Electronic Engineering

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

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

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

Vol. 22. No. 264 **February, 1950**

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Retrospect

THE year which has just passed is unlikely to go down to history as a notable year, and in attempting a review of 1949 it would be excusable if our opinion were prejudiced by the impact of 1949 upon us as private citizens.

The private citizen will undoubtedly look back on 1949 as another year of austerity, exhortation and political wrangling and he is not likely to forget the discomforts—in many cases, acute discomforts—caused by strikes and go-slow tactics.

However, we feel no such pessimism about our own industry and 1949 can be fairly represented as a year of achievement.

Judged from whatever angle, it has been a successful year, and since success nowadays is measured in terms of export figures, we may reasonably claim to have contributed our share towards Britain's economic recovery as the following table will show:—

	Total value of exports Jan. to Oct., 1949
Radio Sets, Radiograms and Chassis ...	£2,613,000
Radio Components, Testing Equipment, Sound Reproducing Equipment, and Electro-Medical Apparatus	£3,619,000
Transmitters, Navigational Aids, General Communication Equipment, and Industrial Electronic Apparatus	£1,686,000
Valves and Cathode Ray Tubes	£1,600,000
Total for Jan.-Oct., 1949	£9,518,000

This table, which is based on figures supplied by the Radio Industry Council, gives the total export figure for the first ten months of 1949 for the various categories shown. The figures for November and December are not yet available, but, assuming that exports continue at the same rate, the figure for 1949 should fall not far short of £12,000,000, compared with £11,897,000 and £10,272,000 for 1948 and 1947.

From the television point of view, 1949 has been a good year. According to the latest figures, again supplied by the R.I.C., the total production of television receivers in this country for 1949 was 205,500 with a steadily increasing rate of production throughout the year. During December alone the production was over 36,000.

A glance at the following table of television production since the war is very encouraging, even compared with the astronomical American figures.

	1946	1947	1948	1949
Annual production of Television Receivers ...	6,500	28,200	90,800	205,000

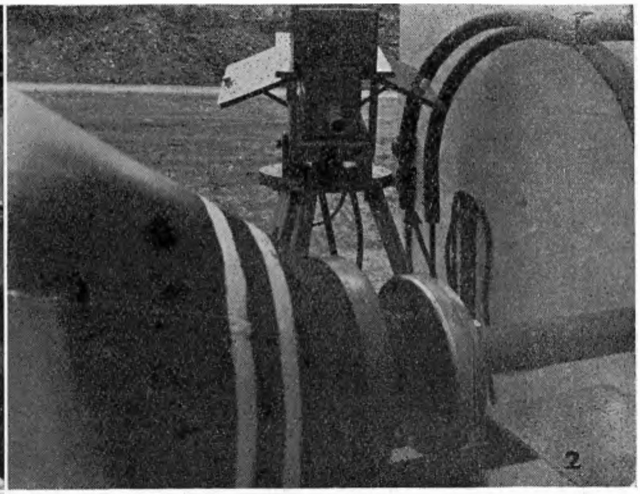
Production has, of course, been stimulated by the opening of the B.B.C.'s second television transmitter at Sutton Coldfield on December 17, which will be followed by a third transmitter during 1951 according to the latest proposals.

Ignoring the initial teething troubles and minor breakdowns which in our opinion have received far too much attention from the Press, Sutton Coldfield has done all that was expected of it, and its range has so far exceeded the B.B.C.'s forecasts that plans for the third and remaining transmitters may have to be modified.

The Sutton Coldfield transmitter and the London-Birmingham Radio Relay Link, both of which we have briefly referred to in earlier issues, are first class examples of all that is best in British engineering practice, and, when the accompanying coaxial cable link is installed as well, Britain may well claim to have an unrivalled system.

1949 has also been a year of exhibitions, the most important of which was Radiolympia. Viewed from this distance it appears to have been a most successful one with attendance figures not quite equalling the 1947 record, but how far attendances were influenced by the delay in opening caused by the strike of the stand fitters is difficult to say.

If 1949 was, from our point of view, a year of too many exhibitions, it is at least indicative of the growth of our industry and of the manner in which it is penetrating into other industries. Apart from the normal exhibitions which have been reviewed in these pages over the last year, new exhibitions at which electronic apparatus has made its appearance have been the British Industries Fair and the Society of British Aircraft Constructors.



INDUSTRIAL TELEVISION

The U.S. Ordnance Department has adopted the " Vericon " system of closed-circuit television for remotely supervising the dismantling of dangerous explosive items. A viewing screen is placed in the reinforced concrete control point, from where the dismantling tools are manipulated.

A single camera unit (weight $31\frac{1}{2}$ lb.) can transmit to as many as ten separate receivers. An extension viewer (weight $42\frac{3}{4}$ lb.) may be a mile away from the master viewer, which in turn may be up to 1,000 feet from the camera. The power unit (weight 49 lb.) may be as much as 100 feet from the camera. All units are portable and can be operated by unskilled users.

1. Operator manipulating defusin. controls. Note pulse power unit at left, also in protective shelter.
2. Close-up of coupling wrench, which is also seen in Fig. 3.
3. Heavy bomb is slid into position before Vericon camera prior to defusing.
4. The camera is light and portable, can be mounted in any position and does not require specially trained operators.
5. Extension viewer with 42 square inch screen. As many as ten of these may be wired through the master viewer.

Electronic Instruments in Diagnostic Medicine

Part I

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

IN a previous paper[1] the author has reviewed the application of electronic forms of instrumentation to dosage measurements in one particular branch of therapeutic medicine, radiotherapy. Continuing the survey of electronic aids in medicine, this article will deal with a much wider field, that of diagnostic medicine. The main fields for electronic development of diagnostic instruments are those connected with the measurement of phenomena associated with the nervous, cardio-vascular and respiratory systems. This includes electroencephalography (E.E.G.), muscle and nerve excitation and electromyography (E.M.G.), electrocardiography (E.C.G.), and auscultation. Certain measurements on physico-chemical and biochemical properties of biological material which are necessary for clinical diagnosis also lend themselves to electronic methods.

Electroencephalography

(1) General Principles

The history of encephalography goes back to the discovery in 1874 by Caton that there is electrical activity in the brains of living animals. Other workers during the same century substantiated his finding and related this activity to central nervous function. All early work in this field was done with capillary electrometers and moving coil galvanometers. In 1906 the Einthoven string galvanometer became available and was a great improvement on previous types of recording instruments. By 1934, Berger had the following findings to his credit: (a) the brain of man has an electrical beat which comes from the nerve cells and not from blood vessels or connective tissue; (b) this beat changes with age, sensory stimulation, disease, and various changes in the physiochemical state of the body; (c) the beat is a mixture of more or less sinusoidal fluctuations in voltage with a frequency of 0.3-70 c/s. and a dominant rhythm of approximately 10 c/s.; (d) the maximum trough-to-peak voltage in normal adults while awake is approximately 100 μV, and the usual waking adult record has an amplitude of 10-50 μV.

Thermionic valve amplification immediately suggested itself for this work. Any apparatus which is used to record brain potentials satisfactorily should possess the following properties: (a) it should be able to handle signals having a wide range of amplitudes and frequencies (as indicated in the previous paragraph) without distortion; (b) the signal should be amplified sufficiently so that it can be recorded on a relatively insensitive inkwriting recorder; (c) the signal to be recorded is between two points on the scalp rather than between one point and an arbitrary earth point, and hence the apparatus should amplify only those signals between the two points (out of phase, or differential signals), and not those appearing equally between each point on the scalp and earth (in-phase signals), e.g., mains interference; (d) the accuracy of the apparatus should be unaffected by variations in operating voltages or valve characteristics. A more complete survey of these special requirements for E.E.G. is given by Parr and Grey Walter.[2]

(2) Choice of Amplifier Circuit

Bearing in mind the requirements (a) and (b) stated in the preceding paragraph the use of a multistage push-pull circuit is immediately indicated. Push-pull circuits can handle signals of greater amplitude than can a single cascade amplifier using identical operating voltages. Matthews[3] described such an amplifier in 1934. The push-pull circuit will discriminate between in-phase and out-of-phase signals if a common cathode resistor is employed, as was shown by Offner[4] in 1937; a degenerative voltage is developed across the resistor for in-phase, but not out-of-phase signals. By amplifying this degenerative voltage, i.e., by applying the feed-back over more than one stage, the discrimination factor can be increased; thus, requirement (c) can be fulfilled. Another type of differential input circuit has been devised by Toennies,[5] and independently by Matthews.[6] This particular circuit is shown as the input stage of Fig. 1, an E.E.G. amplifier described by Parr and Grey Walter.[2] As can be seen from the diagram, the output of the stage can be fed on to a single-sided amplifier. This type of circuit is remarkably stable to changes in operating voltages and valve characteristics, thus fulfilling requirement (d). A detailed discussion of input circuits for biological amplifiers has been set out by Saunders,[7] who concludes that the Offner-type circuit is preferable to that of Toennies from the point of view of ease of adjustment.

The Offner input circuit must be followed by push-pull stages. For some reason British practice has been to use single-sided stages, and the complete circuit diagram of such an apparatus is shown in Fig. 1. The first two stages are battery operated to improve stability, but the last two are supplied from a conventional stabilised mains power pack. An overall gain of 10^7 can be obtained with the circuit.

Most American E.E.G. equipments have been constructed using push-pull stages throughout. The circuit diagram of one such amplifier is shown in the paper by Parr and Grey-Walter.[2] If a fairly narrow frequency band is of particular interest, resistance-capacitance coupling between stages is quite

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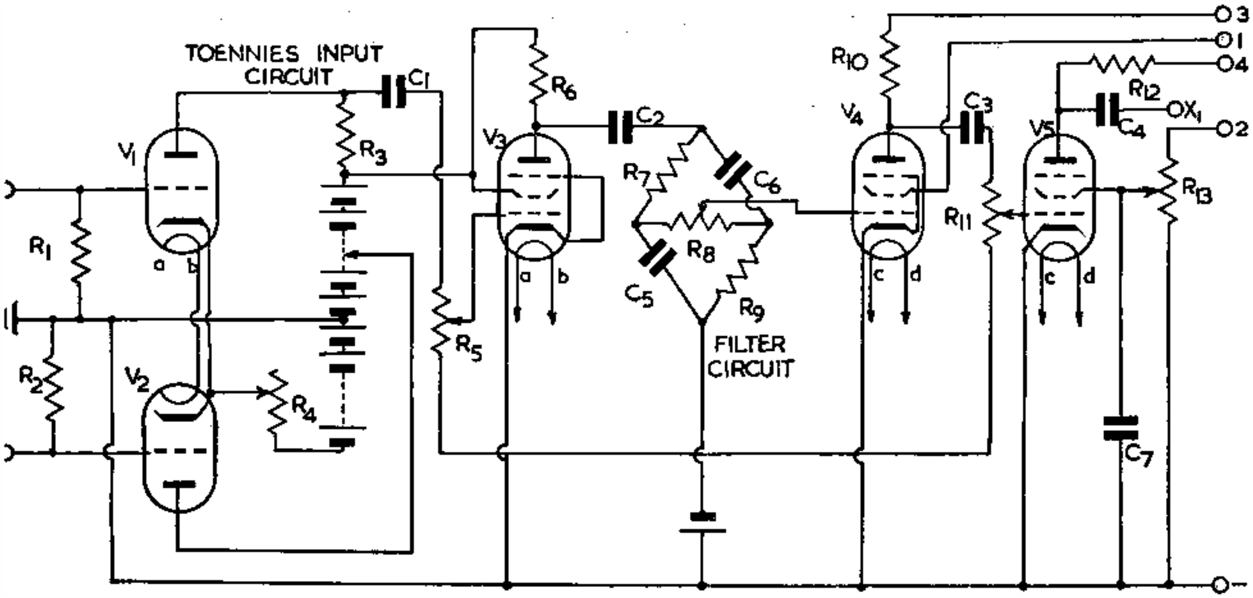
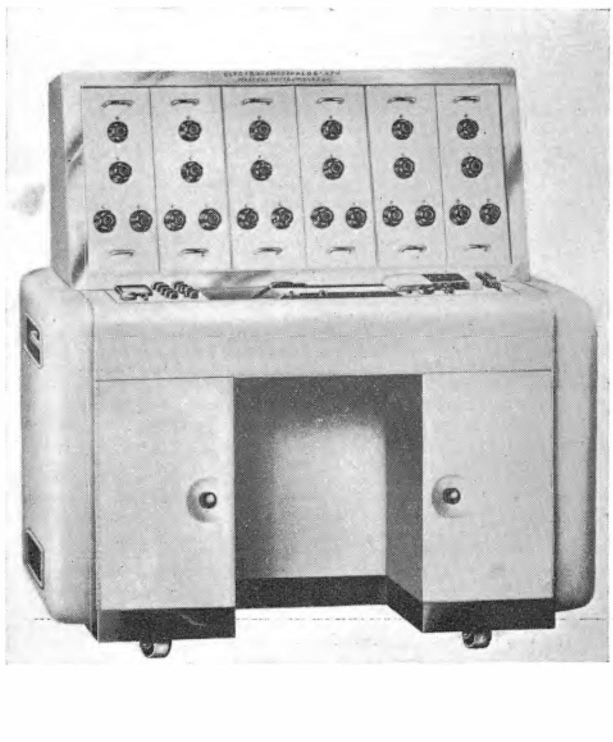


Fig. 1. Amplifier with balanced input used for electroencephalographic recording

satisfactory. This is the usual practice, time constants being about half-a-second. In wide-band investigations a D.C. amplifier is more satisfactory. Such an amplifier is described by Forbes and Grass.[5]

An E.E.G. equipment is manufactured by Marconi's which has three double channel balanced amplifiers and a three-speed recorder. A photograph of the apapratus is shown in Fig. 2. A complete description of the apparatus has been given by Johnston[6] in a paper which includes an excellent bibliography dealing with circuit techniques involved. The amplifier has push-pull stages throughout, and the high discrimination factor of the Offner input circuit is increased by using as the cathode-coupling load the dynamic anode impedance of a pentode. The reason for this increase is that the discrimination factor is increased in proportion to the value of the cathode load; this can be made quite high using a pentode valve, the potential across which is only about 70 volts, while the use of a resistor of the same effective value would necessitate a much larger and inconvenient potential. Spurious output voltages due to space-charge variations and changes induced by heater current changes are reduced to a minimum by using a type 6SC7 valve in the first two stages, which has an indirectly heated cathode common to both triode sections. This type of valve has an under-grid connexion which minimises electromagnetic and electrostatic induction. From the point of view of low-frequency response, the coupled cathode circuit is extremely good, and an overall time-constant of one second is obtained. The small number of capacitance couplings between stages (of the four, two are direct) is also of assistance in producing this relatively long time-constant, without the use of unusually large coupling capacitors and grid leaks. A disadvantage of amplifiers with long time-constants is their tendency to "block" for a long period if an excessive input is applied or an input lead is disconnected. In order to overcome this, arrangements are made so that when an adjustment is made an *On/Off* switch turned to *Off* will short circuit the grids of the fourth and third stages, thus dispersing the blocking impulse rapidly. A simplified basic circuit diagram is shown in Fig. 3.

