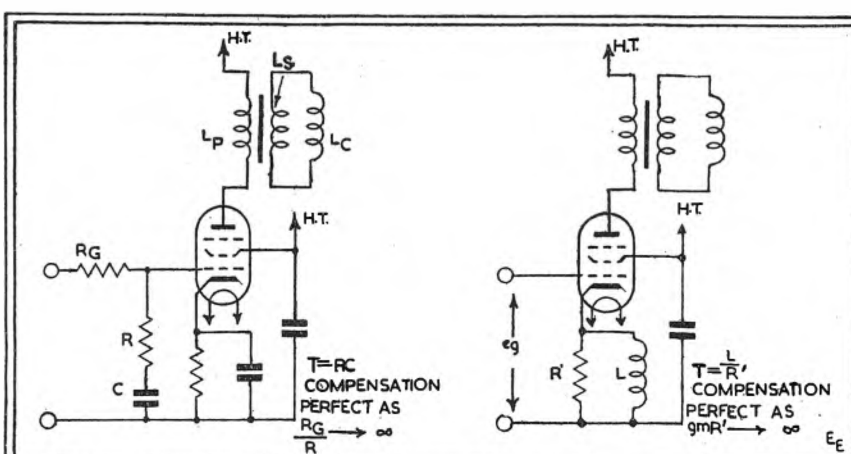


Fig. 8. Practical forms of the compensating circuits of Fig. 7.

It is shown in Appendix 2 that the overall linearity obtained with this compensation will be equivalent to that which would have been obtained had the primary inductance of the transformer been increased from 34 H to 665 H. The maximum departure from linearity expressed as a percentage of the total scan will be of the order of 2 per cent, and such small distortion has been found to be imperceptible by subjective tests, provided it is distributed smoothly over the whole scan, as is true in this case.

Conclusions

The most likely cause of distortion in an electromagnetic frame scanning system is bass loss in the scan of 0.1 db. at the fundamental component due to such a transformer is sufficient to cause serious distortion, and for laboratory or transmitting apparatus considerably less attenuation must be achieved. It is uneconomical to increase the size of the transformer beyond a certain limit and linearity is best obtained by means of bass compensation applied before current amplification.

A suitable limit to the size of transformers may be decided by arranging that the sum of the secondary and scanning coil reactances is twice the sum of their total D.C. resistance at the fundamental saw-tooth frequency. The attenuation introduced by such a characteristic may be corrected by a relatively simple compensating circuit, so as to provide a scanning waveform that is fundamentally linear.

Acknowledgment

The author wishes to thank the Chief Engineer of the British Broadcasting Corporation for permission to publish this paper, and Mr. H. L. Kirke, C.B.E., M.I.E.E., Head of the B.B.C. Research Department, under whose direction the work was carried out.

Appendix 1

The Duhamel superposition theorem states that:

$$i(t) = e(o)\,A(t) + \int_o^t A(t-\lambda)\,e^1(\lambda)\,d\lambda$$

in which:—

$A(t)$ is the indicial admittance of a linear network.

$e(t)$ is the time function of an applied voltage.

$E^1(\lambda)$ is the first derivative of $e(t)$ with λ substituted for t.

$e(o)$ is the value of $e(t)$ at $t=o$.

Substituting $\frac{1}{R}e^{-at}$ for $A(t)$

and $-\left[\frac{1}{2}+\frac{t}{\tau}\right]$ for $e(t)$

we have:—

$$i(t) = -\frac{1}{2R}e^{-at} - \frac{1}{R\tau}e^{-at}\int_o^t e^{a\lambda}\,d\lambda \quad \ldots (1)$$

or $e_o(t) = i(t)\,R =$

$$-\frac{1}{2}e^{-at} - \frac{1}{\tau}e^{-at}\int_o^t e^{a\lambda}\,d\lambda \quad \ldots (2)$$

$$= \left[\frac{1}{a\tau}-\frac{1}{2}\right]e^{-at} - \frac{1}{a\tau} \quad \ldots\ldots\ldots\ldots (3)$$

Appendix 2

Referring to Fig. 8(b), if a perfect pentode is assumed, the valve current may be expressed as

$$i_v = e_K g_m \frac{1}{1+g_m Z} \quad \ldots\ldots\ldots\ldots (1)$$

where g_m = valve mutual conductance

$$Z = \frac{j\omega L R^1}{R^1 + j\omega L}$$

We are concerned with the frequency characteristic only, which may be written:—

$$f(\omega) = \frac{R^1 + j\omega L}{R^1 + j\omega L + j\omega L R^1 g_m} \quad \ldots\ldots (2)$$

Substituting T for L/R and rationalising gives:

$$f(\omega) = \frac{1+\omega^2T^2(1+R^1g_m) - j\omega T R^1 g_m}{1+[\omega T(1+Rg_m)]^2} \quad \ldots\ldots\ldots\ldots (3)$$

In polar co-ordinates we have

$$f(\omega) = \sqrt{\frac{[1+\omega^2T^2(1+R^1g_m)]^2+\omega^2T^2R^2g_m^2}{1+[\omega T(1+R^1g_m)]^2}}$$

$$\angle -\tan^{-1}\frac{\omega T R^1 g_m}{1+\omega^2T^2(1+R^1g_m)} \quad \ldots\ldots\ldots\ldots (4)$$

If the products R^1g_m and ωT are both large compared with unity we have in the limit

$$f(\omega) = \frac{1}{R^1g_m}\sqrt{1+\frac{1}{\omega^2T^2}}\;\angle -\tan^{-1}\frac{1}{\omega T} \quad \ldots\ldots\ldots\ldots (5)$$

By substituting in (4) the values quoted in the section headed "Steady-state Response of Network to Saw-tooth Waveform" which give

$$T = 7 \times 10^{-3} \text{ secs.}$$
$$R^1g_m = 19$$

we obtain for the normalised response of the correction circuit at 50 c/s.

$$f(\omega) = 1.0981 \;\angle -23.2^\circ$$

The response due to the transformer according to Equation (2) of the text is

$$f(\omega)\frac{1}{1.0984}\;\angle\, 24.45^\circ$$

whence the overall response is given by the product

$$F(\omega) = f(\omega).f^1(\omega) = \frac{1.0981}{1.0989}\;\angle\, 24.45^\circ - 23.2^\circ$$

$$= .99973 \;\angle\, 1.25^\circ$$

This response is approximately equivalent to that which would be obtained with a bass-loss circuit as in Fig. 7 with a time constant of 0.137 secs. The actual response of such a circuit at 50 c/s. would be

$$.99973 \;\angle\, 1.3^\circ$$

and it is justifiable, therefore, to consider the latter as an equivalent circuit over the frequency range concerned. In order to obtain a similar result without compensation, it would be necessary to increase the transformer primary inductance by a factor equal to the ratio of the time constants, i.e., to a value of:

$$34 \times \frac{0.137}{7 \times 10^{-3}}\ \text{H} = 665\ \text{H}$$

assuming that the scanning coil resistance was still predominant, and this would be quite impracticable.

If the effective value of time constant is substituted in Equation 12, it is found that the overall steady state response of the compensating circuit to a saw-tooth waveform is such that the maximum departure from linearity does not exceed 2 per cent.

The Design and Limitations

By E. J. Harris, Ph.D., D.I.C.* and

PROGRESS in the use of valves for the linear amplification of small direct currents has resulted from the use of more and more stable supplies, together with various compensating devices to minimise the effects of changes in the external conditions. The degree of approach to the theoretical limit of sensitivity set by the thermal agitation noise of the input resistance depends upon the nature of the response which the amplifier is required to provide, and upon the impedance of the source. Before discussing valve amplifiers it is important to consider alternative arrangements, which may include valves used as impedance changing stages. This is because in many cases additional noise arises in amplifying valves, which makes the performance of a valve amplifier inherently inferior to that of a non-thermionic amplifier. It is possible to calculate the intensity of part of the noise introduced by the valves, but unfortunately, one of the important noise sources at low frequency, commonly known as "flicker noise", described by Johnson[1] and the subject of theoretical treatment by Schottky[2] and Macfarlane,[3] has not so far been susceptible to calculation. One of the writers of this article has been concerned with a number of measurements of "flicker noise" which will accordingly receive a more extensive treatment than is usual. One of the most successful systems for magnifying small direct currents is that developed by A. V. Hill,[4] who passes the current through a specially designed galvanometer having a quick response and high mechanical stability, and magnifies the deflexion optically by means of a differential photo-cell, followed by conventional valve amplification. The residual noise of the arrangement is only 50 per cent greater than the thermal noise of the resistance of the galvanometer coil. This arrangement is limited by the damping of the galvanometer to use below about 100 c/s. as long as maximum sensitivity is required, and as the resistance of the coil cannot be made much more than 1,000 ohms it becomes very inefficient when the current source has a high resistance. The latter difficulty can be overcome by interposition of a cathode follower between the source and the galvanometer, so as to match the source to the coil resistance. This expedient does not appear to reduce the performance appreciably. Apparently, the valve does not add more than the shot noise, which is made up of the sum of the grid current shot noise, and the anode current shot noise, the former being much the more important when a high impedance source is used. It is important then to keep the grid current at a minimum. There is no evidence that the cathode follower generates flicker noise; a cathode follower stage preceding a direct coupled amplifier certainly does not impair performance.

It is not proposed to discuss magnetic amplifiers here, because they have been treated recently.[5] It is relevant to point out that the limiting sensitivity of such apparatus is determined by the effects of thermal agitation on the magnetic properties of the core, in addition to the resistance noise of the pick-up coil, which may be negligible. It is possible that a magnetic system could be developed, having a high sensitivity, approaching the theoretical, but no claims of this order appear to have been made so far. It is clear that the magnetic system will be limited in its frequency response, on account of the inductance of the coils.

When it is desired to use a system capable of magnifying an E.M.F. derived from a high impedance, and having a frequency response up to several Kc/s., some form of valve circuit must be used. The direct coupled amplifier is the most obvious example, but in the past so much difficulty has been experienced in obtaining a reasonable stability that the methods of interrupting the unknown E.M.F. at intervals, or reversing it, have been applied in order to obtain periodic readings for the zero. The frequency and duration of interruption obviously must be chosen so that the waveforms to be observed are not interfered with. In biological applications, where a waveform is to be observed, rather than a steady potential, it is not sufficient to couple the source and interrupter to a capacitor-coupled amplifier, because this would introduce a time constant which would distort the waveform following each interruption or reversal. The arrangement is useful, however, when a time integral of the input is required, and it has been applied, for example, to amplification of thermo-electric E.M.F.'s. When the interruption system is applied to a direct coupled amplifier the result is not as simple as may at first sight be imagined. If the limitation to the stability of the apparatus arises in the first valve, and not in the source resistance, the "zero" read with input shorted (or disconnected) is as much subject to fluctuation as is the reading with source present, and there is no justification for referring a reading on the output record at, say, A (Fig. 1), to either of the zeros, z or Z in preference. It is only when the zero is subject to a continuous change which can be shown to be non-random that it becomes justifiable to join up the instantaneous zero readings to form a base line.

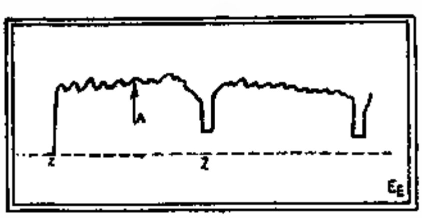


Fig. 1. Specimen record of an amplifier output

Another expedient that is often met in the search to avoid use of direct coupled circuits is that of modulating a carrier with the unknown E.M.F. A simple and effective method is to impress the E.M.F. on a small capacitor whose capacitance is varied periodically. This method is said to give good results if care is taken to avoid contact potentials between parts of the capacitor, but it is limited in scope by the frequency at which the capacitance can be varied, which sets an upper frequency limit, and also by the time constant of the capacitor source (or grid leak) combination. Most other modulation methods use noisy components, i.e., valves or semi-conductor rectifiers. If the signal

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# of D.C. Amplifiers —Part I

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to noise great ratio is to be kept as high as possible the signal applied to the modulator must be of great amplitude compared with the noise introduced by the modulator, so that the most difficult step in amplifying the signal must be taken before the carrier system is applied. As there is no special difficulty in amplifying the signal further, the use of a carrier seems unnecessary, especially as it is liable to lead to non-linear response unless careful attention is paid to the design.

We can now outline the various sources of noise and disturbance which have to be taken into consideration when trying to assess the minimum signal which can be observed with the apparatus.

Sources of Random Fluctuations

(1) Shot noise of the anode current, and partition noise.
(2) Shot noise of the grid current.
(3) Flicker noise.
(4) Leakage and positive ion currents.

Sources of Disturbance Related to External Parameters

(5) Variation of anode supply potential.

(6) Variation of heater temperature.

(7) Variation of temperature of components.

(8) Microphony.

(9) Pick-up of electro-static and electro-magnetic fields.

A useful summary of sources of electrical noise, with many references, has been given by D. K. Macdonald.[6] All the disturbances listed above are most conveniently expressed in terms of the equivalent signal which would have to be applied to the input in order to give the same power output as the disturbance. Difficulties arise when discussing the noise produced in direct coupled amplifiers, for many, if not all, authors have arbitrarily separated the fluctuations at the output into "noise" and "drift." The latter is attributed to the variation of some parameter, which, at least in theory, could be controlled. The plot of the output of the apparatus against time, instead of fluctuating about zero, departs at a more or less constant rate from the original zero, and this is called "drift." The distinction is really quite arbitrary unless simultaneous observation of the source of the drift is made and shown to be correlated with it. In the event of the correlation existing it should be possible to eliminate the drift by feeding the variable factor into an appropriate point of the amplifier. When there is no correlation the "drift" must be regarded as noise of low frequency, and must be included as such when quoting the performance of the apparatus. The crucial figure in assessing the performance is the mean square value of the departure from the original zero, taken over a long period. It can be argued that only by observation for an indefinite time can the true mean be found, but in practice an arbitrary time, long in comparison with the proposed duration of the experiments for which the apparatus is to be used, has to be chosen. On account of the nature of the output the only satisfactory way to obtain a mean value of the output is to employ an integrating watt-meter, such as an ice calorimeter containing a resistance coil.

The separate noise components, 1-4 above, may be estimated.

(1) The shot noise set up by the anode current can be calculated if a value for the space charge screening factor G be assumed, (it is 0.01 to 0.1 under normal conditions, but may be higher if the valve is underrun). With an anode load of magnitude $|Z|$ and valve amplification factor μ the input equivalent to the shot noise of anode current I_a, integrated in the band 0-f c/s. has mean square amplitude: (e = electron charge)

$$\overline{\Delta V_a^2} = 2eI_a\int_0^f G.\frac{(R_a + |Z|)^2}{\mu^2}\ df$$

When a cathode follower stage is in question, with output impedance of magnitude $|Z_o|$ the equivalent input is $\overline{\Delta V_a^2} = 2eI_a^2\int_0^f G.|Z_o|^2 df$

When a multi-electrode valve is used the shot noise is increased by the variable partition of the current between the electrodes and for the amplifier the noise becomes equivalent to

$$2eI\frac{(GI_a + I_s)}{I_a + I_s}\int_0^f |Z_o|^2\ df$$

where I_s is the screen current. (Bakker)[7].