Fig. 2. Marconi electroencephalograph, Type OA180A



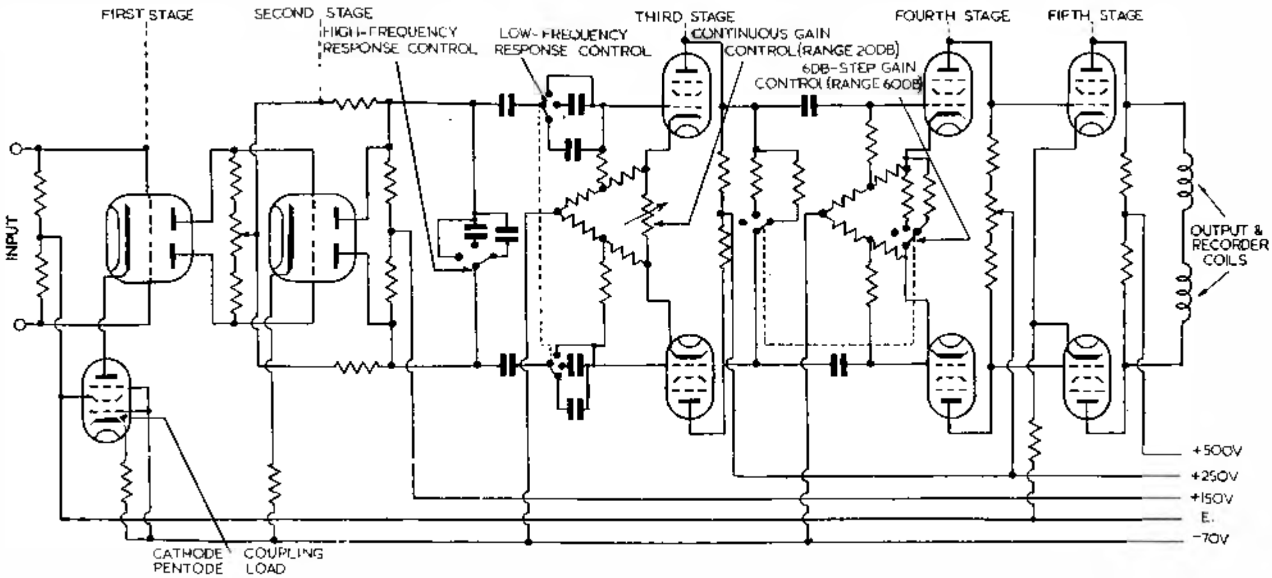


Fig. 3. Basic circuit of Marconi E.E.G. amplifier (Johnston differential type)

(3) Electrical Noise

The amount of useful amplification of any apparatus depends upon the smallest potential change that can be detected; this in itself obviously depends upon the magnitude of spurious background signals or "noise." As the subject of noise is of importance in the design of all biological amplifiers it is worth while discussing it fully at this stage. The sources of noise will be first enumerated, and relations giving the magnitude of the noise voltage stated. It is convenient to quote the equivalent noise R.M.S. voltage at the input to the amplifier which would give rise to the observed noise, because this value can then be compared directly with the input pulse. It may be assumed that a pulse of peak value V, equal to the R.M.S. noise voltage can just be observed. It can also be shown that a pulse four or five times this size is necessary to ensure comparative immunity from spurious measurements due to occasional peaks of noise.

(a) *Resistance Noise.* This may be troublesome in the input circuit and it arises from thermal agitation of the electrons and may be accurately calculated using the formula.[10]

$V = 1.28 \times 10^{-10} \sqrt{R.\Delta f}$ at room temperature where Δf is the bandwidth, and R circuit resistance.

(b) *Anode Current Shot Noise.*[11] This arises from the statistical fluctuations at any moment in the number of electrons constituting the anode current. The noise can be represented by an equivalent noise resistance R_E given by K/g_m where K for triodes of mutual conductance g_m is of the order of 2.5. The noise voltage can then be calculated as above. The use of a screen grid valve in the first stage introduces further shot noise due to random variation in the partitioning of the cathode current between the anode and the screen.

(c) *Grid Current Shot Noise.* Random fluctuations in grid current may, in high impedance grid circuits, give rise to a noise voltage in excess of any other. The shot current noise is given by $i^2 = 2ei_g\Delta f$, where e is the electronic charge, Δf the bandwidth, and i_g the numerical sum of electron and positive ion grid currents which may be guessed from the shape of the grid current curve (N.B. the measured grid current is the algebraic sum of the two components, and thus is less than the numerical sum, which actually contributes to noise). If the input circuit through which the noise current flows is capacitative over the bandwidth of the succeeding amplifier, the noise voltage is given by:

$$V^2 = [2ei_g/4\pi^2C^2]\ [1/f_1 - 1/f_2]$$

where C is the input capacitance and f_1, f_2 the frequency limits of the amplifier. V will be reduced if the capacitance is shunted by a grid leak of resistance R to:

$$V^2 = 2ei_gR^2\ (f_2 - f_1)/[1 + 4\pi^2 f_1 f_2 C^2 R^2]$$

(d) *Flicker Noise.*[12,13,14] A further source of noise arises in the constantly changing composition of the emitting surface within the valve. For a number of valve types it has been found that $V^2 = 0.1 \log_e f_2/f_1$ where V is microvolts and f_2 and f_1 are again the upper and lower frequency limits of the amplifier bandwidth respectively.

Consider an amplifier with the following circuit constants:

Bandwidth 1 – 100 c/s.
Input valve, g_m = 1 mA/volt.
Input capacitance = 30 pF.
Input circuit resistance = 1 MΩ.
Arithmetic grid current = 6×10^{-10} amps.

The separate noise voltages are as follows:

Anode shot noise, Nil
Grid shot noise 0.13 μV
Flicker noise 0.7 μV
Resistance noise 1.3 μV
Total noise 1.4 μV.

As is shown in the above example, for amplifiers having a bandwidth from 1-100 c/s. with input circuits as commonly used in biological work, resistance

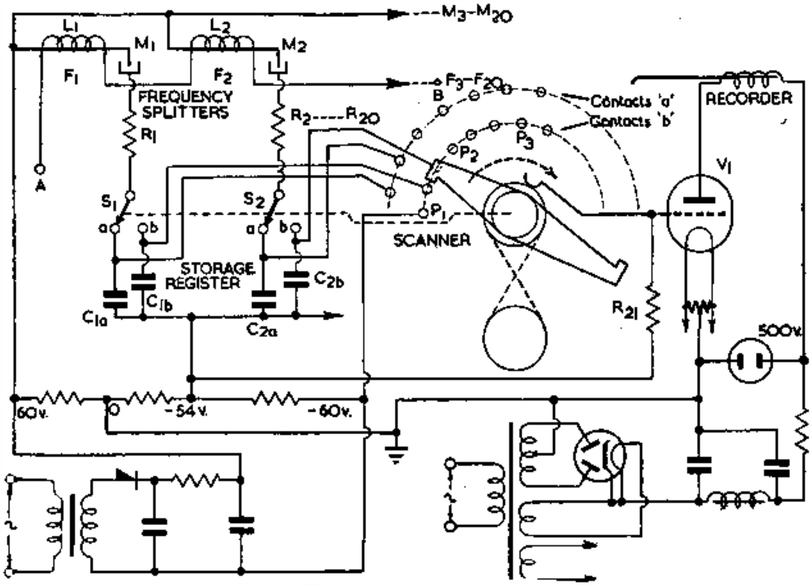


Fig. 4. Circuit of tuned-reed analyser, showing change-over switch and scanning device

noise predominates. If this can be reduced without reduction of signal, flicker noise will contribute most of the total. Johnston[9] discusses these sources of noise in his amplifier. His choice of a 6SC7 valve for the input circuit has already been mentioned; however, he also points out that only a proportion of these valves is suitable, and that initial ageing under operating conditions must be carried out to ensure optimum performance. Further reductions in flicker noise may perhaps be achieved if improvements in cathode coating textures can be made. If the biologist requires to measure potentials with frequencies of the order of 1 per minute, then development on these lines is essential.

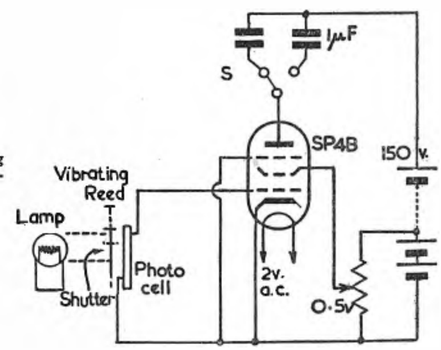
E.E.G. Frequency Analysis

The value of the electroencephalogram in correlating brain activity with changes in central nervous function lies in the analysis of its frequency components. The significance of the record is obtained from a subjective evaluation of its form, amplitude and the most dominant frequency; which is described only as slow, moderate or fast. It is desirable to supplement such a subjective procedure by more quantitative objective methods, and to describe the curves in terms of numerical physical factors such as the amplitude and frequency of the underlying components.

Analytical procedures on E.E.G.'s have had only limited success, and mathematical or graphical methods are tedious. Gibbs and Grass[15] have devised an ingenious procedure for stepping up the frequencies so that they can be treated by a standard audio-analyser; it has the disadvantage that a small section of the record must be arbitrarily selected for analysis. Harmonic analysers are either inaccurate, or while useful at audio-frequencies are inapplicable in the range 2-25 c/s. A number of workers have used tuned recorders,[16,17,18] each tuned element responding to a particular frequency. Among these is an instrument described by Grey-Walter[16] which consists essentially of a bank of tuned reeds which act as frequency splitters, each reed being tuned to a frequency in the band to be studied. Each reed is provided with a fine steel contact wire which dips in and out of a mercury cup when the reed vibrates, but is just out of the mercury when the reed is at rest. A high resistance, a source of E.M.F. and a capacitor are in series with this mercury reed-contact. The capacitor is charged up to a potential which is a function of the total duration of the contact time, and therefore of the amount of energy at the reed frequency during the specified time. An amplifier is connected to each capacitor in turn by a motor-driven rotary switch, and this amplifier works a wide arc recording pen across the recording paper on which the original encephalogram is at the same time being traced. The summation epoch is chosen to be 10 seconds, so that each 10-second stretch of record has traced over it a histogram of its spectrum. The analysis is performed automatically every 10 seconds. A circuit diagram of the arrangement is shown in Fig. 4.

The weak point in the above instrument was the use of a mercury contact and high resistance as the charging circuit for the storage circuit. A modification described by Grey-Walter[19] uses photo-cells as a medium for integrating the energy of the frequency

Fig. 5. Integrating circuit using photo-cell and valve



components. He arranged a barrier layer cell in combination with a pentode so that the properties of a vacuum photo-cell were obtained, i.e., the cell cut off sharply in the dark, and gave rise to a current proportional to the illumination over a wide range. A shutter attached to the vibrating reed then controlled the current to the storage capacitor. The arrangement is shown in Fig. 5. Fig. 6 shows a photograph of the commercially produced instrument, described by Baldock and Grey-Walter.[20]

An ingenious analyser described by Barbour[21] uses the heterodyne principle with a fixed filter, the variation in the frequency being achieved by control of the frequency of an oscillator whose output is mixed with the signal before filtering. This allows a filter of higher resonant frequency to be used,

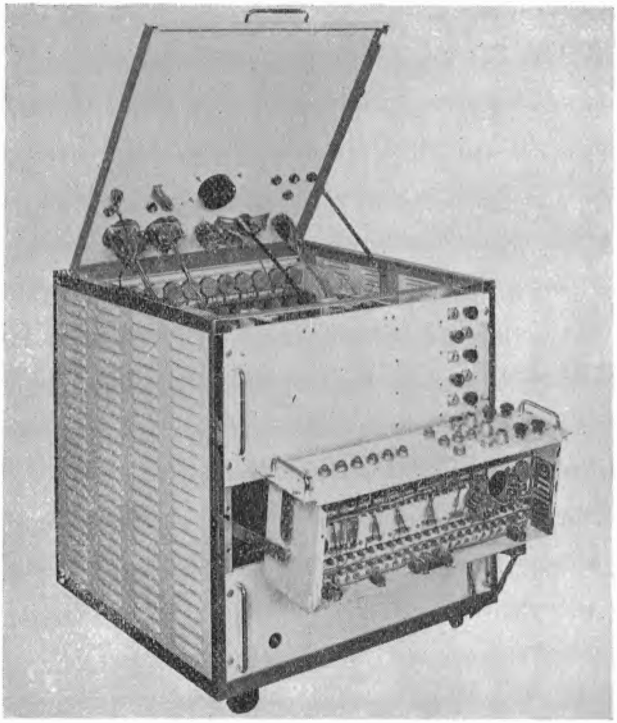


Fig. 6. Electronic analyser made by the Edison Swan Electric Co. Pre-set controls can be seen at the rear of the cabinet. The main relay and averager unit is shown withdrawn

though a greater relative sharpness is required. Again, a vibrating bar serves as a tuned filter element. One advantage which this circuit possesses is that the value of any frequency component is measured in a time short compared with the variations in the amplitude of that component. A block schematic of Barbour's analyser is shown in Fig. 7.

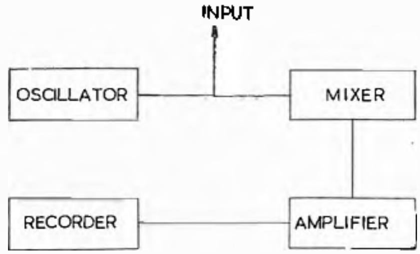


Fig. 7. Diagrammatic circuit of Barbour's frequency analyser

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(To be continued)

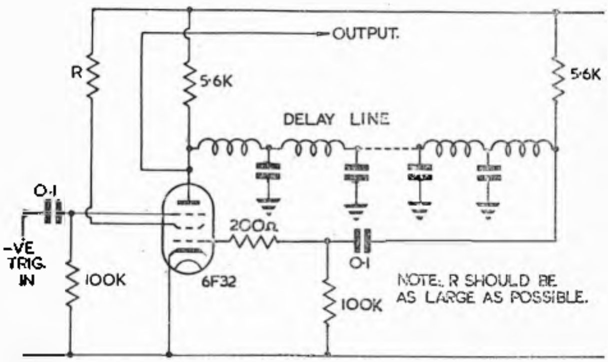
Time Calibrator for C.R.T. Traces

By C. H. BANTHORPE*

IN many uses of the C.R.T. it is desirable to be able to determine the time interval between points on the trace. The circuit described has been used and found useful by the writer and may be of interest to others.

A common method is to apply marker "pips" to the trace either as blackout dots or as a deflexion; in either case the oscillator which generates the pips must be phase locked to the trace. If the trace recurrence frequency is not a sub-multiple of the pip frequency (a most unusual condition), the pips oscillator has to be started at the beginning of the trace, and stopped at the end. The oscillator used in the circuit described here uses a single valve only and fulfils all the requirements.

It will be seen to consist of a valve, in the anode of which there is a delay line. When the valve is switched on by its suppressor grid, the anode takes current, and the anode potential falls. This negative front travels along the delay line and eventually reaches the control grid and biases off the valve. The anode then ceases to take current, its potential rises, and this positive front travels along the delay line and eventually switches on the valve. This cycle of events is repeated until the suppressor is biased off, when the oscillator will at once cease to function.



It will be seen that it starts instantly, generates square waves of constant amplitude which may be differentiated to provide pips, and will stop instantly.

The triggering waveform which is applied to the suppressor can be obtained by differentiating the sawtooth voltage waveform used for the C.R.T. deflexion, or from the brightening pulse of the C.R.T.

One interesting application of this circuit is the generation of vertical bars in a television waveform generator. Negative line sync. pulses are applied to the suppressor, and during the line scan, square waves of suitable duration are generated and mixed with sync. pulses.

* Central Equipment Ltd.

A D.C. Amplifier Using an Electrometer Valve

By D. H. PEIRSON, B.Sc., A.Inst.P.

RECENT reports indicate that a departure from orthodox D.C. amplifier technique to one that uses some form of mechanical modulation, such as a "chopper[1]" or varying capacitor,[2] can produce great improvements in amplifier stability. In this paper, however, only conventional methods of D.C. amplifier design are considered.

The limit to the accuracy of measurement of small currents from ionisation chambers is set by the stability of the D.C. amplifier used to amplify the voltage across the chamber load. The stability or freedom from "drift" of such an amplifier is dependent upon the voltage stability of the power supply and the temperature stability of the immediate surroundings.

Various known methods of reducing drift have been considered and the performances compared. In order to reduce the amplifier controls to the minimum required for zero-balancing, the present approach to the problem discards any means of circuit compensation for supply voltage variation and relies upon a highly stable power pack. Under these conditions external temperature variation exerts a limiting effect upon amplifier stability.

Causes of Drift

Much effort has been directed towards eliminating objectionable zero-shift in valves used as D.C. amplifiers. This zero-shift, more commonly known as drift, is caused by variations in the supply voltage and in the temperature of the immediate surroundings. Changes in H.T. voltage at the anode may be referred to an equivalent change at the grid; changes in L.T. voltage and external temperature produce a change in space current which arises from changes in contact potential or cathode emission. The effect of these variations may be expressed as an equivalent change in grid or input voltage.

A form of discontinuous zero-shift in amplifiers is caused by mechanical disturbances such as vibration or shock. This phenomenon may be attributed to an irreversible shift or creep in the electrode structure and except in extreme cases is easily avoided by suitable flexible mountings and soundproofing.

The application of D.C. amplifiers to the measurement of small ionisation chamber currents requires that input grid current variations should be inappreciable compared with the currents to be measured. The achievement of sufficiently small grid currents ($10^{-13} - 10^{-15}$ A) inovlves use of the electrometer type valve operating at reduced electrode potentials and emission.[3] This restriction complicates the maintenance of stable conditions within the valve because of the reduced space charge and of the increased importance of contact potential variations in relation to the low electrode potentials.

Methods of Reducing Drift

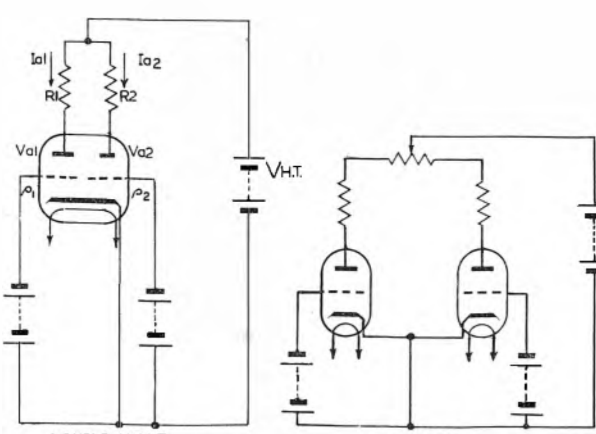
As there are numerous methods devised to minimise drift a complete account[4] of work in this field will not be attempted. It is possible to classify most of the methods under two headings: (a) those which compensate for drift by applying correction through an auxiliary electrode or circuit and (b) those which balance in a bridge circuit the drift effect in the amplifying valve against effects due to a similar component in an opposing arm.

Compensation methods of reducing drift in D.C. amplifiers have been discussed in some detail in an unpublished paper, but for the present purpose the principle of compensation was rejected in favour of the balanced bridge method, for reasons developed below.

Balanced Amplifiers

The basic bridge circuit illustrating the principle of balancing drift effects is shown in Fig. 1. Assuming the two valves remain identical in characteristics, balance may be maintained over a wide range of supply voltages. Potentially the balanced pair of valves provides an inherently stable arrangement: unfortunately it is difficult practically to produce great similarity in the pair without great care in manufacture and selection. Additional compensation may be applied by circuit adjustment, the conditions for which have been analysed by Wynn-Williams and by Nottingham,[5] but the improvement thus gained is off-set by the increased complexity of setting-up and initial adjustment, particularly as the activity of an oxide-coated emitter tends to vary with changes during adjustment of electrode potentials and emitter temperature. Thus

Fig. 1. Balanced Circuit



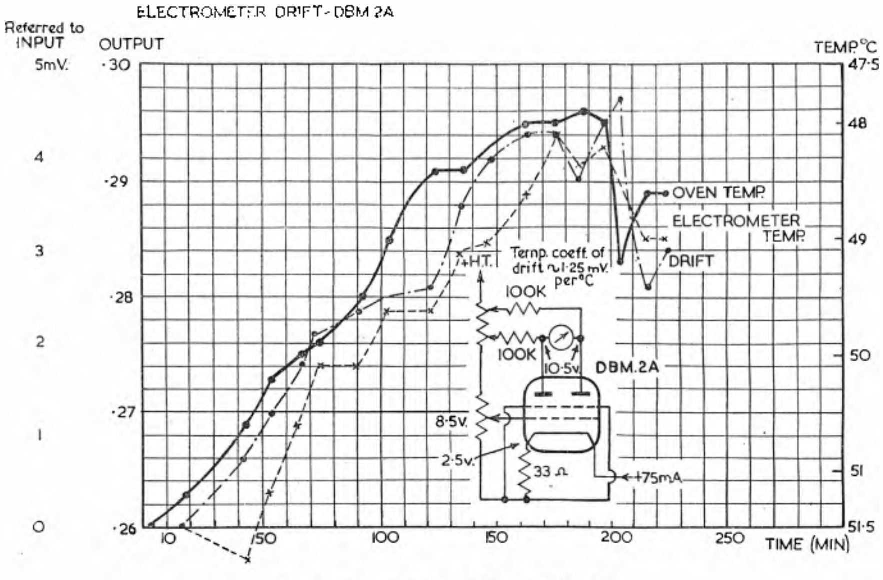


Fig. 2. Temperature Curve—DBM2A

the split FP-54 double electrometer valve (a logical development of the balanced pair), described by Yafferty and Kingdon[6] and which has a reported long-term drift of less than 1 mV when operated from batteries, requires considerable circuit adjustment and manipulation before use.

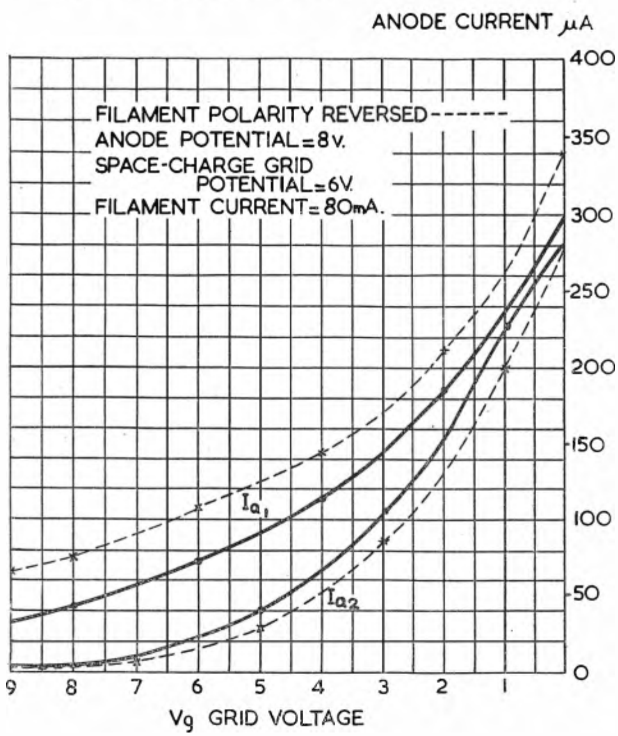
Present Approach

For certain ionisation chamber measurements a long-term drift of not greater than 1 mV seemed both desirable and possible. It was felt that the most promising and convenient means of reducing long-term drift was provided by the double valve balanced bridge circuit, since the symmetry of the arrangement promised stability with a minimum of complexity and adjustment. The Ferranti double tetrode electrometer valve was available for this purpose. An early version (type DBM2A)[7] operates from a common filament and space charge grid, an arrangement that will equalise emission variations due to supply or temperature changes in each section of the valve. In addition, the proximity to one another of the two sections will reduce the effect of stray interfering fields. A grid current of 10^{-15} A may be achieved by operating at suitably reduced electrode potentials and filament current at high grid insulation, and within a light-tight box. In a double valve bridge circuit the effect of grid current upon drift depends upon the differential variations in current at the two grids. If the two sections of the valve operate from a common cathode, these variations should be considerably smaller than in the single valve. Since the gain of the electrometer stage is usually less than unity, the effect of drifts in the following amplifying stage is important. The valve used in this and the succeeding stage is the twin-triode common cathode 12SC7, having a rated heater current (150 mA) low enough for convenient stabilisation. The Osram VX3088, having a 19V, 100 mA heater, has proved a suitable alternative to the American 12SC7. Since it may be necessary to operate the amplifier continuously for days and weeks, the presence of common filament or cathode should tend to equalise, in the two sections of the several valves, slow emission changes due to ageing.

Drifts due to Supply Variations

Close identity of the two sections of a double valve would be fortuitous in the present stage of manufacturing technique, so it was decided to investigate theoretically and experimentally the conditions required for compensating a lack of similarity in the bridge circuits. With reference to Fig. 1 and the appendix: it will be seen that the conditions for independence of H.T. and filament (or heater) variations may be fulfilled by adjusting two of the quantities anode load, grid bias, filament (or heater) current or individual H.T. voltage. By carrying out these adjustments successively, with considerable care it is possible to reduce the drift from 20 mV (in the case of a badly uncompensated valve) to 70 microvolts for 1 per cent change in either H.T. voltage or

Fig. 3. Effect of Reversing Filament Polarity in DBM2A



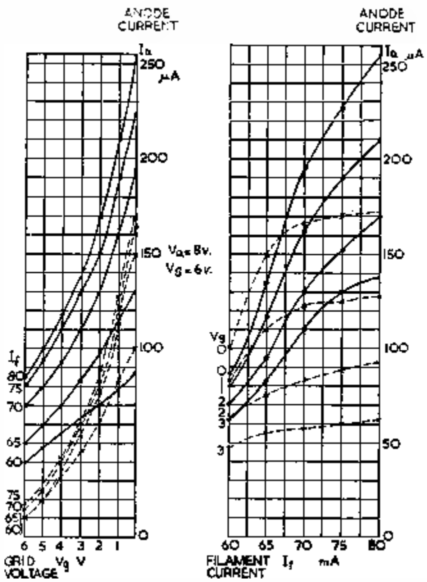


Fig. 4. Characteristic curves of (DBM2A) directly heated electrometer valve

filament current. In view of the complexity of fulfilling one or both of these compensation conditions (H.T. compensation could possibly be dispensed with), together with the condition of zero balance of output voltage, it was decided to simplify setting-up, operation and maintenance by adopting high stability in supply voltages in preference to any means of circuit compensation in the amplifier, as the latter involves quite critical adjustments, possibly varying slowly with age. Accordingly a power pack supplying 170 mA at 200V (for H.T. and heater current) was designed to vary not more than 1/40 per cent for 10 per cent changes in the 230 V mains, a 120 V battery being used as a reference level. The temperature coefficient of the battery is approximately 1/50 per cent per °C.

Temperature Effects

Operating the amplifier from this power pack and allowing some 20 hours to reach equilibrium, there occurred slow drift in output that could not be attributed to supply variations. By enclosing the electrometer stage in an oven and applying variations in temperature measured by thermo-couple, an appreciable temperature coefficient of drift was observed in the electrometer valve. Fig. 2 illustrates a typical example, giving a coefficient of 1.25 mV/°C., the bridge circuit being balanced for zero output but otherwise uncompensated. For comparison a typical 12SC7 or VX3088 stage has a coefficient of from 0.1 mV/°C. to 0.2 mV/°C. In each case it was established that the circuit components contributed inappreciably to the effect measured.

As an explanation of this undue dependence upon temperature it is suggested that owing to the probable geometrical asymmetry in the valve a change in external temperature causes:

(a) a change in emission but unequal changes in anode current.

(b) unequal changes of valve characteristics due to unequal expansion or contraction of the electrode structure;

(c) unequal changes in electrode potentials due to contact potential variations.

Lack of symmetry is suggested by the filament arrangement of the DBM2A in which each section of the valve is operated from opposite "legs" of a V-shaped filament. Further evidence is obtained from Fig. 3, where the effect upon the valve characteristic of reversing filament polarity is illustrated.

It has been difficult to investigate experimentally the effect of external temperature change owing to lack of suitable specialised equipment. Hypotheses (a) and (b) are unsupported, as far as is known, by any published literature. The possibility of an effect due to (c) is supported by an investigation of valve to valve variation of contact potential, due to contamination of the grid by cathodic barium, by Liebmann[5] and a discussion of the temperature variation of work function by Jones.[9]

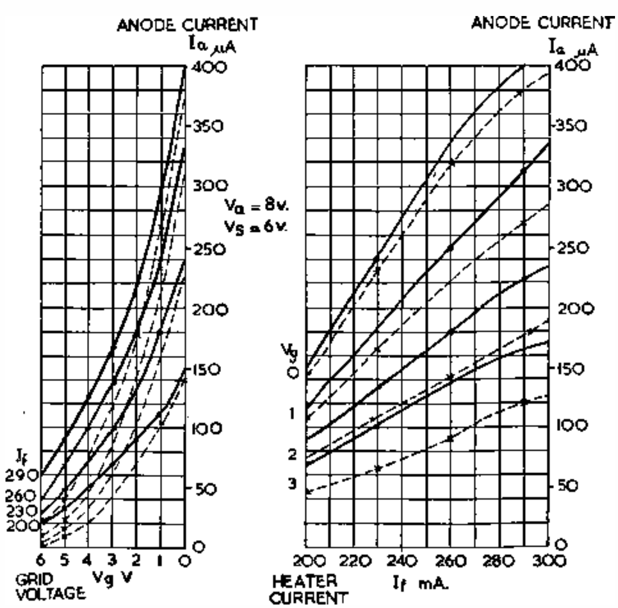
Since temperature has a limiting effect upon amplifier stability and since temperature control and any attempt at circuit compensation for temperature change was considered tedious and inconvenient it was thought to be more profitable to reduce the effect by attention, in design and manufacture, to increasing the similarity between the two sections of the double valve.