(2) The shot noise of the grid current can be more serious than that due to the anode current, when the grid-cathode impedance is high and the system responds to low frequencies. There is no space charge screening. The mean square input E.M.F. equivalent to the grid current noise is

$$\overline{\Delta V_g^2} = 2eI_g\int_0^f |Z_g|^2 df$$

where $|Z_g|$ is the magnitude of the grid cathode impedance. In electrometer amplifiers the effective bandwidth is fixed by the time constant of the grid circuit itself. Milatz and Keller,[8] Keller and Vesseur[9] and Keller[10] have published results of noise measurements made under floating grid conditions, which show that the total noise intensity corresponds to the sum of the anode current shot noise, and a frequency dependent factor calculable by taking the total grid current as twice the measured negative grid current. It does not follow that there was no flicker noise, for the latter has been found to have a value of 50-100 μV (Harris and Bishop)[11] and is concentrated below the lowest frequency at which the above measurements were made.

(3) The authors of this summary consider that flicker noise originates in fluctuations of the contact potential between the material of the heated cathode surface and the grid wires. The cathode coating is a semi-conductor, and the number of current carriers per unit volume is subject to statistical fluctuations, giving rise also to the current noise found in semi-conducting resistors,[12] and at the same time setting up a varying contact potential, which modulates the anode current. It is on account of the variation in contact potential with cathode temperature that the anode current is so dependent upon the cathode tem-

perature. It is important to observe that in the cathode follower there is an automatic compensation which removes part of the flicker effect, because more current will flow to those parts of the cathode which are least positive with respect to the grid, and the product of the local extra anode current and the local cathode coating resistance represents an E.M.F. which is in opposition to the flicker E.M.F. It is unlikely that any improvement in noise performance will acrue by application of the Miller[13] circuit, designed to reduce dependence of anode current on cathode temperature, for the flicker noise arises locally and will not be subject to cancellation by use of two cathodes, or a common cathode.

The intensity (mean square amplitude) of the flicker noise varies inversely as frequency between 10-1,000 c/s., above 1,000 c/s. it is negligible.[12] Below some low frequency its intensity is supposed to be uniform.

For most purposes we may take the integral between zero and 1 c/s. as 2,500 $(\mu V)^2$ noting that it is independent of the grid circuit impedance, but this value is subject to revision in the light of experiment. For the range 1-F cycles the integral is

$$\int_1^F \overline{\triangle V_f^2}\, df = 0.016\ e\ n\ F'\ \mu V^2$$

for the best valves. These formulæ express experimental results, but make use of the fact that all theoretical treatments agree that at some low frequency the flicker noise intensity does become constant, for it is related to frequency as $\frac{1}{1+\omega^2\tau^2}$ in which τ is some characteristic time of disturbance, so that when $(\omega\tau)^2$ is small compared to unity the intensity ceases to be frequency dependent.

(3) The other sources of noise originating in the valve, attributable to leakage current, and to the presence of positive ions depends on the processes of manufacture, and can be minimised by selection.

Before passing on to the avoidable disturbances, let us calculate the noise to be expected in two valve applications, one being the cathode follower unit referred to in the section concerning the use of the galvanometer photo-cell amplifier, and the other a valve amplifier such as may be used in electro-biological work, with a high amplification, and an upper frequency limit of some 5,000 c/s. For the electrometer cathode follower the following constants will be used: $Rg = 10^{12}$ ohm, grid capacitance $C = 5.10^{-12}$ farad, anode current 0.01 mA, grid current 10^{-12} amp., screening factor of space charge 0.1. The thermal noise of the resistive part of the grid circuit impedance, taking the upper frequency limit as that determined by the grid circuit itself is:

$$\overline{\triangle V_g^2} = 4\,kT\int_0^\infty (\text{r.p. of } z)df = \frac{kT}{C}$$

which is 8×10^{-10} $(\mu V)^2$, i.e., 28 μV r.m.s. The shot noise of the anode current developed in the output impedance, say, 5,000 ohms for a cathode follower with cathode load, is:

$$\overline{\triangle V_{sa}^2} = 2eI_aG\ R_o^2 \text{ volts}^2 \text{ per cycle.}$$

So for a 1 cycle bandwidth, set by the circuit following:

$$\overline{\triangle V_{sa}^2} = 7.5 \times 10^{-11} \text{ volt}^2.$$

The shot noise of the grid current, set by the bandwidth of the grid circuit is

$$\overline{\triangle V_{sg}^2} = 2eI\int_0^\infty |Z_s|^2 df = \frac{eI_gR_g}{2C_g}$$

and for our conditions $\overline{\triangle V_{sg}^2} = 1.6 \times 10^{-8}$ volt, so the shot noise of the grid current is the principal limitation and is 1.3×10^{-4} V = 130 μV.

For the second example, in which the valve is used as an amplifier, the following constants have been applied: amp. factor 10; R_a 50 K ohm; R 100 K ohm; space charge screening factor, 0.1; I_a 10^{-3} amp.; G 5×10^{-12} farad; I_g 10^{-9} amp. and R_g 6×10^6 ohm (making the response half in power at about 5,000 c/s.).

The thermal noise of the grid resistance $\overline{\triangle V_g^2} = \frac{kT}{C} = 8 \times 10^{-10}$ volts², i.e., 28 μV as before.

The shot noise of the anode current, expressed as equivalent input is

$$\overline{\triangle V_{sa}^2} = \frac{2eIG}{\mu^2}\int_0^f (R_a + R)^2 df$$
$$= 3.6 \times 10^{-11} \text{ volts}^2$$

assuming a square 5 Kc/s. bandwidth in the anode circuit. The shot noise of the grid current is

$$\overline{\triangle V_{sg}^2} = \frac{eI_gR_g}{2C} = 9.6 \times 10^{-11} \text{ volts}^2.$$

The flicker noise will be as much as 50 μV, so it is now the factor which will limit the performance.

In addition to the specific effect of cathode temperature, the temperature coefficients of the various components making up the apparatus are of importance.

General Design Features

A wide variety of single stage circuits suitable for coupling a very high impedance source to a sensitive galvanometer have been developed. Most of the methods of compensation for changes in the external conditions which have been applied in such circuits have not, however, found application in multi-stage amplifiers suitable for working into a cathode-ray tube. The usefulness of an amplifier is so much enhanced by the possibility of displaying the output that a general purpose apparatus must necessarily include this feature. The design of direct coupled amplifiers is so much simplified by the use of a balanced system that the remarks following will be almost completely concerned with balanced stages. It is a simple matter to convert an unbalanced input into a balanced signal by the use of a stage having a high impedance common cathode load. In addition to a discussion of general problems concerning circuit design, some remarks on the power supplies, valves and resistors may be of assistance. The following references contain descriptions of direct coupled amplifiers of the type considered in this summary.[13,26,28,31-38]

Stability of H.T. and L.T. Supplies

When we are evaluating the necessary stability of the supplies which determine the effect of numbers 4-6 in the list of sources of disturbance it will be necessary to ensure that the additional disturbance arising from the supplies is less than, say, 50 μV, for the cathode follower unit, and less than 20 μV for the amplifier. The necessary stabilities in the supplies as far as short term fluctuations (i.e., of periods shorter than 10 secs.) is concerned, is relatively easy to achieve, and if alternative capacitor coupling is provided in the amplifier design the effect of supply voltage fluctuations should be reduced to less than about 2 μV equivalent signal at the input. Long term stability in the power supplies is somewhat harder to achieve, but the standard required is some ten times less stringent.

It has been found quite practicable to obtain the necessary supplies from the A.C. mains. In order to give some idea of the stability required of the power supplies, some figures are given of the respective signals equivalent to a 1 volt change of the positive and negative supplies to an amplifier. It will be appreciated, of course, that these figures are relevant to a particular design and do not necessarily have any general application. The figures given are for an actual 1 volt change in H.T. and thus the signals equivalent to a smaller change (i.e., 7 mV and 1 mV) may be less than quoted as there is less likelihood of non-linearity.

Supply	Voltage change		Equivalent signal
450 V positive	1V	≡	300 μV
	7 mV	≡	2 μV
105 V negative	1V	≡	1.7 mV
	1 mV	≡	2 μV

As the supply to later stages need not be so stable it is often convenient to apply special stabilisation to the potential required for the first stage, the current requirement of which is very small.

It has been found that the effect of changes in heater voltage for a balanced stage using 6J6 valves is such that a fluctuation of 13 mV is approximately equivalent to a 1 mV change at the input. This means that the L.T. supply must be stable to about 0.65 mV if the residual instability caused by it is not to exceed 50 μV. This can only be achieved by the use of lead accumulators of large storage capacity and then only over a relatively limited time. The requirements can be eased by the use of the compensating arrangements discussed later. As a sudden change in heater voltage takes several seconds to effect a change in anode current the requirements for the stability of the L.T. supply can be greatly lessened when alternative capacitor coupling is used in the amplifier.

The Power Supplies

Following what has been said, it will be realised that the practical performance of any design of direct coupled amplifier is ultimately decided by the stability of the power supplies. Those circuit arrangements which provide a self-stabilised system will, from equally good supplies, give the best performance and are to be preferred accordingly. Amplifiers with few stages can be operated from batteries, but in this event it is still necessary to pay attention to the temperature constancy at which the batteries are maintained. It is quite easy to design a high-tension supply which is stable to a millivolt provided the reference potential is stable and the shunt valve is heated from a stable source. If a gas discharge tube is used as a reference source it should be run at a low current, and resistance capacitance smoothing should be applied between it and the control valve. A battery has a higher temperature coefficient than the discharge tube and if used as a reference source requires temperature control to a fraction of a degree in order to maintain its E.M.F. constant to a millivolt. The temperature coefficient of a dry cell[14] is +0.0016 volt per degree per cell in the range 15° C. to 20° C. Thus a 120 V battery will have a temperature coefficient of the order of +100 mV per ° C. Gas discharge tubes have a temperature coefficient of about 5 mV to 20 mV per °C.[15]; the value depends upon the type of tube, as does the sign of the coefficient. Special cells of low temperature coefficient can be obtained, but for a high tension supply the reference potential should be at least 20 V or 1/10 of the H.T. level, for otherwise the instability arising in the shunt amplifier valve on account of changes of contact potential in the valve become noticeable.

The provision of a very stable source of heater current is difficult. It is most convenient to run a number of low current valves (i.e., up to 150 mA) in series across an electronically stabilised supply and this can be applied to most of the stages of the amplifier and to the shunt valve of the H.T. supply. Unfortunately when the 6J6 is used in the initial stages its higher heater current (0.45 A) precludes the use of this method and a 6 volt supply must be made. This can be obtained from an accumulator which is continuously charged at a rate equal to its discharge rate. Even the charging current requires stabilising for the acumulator E.M.F. is dependent upon the net charge rate.

Valves

There are several factors which bear on the choice of valves for direct coupled amplifiers. It is only the valve used in the input stage, commonly a cathode follower, or a suppressor controlled multi-grid valve used as an impedance transformer, which must have a particularly low grid current. Suitable valve types and operating conditions are described later. For the early amplifying stages the best performance is provided by those valves which remain as balanced as possible when the heater current varies, and for this it is best that twin valves, with a cathode shared between the two assemblies, should be used. The 6J6 (ECC91) is an example, and another possible type is the UA55, which is a twin tetrode. The latter valve has the advantage that the D.C. balance of a stage can be obtained by varying the screen potentials, and it has a lower input capacitance. It is, however, more microphonic than the 6J6. It is also difficult to make reliable contact with the pins, a point which always requires attention. The high gain obtainable from the UA 55 allows an ampifier with a gain of about 75 db. to be made, using only two stages. The 12SC7 is frequently specified for use in direct coupled circuits. It is a twin triode, but the electrode arrangement is such that the two assemblies are one vertically above the other, and there is no reason why the performance should be superior to that when a valve with independent assemblies in one envelope is employed.

The following table of the average characteristics of 10 6J6 valves under operating conditions suitable for the early stages of amplification may be found useful.

Vg	Va	gm MA/V	Ra
−0.9V	30V	1.7	19000
,,	42V	2.5	13000
,,	54V	3.3	10300
−0.5V	30V	2.5	13000
,,	42V	3.0	11000
,,	54V	3.5	9400

(*To be continued*)

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A view of the stands in the centre of the hall, showing (foreground) a metal detector equipment and other Cintel exhibits.

The Fourth Manchester Electronics Exhibition

The Fourth Annual Exhibition of Electronic Devices organised by the N.W. Branch of the Institute of Electronics was held at the College of Technology on July 19-21, 1949. A brief account of some of the exhibits is given below.

Evaporating and Sputtering Plant

EDWARDS Model E.3 Evaporating Plant. This is a completely self-contained unit comprising: Work chamber, exhausting system, vacuum gauges, discharge cleaning equipment, evaporating sources and controls, safety controls, auxiliary apparatus to suit particular application, *e.g.*, work holder, radiant heaters, moving work-table, etc., operated through vacuum seals.

The stainless steel or Pyrex glass work chamber is 46 cms. diameter by 60/75 cms. high and rests on a plated steel base. The pumping system consists of high speed diffusion pumps (500-1,000 litres/second oil of mercury operated) backed by a 10 litres/second mechanical pump.

Robust and simply operated electrical vacuum gauges are fitted—a Philips type reading to 10^{-5} m/m Hg. for the work chamber and a Pirani type for checking pump performance and vacuum tightness.

A typical time schedule for metallising glass mirrors in the Edwards E.3 Unit would be: Evacuation to 10^{-5} m/m. Hg., 5 minutes; discharge cleaning, 10 minutes; evaporation, 1 minute.

Edwards S.5 Sputtering Plant. A self-contained unit for research or industrial sputtering of small objects. The glass work chamber is 24 cms. dia., and 17 cms. high. A 20 cms. dia. cathode (plated with the sputtering metal) is mounted on insulators and makes electrical contact with a chamber electrode when the cover is closed. The work is held under the cathode on a glass table having adjustable legs. A typical time schedule to produce opaque gold coatings is 2/3 minutes pumping to reach sputtering pressure, 15 minutes sputtering.

W. Edwards & Co., Kangley Bridge Road, S.E.26.

The Ferranti Servodyne

AN electronic equipment, intended for use as a controlling medium for processes which require positive and precise control. The equipment is, in effect, a self-balancing servo-mechanism which will observe an error in any of several forms, and provide the necessary correction via the motor unit which supplies the mechanical output power of the mechanism. The input to the equipment depends upon the particular application it may, for example, be the small power derived from a photo-electric cell, the unbalanced power of a bridge circuit, the output of a piezo-electric crystal, etc. The equipment has been successfully applied as a web-guider on slitting machines in the paper converting industry, and a cloth-guider on pin-stenters in the textile finishing industry.

Ferranti, Ltd., Hollinwood, Lancs.