Electrometer valve type DBM8A

An indirectly heated version of the Ferranti double electrometer valve has shown decided improvement and a later version, the DBM8A, operates from 120 mA heater current, exhibits a temperature coefficient of not greater than 0.2 mV/°C., grid current of approximately 3.10^{-15} A and grid insulation, when carefully cleaned, usually greater than 10^{16} ohms.

As summarised by Little,[10] the advantages asso-

Fig. 5. Characteristic curves of indirectly heated electrometer valve (DBM6A)



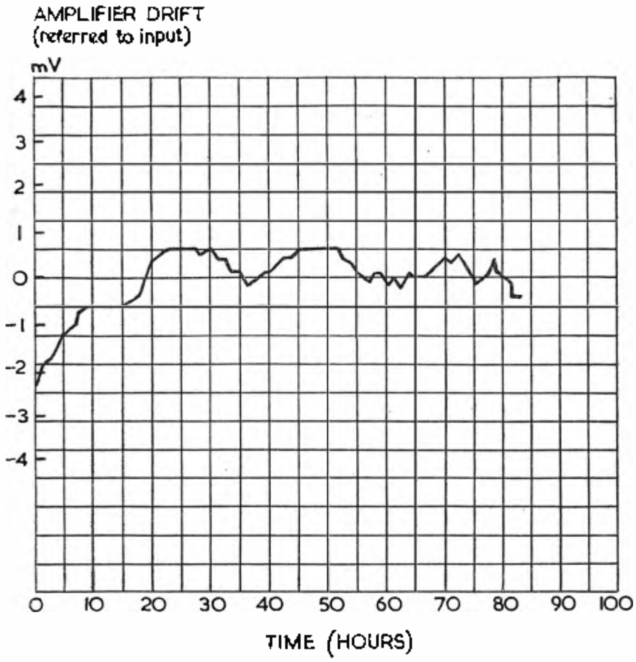


Fig. 6. Record of DBM8A amplifier drift. Grid resistors $10^{10}\,\Omega$

ciated with the indirectly heated cathode compared with a filament are:

(a) greater ease of symmetrical location;

(b) the cathode is maintained at a uniform potential;

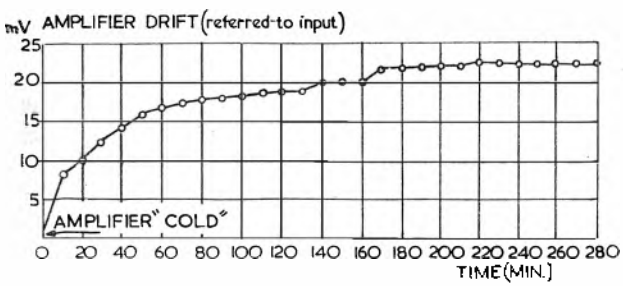
(c) less microphonic disturbances owing to greater rigidity of mounting;

(d) a larger space charge which decreases the effect at the anodes of a change at the cathode.

Characteristic curves of directly and indirectly heated electrometer valves are shown in Figs. 4 and 5, from which the similarities of g and $\delta Ia/\delta If$ between the two sections of each valve may be compared.

Drift in a push-pull or balanced amplifier incorporating a DBM8A in the electrometer stage followed by two 12SC7 stages showed a maximum fluctuation of ± 1 mV over periods of several days, the amplifier being enclosed within a large thermally insulated box. Fig. 6 shows a typical amplifier drift record after suitable warming-up. The fluctuations may be attributed to residual effects of temperature and mains variations, the relative magnitudes of which are difficult to estimate, and possibly to the effect of slow variations or ageing in valves and components. Fig. 7 demonstrates drift during the warming-up period, the amplifier approaching equilibrium conditions after about three hours running. Both amplifier and power pack were switched on from the cold state at the commencement of the record. (As comparison, a DBM2A amplifier requires from 10 to 20 hours warming up, as may be expected from the larger temperature coefficient of drift.) As an indication of the effect of grid current variation Fig. 8 compares records of drift with grid resistors of 10^{12} and zero ohms in each grid circuit. Thus a difference of 0.25 mV variation in drift would represent a difference of 0.25×10^{-15} A in grid current between the two sections of the valve.

Fig. 7. Record of DBM8A amplifier drift. "Warming Up" drift



The gain of the amplifier is approximately 700 and provides a maximum undistorted output of ± $3\frac{1}{2}$ volts. Negative feedback from 0 to 100 per cent may be applied.

Circuit Details (Fig. 9)

The power pack is of conventional design, consisting of a positive voltage regulator CV345 ($V8$) series valve controlled by $V10$, which is in turn balanced for heater variations by $V11$. The anode of $V10$ is connected to the screen of $V8$ which is fixed at a potential equal to the sum of a regulated positive

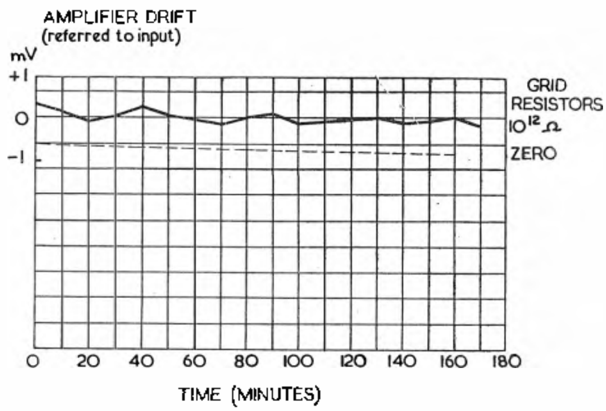
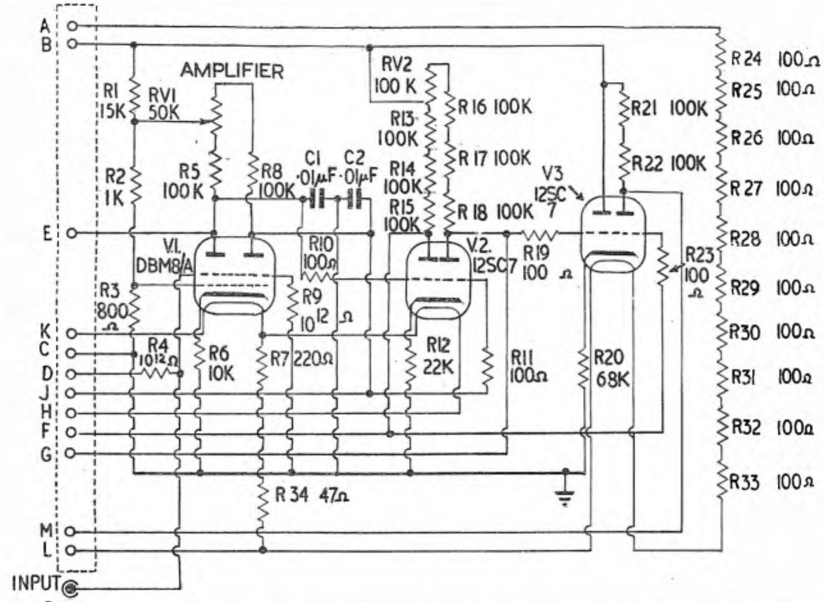


Fig. 8. Record of DBM8A amplifier drift. Effect of grid current variations

output voltage (200 V) and the voltage across the neon stabiliser $V9$. The power pack is balanced for mains variations by the controls $RV3$ and $RV4$. Where mains variations larger than ± 20 per cent are expected it has been found advisable to place a constant voltage transformer before the power pack.

The 100 volt negative supply is stabilised by neons $V5$ and $V6$ and varies 0.05 per cent for 10 per cent variations in mains. The regulated 200 volts positive supply varies not more than 1/40 per cent for mains changes of 10 per cent in voltage or 6 per cent in frequency. A 120 volt battery is used as a reference level.

In the amplifier the electrometer ($V1$) and first 12SC7 ($V2$) stages may be balanced by controls $RV1$ and $RV2$ in the respective anode circuits. The heaters of $V1$, $V2$ and $V3$ are connected in series from the regulated 200 volt supply; overheating the droping resistors is avoided by suitable under-running. The order of connexion is such that the electrometer heater is biased positively with respect to



Resistors

R1	15K	6W
R2	1K	6W
R3	800 Ω	6W
R4	10^{12} Ω	6W
R5	100K	6W
R6	10K	6W
R7	220 Ω	6W
R8	100K	6W
R9	10^{12} Ω	6W
R10-11	100 Ω	½W
R12	22K	6W
R13-18	100K	6W
R19	100 Ω	½W
R20	68K	6W
R21-22	100K	6W
R23	100 Ω	½W
R24-33	100 Ω	10W
R34	47 Ω	6W
R35	1.5K	10W
R36	680 Ω	10W
R37	500K	10W
R38	47 Ω	6W
R39	47K	10W
R40	68K	10W
R41	1K	10W
R42	20K	10W
R43	100 Ω	½W
R44-47	500K	10W
R48	100 Ω	½W
R49	10K	10W
R50	10K	10W
R51	50K	10W
R52	100K	10W
R53	1M	10W

Variable Resistors

RV1	50K	
RV2	100K	
RV3	2.5K	
RV4	50K	
RV5	5K	25W
RV6	100K	10W
RV7	To suit output meter	

Inductors

L1	20H	120mA
L2	20H	120mA

Capacitors

C1	0.01 μF	
C2	0.01 μF	
C3	4 μF	250V
C4	8 μF	600V
C5	8 μF	600V
C6	4 μF	250V
C7	0.5 μF	
C8	0.01 μF	

Transformers

T1-2

Switches

S1-2

Fuses

F1	2A
F2	2A
F3	60mA
F4	500mA

Meters

M1-2

Jacks

JK1-7

Valves

V1	DBM8/A
V2	12SC7
V3	12SC7
V4	CV1863
V5	CV216
V6	CV686
V7	CV575
V8	CV345
V9	CV686
V10	CV1056
V11	CV1056
V12	CV1933

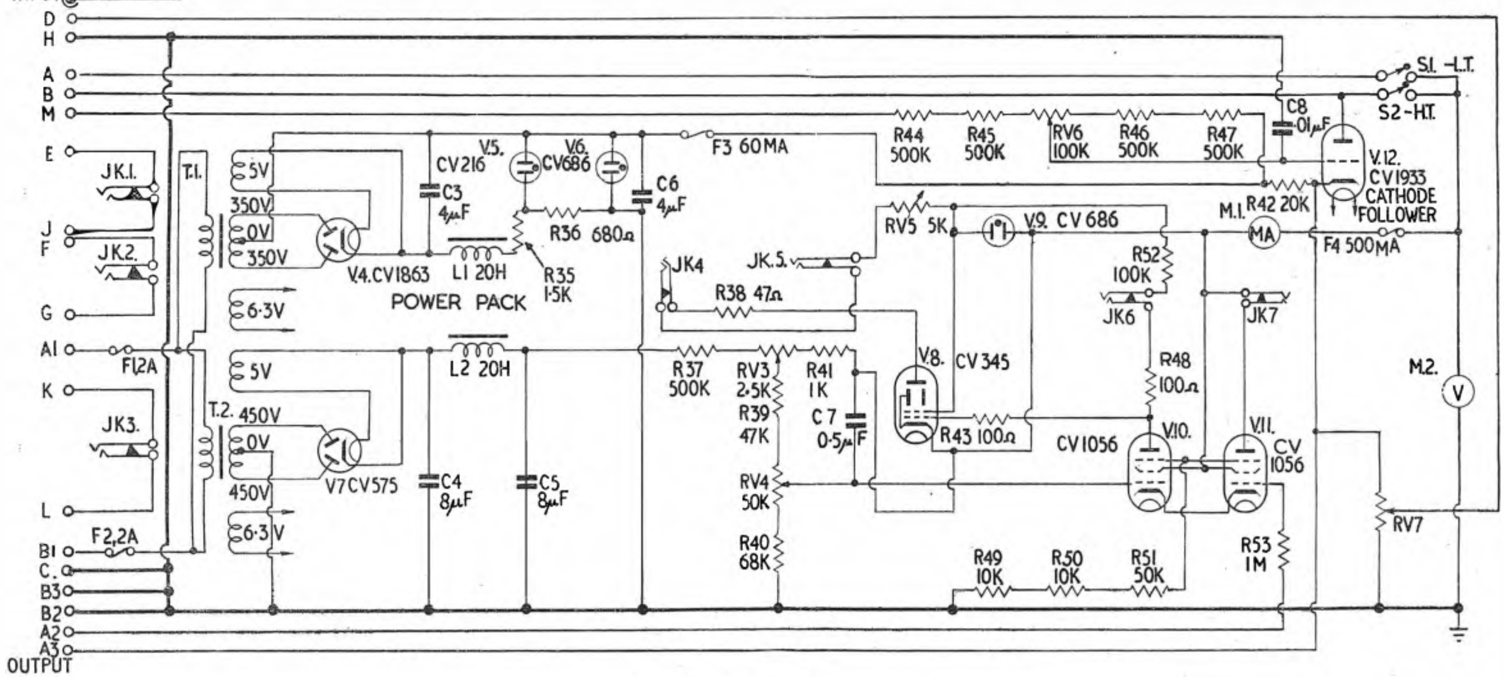


Fig. 9. Low-drift D.C. amplifier circuit diagram

cathode as this measure has been found to decrease long-term grid current fluctuations. The earth return from the heater line is taken to the power pack chassis by a lead separate from the main earthing connexion. The stage $V3$ converts from push-pull to single-ended operation for connexion to the cathode follower ($V12$) output stage mounted in the power pack chassis. The electrometer is operated with grid resistors in both sections to balance grid current effects. These large value resistors must be of high quality and stability, such as the "Victoreen." The operative grid resistor is returned to the feedback control $RV7$.

To overcome oscillatory instability it was found necessary to shunt both anodes of $V1$ and the grid of $V12$ by silver mica capacitors; also, grid and anode "stoppers" are used where necessary in the amplifier and voltage regulator circuits. Throughout, great care must be taken with screening and with soldered and plug-socket contacts. No special provision has been made against vibration and mechanical shocks. The amplifier can with advantage be mounted on rubber supports.

Acknowledgments are due to the Director, Atomic Energy Research Establishment, Harwell, for permission to publish this paper. Much of the work described is due to the advice and assistance of Mr. D. G. A. Thomas, Mr. F. B. Whiting and Mr. N. F. Wood, of A.E.R.E., and of Dr. J. A. Darbyshire and his colleagues at Messrs. Ferranti.

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APPENDIX

Compensation conditions in the double valve bridge circuit

With reference to Fig. 1, it is required to establish operating conditions that minimise effects due to H.T. and heater variations.

In the two sections of the valve, let

R_1, R_2 represent the respective anode loads

Ia_1, Ia_2 represent the respective anode currents

ρ_1, ρ_2 represent the respective anode impedances

Va_1, Va_2 represent the respective anode potentials

V_{HT} represent the common H.T. potential

I_f represent the common heater current.

Changes in anode current and voltage due to increments δV_{HT} and δI_f are given by

$$\delta Ia_1 = \left(\frac{\partial Ia_1}{\partial I_f}\right) . \delta I_f + \left(\frac{\partial Ia_1}{\partial Va_1}\right) \delta Va_1$$

$$= \left(\frac{\partial Ia_1}{\partial I_f}\right) . \delta I_f + \frac{\delta Va_1}{\rho_1}$$

$$\text{and } \delta Va_1 = -R_1 \delta Ia_1 + \frac{\rho_1}{R_1 + \rho_1} . \delta V_{HT}$$

$$\left.\begin{aligned} &\text{Eliminating } \delta Ia_1,\ \delta Va_1 = \frac{\rho_1^2}{(R_1+\rho_1)^2} . \delta V_{HT} - \frac{R_1\rho_1}{R_1+\rho_1}\left(\frac{\partial Ia_1}{\partial I_f}\right) . \delta I_f \\ &\text{Similarly, } \delta Va_2 = \frac{\rho_2^2}{(R_2+\rho_2)^2}\ \delta V_{HT} - \frac{R_2\rho_2}{R_2+\rho_2}\left(\frac{\partial Ia_2}{\partial I_f}\right) . \delta I_f \end{aligned}\right\} . \quad (1)$$

The condition that the push-pull output should be independent of H.T. variations is:—

$$\delta Va_1 - \delta Va_2 = 0$$

$$\frac{\delta Va_1}{\delta V_{HT}} = \frac{\delta Va_2}{\delta V_{HT}} \quad (\delta I_f = 0)$$

i.e.,

$$\frac{\rho_1^2}{(R_1+\rho_1)^2} = \frac{\rho_2^2}{(R_2+\rho_2)^2}$$

$$\therefore \quad \rho_1/\rho_2 = R_1/R_2 \quad \ldots\ldots\ldots\ldots \quad (2)$$

The condition for independence of heater variations is

$$\frac{\delta Va_1}{\delta I_f} = \frac{\delta Va_2}{\delta I_f} (\delta V_{HT} = 0)$$

$$\text{i.e., } \frac{R_1\rho_1}{R_1+\rho_1}\left(\frac{\partial Ia_1}{\partial I_f}\right) = \frac{R_2\rho_2}{R_2+\rho_2} . \left(\frac{\partial Ia_2}{\partial I_f}\right)$$

Using (2)

$$\frac{R_1}{R_2} = \left(\frac{\partial Ia_2}{\partial I_f} \Big/ \frac{\partial Ia_1}{\partial I_f}\right) \ldots\ldots \quad (3)$$

The compensation conditions (2) and (3) may be fulfilled simultaneously by adjustment of two of the three quantities R, ρ and $(\partial Ia/\partial I_f)$, ρ and $(\partial Ia/\partial I_f)$ being altered by adjustment of grid bias, I_f or individual H.T. potentials. In practice, each of the two selected variables is adjusted successively to reduce the change in push-pull output for small changes in the corresponding supply voltage. This is repeated until sufficient compensation is obtained. This state will hold for the range of supply voltage variation over which the valve may be considered to operate in a linear manner.

It is interesting to note that a combination of (3) and the condition for zero-balance of output potential gives

$$\left.\begin{aligned} &\frac{R_1}{R_2} = \left(\frac{\partial Ia_2}{\partial I_f}\right)\Big/\left(\frac{\partial Ia_1}{\partial I_f}\right) \\ &\text{and } Va_1 = Va_2, \text{ i.e., } \frac{R_1}{R_2} = \frac{Ia_2}{Ia_1} \end{aligned}\right\}$$

$$\text{i.e., } \frac{Ia_2}{Ia_1} = \left(\frac{\partial Ia_2}{\partial I_f} \Big/ \frac{\partial Ia_1}{\partial I_f}\right)$$

By consideration of the anode current and heater current characteristics for each section of the valve, this last expression may be interpreted as meaning that in the fully compensated and zero balanced condition the tangents at the operating point on each curve should intersect on the I_f axis. By inspection of the characteristic curves, this furnishes a means of determining the correct operating point and also whether it is practically possible to satisfy simultaneously the three conditions for the particular valve.

A Beam-Switching Circuit

By ROSEMARY MILNE, Grad.I.E.E.*

THERE are several well known methods of photographing cathode ray tube traces. Transients are usually recorded on moving film without a time-base, or on stationary film by means of a single sweep time-base. Recurrent phenomena are displayed on a time-base synchronised with the signal, and several cycles are photographed.

Some phenomena, however, cannot be properly classified as either transient or recurrent. The problem which gave rise to the design of this circuit was the measurement of the distortion of a brake drum during a brake application. The distortion at any one point on the surface of the drum can be considered as a transient with respect to time, and the distortion at any one instant can be considered as recurrent with respect to distance, one cycle corresponding to the circumference of the drum. In order to measure the distortion all round the drum during the whole of the brake application, the surface of the drum may be rapidly "scanned" by the pick-up device. The output from the pick-up then has a definite repetition frequency, but successive cycles differ slightly from each other.

It is, of course, possible to photograph the whole of the distortion record on continuous film, but this method uses large quantities of film and provides an unnecessary amount of information. A more economical method is to synchronise or couple a time-base to the repetition frequency and to photograph a number of single sweeps of the time-base. The phenomenon can be continuously observed on a monitor tube.

A suitable circuit for coupling a time-base to a rotating shaft was described in this journal recently[1], and a circuit is described below which switches on the electron beam of the C.R.T. during the sweep of the time-base following the sweep during which the operating push-button was released.

General Description

The first stage is the "pulse-shaping unit" which produces negative pulses from the flyback of a saw-tooth time-base voltage.

The next stage is an Eccles–Jordan trigger circuit which is triggered by pressing a button, and returned to normal at the end of the time-base sweep during which the button was released.

The last stage, also an Eccles-Jordan trigger circuit, is triggered by the return of the first trigger circuit and is returned to normal after one sweep of the time-base. The output voltage of the circuit is taken from the anode of one of the valves of this last stage and applied to the grid of the C.R.T.

The waveforms at various parts of the circuit are shown in Fig. 1 and a complete circuit diagram in Fig. 2.

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Circuit Details

Pulse-shaping Unit.

This beam-switching circuit was designed to work primarily with the time-base circuit described by Marshall,[1] which requires an input impedance of the order of 100 megohms. The maximum permissible grid leak for a normal valve is 3 megohms. However, if a very small coupling capacitor is used, the input impedance will be effectively high during the sweep time and the sweep voltage will not be much reduced. $C1$ has, therefore, been made 10 pF and $R1$ 3 megohms. This arrangement produces small positive pulses during the flyback time and smaller negative pulses during the sweep time from the

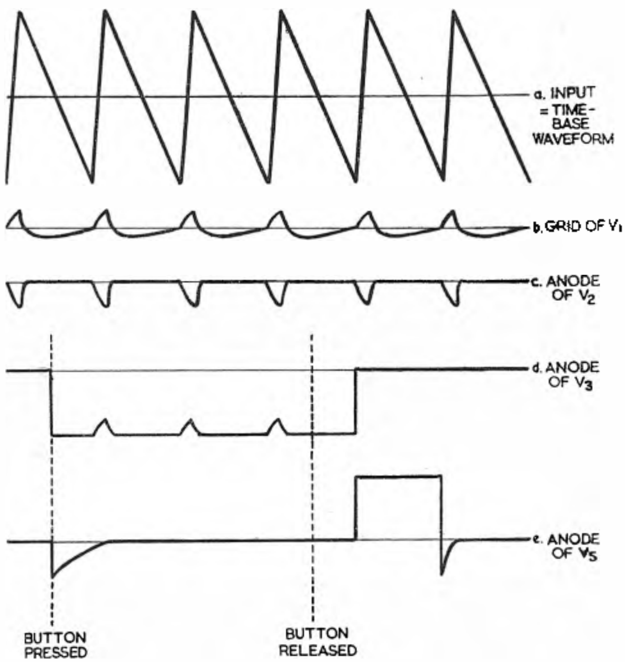
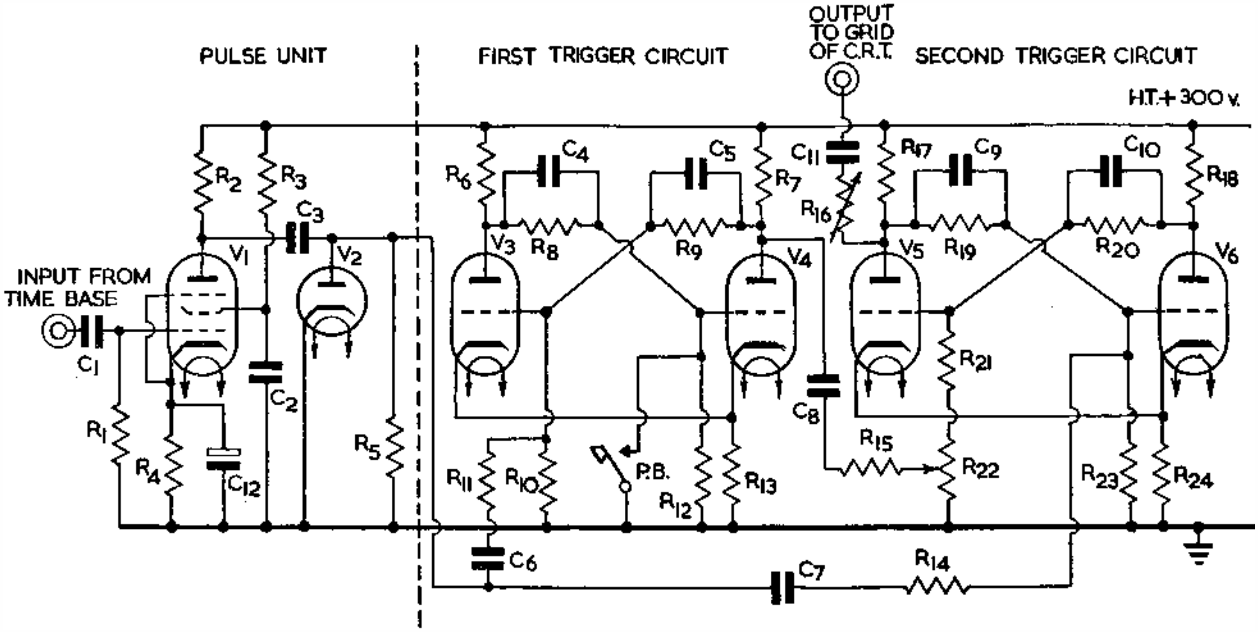


Fig. 1. Waveforms (not to scale)

waveform shown in Fig. 1(a). The relative amplitudes of the pulses depend inversely on their relative durations because the mean voltage is zero, i.e., in Fig. 1(b) the area above the zero line equals the area below.

The pulses are amplified by the pentode $V1$. The flyback pulses, which are now negative, are passed by $C3$ and $R5$, but the sweep pulses are short-circuited by the diode $V2$. The waveform is then as in Fig. 1(c) and is suitable for operating the trigger units. If the time-base waveform is of opposite polarity to Fig. 1(a), the diode $V2$ should be reversed and positive pulses will then be produced.



R_1 3M	R_5 1M	R_9 10M	R_{13} 10K	R_{17} 10K	R_{21} 2M	C_1 10pF.	C_5 50pF.	C_9 50pF.	V_1 EF36 (Mullard)	V_5 L63 (Osram)
R_2 100K	R_6 15K	R_{10} 3M	R_{14} 5M	R_{18} 10K	R_{22} 1M	C_2 0.5μF	C_6 0.1μF.	C_{10} 50pF.	V_2 EA50 ,,	V_6 L63 ,,
R_3 390K	R_7 15K	R_{11} 5M	R_{15} 2M	R_{19} 10M	R_{23} 3M	C_3 0.1μF	C_7 0.1μF.	C_{11} 0.1 μF.	V_3 L63 (Osram)	
R_4 1.5K	R_8 10M	R_{12} 3M	R_{16} 1M	R_{20} 10M	R_{24} 8K	C_4 50pF.	C_8 1000pF.	C_{12} 25μF.	V_4 L63 ,,	

Fig. 2. Circuit diagram with component values

Alternative Sources of Operating Pulses.

When a Puckle[3] time-base is used, negative pulses may be obtained from the anode of the triode and applied to the junction of *C*6 and *C*7, the pulse shaping circuit being omitted.

If the time-base circuit used does not require an input impedance of more than 4 megohms, and if the flyback time is short compared with the sweep time, the pulse-shaping unit may again be omitted and *C*6 and *C*7 reduced in capacitance so that with *R*11 + *R*10 and *R*14 + *R*23 respectively, they act as differentiating circuits and produce positive pulses.

Polarity of Operating Pulses.

The connexions shown in Fig. 2 are for use with negative pulses. If, however, the circuit is to be operated by positive pulses, the push-button and the capacitor *C*8 should be connected to *V*3 instead of *V*4 and the output should be taken from the anode of *V*6.

Trigger Circuits.

These are two similar Eccles-Jordan circuits. Negative pulses from the pulse-shaping unit are fed, via decoupling resistors *R*11 and *R*14, to the grids of *V*3 and *V*6. Normally, therefore, *V*4 and *V*5 are conducting. The operating push-button (P.B.) connects the grid of *V*4 to earth and so triggers and holds the first trigger circuit. This action passes a positive pulse, via *C*8 and *R*15, to the grid of *V*5, but it has no effect, since *V*5 is already conducting. When the button is released, the circuit remains stable until the end of the time-base sweep, when a negative pulse reaches the grid of *V*3, and returns the first trigger circuit to normal (i.e., *V*4 conducting). A negative pulse is passed to the grid of *V*5 (see Fig. 1(d)), and since this pulse is made longer than the pulse which arrives simultaneously at the grid of *V*6 from the pulse unit, the second circuit triggers (i.e., *V*6 conducts). After one sweep of the time-base, the negative pulse on the grid of *V*6 returns the circuit to normal (see Fig. 1(e)). The anode voltage of *V*5 is applied, via the output control *R*16 and a blocking capacitor *C*11, to the grid of the C.R.T. A grid resistor of 500 k is suitable. The grid bias of the C.R.T. is adjusted to keep the beam normally switched off.