Metro-Vick. Miniature Oscilloscope

POSSESSING all the facilities of a full-size oscilloscope, this instrument measures only 4½ in. × 5¾ in. × 11½ in. The tube (2¾ in. screen), has a green fluorescence, negligible afterglow, and a sensitivity of the order of 40 V. per cm. Sharp focusing of the beam and the complete absence of astigmatism are notable features. One X-plate is brought out, permitting horizontal deflexion to be effected by means of an exterior source. The linear push-pull time base has a repetition frequency range of 10 c/s. to 50 Kc/s., and provision for synchronisation. Also included is a single-stage amplifier with high frequency compensation, switched to select a gain of either 35 db. flat within ± 1 db. from 50 c/s. to 150 Kc/s. or 20 db. + 2 db. from 50 c/s. to 3.5 Mc/s.

Metropolitan-Vickers, Ltd. Trafford Park, Manchester.

Educational Valve Test Panel

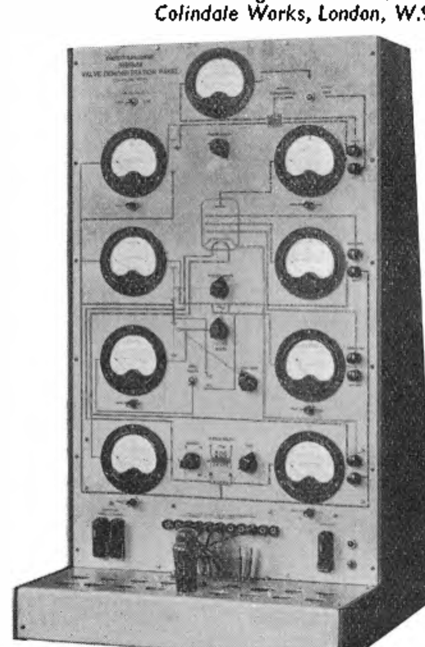
THIS instrument (illustrated below) is designed for laboratory use to facilitate the plotting of valve characteristics under properly regulated conditions with D.C. potentials for the electrodes.

It comprises an A.C. power pack, with voltage controllers, voltmeters and ammeters in each circuit. The valve-holder board contains fifteen valve holders, which can be wired to the power pack by means of a plug and socket system with retracting connectors.

The employment of a single equipment, with built-in meters and power supplies saves times in wiring up and so enables more experiments to be carried out in a given time. It also reduces the risk of damage to valves and meters.

The circuit diagram, which is shown on the front of the panel, ensures that the student understands the function of the equipment.

Everett Edgcumbe & Co., Ltd., Colindale Works, London, W.9.



The Plessey-Electron Microscope

THIS is a vertical two-stage electromagnetic instrument operating at 50 KV and giving magnifications of 100 diameters and 400-20,000 diameters. The resolving power is better than 50 A.U., the actual resolution achieved varying with the type of specimen used in the customary manner.

The seven major units forming the column are bolted together with machined metal flanges in contact, to provide a rigid assembly and minimise the effects of vibration. Neoprene gaskets compressed between special contours machined in the metal flanges are used for vacuum sealing.

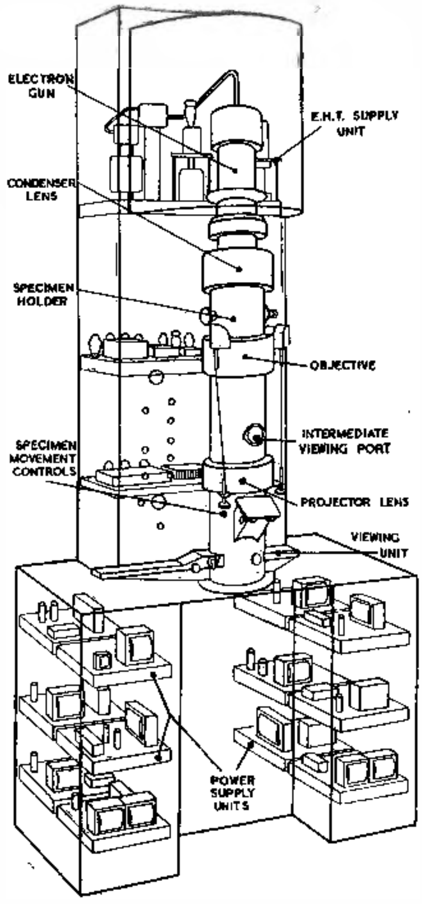
The specimen under examination may be supported on a 200-mesh grid formed by etching holes approximately .003 in. square in copper foil, although other types of support are also commercially available. No additional support is required for certain powders and smokes, but other specimens may require an electron-transparent film between 100 and 1,000 A.U. thick. Cellulose nitrate is widely used for this purpose.

The specimen grid is attached to a cylindrical insert which is placed in one of two positions in the specimen holder, depending on the range of magnification required. The holder is introduced to the microscope column through an air lock fitted with two valves.

By adoption of the "unit" principle throughout the maintenance of the instrument has been much simplified. Both the vacuum system and the electrical system can be easily separated into a number of small units for individual testing.

Height	8 feet 5 inches
Floor area	3 feet 6 inches wide
	2 feet 4 inches deep
Back clearance required	3 feet 6 inches wide
	1 foot 6 inches deep

The Plessey Company, Ilford, Essex.



This skeleton sketch shows the component parts of the Plessey Electronic Microscope.

New "Cintel" Equipment

Mutual and Self Inductance Bridge

THIS bridge has a very wide range and high accuracy. Mutual and self inductance can be measured without complicated circuit changes, and the phase sensitive detector used ensures that the inductance and resistance controls are largely independent. Special circuit arrangements are made to eliminate the effect of the series resistance of leads and contact resistance of clips so that full use may be made of the low resistance ranges. Particular applications of the instrument include the measurement of the inductance of tuning coils, loading coils, transmission lines and cables and stray inductances in circuit wiring. Since it is sensitive down to 0.001 microhenries it is specially useful for the latter purpose in television and radar circuits.

Range of Inductance Measurement: 0.03 microhenries to 30 millihenries: first scale reading 0.01 microhenries.
Number of Ranges: 12.
Range of Resistance Measurement: 0.003 ohms to 3,000 ohms.
Accuracy: $\pm$ 1 per cent on all ranges.
Frequency: 1,592 c/s.
Power Supply: 100-250 V, 40-60 c/s.

Microsecond Counter Chronometer

This instrument is designed primarily for the accurate measurement of short intervals of time. It is used for the very accurate measurement of projectile velocities, landing speeds of aircraft, etc., and also serves as a frequency meter, pulse generator and as a simple counter. It consists essentially of an oscillator, whose frequency is accurately controlled, an electronic gate and six electronic decade counters.

Initially the oscillator is operating, the gate is shut and all the decades are set to zero. On receipt of a start pulse the gate opens and the decades begin to count individual cycles of the oscillator. When a stop pulse is received the gate shuts leaving indication on the panel of the number of cycles executed by the oscillator in the time elapsed between the two pulses.

Standard Frequency: 1 Mc/s.
Range: 1 μs. to 1 second in steps of 1 μs.
Accuracy: $\pm$ 0.01 per cent. $\pm$ 1 μs.
Start and Stop Pulses: 10-40 V $\div$.

Other frequencies available are 100 Kc/s., and 100 c/s., giving maximum timing ranges of 10 seconds, and 10,000 seconds. 1,000 c/s. available to special order.

Cinema-Television, Ltd., Sydenham, S.E.26.

Other Noteworthy Items in Brief

"Caslam." A soft magnetic core material in three grades, with a finely laminated structure for use at frequencies from 50 c/s. up to at least 10 Kc/s., Grade 1 is a low-density material with a maximum permeability of 860. Grade 2 is denser, with a maximum permeability of 1,000. Grade 3 is similar to Grade 1 in magnetic characteistics but has remarkable machining qualities and high strength.

The method employed to determine engine speed is an integral count of the number of revolutions occurring in a standard time-interval, derived from a 1,000 c/s. valve-maintained tuning fork.

The Plessey Company, Ilford.

Electro-Encephalograph Type 0A180A. The design of this transportable six-channel Electro-encephalograph conforms to the recommendations put forward by the Technical Sub-Committee of the E.E.G. Society. The equipment includes a fourteen-electrode headset with stand, three double-channel balanced amplifiers and a three-speed recorder carrying six signal pens and two marker pens. The main amplifier and recorder assembly is in the form of a wheeled desk-trolley with full provision for the storage of cables, accessories and spares.

Potentiometer Test Set TF 776. A voltage-source calibrated in pH and millivolts for checking the scale accuracy, grid-current, input-impedance and temperature compensator efficiency of pH equipment.

Marconi Instruments, Ltd., St. Albans, Herts.

Acoustic Strain-Gauge, D-423-A. This robust instrument measures mechanical strain in structures in terms of the change in frequency of a vibrating wire. It is portable, and lends itself to measurements in exposed locations and on rough surfaces.

Audio-Frequency Oscillator. Among the range of resistance-capacitance oscillators shown is the D-330-B., which covers 20-20,000 c/s. in three overlapping ranges. No reaction-control is necessary, and the output is constant within $\pm$ 1 db. over the entire range. A built-in output attenuator and a reference-level indicator are fitted.

Impedance Bridge, D-197-A. This versatile instrument is completely self-contained except for a pair of phones which are required for A.C. measurements. It is capable of measuring inductance, capacitance and resistance over a wide range with good accuracy. Other more specialised bridges are also displayed.

Muirhead & Co., Ltd., Elmers End, Kent.

"Avo" Heavy-Duty Meter. A modified version of the "Avo" Universal Meter at 1,000 ohms per volt. Cased in heavy cast-aluminium. Fitted with cut-out, mirror and $3\frac{1}{2}$ in. scale.

Automatic Coil Winder Co., Ltd., London, S.W.1.

A WIDE-BAND AMPLIFIER

(100 c/s. to 20 Mc/s.)

By J. C. PLOWMAN, B.Sc.(Hons).*

IT is well known that the upper frequency limit of simple, choke compensated (series or shunt) amplifiers cannot be extended beyond about 10 Mc/s., without considerable loss of gain per stage. This necessitates several stages of amplification when band-widths of this order are desired.

The essential features of a wide-band amplifier are that it should give a constant amplification throughout the desired band-width, and also that either no phase shift, or a phase shift directly proportional to frequency, should be produced. These considerations suggest the use of a low-pass filter networks as the anode loads of the amplifying stages and a great deal of work has been performed during the last ten to twelve years upon this subject.

This article deals with a two-stage, filter-coupled amplifier with a cathode follower output, and although no original principle is involved, it does show the potentialities of the filter coupling method, when used in conjunction with modern video-amplifying valves.

The Filter Circuit

Certain m-derived filters have an approximately constant impedance over a wide frequency range—the impedance being nearly equal to a pure resistance in the pass-band, and hence such a filter can be terminated by a resistance of the correct value without affecting its properties within the desired pass-band.

A shunt derived π type filter has the above property when $m = 0.6$.

A convenient form of filter to be employed in an amplifying stage is as shown in Fig. 1.

The values of the constituent components of this filter network can be calculated (see Appendix 1) and are found to be given by the following expressions:

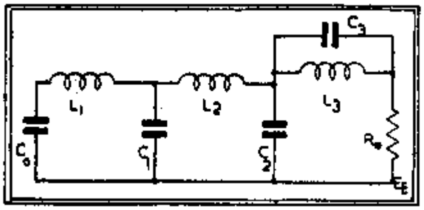


Fig. 1. The Filter Network

$$L1 = L2 = \frac{R_0}{\pi f_c} \text{ and } L3 = 0.3L1$$

$$C0 = C1 = \frac{1}{\pi f_c R_0} \text{ and}$$

$$C2 = 0.8C1, \; C3 = 0.53C1,$$

where f_c is the cut-off frequency of the filter. The actual upper frequency limit of the band-width of the amplifier is about $0.8 f_c$.

When the filter is in circuit it is arranged that the stray capacitances of adjacent stages are utilised as the shunt capacitances of the filter—thus in effect $C1$ is fixed and the other parameter, from which filter component values are calculated, can be either the upper frequency limit of the band-width, of the value of R_0.

The Amplifier

The amplifier is designed on the principles outlined above. Valves with high value of mutual conductance (g_m) are required. It is also necessary that these valves should have low input and output capacitances and desirable that they should have a wide grid base—particularly when more than one stage is employed. Valves type EF55 (Mullard) are very satisfactory. These valves are designed for video-amplifying, and although they have a rather high anode-cathode capacitance compared with the EF50 (Mullard), their high g_m (approximately 12 mA/V) more than counterbalances this disadvantage.

An EF50 type valve connected as a triode is employed as cathode follower output stage, primarily because of its low input capacitance.

In the construction of the amplifier it is essential to keep the stray capacitances down to an absolute minimum. It is necessary to measure the anode-earth capacitance of the first stage, the grid-earth, and then anode-earth capacitances of the second stage and also the input capacitance of the cathode follower.

The anode-earth stray capacitance of the second stage (Fig. 2) is employed as $C1$ in the filter network, and the cathode follower input capacitance as part of $C0$—the remainder of $C0$ being provided by small trimming capacitors from grid to earth.

Taking, for example, a desired band-width of about 20 Mc/s.; an anode-grid capacitance of V_2 = 14 pF and an input capacitance of V_3 = 6 pF.

If R_0 = 1,000 ohms then

$$f_c = \frac{1}{\pi C1 R_0} \simeq 24 \text{ Mc/s.}$$

20 Mc/s. is approximately $0.8 f_c$ and thus the desired band-width will be obtained.

The voltage gain of the stage will be:—

$$g_m \times R_0 = 12 \text{ mA/V} \times 1{,}000 \text{ ohms} = 12.$$

The values of $C2$, $C3$, $L1$, $L2$, $L3$, are all easily calculated.

The amplifier is constructed with consideration for low frequency characteristics, which it is not proposed to discuss in detail. It will suffice to note that phase shift is of primary importance at low frequencies and proper attention must be paid to the time constants of the grid, anode, cathode and screen circuits. Optimum conditions must be aimed at, for in view of the number of interdependent parameters it is not possible to achieve ideal low frequency compensations in a practical circuit.

The design of the first and second stages of the amplifier is similar, although component values in the

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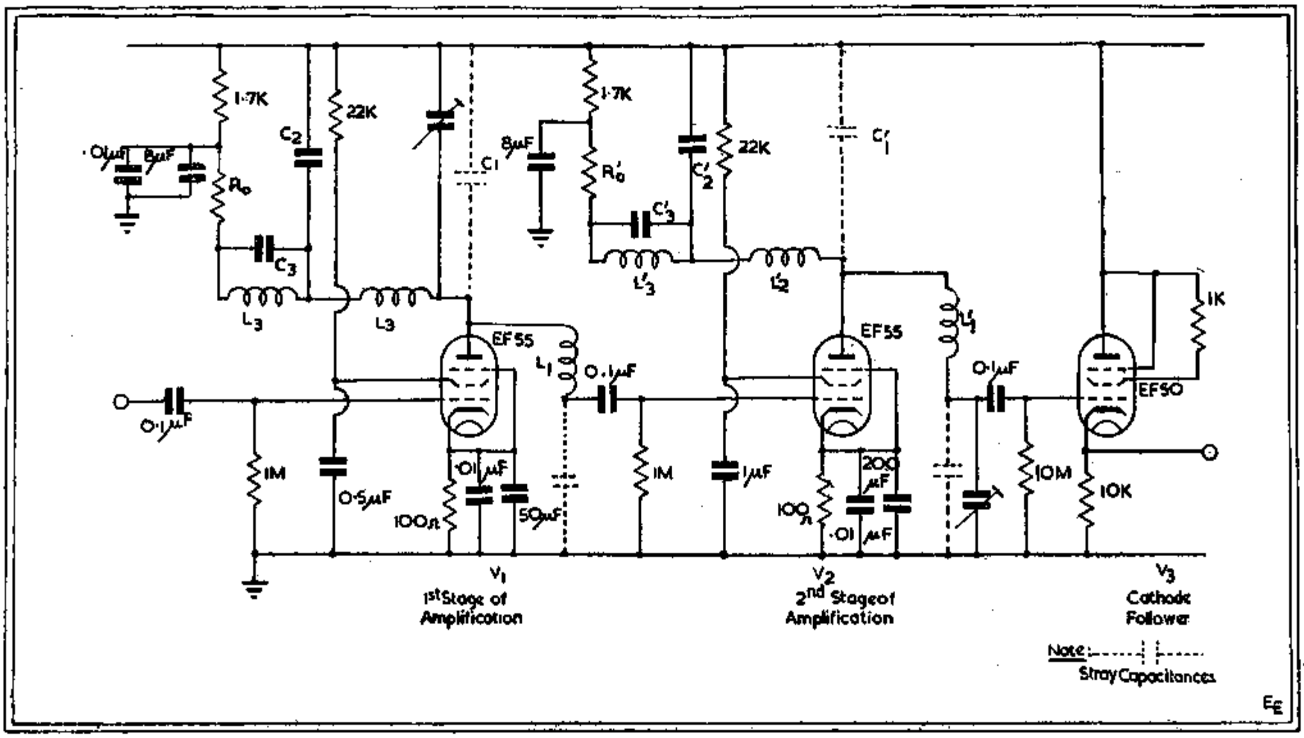

Fig. 2. Wide Band Amplifier Circuit.

filter network may differ slightly due to differences in stray capacitance. A padding capacitor of suitable value is required across the cathode follower input, since the input capacitance of this stage is small compared with the anode-earth capacitance of the preceding stage.

A circuit of a complete amplifier is shown in Fig. 2, and the overall frequency response characteristic in Fig. 3.

The response of the cathode follower stage does not fall appreciably within the frequency range considered, provided that the input capacitance of any stage into which it feeds is not greater than 20 pF.

Construction

The amplifier may be built upon a chassis 12 in. × 7 in. × $2\frac{1}{2}$ in.

There are two main points to observe in the construction:—

(a) Low input and output capacitances must be aimed at, this being achieved by mounting the input and output terminals on paxolin strips, which are then fixed across holes of about $1\frac{1}{4}$ in. dia., in the ends of the chassis.

(b) It is advisable to mount each filter unit on separate tag boards, which are then mounted on the underside of the chassis, supported about an inch from the base.

The coils can be conveniently wound on $\frac{1}{2}''$ hollow paxolin tubing using 34 or 36 S.W.G. copper wire—the number of turns required can easily be derived from standard coil data.

Conclusion

The overall gain within the pass band (100 c/s. to 16 Mc/s.) of the amplifier is 38 db. The output falls by one db. at 17 Mc/s., and by 3 db. at 18 Mc/s.

Both stages are operated in Class A condition with an H.T. of 300 volts (preferably stabilised).

Thus we see that, by suitable use of modern valves in conjunction with filter networks, the upper frequency limit of wide band amplifiers can be considerably extended.

The main use of such amplifiers is in the radar field when it is desired to amplify narrow pulses prior to application to a viewing tube or inspection circuits.

The author wishes to thank Dr. R. Feinberg for his encouragement in the preparation of this paper.

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Appendix I

Theory of the Filter Network

The filter network as shown in Fig. 1 can be sub-divided into its constituent sections as shown in Fig. 4:

Where $C' = C_0 - \frac{C_1}{2}$ and $C'' = C_2 - \frac{C_1}{2}$

Looking into the filter from AB, we see the input impedance of a constant k, π section.

From EF looking into the end section, which is an m-derived half section, it can easily be shown that we again see the input impedance of a constant k, π section. This impedance can, by suitable choice of components, be made identical with that observed from AB and thus the sections are matched, provided the terminating resistance, R_0, is of the correct value.

The correct value of R_0 can be shown to be equal to the value of k, in the constant k sections, as is briefly indicated below.

Consider the half section shown in Fig. 5:

Looking from EF into the section and assuming an impedance Z_2' across GH, we see an impedance Z_1', where

$$Z_1' = \frac{Z_b(Z_a + Z_2')}{Z_a + Z_b + Z_2'} \qquad (1)$$

Similarly looking to the left from

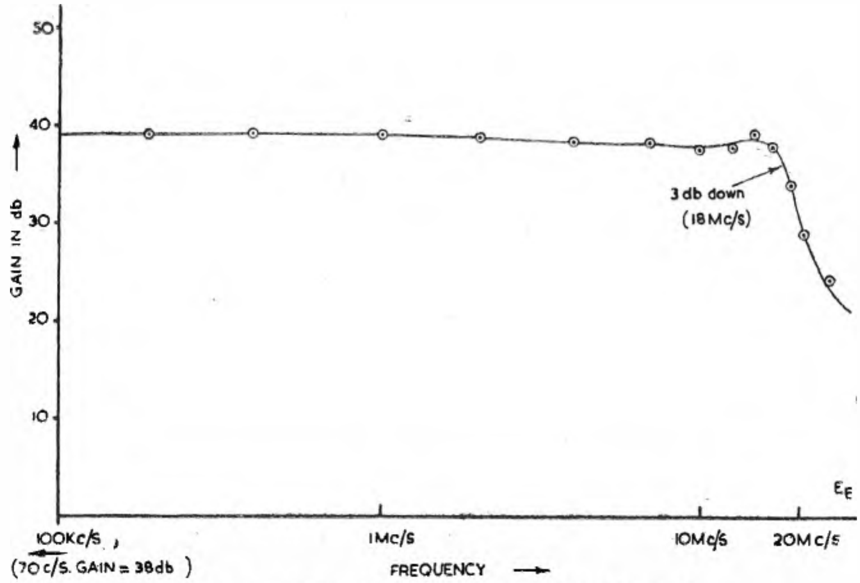


Fig. 3. Frequency reference curves for the wide band amplifier

GH—assuming impedance Z_1' across EF, we see

$$Z_2' = Z_a + \frac{Z_b Z_1'}{Z_b + Z_1'} \quad \ldots\ldots\ldots\ldots (2)$$

From these equations

$$Z_1' = \sqrt{\frac{Z_a Z_b}{1 + \frac{Z_a}{Z_b}}}$$

$$Z_2' = \sqrt{Z_a Z_b\left(1 + \frac{Z_a}{Z_b}\right)} \quad \ldots\ldots\ldots\ldots (3)$$

Z_1' and Z_2' are known as "image" impedances.

Now since the filter considered is an m-derived half section, then in terms of the constant K, prototype filter, from which this section is derived we have:

$$Z_a = \frac{2m\dfrac{L_1}{C_1}}{j(1 - m^2)\omega L_1 - j\dfrac{4}{\omega C_1}} \quad \ldots\ldots (4)$$

and $$Z_b = -j\frac{2}{m\omega C_1} \quad \ldots\ldots\ldots\ldots (5)$$

Substituting in the expressions for Z_1' and Z_2', we obtain:

$$Z_1' = \sqrt{\frac{L_1/C_1}{1 - \dfrac{\omega^2 L_1 C_1}{4}}} \quad \ldots\ldots\ldots (6)$$

$$Z_2' = \sqrt{\frac{L_1/C_1}{1 - \dfrac{\omega^2 L_1 C_1}{4}}} \times \left[1 - (1 - m^2)\frac{\omega^2 L_1 C_1}{4}\right] \quad \ldots\ldots\ldots (7)$$

Thus:—

(a) Z_1' = Input impedance of the constant k, π section, as previously stated.

(b) Z_2' can be shown to be approximately equal to $\sqrt{\dfrac{L_1}{C_1}} \equiv k$ throughout the major part of the pass band, when $m = 0.6$. k is resistive and thus when $R_0 = \sqrt{\dfrac{L_1}{C_1}}$ the filter is matched.

It can easily be shown that the cut-off frequency, $f_c = \dfrac{\omega_c}{2\pi}$ for this filter network, is given by:—

$$\omega_c = \frac{2}{\sqrt{L_1 C_1}} \quad \ldots\ldots\ldots\ldots (8)$$

(This expression is obtained from Z_1' or from the input impedance of a π section.)

Now we have seen that the input impedance of a constant k, π section is observed from AB.

This impedance is given by

$$Z\pi = Z_1' = \sqrt{\frac{L_1}{C_1}}\,\frac{1}{\left(1 - \dfrac{\omega^2 L_1 C_1}{4}\right)^{\frac{1}{2}}}$$

$$= \frac{k}{\sqrt{1 - \left(\dfrac{\omega}{\omega_c}\right)^2}} \quad \ldots\ldots\ldots\ldots (9)$$

Thus $Z\pi$ is not independent of frequency.

However, if a small capacitor C' is connected across AB, and, by consideration of the resultant impedance of $Z\pi$ and this capacitor in parallel, it is found that when

$$C' = \frac{1}{\omega_c R_0} \quad \ldots\ldots\ldots\ldots (10)$$

then the resultant input impedance is:—$Z_{(in)} = k \equiv R_0$ .... (11)

It can also be shown that the phase shift varies linearly with frequency throughout the major part of the pass-band.

The values of constituent components of the network are obtained as below.

$C_2 = \dfrac{C_1}{2} + C''$, and $C'' = \dfrac{mC_1}{2}$

Thus $C_2 = \dfrac{C_1}{2}(1 + m)$

Now $2\pi f_c = \dfrac{2}{\sqrt{L_1 C_1}}$ and $R_0 = \sqrt{\dfrac{L_1}{C_1}}$

Thus $C_1 = \dfrac{1}{\pi f_c R_0}$ and $C_0 = C_1$

For optimum conditions $m = 0.6$.

Therefore $C_2 = \dfrac{1.6}{2\pi f_c R_0} = 0.8C_1$

Also $L_1 = L_2 = \dfrac{R_0}{\pi f_c}$

From consideration of the m-derived half section we have:—

$C_3 = \dfrac{(1 - m^2)}{2m}\,C_1 = 0.53C_1$

and $L_3 = \dfrac{mL_1}{2} = \dfrac{0.3R_0}{\pi f_c}$

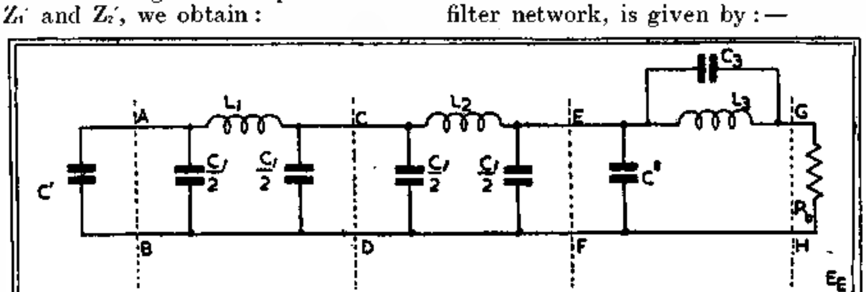
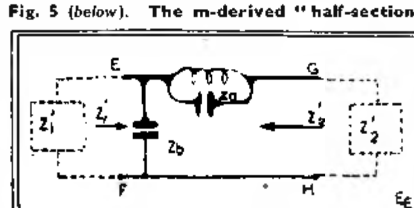


Fig. 4 (left). The filter network subdivided into its constituent sections

Fig. 5 (below). The m-derived "half-section"



An Automatic Timer for Radioactivity Measurements

By BRIAN D. CORBETT, B.A., and A. J. HONOUR*

IN research work involving radioactive isotopes it is usually necessary to make frequent determinations of the rate at which a Geiger counter tube or other device generates voltage pulses in response to ionising radiations. In a typical case it might be necessary to detect pulses of a few volts amplitude and 100 microseconds duration and to determine the "counting rate" over a range of perhaps 10 per minute up to several thousand pulses per second. An electronic pulse rate meter may often be used, particularly when the counting rate is high; but, owing to the random distribution of the pulses in time, such an instrument is usually not convenient for use at low counting rates, since either the reading must be imprecise or else the integrating time must be made inconveniently long. For a great body of work, therefore, use is made of an electronic counter or "scaler" together with a timing instrument such as a stop-watch.

The instrument described in this article has been designed to carry out this timing operation automatically in conjunction with one of the commercially available electronic scalers developed by the Atomic Energy Research Establishment. These scalers embody two hard-valve scale-of-ten circuits; units and tens are normally displayed by means of neon lamps, while the hundreds are displayed on a non-resettable electro-mechanical register of the telephone call-meter type; alternatively, a scale switch enables tens or units to be fed into the register. The relay which drives the hundreds register is wired so that its contacts may be used for the operation of external apparatus. The timer embodies the following features:—

(a) Hundreds of counts, minutes and seconds are displayed on three dial-type registers which may be rapidly reset by means of a key. The same key is used to reset the scaler neons.

(b) The duration of an operation may be preset in the range 1 second—200 minutes.

(c) Alternatively, the number of counts may be preset in the range 100-20,000 counts.

(d) An operation may be terminated at any time by means of a press-button switch.

(e) An alarm bell may be connected to ring at the conclusion of an operation.

(f) A bell may be made to ring at intervals of 12.5 seconds or at intervals of 25, 250 or 2500 counts. This facility is useful when it is required to take continuous readings of a slowly changing counting rate.

(g) The clock which is used to control the timing and drive the two time registers is a high-grade spring-driven instrument provided with a pair of contacts closing every half-second. Alternatively, or in addition, an electrically-controlled stop-clock may be employed.

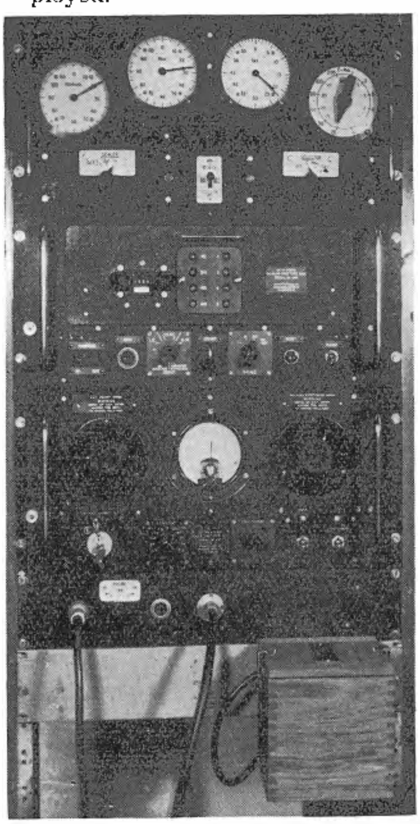


Fig. 1. The complete equipment, rack-mounted

(h) The complete timer equipment may be constructed for a few pounds from components currently available in the surplus market.