Performance and Adjustment

It is difficult to quote figures for performance, since the operation of the circuit depends on both the time-base frequency and the flyback time. If the latter is too long, the pulses produced by the pulse unit are not of sufficient amplitude to trigger the trigger units. If the sweep time is too short, the pulse from *V*4 holds the grid of *V*5 negative longer than one sweep time and more than one sweep is brightened. However, the circuit will operate over a very wide range without adjustment. With the component values shown in Fig. 2, the circuit has been found to operate with flyback times of more than 1/100 sec. and sweep frequencies higher than 50 per second. The flyback time in these tests was about one-seventh of the sweep time.

In order to use the circuit with flyback times longer than 1/100th second the time constant $R1$, $C1$ should be increased and, for very much longer times, it may be necessary to increase $C8$, $R15$ as well. Conversely, for sweep frequencies higher than 150 per second, $C8$ ($R15$ + part of $R22$) should be reduced and, if the flyback time is less than 300 micro-secs., $C1$, $R1$ should be reduced as well. The time constant $C11$ ($R16$ + grid resistor of C.R.T.) should be about 20 times as long as the longest sweep time to be used.

The bias on the C.R.T. grid should be just sufficient to switch off the beam and the brightness of the single brightened sweep should then be controlled by $R16$. This arrangement gives the most even brightening of the trace.

The potentiometer $R22$ adjusts the amplitude of the pulse passed from $V4$ to $V5$ and should be adjusted so that the circuit will just operate.

Further Uses

In addition to the use for which this circuit was originally designed, namely brake testing, there are a number of others. The indicator diagrams of engines running at varying speeds and electrocardiographs are typical examples of phenomena of which successive cycles are similar but not identical. When observing a recurrent phenomenon on a C.R.O., it is usual to employ a synchronised time-base. If this is not convenient, the circuit described above can be used in conjunction with a long-delay screen and the phenomenon can be observed without a synchronising signal and its attendant difficulties.

The author wishes to express her thanks to Mr. William Smith, Chairman of Messrs. Ferodo, Ltd., for permission to publish this paper.

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[1] P. R. Marshall. *Electronic Engineering*, March 1949.
[2] O. S. Puckle see " Time Bases " (Chapman & Hall) 1943, p. 30.

MAGNETIC TRANSDUCING HEADS

A MAGNETIC transducing head, with a ring-shaped magnetic core, as illustrated, is well known, and in magnetic sound recording and reproducing apparatus three such heads are commonly used for erasing, recording and reproducing respectively, although it is also known to combine the two latter functions in a single head.

In certain kinds of magnetic transducing apparatus, it would be of advantage to use only a single head which could fulfil all three functions as and when required, thereby reducing the size and cost of the apparatus. Such an arrangement is eminently suitable for use in a sound reproducing device utilising an endless loop of magnetic tape on which a message, once recorded, is required to be repeated a large number of times, and the operations of erasing and re-recording are likely to be carried out only at infrequent intervals. For example, in a device of this kind for use in conjunction with traffic lights a directive message recorded on the tape is required to be reproduced cyclically in timed relation with the visual signals, such that suitable audible warnings or instructions are issued to pedestrians at appropriate times.

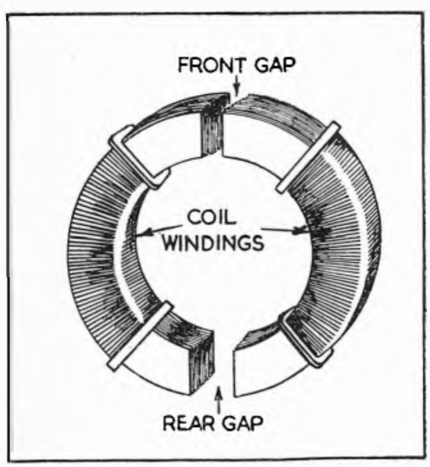
Referring to the illustration, it will be seen that the ring-shaped magnetic core has two diametrically opposed non-magnetic gaps, the widths of these gaps being of the order of 0.001 in. and 0.010 in. respectively. The magnetic tape is adapted to run tangentially past the narrower or operative gap, and the rear gap is provided to reduce remanence in the core and to ensure that the core does not become magnetically saturated.

In accordance with the present proposal the rear gap is also utilised to effect erasure of a previous recording. For this purpose the tape may be lifted from the front to the rear of the head to bring it into operative relation with the rear gap, or, alternatively, the head may be mounted so as to be rotatable through 180°.

In a compact sound reproducing device embodying the present proposal and suitable for use in conjunction with traffic lights, etc., it is necessary to include only a suitable reproducing amplifier and loudspeaker. When it is desired to make a new recording the transducer head windings are connected to an external source of H.F. oscillations or other erasing means and, with the tape operatively associated with the rear gap, the old recording is erased. After having returned the tape, or the transducer head as the case may be, to its original position, a new recording can then be impressed upon the tape with the aid of suitable external recording equipment. If, as is usual, the recording is effected with an accompanying H.F. bias, a common source of H.F. oscillations may be used to provide both the erasing currents and the recording bias.

To facilitate connexion of the transducer head windings for reproducing, recording or erasing, the windings may be associated with a suitable selector switch.

—*Communication from E.M.I., Ltd.*



Magnetic transducing head of a type suitable for recording reproducing or erasing

Noise in Fixed Resistors

By FRANCIS OAKES

FLUCTUATION voltages, generally referred to as "noise," appear across the terminals of any resistor whether this resistor carries a current or not. These fluctuation voltages derive their more common name "noise" because of the interference they cause in certain radio circuits, which is of an audible kind. Apart from their adverse effects in raising the level of background hiss, noise voltages are the cause of undesirable phenomena in other electronic circuits, such as high gain audio and wide-band amplifiers, sensitive valve voltmeter circuits and similar devices where small signals are handled by high resistances carrying a relatively large D.C. voltage at the same time.

The fluctuation voltages can be due to various causes which may be responsible separately or simultaneously for the phenomena produced. The best known of these causes is the thermal agitation of electricity, also called the Johnson effect. Thermal agitation noise is particularly well known in connexion with high frequency amplifiers. The lesser known resistance fluctuation noise is more pronounced at lower frequencies and, at normal temperatures, can be responsible for much larger noise amplitudes than most other sources of resistor noise; for this reason, and because little has so far been published on this subject, resistance fluctuation noise will be discussed in detail. The remaining phenomena are chiefly due to chemical action, ageing and mechanical effects such as microphony. They occur less frequently and, in general, should not produce very large noise amplitudes in an ordinary radio resistor.

Thermal Agitation

Whatever the type, material or resistance of a resistor, thermal agitation is always present. It is explained by the action of free electrons in the resistor which are in random motion in equilibrium with the thermal motion of the molecules of the resistor; the charges of the electrons are in motion relative to the molecules, and in this way they constitute element-currents flowing within the resistor. The P.D.s which they produce add up to the fluctuating voltage apparent across the terminations of the resistor. The fluctuation frequencies of thermal agitation voltage are uniformly distributed over the whole spectrum. Their R.M.S. value is dependent only on the temperature of the resistor and on the resistance value. R.M.S. values over different band widths and for resistance values from 100 ohms to 1 megohm are shown in Fig. 1. The diagram represents values at room temperature, approximately 20° C. When a current flows through the resistor the electrons thereby moved produce a small increase in the magnitude of thermal agitation noise. It is an important distinction between this and all other types of noise, that the magnitude of thermal agitation voltage can be calculated mathematically from the resistance, temperature and current data alone, being independent of the other properties of the resistor, and that its frequency distribution is uniform over the whole spectrum.

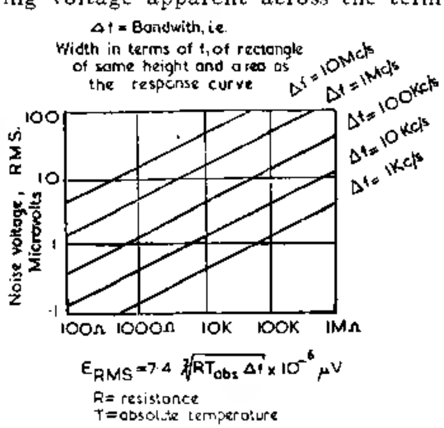


Fig. 1. Thermal agitation noise

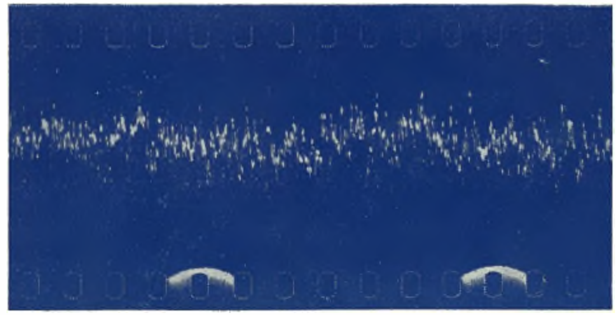
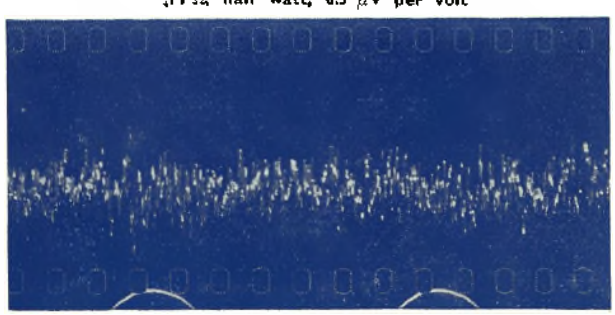


Fig. 2. Noise trace of composition resistor (solid) under load ½M Ω half watt, 2.5 μV per volt

Resistance Fluctuation Noise

In the case of certain resistances, especially such as carbon composition and carbon track resistors, a much greater increase in noise voltage can be observed with increasing current than could be accounted for by the action of thermal agitation. The frequency distribution of this additional noise component is no longer uniform, but from about 30 cycles per second upwards decreases with increasing frequency. Therefore the major portion of the energy is produced within the audio-frequency band. Furthermore, the amount of noise produced is no

Fig. 3. Noise trace of composition resistor (solid) under load ½M Ω, half watt, 0.3 μV per volt



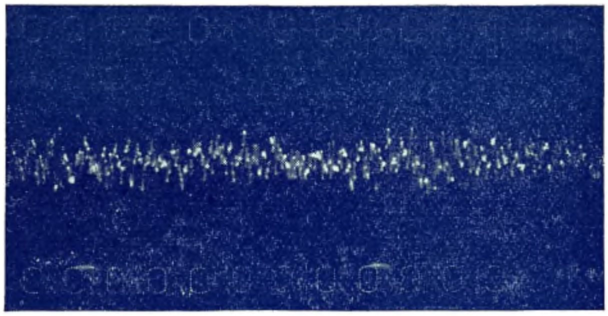
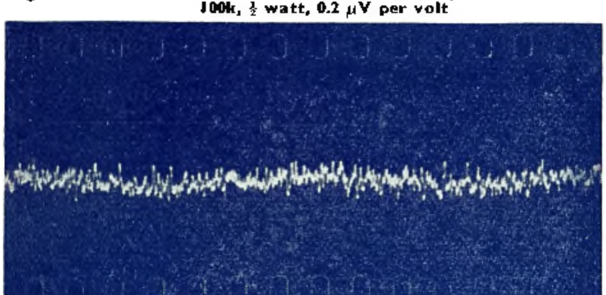


Fig. 4. Noise trace of carbon track resistor (Film) under load 1 MΩ, ½ watt, 20 μV per volt

longer independent of physical properties other than temperature and resistance value. It varies according to the material and construction of the resistor, and even within one particular type and size, the amount of resistance fluctuation noise can vary considerably from sample to sample. Especially for high-valued carbon composition and carbon track resistors, from about 50,000 ohms upwards the noise due to resistance fluctuation under normal loading currents can reach many times the value of thermal agitation noise. The R.M.S. value of resistance fluctuation noise increases with increasing current, and, within limited ranges of current, is fairly accurately proportional to the loading current. For this reason, the term " resistance fluctuation noise " has been found applicable; in other words, the ratio between loading current and noise voltage produced is constant. Since the resistance value is essentially constant, the fluctuations amounting to something of the order of its millionth part, the " noise value " of a resistor can be expressed in microvolts noise generated per volt applied. Figs. 2, 3, 4 and 5 show oscillograms of representative noise traces produced by different radio resistors. From these, certain conclusions can be made as to the nature of the cause of resistance fluctuation noise. On close inspection of the noise traces, it will be seen that they are composed essentially of a series of dots joined by almost vertical thin lines. This means that the voltage, and the resistance, are changing abruptly from one value to another, these changes taking place in rapid succession. This tallies well with one theory which seeks an explanation in contact phenomena, according to which resistance paths within the resistor are constantly put into and out of operation. Noise traces of this kind can be produced by photographic recording of cathode-ray

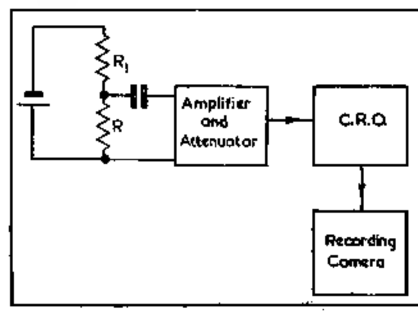
Fig. 5. Noise trace of carbon track resistor (Film) under load 100k, ½ watt, 0.2 μV per volt



oscillograms, with an arrangement such as shown in Fig. 6, consisting of a battery for loading the resistor with a direct current, a series resistor which must be free of resistance fluctuation noise, a high-gain wide-band amplifier, of approximately 100 to 120 db. gain from the lower end of the audio-frequency band up to 10 kc/s. or more and including a suitable attenuator. (As the amplifier used for the recording of the noise traces shown has a response falling off at the lower audio-frequencies, this should be taken into account when looking at the oscillograms). A cathode-ray oscillograph without a time base, and a rotary drum camera were used; (but a still camera can be used in connexion with a single-stroke time base, to obtain adequate oscillograms of noise traces). For these illustrations, a 50-cycle timing wave was simultaneously produced on the oscilloscope tube screen, and allowed to protrude with its wave crests into the recording field. The time marks at the bottom of each oscillogram were obtained in this way, the distance from crest to crest denoting 1/50 second.

Most carbon composition high resistors show an appreciable amount of resistance fluctuation noise under even a small load, with an increase in the noise value (microvolts noise per volt load) as the resistance increases. But a number of carbon composition resistors have been designed recently with a remarkably low noise level as compared with the more conventional types. With carbon track resistors, especially the helical track and the cracked carbon types, low noise levels have been achieved for a long time and even with conventional designs; on the other hand, some carbon track resistors are very noisy and, in fact, the noise trace in Fig. 4 has been taken from a carbon track type resistor, whereas the one in Fig. 3 is that of a conventional miniature composition resistor. The illustrations are given as representative examples, but must not be taken as a basis for generalisation so far as types or actual noise voltage magnitudes are concerned. It must also be stressed that, obviously, an attenuator had to be used to reduce the traces to a size suitable for recording, therefore only the time values in the x-direction are directly commensurable, but not the noise amplitudes in the y-direction.