Fig. 1 shows a general view of the counting equipment rack with the timer panel at the top and the clock in its wooden box at the bottom right. Fig. 2 shows the timer panel and Fig. 3 is a plan view of the chassis, which includes an A.C.-operated power supply. Referring to Fig. 2, the functions of the timer panel controls are as follows:—

In the *Scaler Out* position the timer and scaler may be operated independently of one another. In the *Scaler In* position the scaler "hundreds out," "reset neons" and start-stop circuits are linked with the timer and the timer is prevented from starting until the scaler key is on.

System switch (S_3, bottom right). In the position marked *C* the number of counts may be preset; in the positions marked *Min* and *Sec* the time may be preset. In the *Selector Off* position the count may only be terminated manually.

Selector switch (S_4, top right). This selects a uniselector contact which determines the duration of the count. It is permissible to alter the setting of this control after counting has begun.

Reset key (S_6, bottom centre). A downward movement (non-locking) resets the scaler neons. In the upward position of the key the three dial pointers zero themselves, if not already reset, at a rate of one revolution in 2.4 seconds. On restoring the key to the central position the amber lamp in the centre of the panel lights up to indicate that another timing operation may begin.

Start button (S_2, on right of key). This is coloured green. Counting and timing begin when this button is pressed provided the amber lamp is alight (indicating that the dials have been reset), and provided that the scaler count key is on and the clock is running. The green lamp (immediately above the start button) lights up on starting and remains alight until the conclusion of the count.

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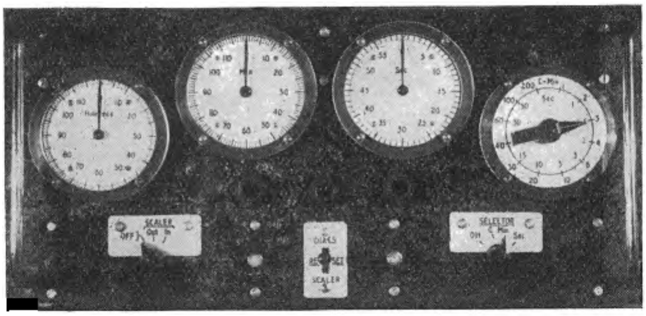


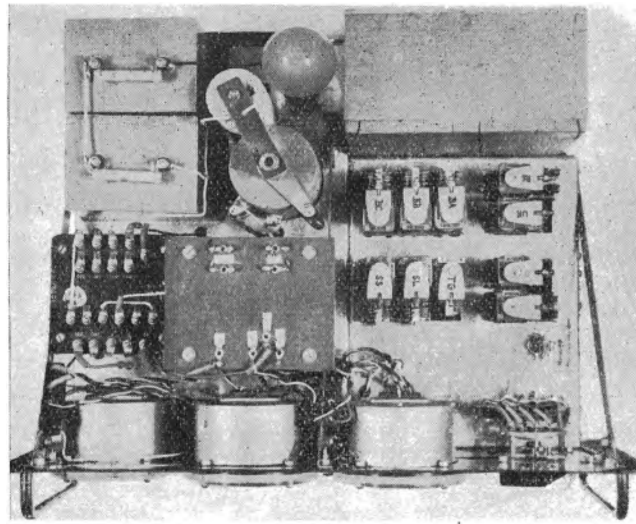


Fig. 2 (above). Timer control panel-switches set to stop count after 3 minutes

Fig. 3 (right). Plan view of timer chassis with relay cover removed

Stop button (S_3, on left of key). This is coloured red and may be used in order to terminate a count at any time. Above this button is a red lamp which is normally alight except when timing is in progress.

Clock and Register Circuits (Fig. 4)

Both the clock and the dial-type registers are currently available in the surplus market; the Service nomenclature is Contactor, Master, Type 2 (24 volts) and Contactor, Remote, Type 4 (24 volts) respectively. The clock or master contactor runs for about seven hours at a winding and is provided with three terminals, namely:—

(1) Half-second contact.
(2) 24-volt thermostatically controlled heater (requires about ½ amp.).
(3) Common.

A built-in filter provides interference suppression at all three terminals. The clock is connected to the timer through pins 1, 2 and 3 of a five-pole socket (SK_1 or SK_2, see Fig. 4). This socket is duplicated in order to provide for simultaneous connexion of an electrically-controlled stop-clock if desired; the stop-clock solenoid is connected to pins 4 and 5 for operation on a break-to-start and make-to-stop basis.

The control circuits are described below and it will suffice for the present to state that after the *Start* button has been pressed the first earth pulse provided by the master clock at pin 1 of SK_1 or SK_2 is utilised to operate a relay trigger circuit; this causes relay contacts SS_6 to close at the trailing edge of the pulse and to remain closed for the duration of the count. The solenoid of relay *HS* (1,000 ohms, 4 contact units) is consequently energised through the clock half-second contacts while timing is in progress. Contacts HS_1 of relay *HS* are connected across SS_6 in order to prevent clipping of a clock pulse when SS_6 opens at "stop."

The three registers (see Fig. 4) are fitted with 200-ohm solenoids and operate their pointers through wheels having 120 teeth; hence the seconds register makes a complete rotation in one minute when operated from the contacts of the master clock. Movement of the pointers takes place at the leading edge of the energising pulse. The registers, as supplied, are fitted with a pair of cam-operated contacts which make over one quadrant of the dial. For homing purposes each cam was replaced by an insulating disk of the same diameter bearing silver segments disposed so as to provide a circuit between two silver brushes at the zero position of the pointer only. The seconds register was provided with an additional segment and brush (marked "min" in the diagram) providing a circuit to earth in the 119th position of the pointer for operation of the minutes register. The circuit is such that the brushes never carry current at the moment of break.

The reset key utilises six change-over contacts (S6/1 - 6) for homing the registers in the upward position and one make and one break contact (S6/7 - 8) for resetting the scaler neons in the downward (non-locking) position. In the normal (i.e. central) position of the key the two time registers are operated from the main 50-volt D.C. supply through 200-ohm dropping resistors. The hundreds register is permanently connected to one side of a 240-volt half-wave rectified supply with a 32 μF reservoir capacitor and an ordinary 230 V 40 W incandescent lamp as dropping resistor; this arrangement reduces the resolving time of the hundreds register to approximately 15 m.sec. when it is actuated by a suitable driving circuit. When all three registers are zeroed and the reset key is in the normal position there is a circuit from the point marked "X" in Fig. 4 to earth by way of the register "commutators"; this circuit is utilised in order to light the amber "ready" lamp and also to render the start button operative.

When the reset key is raised the three registers are connected in series with one another and with the 240 V half-wave supply; simultaneously each solenoid is connected with the appropriate homing brushes and the smoothing action of the 32 μF electrolytic capacitor is removed by the operation of S6/6. The dial pointers move simultaneously at a rate of 50 impulses per second and each pointer stops dead in the zero position when the associated solenoid is shorted out. Successful operation of the registers in this manner depends not only upon the reduction in L/R afforded by the high series resistance but also upon the careful adjustment of the pawl-stop and back-stop grub-screws. Moreover, it was found necessary in most cases to make the air-gap at "make" adjustable by means of an additional screw tapped through the armature. The registers were further modified by fitting enlarged dials of Bristol board with clock hands and "Perspex" rims and covers.

In the downward position of the reset key an earth is applied to terminal PL_2 which is connected to terminal No.2 on the scaler to reset

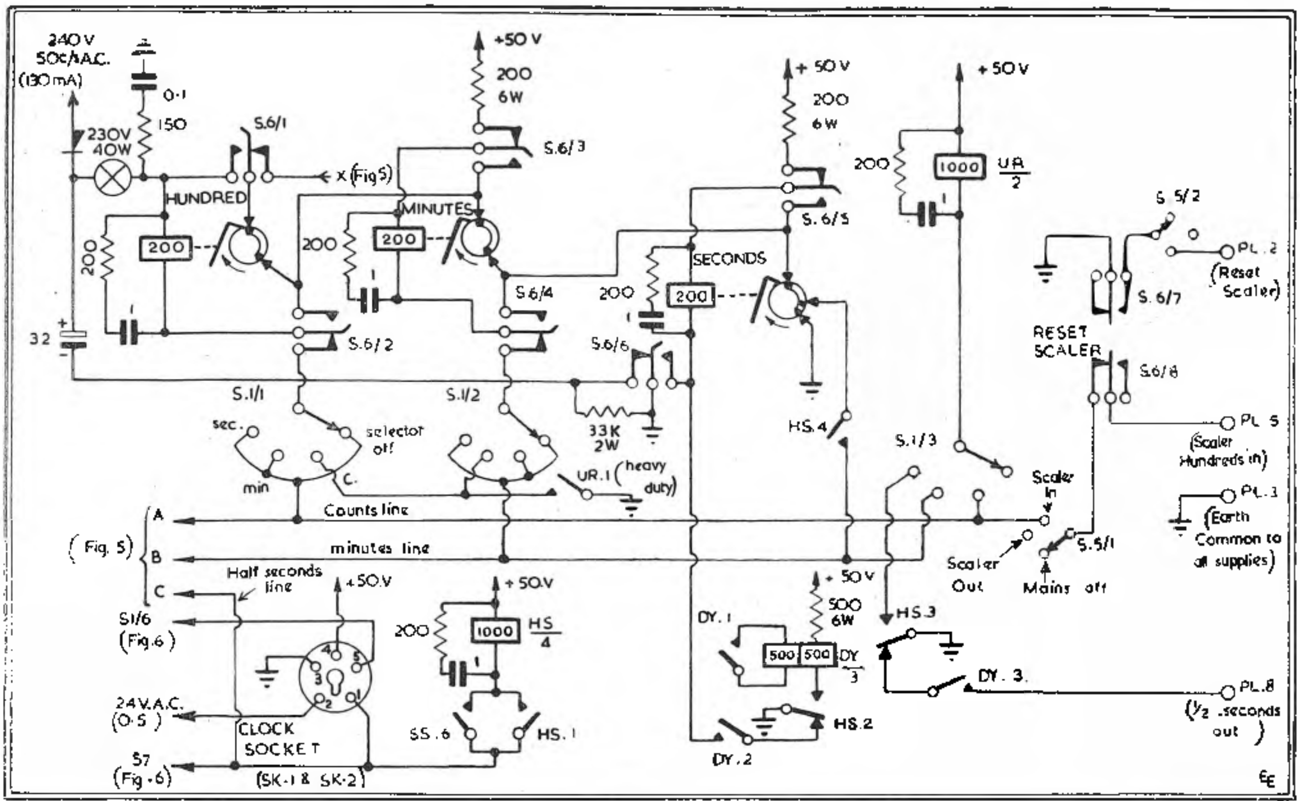


Fig. 4. The clock and register circuits. (Terminals marked PL are on a 10-way terminal strip for external connexions)
Clock socket: *Pin 1—1-second contacts Pin 2—Heater Pin 3—Common Pins 4-5—Stop-clock solenoid*
Switches: S_1—*System switch* S_5—*Main switch* S_6—*Reset key. Note : S_6/1-3-5 should make before S_6/2-4-6. S_6 is ganged to a D.C. toggle switch in the A.C. mains supply.*

the neons. Simultaneously the connexion to the scaler " hundreds " relay via terminal PL_5 is broken by $S6/8$, thus preventing the unwanted operation of the hundreds register.

When timing is in progress relay *HS* is energised at half-second intervals by the clock contacts as stated above. The arrangement adopted for generating the minute impulses requires that the movement of the seconds pointer should be delayed until the trailing edge of each clock pulse is reached; this is effected by means of relay *DY* (500 + 500 ohms, 3 contact units) in the following manner. *DY* is energised when contacts HS_2 are at " make", DY_1 then closes and this makes *DY* slow to release. At the trailing edge of the clock pulse HS_2 goes back to the " break " condition, but DY_2 remains at " make " for a short additional period during which the seconds register is energised.

When the seconds pointer reaches the 59½ seconds position an earth appears on the minute brush of the seconds register and consequently when *HS* again operates the minutes line is earthed through HS_4 in synchronism with the clock contacts. Due to the delay in the operation of the seconds register this earth pulse is not interrupted by the movement of the seconds " commutator."

The system switch (S_1) is wired so that the hundreds register is operated by the scaler " hundreds " relay except when the number of counts is to be preset. In the latter condition the scaler relay energises relay *UR*; this has two heavy-duty contacts, one of which operates the register while the other operates the uniselector.

Pulse-Counting Circuits (Fig. 5)

The basis of the pulse counting circuits is a standard eight-bank uniselector with two 37.5-ohm coils wired in series for operation from the 50 V D.C. supply line. In addition three relays are employed in order to raise the counting capacity of the uniselector from 50 to 200 pulses; these are the relays *BA*, *BB* and *BC* in Fig. 5.

In the " stop " condition of the timer, which holds when the apparatus is first switched on or immediately after the conclusion of a count, relay contacts RS_6 and RS_7 are closed and the uniselector is connected through its interruptor contacts to the homing banks (7 and 8); this causes the uniselector to home automatically.

When the start button is pressed *RS* and RS_7 open and SL_3 closes shortly afterwards; the uniselector is then operated by relay contacts UR_2 which are closed by the half-second, minute or count pulses in accordance with the setting of the system switch.

The numbered terminals at the top of Fig. 5 are connected to the selector switch (S_4 in Fig. 6) and this switch connects one of them to the relay trigger circuit. It is the function of the uniselector to provide a circuit between the trigger and one of the points marked A, B and C in Fig. 5 when the Nth pulse is due to arrive (where *N* is the number of pulses to be counted). Relay contacts SL_3 open at the leading edge of the pulse operating the trigger circuit, so that the almost simultaneous closure of UR_2 causes no movement of the uniselector at the Nth pulse. This means that when only one pulse is to be counted, as when the time is preset at one minute, the wipers do not move from their starting position and the sound of the uniselector homing is not heard on " stop ".

Bank 6 of the uniselector is used for operating the relays *BA*, *BB* and *BC* during successive rotations of the uniselector wipers when *N* is selected to be more than 40. While timing is in progress relay contacts RS_5 are closed. When the wiper of Bank 6 reaches the contact numbered 48 (this occurs at the 47th pulse operating the uniselector) relay *BA* locks on through its contacts BA_1. Thereafter there is no circuit for the stop signal through Banks 1 and 2 of the uniselector. During the next rotation of the wipers a circuit is established through Bank 3 via BA_4 and BB_5 to the terminal marked 60. Meanwhile relay *BB* has been prepared by the closing of BA_2 and when the Bank 6 wiper reaches contact 46 on the second rotation *BB* locks on. Operation of BB_2 causes *BA* to release owing to the differential connexion of its twin windings; this enables relay *BA* to be utilised a second time. *BC* locks on during the third rotation and *BA* locks again during the fourth rotation when the Bank 6 wiper reaches contact 42. The circuit for the stop signal is so arranged that the timer will run indefinitely if the selector switch is turned back to a value of *N* which has already been exceeded; hence there is no danger of premature stopping when the switch setting is changed while timing is in progress. At "stop," relay contacts RS_5 open and the three counting relays are released so as to be ready for another cycle.

Bank 5 of the uniselector is utilised in order to make the homing of the uniselector a necessary precondition for the lighting of the amber "ready" lamp and operation of the start button. Terminal PL_7 is used when it is required to operate a bell at every 25th uniselector driving pulse.