Fig. 6. Arrangement for recording noise traces



In the range from 50,000 ohms up to about 2 megohms, noise values from a fraction of one microvolt per volt up to more than 20 microvolts per volt can be found in commercial radio resistors. For ordinary radio receiver circuits this noise does not provide a great problem as the signal strength in the audio-circuits is large and the resistance fluctuation noise component in the radio circuits is naturally

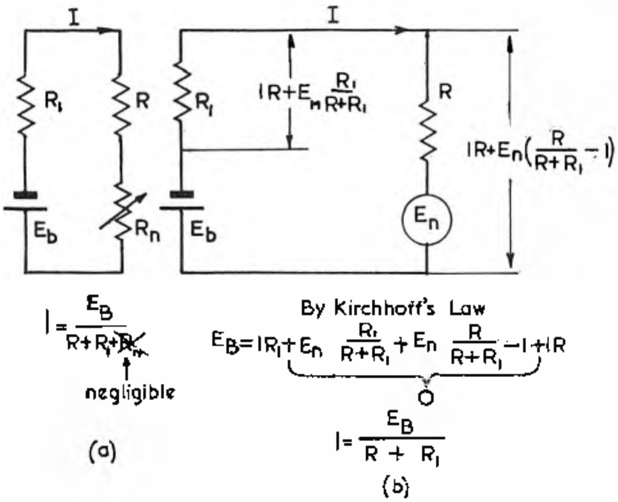


Fig. 7. Simple circuit containing noisy resistor, with equivalent circuit (*at right*) **for purposes of analysis**

negligible. The position is quite different so far as high gain amplifiers of the wide band type are concerned because there a small signal can be handled by a high resistance carrying a D.C. current, such as would be the case in the first stage of a resistance coupled voltage amplifier.

In order to study the conditions in an amplifier circuit, the resistor will first be regarded on its own in a circuit containing also a D.C. source and a resistor not producing noise due to resistance fluctuation. Such a circuit is shown in Fig. 7(a). R represents the noise generating resistor, Rn the resistance fluctuation (a small variable quantity) R and $R1$ are of the same order of magnitude, whereas Rn is a fraction of one-thousandth of this magnitude.

$$R \approx R1$$
$$Rn << R$$
$$I = \frac{Eb}{R + R1 + Rn} \quad \ldots Rn \text{ negligible}$$
$$\therefore I \approx \frac{Eb}{R + R1} \text{ and } En = IRn$$

where Eb . . . battery voltage
En . . . noise voltage
I . . . . circuit current

With these simplifications, sufficiently accurate analysis can be carried out, aided by the equivalent circuit shown in Fig. 7(b). From these considerations, it is evident that the noise voltage produced in the resistor R appears partly across R and partly across $R1$, in ratio to their fraction of the total series resistance of the circuit, i.e.:

$$En_R = En\frac{R}{R + R1} \text{ across } R$$
$$En_{R1} = En\frac{R1}{R + R1} \text{ across } R1$$

Figs. 8 and 9 show resistance coupled voltage amplifiers with a triode and a pentode valve respectively, together with their equivalent circuits. So far as the noise produced by the anode load resistance is concerned, there is no difference in the equivalent networks except for the actual values of the parameters. And it is the anode load which contributes the resistance fluctuation noise to the output passed on to the next valve, because the grid leak, the only other resistance also carrying signal, does not carry an appreciable load. The noise voltage generated in the anode load by virtue of the anode current carried distributes itself over the circuit in a manner similar to that shown in Fig. 7, if the fact is taken into account that $R1$ is replaced by the combination value of the plate resistance of the amplifying valve and the grid leak resistance of the following valve in parallel. (For accurate analysis the inter-electrode and other capacitances would also have to be taken into consideration.) The noise passed on to the next stage is represented by:

$$E'n = En\frac{\dfrac{Rp\,Rg}{Rp + Rg}}{R_L + \dfrac{Rp\,Rg}{Rp + Rg}}$$

Introducing representative numerical values, approximately $E'n = \frac{1}{2}En$ for triode amplifiers and $E'n = \frac{3}{4}En$ for pentode amplifiers is transmitted to the grid of the following valve. If the signal at the input grid is below 1 millivolt the signal to noise

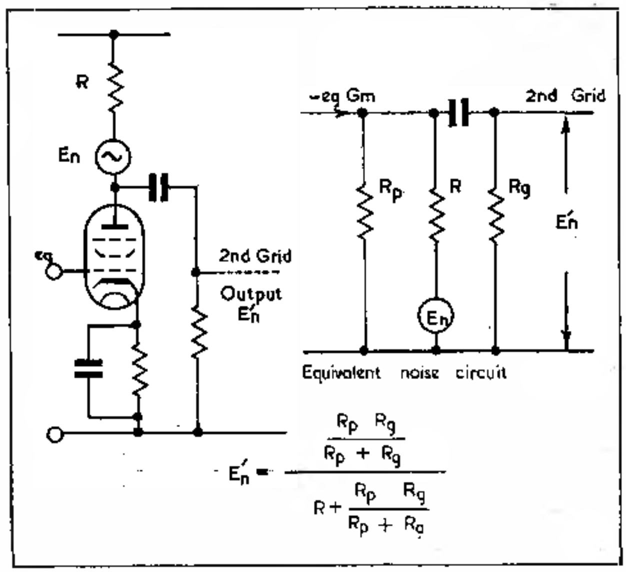


Fig. 9. Pentode amplifier

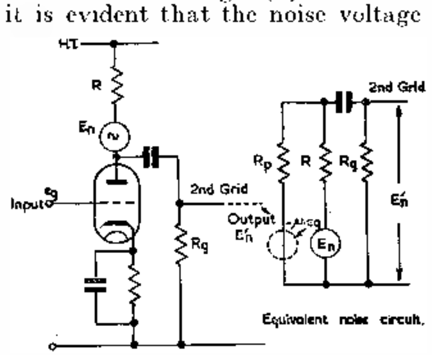


Fig. 8. Triode amplifier

ratio can suffer considerably unless a silent resistor is used as anode load resistor. It is for this reason that wirewound resistors are often used for high gain audio-frequency amplifiers handling small input voltages. But a well designed carbon resistor of the low-noise type can be used in most cases, and is often preferable because of its small dimensions, non-conductivity and smaller hum-pickup.

A triode amplifier, for instance, with 100 volts across a 200,000 ohms anode load of 0.1 microvolts per volt noise value, and a stage gain of 50, would deliver an output of 50 millivolts from 1 millivolt input, with a noise level of about 0.005 millivolts due to resistance fluctuation noise, i.e., of the same order as the contribution of thermal agitation. A signal to noise ratio of 5,000 would not provide any complications. On the other hand, a poorly designed load resistor with 50 microvolts per volt noise value, would already give a 1 per cent contribution of noise in terms of total output, with an input voltage of 1 mV. Serious problems would arise with smaller signals or noisier resistors.

It can be said that, generally, resistors made of low-conductivity material, or of materials used in very thin films or distributed in granulated form are more prone to resistance fluctuation noise than those made from high-conductivity and homogeneous materials. Wire wound resistors do not suffer from resistance fluctuation noise. If they do appear noisy, this is usually due to faulty connexions or corrosion of soldered joints.

Chemical Action, Ageing and Mechanical Effects

Whereas the Johnson effect and resistance fluctuation are commonly present in commercial carbon resistors, noise from chemical action, ageing and mechanical effects should not occur to a noticeable extent in a well designed resistor. The first two causes are due to undesirable curing effects, influence of atmosphere moisture and temperature, and reaction of the carrier, binder and other substances with the conducting materials with which they are in contact.

Mechanical sources of noise are generally due to bad contact between the resistance element and the termination wire or termination caps. Such faults sometimes show microphonic characteristics. When movements occur, or the resistor is mechanically agitated by vibration, the current carried by the resistor is modulated by the vibrations, and noise voltages are generated. In all cases of bad contact the noise voltage increases with increasing load current.

Chemical and thermal E.M.F's set up in circuits carrying small signals are well known, they are not essentially affected by load currents and their mechanism is too familiar to be discussed here.

Conclusions

Noise generated in resistors carrying a current is due to several causes, at least one of which, thermal agitation, cannot be eliminated. The magnitude of the noise fluctuations is relatively small and can be ignored in most L.F. circuit applications. Where high resistance values are involved and small signals are to be superimposed on steady D.C. voltages, the noise level must be kept low by the choice of suitably designed resistors. For the manufacturer, resistance fluctuation noise provides a design problem of considerable importance. Modern carbon track and carbon composition resistors are, however, so designed that in most cases their noise level remains low for currents up to their full loading capacity.

Interference from Incandescent Lamps

LAMPS on the point of burn-out have long been known as sources of interference to radio and television, the cause being sparking across a small gap in the filament wire. But interference from a lamp in good working order had not been suspected. One of the first to track down such a source was the Communication Committee of the Pennsylvania Electric Association. This group found that the Mazda A lamp, an old type containing a straight-wire filament in the form of a cage supported on a glass post, was a notable offender. Television service organisations have detected similar interference from modern inside-frosted bulbs.

To study the effect at first hand, the editors of *Electronics* requested the Pennsylvania Electric Association to provide a lamp known to cause the interference. The lamp was installed in an open holder about 2 ft. away from a model 630TS television chassis. The lamp and receiver were fed from the same power point.

The interference observed in the test takes the form of one or more horizontal bars, covering from three to 50 lines of the image (corresponding to interference pulses of 200 to 3,000 micro-seconds duration). The wider bars display beat-pattern striations having from 50 to 150 black and white segments in the width of the picture. The beat pattern indicates that each pulse of interference has a frequency in the vicinity of the picture carrier, and is frequency modulated over a range of 1 to 2 Mc/s. The possibility that the interference occurs at intermediate frequencies is ruled out by the radical change in its character as the receiver is tuned from one channel to another.

With the particular lamp tested, the strongest interference was experienced on channel 4 (66-72 Mc/s.), and took the form of two bars. On channel 2 (54-60 Mc/s.) the interference was less intense and appeared in three bars.

When the lamp was removed about 20 ft. from the receiver and connected to a separate outlet (joining the receiver outlet through the fuse box over 40 ft. of cable), the interference disappeared on all channels except channel 4, and then only one bar was of sufficient strength to be visible.

These observations indicate that the source of the interference is shock excitation of one or more tuned circuits whose natural resonances are in the 50-100 Mc/s. band. One possible explanation is thermal vibration, causing the filament wire to make intermittent contact with its supports. This would cause the natural period of the filament wire, considered as a resonant system made up of its own self-inductance and distributed capacitance, to change synchronously with the alteration in voltage drop along the wire. If the duration of the intermittent contact is brief, the resonant circuit formed by the wire receives a brief pulse of excitation from the power line once each cycle. —" *Electronics* ", *December*, 1949.

A Self-Adjusting Time Base Circuit

By H. ASHER, B.A.*

THERE are two kinds of time base which are widely known, the free running time base and the triggered time base. In the case of the free running time base the frequency must be adjusted manually to be an exact sub-multiple of the wave form to be examined, and this involves the use of two controls, the frequency control and the synchrony control. The triggered time base requires no such controls, the frequency of the time base being automatically locked to the signal which

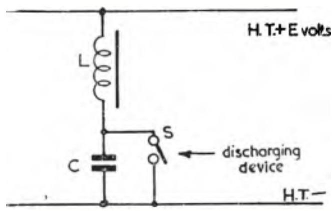


Fig. 1. Diagram to show the essential parts of the time base circuit

is to be examined. However, it is not suitable for a general purpose oscilloscope, because the amplitude of the time base is then a function of the frequency of the wave form to be examined. There is, however, a less well known type of time base which combines the advantages of the triggered and the free running types. References to it are to be found in Patents literature.[1,2] It is a triggered time base in which the amplitude of the sweep voltage is automatically maintained at a constant voltage, whatever the frequency of the wave form to be examined may be. If this time base is incorporated in a general purpose oscilloscope, then the only control which is required is a range switch, which selects one of three wide frequency ranges, inside which the frequency of the time base will be rigidly locked to that of the signal, and at the same time the amplitude of the time base voltage will remain constant. If a circuit which counts down independently of frequency is included, then the time base may be locked to any convenient sub-multiple of the signal frequency, and thus any convenient number of wave forms may be shown on the C.R.T.

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Description of the Time Base Circuit (Fig. 1)

The time base consists essentially of an inductance in series with a capacitor. A discharge device, (usually a valve), is connected across the capacitor. The discharge device is operated by:

1. The signal itself, or
2. An impulse derived from the signal, and occurring at the same frequency as the signal; or
3. By an impulse derived from the signal, and which has a frequency which is an exact sub-multiple of the frequency of the signal.

S is a switch which represents diagrammatically the discharge device. Suppose that the switch *S* be closed for a very short time, and then reopened. The capacitor will discharge quickly through *S*, and while *S* is closed the full H.T. of E volts is applied across the choke. If the time during which the switch is closed is so short that no appreciable current has been established in the inductance *L* before *S* again opens, then on reopening the switch *S*, oscillations of peak amplitude E volts will occur across *C* (see Fig. 2).

Suppose now that the switch be closed momentarily every time that the top plate of the capacitor reaches a potential of 2E volts. At this moment the current in the inductance *L* is zero, so that the momentary closure of the switch re-establishes the conditions which obtained at the first closure of the switch, and the wave form shown in Fig. 3 will be obtained at the junction of *L* and *C*.

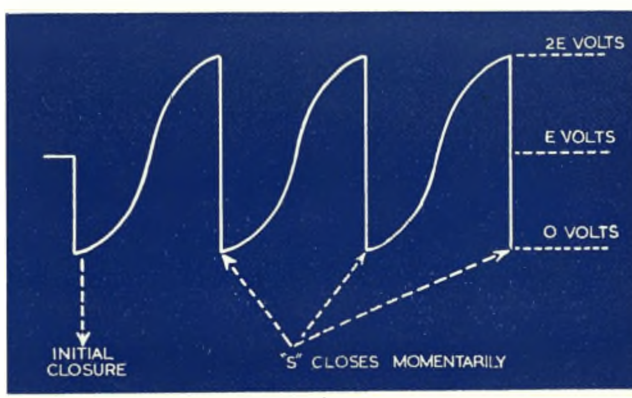


Fig. 3 (above). Shows what the wave form would be like if the period between successive discharges of the capacitor were equal to half the natural period of oscillation of the LC circuit

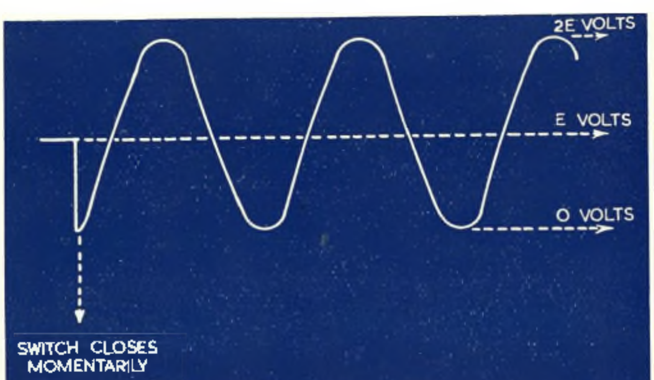


Fig. 2 (left). Shows the pure sine wave which would be generated if the circuit were shock excited by momentary closure of S in Fig. 1

The result is a *non*-linear time base of amplitude 2E volts. Suppose now that the second closure of the switch *S* occurs at some time before the voltage at the junction of *L* and *C* has reached the value 2E volts. Let the rate of change of voltage at the instant immediately before the second closure of the switch be dV/dt. Then at the moment of the second closure the current in the inductance *L* will be equal to

$C.dV/dt$, and if it is supposed that the time of closure of S is so short that there is no time for an appreciable change of current in L, then when S re-opens, a current of $C.dV/dt$ flows into the capacitor. It follows that the rate of change of voltage immediately after the second closure of the switch is equal to the rate of change of voltage immediately before this closure of the switch. It follows that by the time the switch closes for the third time the voltage will have risen to a higher value. After the next closure it will rise to a still higher value, until finally the voltage 2E is reached. This process is shown in Fig. 4.