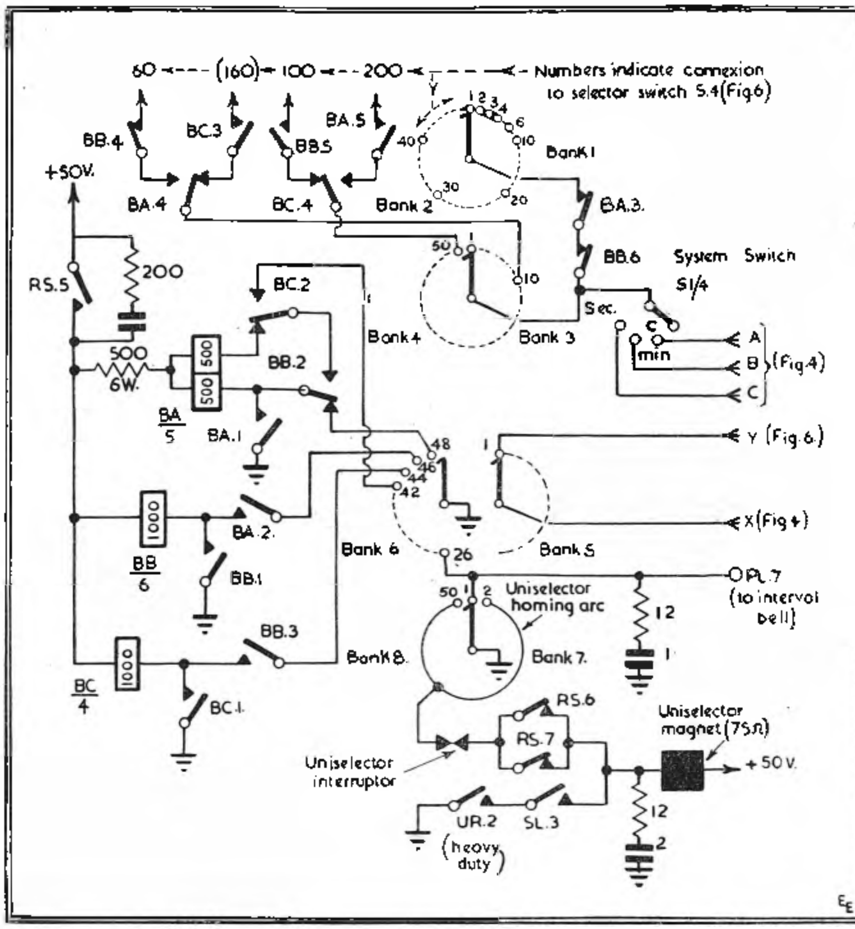


Fig. 5. Pulse-counting circuits

Timing and Control Circuits (Fig. 6.)

Before timing can begin the master relay *RS* (see Fig. 6) must first be locked on. As long as this relay is released the red "stop" lamp will be alight. When the main switch (S_5) is in the *Scaler In* position terminals PL_4 (50 V D.C. out) and PL_5 must first be linked by operation of the scaler count key, which is provided with spare contacts for such a purpose. When this has been done the scaler will still be inhibited from counting owing to the fact that an earth is applied to scaler terminal No. 1 via terminal PL_3 in the timer. Removal of this earth will break a high-speed relay circuit in the scaler which allows pulses to enter the first electronic scale-of-ten.

After the three registers have been reset and provided the uniselector has homed (an automatic process) there is a circuit to earth from the point marked "Y" in Fig. 6 by way of uniselector Bank 5 (Fig. 5) and the register "commutators" (Fig. 4). Consequently the amber "ready" lamp will be alight as long as the start button (S_2) is not pressed. Under these conditions relay *RS* will lock on through contacts RS_1 when S_2 is operated and the red and amber lamps will be extinguished.

The relay trigger or scale-of-two circuit, which performs the start-stop switching operations, is made up of the three relays *TG*, *SL* and *SS*. The behaviour of this device is somewhat similar to that of an electronic trigger circuit. Initially the three relays are released but they become ready to function when contacts RS_3 and RS_4 are operated after pressing the start button Relay *TG* utilises two make-before-break contacts (TG_1 and TG_2). The first pulse to operate *TG* normally comes from the half-second contacts of the master clock through the switch S_7. At the leading edge of this pulse relay *SL* operates through SS_2 and TG_1 while at the trailing edge *SS* is energised through SL_2 and TG_2; both *SL* and *SS* then remain at "make" until another pulse arrives to operate *TG*. The operation of SS_3, after the passage of the initiating pulse, switches the earth return circuit of *TG* to the movable arm of the selector switch (S_4) and *TG* is prepared to receive the stop signal pulse by way of the uniselector switching contacts. Operation of *TG* by the stop signal causes *SL* to release at the leading edge while *SS* remains operated through SS_1 until the trailing edge of the stop signal is reached.

The switch S_7 is a behind-the-panel control which is only used when it is desired to use a stop-clock alone instead of the master clock; in these circumstances the switch is thrown to the position marked "*Stop-Clock Only*" and the initiating pulse to the trigger circuit is provided by the operation of contacts *S2/2* of the start button.

At the beginning of the timing pulse, which is normally the first master clock pulse to occur after pressing the start button, *SL*, operates and the green "running" lamp lights up. At the conclusion of the timing pulse SS_6 in Fig. 4 closes and relay *HS* is then ready to be operated by the subsequent half-second pulses. Simultaneously SS_4 (Fig. 6) removes the "stop" earth from the scaler start-stop relay via terminal PL_1, and SS_5 performs a similar function on the stop-clock when one is used. The operation of SS_4 also applies a holding earth to the master relay *RS*, so that this relay remains operated when the circuit to earth from the point "Y," through the uniselector and register "commutators," is broken by the movement of these devices.

After the stop signal has operated *TG*, the circuit of relay *HS* is broken so that the two time registers cease to function. The "running" lamp is extinguished at the beginning of the stop signal by the action of SL_4; *RS* releases at the end of the signal due to SS_3, so that the "stop" lamp lights up. When the time is preset the "stop" earth is re-applied to the scaler and stop-clock at the trailing edge of the appropriate master clock pulse by the action of SS_4 and SS_5 respectively. When the number of counts is preset the "stop" earth to the scaler and stop-clock is applied at the leading edge of the *N*th "hundreds" pulse by the release of SL_4 and SL_5 acting respectively through sections *S1/5* and *S1/6* of the system switch.

When it is desired to control the duration of a count manually, or when a count is to be terminated prematurely, the stop button S_3 may be operated. This causes relay *RS* to release, and immediately afterwards any other relays which may be energised are also released. The dials may then be read and a new count started as soon as the registers have been reset. When the scaler is switched in via S_5, raising the scaler count key will also perform this stopping function.

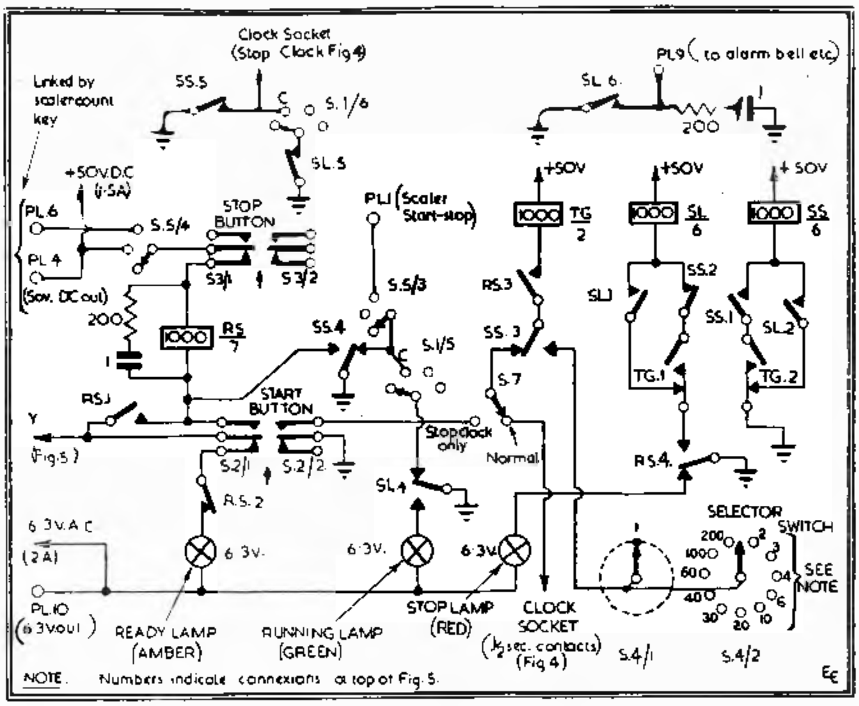


Fig. 6. Timing and control circuits

Timing Accuracy and Other Considerations

The master clock (Contactors, Master, Type 2) has been proved to give very consistent results and, provided its thermostatically controlled heater is employed, an accuracy considerably better than that of most stop-watches is to be expected. In these circumstances it is desirable to consider the timing errors which may be introduced by the finite operating and release times of the relays. The writers used standard P.O. 3,000 type relays throughout, since these relays are robust and are readily available in a multiplicity of different contact arrangements. Although the circuit diagrams only show the relay contacts that are actually necessary, in practice it is convenient to use relays having four "make" and four "break" contact units in all cases with the exception of *TG* (two make-before breaks) and *UR* (two heavy-duty "make" contacts).

When the number of seconds is preset relay *TG* is connected to the clock contacts both at "start" and at "stop," and the start-stop switching is effected by the release of *TG* in both cases. At the beginning of the count there is a further time interval before the scaler starts to function: this is made up of the operate time of relay *SS* and the release time of the scaler high-speed relay. At the end of the count there is a corresponding time lag which is made up of the release time of relay *SS* and the operate time of the scaler high-speed relay. The total switching error, which is the difference between these two time intervals, is not more than one hundredth of a second. A similar result is obtained for the minute-controlled condition of operation. This timing error could be reduced if desired by the connexion of a suitable resistance across the coil of relay *SS* so as to increase its release lag.

When the number of counts is preset there is a delay of the order of 20 milliseconds between the beginning of the selected "hundreds" pulse and the moment when the scaler stops counting. With this mode of operation, however, the timing accuracy is limited by the fact that the time may only be read to the nearest half-second unless a stop-clock is employed.

It should be noted that when the half-second pulses are being counted relays *TG* and *HS* are connected in parallel across the clock contacts when the uniselector reaches its final position. Similarly, *TG* is connected in parallel with *UR* during the stop signal when counts or minutes are present. These three relays, therefore, should be of similar type and should have identical windings in order that surge interaction may be avoided at the moment of break.

BOOK

Microwave Antenna Theory and Design

Edited by Samuel Silver. Radiation Laboratory Series ; volume 12. 623 pages. (McGraw-Hill, 1949), 48s.

THIS book provides a systematic treatment of the basic principles and the fundamental microwave antenna types and techniques. It is an excellent record of the research and development work which has been carried out in this field, and a very high technical standard has been maintained throughout.

The information falls into four principal groups; basic theory, theory and design of transmission lines, theory and design of antenna systems and arrays, and measuring techniques and apparatus.

In the first chapter, a survey is made of the problems encountered in the design of microwave antenna and the general plan of the book is discussed.

Chapter 2 deals briefly and concisely with the theory of the circle transmission line, referring to the use of transformation charts of the circle diagram type, and the equivalent network representation of the antenna.

The discussion of the field equations and general properties of an electromagnetic field in Chapter 3 is rather brief and the reader is referred to Stratton's Electromagnetic Theory for a more detailed treatment: however the outline is given and examples of certain idealised current distributions furnishes a theoretical guide to the design of various antenna feeds.

The subject of wave propagation, approached from the point of view of geometrical optics, makes interesting reading in Chapter 4, and in Chapter 5, the theory of scattering, diffraction and reflexion is developed with reference to general techniques, leading on to aperture systems and antenna radiation patterns in Chapter 6.

This treatment of basic theory occupies about a third of the book and the remaining four hundred pages are devoted to more practical aspects and to the solution of design problems.

Chapter 7, dealing with microwave transmission lines (including waveguides) is very short and reference is made to other publications to enable the student to fill the gaps.

Chapter 8 is a very interesting introduction to the problem of feeding microwave antennas. There is some confusion in section 8.3 in the explanation of dipole-disk feeding and it is felt that the device for feeding a balanced dipole from an unbalanced coaxial line (termed slot feeding) might have been dealt with more fully.

Chapter 9, the work of three authors, supplies a wealth of information on broadside and end fire arrays, slot radiators in waveguides, and the various feeding methods. It includes a very interesting reference to probe fed slots.

The difficult subject of radiation from the ends of waveguides and from electromagnetic horns has been thoroughly examined in Chapter 10, and an account is given of pressurising and matching enclosures.

Dielectric and metal plate lenses receive a very practical treatment in Chapter 11, particular reference being made to the subject of permissible tolerances, demonstrating that to American manufacturers suitability for mass production is regarded as one of the chief virtues.

Chapter 12 deals wth paraboloidal reflectors, parabolic cylinders, cheese and pill box antennas, with experimental results of their radiation patterns, gain factors, impedance characteristics, and some information to help with structural design problems.

Beam shaping techniques are surveyed in Chapter 13, and the questions of feeding the various shaped beam antennas, and of the design of the reflectors, are considered in a very satisfactory manner.

Antenna installation problems are discussed in Chapter 14 and the practices that have been adopted for dealing with them are clearly outlined. The problems associated with scanning antennas on aircraft are particularly investigated and PPI (plan-position-indicator) photographs are used to illustrate defects due to antenna pattern. The relation of the dielectric housing, or radome, to the general installation problem, and the design of radomes are considered from a practical angle.

The last two chapters are devoted to a discussion of measurement techniques and a survey of the equipment required for such measurements. This is an important section and includes the determination of impedance, radiation patterns, and gain of antennas; the experimental setups are well described.

In general it may be said that the aim to provide the antenna engineer with a fundamental microwave antenna reference book has been excellently achieved, but a complete work on such a vast subject in one volume must not be expected. A good mathematical knowledge on the part of the reader is assumed and the writers have made no attempt to provide physical pictures in explaining microwave phenomena. A minor criticism might be made that occasionally the diagrams illustrate points which are not clearly referred to

REVIEWS

in the text; for instance on page 287, Fig. 9.16, a slot marked (a) is shown which is not mentioned elsewhere. Again on page 243, Fig. 8.2 (b) a stub support termination is shown but the text refers to a metal plate which is illustrated in Fig. 8.3 with no linking explanation.

The book is a very valuable contribution to radio literature.

H. V. SIMS

Pulses and Transients in Communication Circuits

C. Cherry. Chapman & Hall. 317 pages. 32s.

A TEXTBOOK dealing with the effect of communication circuits on pulses and transients has long been required and this treatise will go far towards fulfilling that need. The author considers his subject from an engineering angle and although he makes full use of mathematics he is careful always to provide the physical interpretation. In spite of this valuable practical approach, the book will make difficult reading for those not well versed in trigonometrical and exponential integrals, differential equations of first and second order, and Fourier analysis.

There are eight chapters, one short appendix on the Probability Function (used in Chapter 6), a complete list of symbols and a fairly comprehensive index. The first chapter is introductory, starting off with circuit basic principles. The significance of the statement "current waveform × network function = applied E.M.F. waveform" is emphasised, and the physical effect of the variation of the real and imaginary parts of the solution to an *L.C.R.* circuit, whose E.M.F. is suddenly applied, is well illustrated. Reference is made in Fig. 8 to *Q*, but we have to wait for two pages to find its meaning; it would have been more helpful to use $(1/R)\sqrt{L/C}$ instead of *Q*. Chapter 2 is concerned with the frequency spectra of modulated waves, pulses and transients. In illustrating the rotating vector generating a sine wave the author makes the real axis vertical and the imaginary axis horizontal, and this convention may cause confusion to some students. The advantages of the negative and positive frequency spectrum concept in providing information on the phase as well as amplitude of the components is made very clear. In the section on frequency and phase modulated waves the explanation of the need for harmonic sidebands is rather scanty. The author's presentation of Fourier Analysis is less satisfactory than that found in some other textbooks, but its application to pulse waveforms is very well treated, as also are the theorems applicable to transients and their spectra.

The steady state characteristics of networks is the basis of Chapter 3, the most important sections of which are those dealing with circuit duality (useful in electromechanical analogies) and Foster's reactance theorem (one of the most valuable aids to circuit analysis). Some expansion of the last section would be an advantage so as to show how the individual *L.C.* components can be calculated for 4-element arms in addition to the 3-element arms which receive adequate treatment. Chapter 4 follows naturally and concerns the transient response of networks. The separate effects of attenuation and time delay distortion on pulse waveshape are well illustrated and the author indicates the method of attack in dealing with symmetrical and skew symmetrical waveforms. The equivalence of low pass and band pass filter response is also considered. The appetite for the complete working out of a practical example is whetted but never satisfied in this chapter. The omission of brackets from the repeated Equation 114 on page 131 (and subsequently) and the mixture of brackets and solidus may be puzzling to the student. The next chapter on some approximations and idealised results shows the response to be expected from circuits with idealised straight-sided selectivity curves; the fact that "overswing" is associated with a discontinuous selectivity curve is emphasised. A section on the effect of a pulse tuned outside the normal pass band concludes the chapter. The factors affecting the transient response of multistage amplifiers are described in Chapter 6. Signal-to-noise ratio is considered in relation to optimum bandwidth.

About half of Chapter 7 on asymmetric sideband transmission deals with a tone modulated signal and shows how an amplitude modulated carrier is converted into a combined amplitude and phase modulated signal by this asymmetry. The other half treats in a similar manner a signal with square wave modulation and one with transient modulation.

Chapter 8 on reflexion and echo effects in lines and lumped networks is particularly apposite today when there is so much interest in pulse testing, measurement and operation as in,

BOOK REVIEWS *(continued)*

for example, radar and fault location on telephone and power lines.

A valuable feature is the very complete list of references at the end of each chapter; this list gives also titles of works, whose subject matter, though not specifically mentioned in the chapter, would repay study. A minor point of general criticism is the rather large number of asides and notes which are distracting in the earlier chapters and which could often have been avoided at the expense of a little recapitulation.

The author in his preface states that his aim is to give the "essential groundwork using rigorous physical arguments" and he has certainly succeeded. His book can confidently be recommended to graduate engineers, particularly those teaching or practising in radar and television.

K. R. STURLEY

Electricity

By C. L. Boltz, George G. Harrap & Co. Ltd. 311 pages, 150 figs. Price 8s. 6d.

THIS book of about 300 pages is one of a series edited by the author and intended to provide the general reader with the fundamentals of Mathematics, Physics and Biology.

It modestly claims to be useful reading preliminary to taking to textbooks.

The author might also claim it to be of interest to readers who have already taken to textbooks, because it not only tells the reader interesting things about the early men of science usually omitted from textbooks but also relates their discoveries historically one with another so as to show clearly the progress of thought (or non-progress, see p. 154, Sturgeon) through a series of discoveries.

There is some incongruity in the choice of subjects for illustration, particularly in conventional symbols and some diagrams of plugs and switches in the home compared with Georg Ohm's current meter (Fig. 26) and modern D.C. armature (Fig. 95).

I think that the core should have been shown in place in Fig. 62; without it the diagram is misleading on the fundamental magnetic circuit. In the arrangement illustrated in Fig. 70 it would have been better to show the coil *not* along the line of the needle, and in the text it would be better to say "when the circuit is opened the needle returns to its original position of rest."

However, these are small points in a pleasing and stimulating book.

One might join issue with Mr. Boltz about his claims for the virtue of Baltic air (p. 148) and use his own excellent index to confound him and prove the superiority of the air of this island but France might come in and make it a triangular argument. Shall we make it four-sided by bringing in the New World?

H. CLARKE.

Atoms in Action

By George Russell Harrison. Allen and Unwin, 411 pages, 12s. 6d.

THIS book was first published here in 1940, and consists of a series of more or less independent chapters on different branches of applied physics in modern civilisation. It is attractively easy to read and Professor Harrison has shown considerable skill in making at least the main features of abstruse processes comprehensible to the layman. The book is particularly well adapted to the schoolboy at the radio-set-construction stage.

The fundamental nature of energy is well brought out in a variety of ingenious ways but the importance of some of the details treated may have been a trifle overstressed, making appreciation of their order of merit difficult. This also leads to occasional anticlimax, as when a considerable build-up of the genetical effect of X-rays is concluded by saying that a scientist of the General Electric Co. has suggested that orange trees which would resist the cold might be developed this way.

The rapid advance of physics is shown by the fact that one or two chapters, such as "The Doctor and the Physicist" have dated somewhat but three chapters added to the new (third) edition have done a lot to help this. A final chapter discusses future scientific developments and deals excellently with the erroneous but common view that modern wars kill relatively more people than the old ones. They are more expensive, and with more education and better communications people are able to worry about them much more but this is not necessarily bad.

Among the creditably small number of errors may be mentioned the statement (p. 144) that the ionosphere is transparent to wavelengths above 100 metres, and (p. 366) that the graphite in a pile slows neutrons down so as to stick more readily to the U_{238} nuclei—U_{235} of course is intended. There is some confusion in the explanation of the "compression" of intensity range on p. 151.

In spite of these details the book should be extremely successful in popularising an important subject.

J. H. FREMLIN

"Ultrasonics"

By Benson Carlin. McGraw-Hill Book Co. 270 pp. Price 30s.

NOW that ultrasonics has become established as a worthy subdivision of the electronic art the time is ripe for a fairly comprehensive book on the subject and the present volume fulfils this need very well indeed. The keynote of the book is practical engineering and it will be of most value to those who wish to experiment in and explore the subject for themselves. Classical analysis and full mathematical treatments are not included.

The theory of wave motion, particularly where ultrasonic waves are concerned, is dealt with concisely but adequately enough to form an introduction to practical work. Then follow two chapters which in some ways are the most useful in the book. They deal with crystals, one with their choice and technique of usage and the other with crystal holders. This latter chapter is full of valuable information of a type that has hitherto been very difficult to obtain.

Later chapters deal with resonance and reflexion, continuous-wave and pulsed systems, ultrasonic agitation and various practical considerations. It is felt that it might have been better to have dealt a little more fully with certain aspects of the subject, such as the design of oscillators or the interpretation of results, even had this necessitated the omission of other matter, e.g., magnetostriction, which although very interesting is perhaps not of major importance from an industrial point of view.

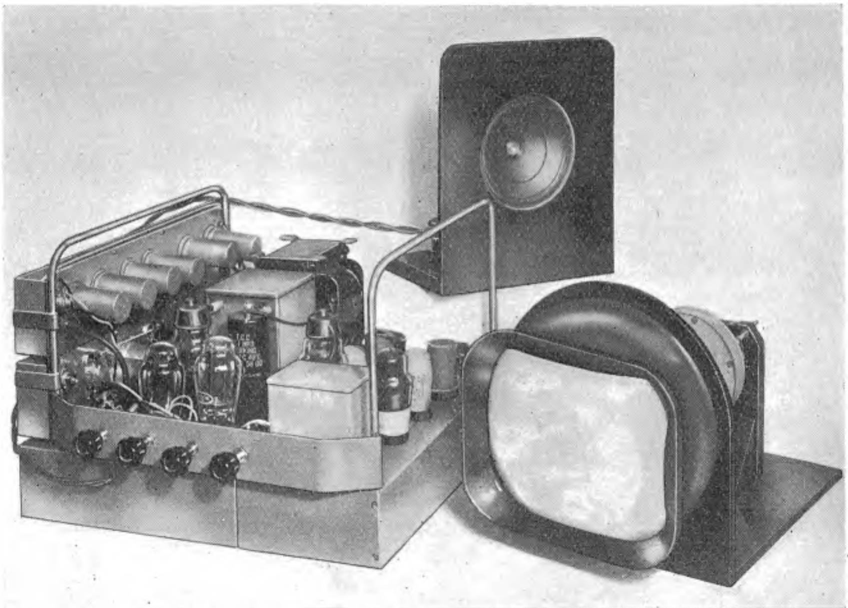
It is, however, nice to have the author's assurance that he has personally carried out most of the practical work and not taken other workers' results for granted.

Towards the end of the book he meets the difficulty that besets many writers; he realises that his allotted space is running out and yet many things have not been discussed. He then proceeds to mention them briefly in the hope that appetites will be whetted. This plan has something to recommend it but can be misleading at times, as in the present volume, where the photography of cathode ray screens is dismissed with less than 200 words.

It would be churlish, however, not to compliment the author on producing a useful book. He has tried to guide his fellow engineers in the foothills of a new and difficult land and, in the opinion of the reviewer, has succeeded.

J. H. JUPE

The "Electronic Engineering" Televisor for the Science Museum



FROM its introduction two years ago the ELECTRONIC ENGINEERING Televisor has proved one of the most popular designs of home constructed television receiver ever published.

The neat layout and workmanlike construction have earned commendation from professional television engineers, and its performance has been surpassed by only a few high-priced commercial receivers.

The publishers of ELECTRONIC ENGINEERING considered that such an example of British design and construction was worthy of being permanently exhibited and accordingly approached the Director of the Science Museum with an offer to instal a working model with an "exhibition" finish in the Radio Section. They are pleased to say that this offer has been accepted and a Televisor will shortly be a feature of the gallery which houses so many important examples of telecommunications progress.

Thanks are due to the suppliers of components who are so willingly collaborating to make the exhibit a noteworthy addition.

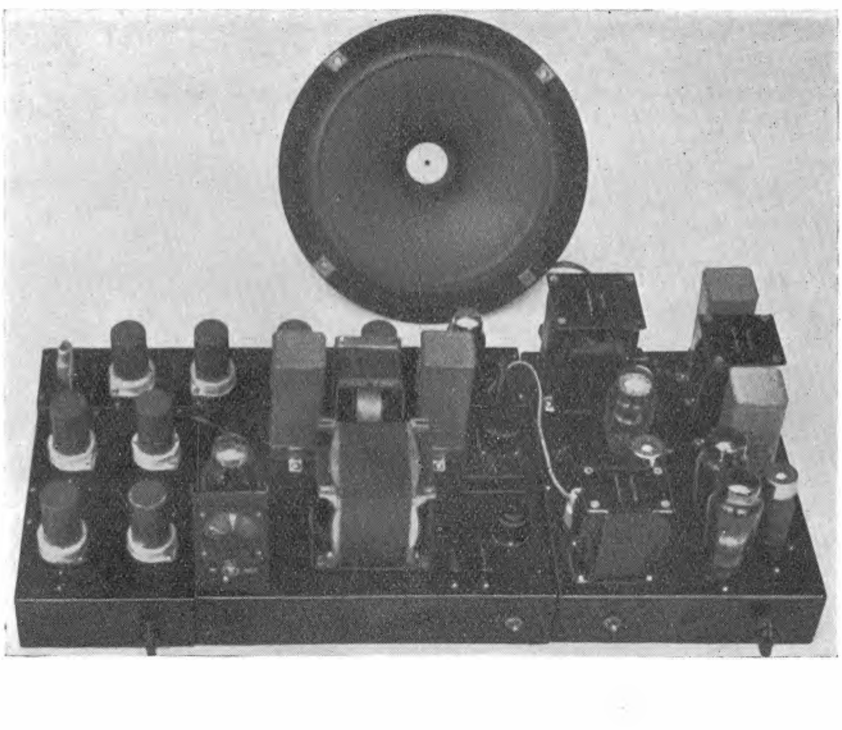
A New Frequency Modulated High Quality Receiver

AS many readers are already aware, the B.B.C. is shortly opening an experimental F.M. transmitting station at Wrotham, Kent, to cover London and the Home Counties. The service will operate on a frequency of approximately 90 Mc/s. with ± 75 Kc/s. deviation.

It is felt that there will be a number of radio enthusiasts ready to explore this new field of V.H.F. high fidelity transmission, and the publishers of ELECTRONIC ENGINEERING have accordingly invited Dr. K. R. Sturley to design a suitable receiver embodying the latest practice and capable of being constructed by the amateur as well as the professional.

A booklet describing the design and constructional features of the receiver is now in hand and it is hoped that it will be available for Radiolympia. The prototype model, which is shown in the photograph, will also be on exhibition at the ELECTRONIC ENGINEERING stand (No. 155), in the Electronics section.

For those wishing to study the theory of frequency modulation, copies of Dr. Sturley's Monograph are obtainable from the Circulation Department at the offices of this journal.



NOTES FROM THE INDUSTRY

Radiolympia 1949—Services and Government Departments to Exhibit

Exhibits of the various Government Departments will be among the most interesting to be shown at the 16th National Radio Exhibition held in London between September 28 and October 8. The Royal Navy, Army and the Royal Air Force will be showing, although strict secrecy has so far been maintained regarding details.

Among the Government sponsored exhibits will be those of the four establishments of the Ministry of Supply. The Radar Research and Development Establishment will have a supersonic experimental tank to show how radar works, and a high velocity measurement device. The "Sonde" radio-controlled balloon will be shown by the Telecommunications Research Establishment. This balloon ascends to 89,000 feet, measuring wind velocity, temperature and humidity.

Great interest is expected to attend one of the exhibits of the Department of Scientific and Industrial Research, who are showing at Radiolympia for the first time. This will be the location of storms by radio. Plotting of storm areas will be carried out in public view, based on readings received direct from Cornwall, Northern Ireland, Scotland and Dunstable. The measurement of the height and density of ionisation, for forecasting the most favourable frequencies to be used in radio communications and broadcasting, will also be demonstrated.

A working model, 10 ft. by 4 ft., of the main runway at London Airport equipped with all the navigational aids, including the new "bar and line" artificial horizon lighting system, will be shown by the Ministry of Civil Aviation to demonstrate the use of lighting and radar in assisting aircraft to land at night or in bad weather. The realistic display will include a model aircraft taking off from the runway in semi-darkness and making a circuit of the airport before coming in to land. The aircraft will be located by GCA radar at a scale distance of about four miles and its position shown to the public on a control console with position indicators arranged for direct viewing. Visitors will thus be able to share the experience of an airport controller.

H.M.V. Television Demonstration

Six standard H.M.V. console model television receivers were installed in the Royal Hotel, Woburn Place, W.C.1, for a demonstration before a thousand delegates to the Esperanto Congress in London. Dr. Wanda Zamenhof, their honorary president, and daughter of the founder of the language, was appearing in the "Picture Page" feature at Alexandra Palace, and addressed the meeting via the medium of television.

Polytechnic Courses in Telecommunications

Telecommunications courses leading up to Ordinary and Higher National Certificate and City and Guilds of London Institute examinations will begin on Monday, September 26 at the Polytechnic, 309 Regent Street, London, W.1. (LANgham 2020). Enrolment dates are September 21 and 22, from 6 to 8 p.m. Enrolment forms will be circulated to qualified students who attended the previous session, to enable them to enrol by post.

H.Q.-to-Ambulance Radio

The first ambulances in the United Kingdom to be fitted with radio for emergency calls and general control are being used in Guernsey. Since the radio apparatus was installed in five St. John ambulances on the island, many lives have been saved, thanks partly to the speed of working achieved by radio communication. Each of the five ambulances is able to keep in touch with the others or with headquarters, and in the 12 months that the radio ambulances have been operating over 4,000 calls have been made.

Electric power to work the radio apparatus on the ambulances comes from Exide-Ironclad or heavy duty, large capacity Exide batteries. The discharge for the radio apparatus is 25 amps, and in addition to this the batteries feed headlights, interior lighting and heating, and a brass warning bell, as well as the starting and other electrical equipment.

A Modern Home-Built

TELEVISOR

ANNOUNCEMENT

Components

Further components have now been tested and approved for use in the Electronic Engineering Televisor. The full list of such components approved to date now reads:—

Scanning Coils
- (i) Haynes Radio Ltd.
- (ii) Scanco Ltd.
- (iii) Allen Components.

Focus Coil
- (i) A.M.C.
- (ii) Haynes Radio Ltd.
- (iii) Porthminster Engineering Ltd.
- (iv) Scanco Ltd.
- (v) Walpole Norman Engg. Co., Ltd.
- (vi) Allen Components.

Line Output Transformer
- (i) Haynes Radio Ltd.
- (ii) Scanco Ltd.
- (iii) Walpole Norman Engg. Co., Ltd.
- (iv) A.M.C.
- (v) Banner Electric Co., Ltd.
- (vi) Allen Components.

E.H.T. Transformer
- (i) Haynes Radio Ltd.
- (ii) Partridge Transformers Ltd.
- (iii) Varley (Oliver Pell Control Ltd.).
- (iv) Vortexion Ltd.
- (v) Woden Transformer Co., Ltd.
- (vi) Gardners Radio Ltd.
- (vii) Allen Components.

H.T. Transformer
- (i) Gardners Radio Ltd.
- (ii) Partridge Transformers Ltd.
- (iii) Varley (Oliver Pell Control Ltd.).
- (iv) Vortexion Ltd.
- (v) Woden Transformer Co., Ltd.

250mA and 80mA Chokes
- (i) Gardners Radio Ltd.
- (ii) Patridge Transformers Ltd.
- (iii) Varley (Oliver Pell Control Ltd.).
- (iv) Vortexion Ltd.
- (v) Woden Transformer Co., Ltd.

A.M.C. have also submitted a satisfactory chassis, complete with valve holders and coil formers.

Ekco-Ensign South African Development

Ekco-Ensign Electric, Ltd., announce the conclusion of an important agreement with African Incandescent Lamps (Pty.) Ltd., of Johannesburg, which is designed to strengthen still further the company's position in the South African market.

The agreement provides for E. K. Cole, Ltd., to become a substantial shareholders in the South African company, and Ekco-Ensign Electric, Ltd., will contribute plant and technical information required for the manufacture of tungsten and fluorescent lamps, fittings and control gear.

A new factory has recently been acquired at Industria, on the outskirts of Johannesburg, and is already in production.

Instrument Service

To repair damaged or faulty Elliott electrical test equipment, Elliott Brothers (London), Ltd., maintain service depots at London, Birmingham, Manchester and Cardiff. Each is staffed by instrument engineers who are readily available for urgent work.

The service has recently been extended to Continental countries, which can now take advantage of the comprehensive service scheme, ensuring regular, expert attention at a fixed annual charge. Details of the scheme are available on request from Elliott Brothers (London), Ltd., Lewisham, S.E.13.

Ekco Works at Hadleigh

E. K. Cole, Ltd., purchased the Public Hall, Hadleigh (near Southend), in 1946 for immediate use as a stores and for subsequent use as a radio assembly factory when circumstances should warrant this. Due to the increasing television production being undertaken at the main Southend factory, it is planned, in the near future, to assemble some radio sets in the Hadleigh premises for the home and export markets. While only a few local employees will be required in the first instance, it is hoped that ultimately this factory will absorb over a hundred employees from the locality.

CORRESPONDENCE

Leakage Inductance

DEAR SIR,—In the article on "Leakage Inductance" published in issue No. 254 of April, 1949, I find something which does not agree with my understanding of that problem.

It is quite right that the maximum leakage flux density, B in Fig. 2 is half that of B in Fig. 1. And also that the total leakage inductance due to winding space will be one-quarter of the previous case. But the leakage inductance due to the space between windings is proportional to the product of leakage flux and number of turns.

The leakage flux in each space between windings has half its original value and the number of turns linked with it also has half its original value. The leakage inductance due to c is then one-quarter of its original value.

There are now two such spaces, and the total component of leakage inductance due to spaces will be one-half the previous case if each space is equal to the single space in the original case.

Thus the whole expression for leakage inductance must be written

$$L = \frac{k.d}{b.N}\left(\frac{a}{3} + N.c\right)$$

instead of Formula (1). $N.c$ represents the total space between windings.

Yours faithfully,

FRANTZ NIELSON

Jyllandsgade 4, Denmark

Mr. Crowhurst replies:

I am thankful to Mr. Nielson for calling attention to the error in formula (1) of my article on leakage inductance. The correct expression becomes.

$$L = \frac{k.d}{b}\left(\frac{a}{3N^2} + \frac{c}{N}\right)T^2$$

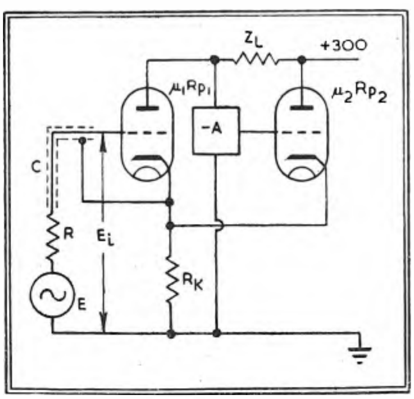
I do not agree that it would be simpler by writing

$$L = \frac{k.d}{b.N^2}\left(\frac{a}{3} + N.c\right)T^2$$

as Mr. Nielson suggests, to take N.c as the total insulation space between windings. A careful consideration of the table of winding arrangements on page 132 will show that this will only be true for about half of the arrangements shown, i.e., one each of those shown against values of N^2 listed as 4, 9, 16, 25 and 36, and those shown against 49, 64, 81, 100, 121 and 144. This is due to the fact that the full maximum leakage flux density does not appear at all the spaces in the other arrangements shown.

In practice the modification made by introducing this correction will only affect designs where mixing arrangements are employed. It is generally sufficient in these circumstances to ensure that the value of leakage inductance is small enough to satisfy external circuit requirements. As the prediction given by using the formula given in the article, and upon which the use of the chart is based, will give a value of inductance that is slightly greater than the true value, the error is seen to be on the safe side. However, if it is desired to use the chart for more accurate computations, the simplest method will be to divide the actual dimension c by the square root of the value of N^2 used at the bottom of the chart, before adding it in the space at the middle of the chart, as shown at figure 7.

For cases where it is desirable to produce a specific value of leakage inductance, winding arrangements of the types represented in figures 4 and 5 of the article are used, which the modification does not affect.



D.G. Lampard: Generalised Circuit.

Reducing the Effect of Capacitance in a Screened Cable

DEAR SIR,—With reference to your recent article on "Reducing the Effect of Capacitance in Screened Cable" by V. H. Attree, March, 1949, it may be of interest to you to know that we have used this system for about eighteen months now for coupling a P.E. cell to its pre-amplifier.

However, by using a special cathode follower circuit we have been able to extend the length of the coaxial coupling cable (UR 1) to 50 ft. or more using a 1 Megohm P.E. cell load resistor. The measured response was less than 0.5 db. down at 20 Kc/s. An outer earthed sheath is not required as the output resistance of our cathode follower circuit is approximately 1 ohm.

Referring to the generalised circuit it may be shown that:—

$$m = \frac{E_k}{E_i} = \frac{\mu_1}{Rp_1 + Z_L}\left(1 + \frac{\mu_2 AZ_L}{Rp_2}\right) \div \frac{\mu_1 + 1}{Rp_1 + Z_L}\left(1 + \frac{\mu_2 AZ_L}{Rp_2}\right) + \frac{\mu_2 + 1}{Rp_2} + \frac{1}{R_k}$$

If the first valve is a pentode and $\mu_2 >> 1$ then:—

$$m = \frac{E_k}{E_i} \approx 1 - \frac{1}{g_m AZ_L}$$

where g_m is the mutual conductance of valve 1.

Similarly the output resistance R_o is:—

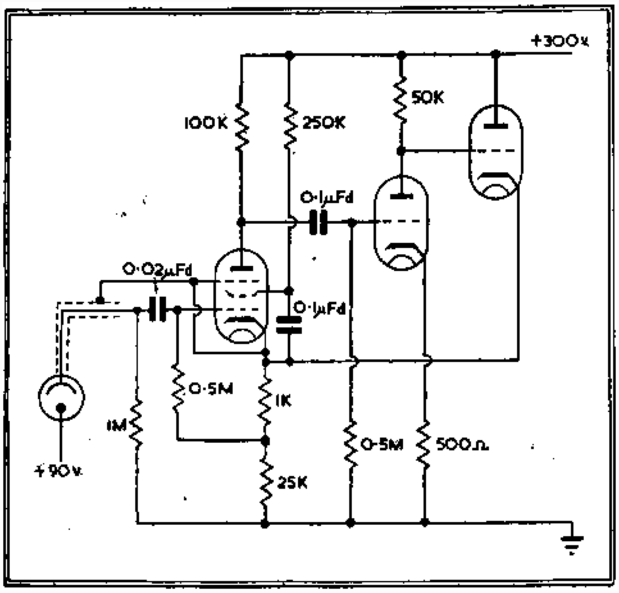
$$R_o = \frac{1}{g_m\left(1 + \frac{\mu_2 AZ_L}{Rp_2}\right) + \frac{\mu_2 + 1}{Rp_2}}$$

Using the values shown in the second circuit we find:—

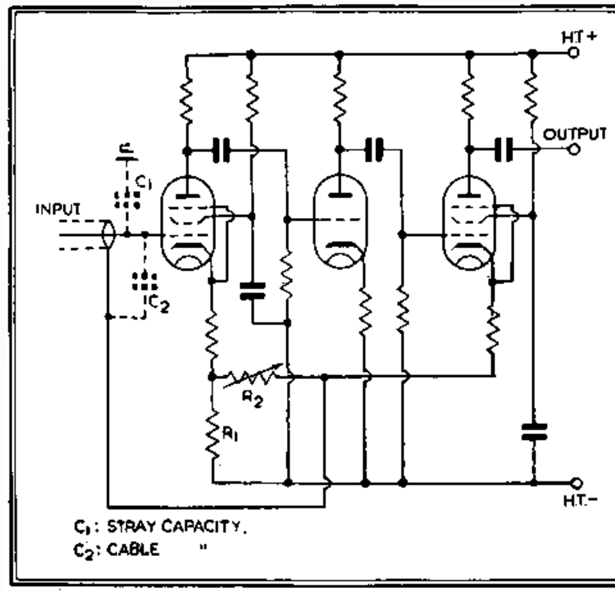
$m = 0.998$ $\qquad R_o = 1.1\Omega$

Now it will be seen that even if Z_L or A is complex (as could be caused by stray capacitance) the resultant phase angle of m is in general very small. Thus it is justifiable to assume that m is purely real. The effective capacitance of the cable is given by:—

$$C_e = C(1 - m)$$



D. G. Lampard: Detailed Circuit.



Circuit referred to in letter from D. W. Boston *(below)*.

but

$$1 - m \approx \frac{1}{g_m A Z_L}$$

and hence for the typical values given above the cable capacitance has been reduced by a factor of $\frac{1}{500}$

The frequency at which the response is 3 db. down is:—

$$f_{3db} = \frac{1}{2\pi RC(1-m)}$$

For a length of 50 ft. of UR1 cable (about 20 pF./ft.) and a P.E. cell load resistor of 1 Megohm we find using the above values that f_{3db} down $\approx$ 80 Kc/s.

Yours faithfully,
D. G. LAMPARD.
University Film Society, Sydney.

DEAR SIR,—In the March issue of ELECTRONIC ENGINEERING there was an article by V. H. Attree in which a circuit for reducing the capacitance of a screened cable by connecting the screen to the cathode of a cathode follower is described and analysed. In the May issue he discusses a circuit for neutralising the effect of stray capacitances on an input circuit. Readers may be interested in a new circuit which combines the advantages of both methods and provides a screened input cable with a very high shunt impedance at audio frequencies.

The simple cathode follower circuit cannot completely remove the effect of input capacitance since the A.C. potential at the cathode is always less than the applied signal; and though a close approximation can be made by using a high cathode impedance, this is shunted at high frequencies by the capacitance of the screen to earth. By means of an amplifier it is possible to obtain a potential identical to the input potential and develop this potential across an impedance which may be relatively low. If a high ratio of feedback be used the arrangement is stable and consistent in performance. Such an arrangement is shown in the Figure. Feedback is developed across R_1 and the potential at the point P may be adjusted by R_2 to give a voltage identical to the input voltage. In this case the effect of the cable capacitance can be completely eliminated. A further slight increase in R_2 can be used to compensate for the effect of stray capacitances; the cable capacitance now acting as the neutralising capacitor as described by Mr. Attree in the May issue.

The new circuit has the advantage that a very high input impedance is obtainable and that the impedance of the screen to earth can be made relatively low. It should be added with regard to all these circuits that they may be considered as circuits with negative feedback, and that the action of the input capacitance is to remove the negative feedback so that the resulting frequency response is unaffected. The removal of the feedback by the shunt capacitance at high frequencies will, of course, increase the effect of valve noise at these frequencies so that where a high ratio of signal to noise is desired it will still be necessary to keep the cable capacitance small.

Yours faithfully,
D. W. BOSTON.
British Acoustic Films, Ltd.

Printed Circuits

DEAR SIR,—We would refer to the letter from Mr. P. P. Hopf in your July issue regarding the publication "New Advances on Printed Circuits" by the National Bureau of Standards of the United States Department of Commerce and should like to point out that this Institution in its library has copies of the national standards issued by all oversea countries; we are able to offer these on sale and can also easily obtain on demand copies of any other publications by these national bodies.

The value of these services to the exporting industries in this country will be readily appreciated and we should be grateful if you can draw it to the attention of your readers.

Yours faithfully,
L. G. WATKINS
Public Relations Officer,
British Standards Institution