It is clear that the voltage must finally oscillate between zero and 2E, because if there is no resistance in the choke then there can be no mean direct voltage across it. The parts of the wave form a, a′, a″, are portions of sine waves. If the period between the successive discharges of the capacitor is made small in comparison with half the natural period of oscillation of the LC circuit, then a, a′, a″, etc., will be very small portions of sine waves, and the time base will be substantially linear.

Suppose now that the discharge device is operated by the signal, or by a pulse derived from the signal, then whatever the frequency of the signal, the time base will be linear provided that half the natural period of oscillation of the LC circuit is very much longer than the reciprocal of the signal frequency. Fig. 5 shows a valve used as the discharge device. The valve is normally biased beyond cut-off. Positive pulses fed into the grid cause the valve to conduct. When this occurs the capacitor $C1$ discharges through the valve.

If no negative supply is available, bias may be derived from the grid current flowing in the grid leak. (See Fig. 6.) If $C2R$ is a long time constant, the positive pulses fed to the grid will be D.C. restored with the positive peaks at earth potential. In this case it will be advisable to include a fairly high value of resistance in the screen circuit, suitably decoupled. The mean screen current will be small, because the valve is only conducting for a small fraction of the total time, so that the screen potential will remain high. Should the input to the grid fail, the screen potential will drop to a low value, thus preventing the valve overheating.

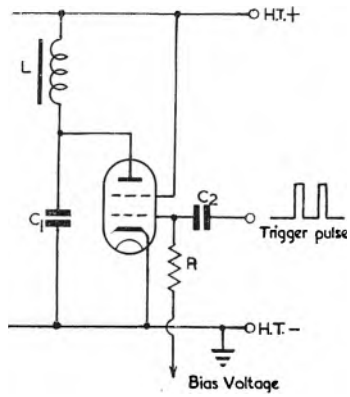


Fig. 5. Shows the circuit when a valve takes the part of S in Fig. 1

Practical Considerations : Design

Before calculating the values of L and C for a given speed of time base, it is necessary to find a formula showing how the mean current i through the inductance L depends upon the value of L, C, and the time base frequency required.

Let $f = 1/2\pi\sqrt{LC}$ be the natural resonant frequency of the series LC circuit.

Then $C = 1/4\pi^2 f^2 L$ .... (1)

Let $f = 1/t'$, so that t' is the period of one complete sine wave of oscillation.

Then $C = t'^2/4\pi^2 L$ .... (2)

Now we have seen that the duration of the time base t must be short compared with t'. Let it be n times shorter.

Then $t' = nt$

$t'^2 = n^2 t^2$ .... (3)

from (3) and (2),

$C = n^2 t^2/4\pi^2 L$.

The quantity n is important, as it determines the linearity of the time base. It will be shown later how n may be determined for a given degree of linearity, but let us assume for the moment that n has been made large enough for the time base to be substantially linear, i.e., values of L and C have been chosen which have a natural resonant period much greater than the time interval between successive discharges. Therefore dV/dt = constant = V/t where $V = 2E$ is the peak to peak time base voltage (see Fig. 7).

The steady current in the choke,

$i = C.V./t$

Therefore $C = it/V$ .... (5)

From (4) and (5),

$$\frac{it}{V} = \frac{n^2t^2}{4\pi^2 L}$$

$$i = \frac{n^2 tV}{4\pi^2 L}$$

It now remains to show how n may be calculated in terms of the required degree of linearity.

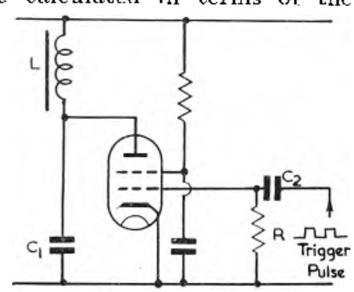


Fig. 6. Bias for the discharge valve is generated in the grid circuit. High resistance in the screen circuit prevents damage to valve should input to grid fail

Let us define the linearity of a time base as :

$$\text{linearity} = \frac{dV/dt\ (\text{minimum})}{dV/dt\ (\text{maximum})} \times 100$$

Fig 8 shows the time base represented as a portion of the theoretical sine wave from which it is derived. The voltage V is plotted against ωt.

$$\text{Then linearity} = \frac{\text{slope at } a}{\text{slope at } b} \times 100$$

suppose $n = 4$

$t' = nt = 4t$

$t = t'/4$

Now $\omega t'/4$ represents 90°, so that the time base represents a portion of a sine wave swinging 45° either side of zero. The parent sine wave is given by $V = V_0 \sin \omega t$

The slope $dV/dt = \omega V_0 \cos\omega t$

The linearity =

$$\frac{dV/dt \text{ when } \omega t = 45^\circ}{dV/dt \text{ when } \omega t = 0^\circ} \times 100$$

$= \cos 45^\circ/\cos 0^\circ \times 100$

$= 70.7$ per cent.

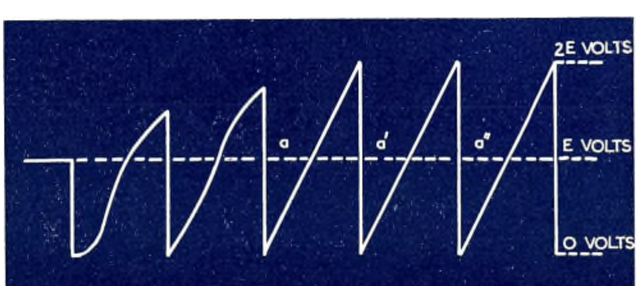


Fig. 4. Wave form when the period between successive discharges is shorter than half the natural period of oscillation

Repeating this calculation for n = 4, 8, 16, 32, we get:

n = 4, linearity = 71 per cent
n = 8, linearity = 92 per cent
n = 16, linearity = 98 per cent
n = 32, linearity = 99.6 per cent

For most purposes a linearity of 92 per cent will be sufficient, so the following calculations are based on the value n = 8. From the formula $i = n^2Vt/4\pi^2L$ we can find how the mean current i through the choke depends on the duration of the time base t and the inductance L. The following table shows how i depends on t and and L. It is calculated for V = 200, n = 8, from the formula given above.

This table shows that the time base is easy to design for frequencies of 1 kc/s. and higher, but that for frequencies as low as 10 c/s. difficulties due to the higher current requirements are met with. If a good ratio of forward stroke to flyback is required, say if the flyback time must be one-ninth of the forward stroke, then the current required to discharge the capacitor must be ten times the mean choke current.

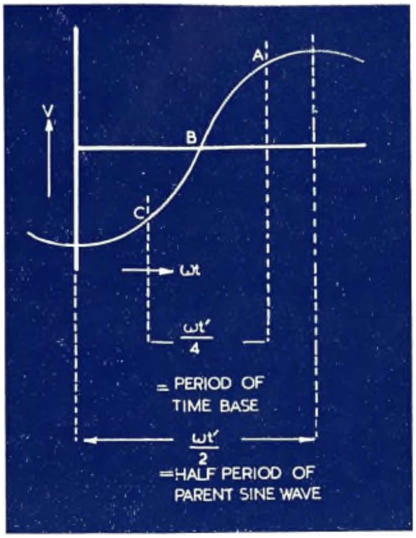


Fig. 8. This shows the parent sine wave as V plotted against ωt

The value of the capacitor is calculated very simply from:

$$t' = nt = 2\pi\sqrt{LC}$$

The lower table shows the value of C for different values of t and L.

Mean current at 100 kc/s. = 16.0 mA.

Say peak current available = 11 × 16 = 176 mA

Then at 10 kc/s. forward/back ratio = 100 to 1,

and at 100 kc/s. forward/back ratio = 10 to 1.

The linearity is 92 per cent at 10 kc/s., and is better than this at the higher frequencies. The time base will follow frequencies higher than 100 kc/s., but then the forward to back ratio will suffer.

Modification of the Circuit for Use at Low Frequencies

It is well known that if an impedance is connected in a cathode follower circuit (see Fig. 9), then the A.C. impedance looking in at terminals 1 and 2 is given by $Z(A + 1)$, where A is defined as the ratio of the voltage change at the cathode to the voltage change between the grid and the cathode. The equivalent input circuit is therefore as shown in Fig. 10. Since at any frequency the impedance of the equivalent choke will be given by $\omega L(A + 1)$, we may replace the choke by one whose inductance is given by $L(A + 1)$. Vc represents the voltage at the cathode in the steady state, in the absence of any D.C.

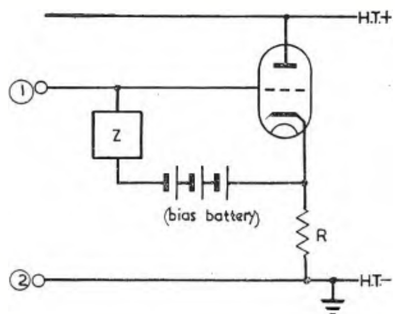


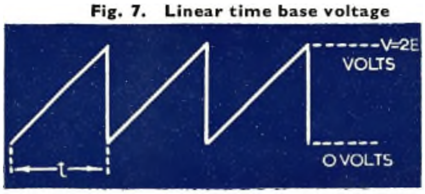
Fig. 9. Method of increasing the effective value of an impedance

f.	t	i when L = 10H	when L = 100H	i when L = 1000H
100 Kc/s.	10 μ sec.	0.32 mA	0.032 mA	0.00032 mA
10 Kc/s.	100 μ sec.	3.2 mA	0.32 mA	0.0032 mA
1 Kc/s.	1 m.sec.	32 mA	3.2 mA	0.032 mA
100 c/s.	10 m.sec.	320 mA	32 mA.	0.32 mA
10 c/s.	100 m.sec.	3.2 amps.	320 mA	3.2 mA

There is no doubt, therefore, that this simple time base, as it stands, is not suitable for use as a slow time base. It is, however, ideally suited to time bases of the order of 100 micro-seconds. The absence of controls should make it suitable for use in the line scan of television receivers.

On following through the derivation of the formula from which the table has been calculated, it will be found that the fundamental reason for the high current requirements at low frequencies is the square root sign in the formula $f = 1/2\pi\sqrt{LC}$. Later it will be shown how by means of one additional valve the circuit may be made to operate satisfactorily at low frequencies.

Fig. 7. Linear time base voltage



These tables have all been based on n = 8, linearity 92 per cent. If linearity 98 per cent, n = 16, is required the values of current and the capacitor values will be four times greater.

Suppose a 100 micro-second time base is required, forward to back ratio 10 to 1, linearity 98 per cent, then the following values will be required: L = 10 H, 12.8 mA, C = 0.0064 micro-farads. The valve must pass a peak current of 11 × 12.8 = 140 mA. If a circuit is required which will automatically follow a frequency from 10 kc/s. to 100 kc/s., then the following values will be correct. They are calculated in the same way as before, for n = 8, linearity = 92 per cent:

L = 20 H, C = 0.0008 micro-farads.

Mean current at 10 kc/s. = 1.6 mA

f.	t.	Value of 'C' in microfarads for:— L = 1H	L = 10H	L = 100H	L = 1000H	L = 10,000H
100 Kc/s.	10 μ sec.	0.00016	0.000016	—	—	—
10 Kc/s.	100 μ sec.	0.016	0.0016	0.00016	0.000016	—
1 Kc/s.	1 m. sec.	1.6	0.016	0.016	0.0016	0.00016
100 c/s.	10 m. sec.	—	1.6	1.6	0.16	0.016
10 c/s.	100 m. sec.	—	—	—	—	1.6

connexion between 1 and 2 of Fig. 9.

It has already been shown that a limitation to the usefulness of the time base circuit at low frequencies was set by the large value of inductance required. We may use the equivalent inductance of a choke connected in a cathode follower circuit instead of an actual choke. The complete time base circuit is now as shown in Fig. 11. $V1$ is the cathode follower. $V2$ is the discharge valve.

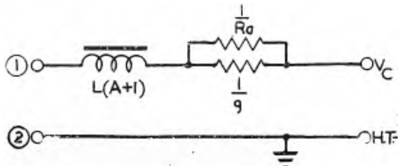


Fig. 10. Equivalent circuit diagram of Fig. 9

The value of r is adjusted so that in the steady state the potential of the cathode is about one-third of the H.T. voltage. Then when the time base is operating, the cathode will swing from a potential which is a few volts above earth potential to a potential equal to two-thirds of the H.T. voltage. There is now only one-third of the H.T. voltage across the valve, and a valve must be chosen capable of supplying the required steady choke current at this reduced voltage. Owing to the fact that the equivalent value of the choke is now very large, this current is very small. In order to obtain a higher value for A, a pentode may be used, but in this case the screen must be decoupled to cathode. When this is done the A.C. cathode load is equal to the resistance of the screen and cathode resistances in parallel.

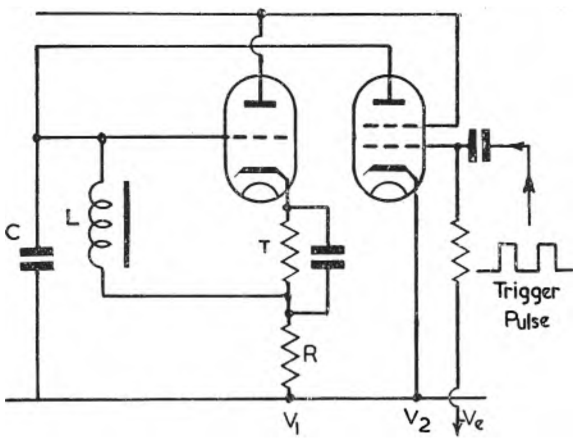


Fig. 11 (left). Time base circuit, using a triode to increase the effective value of the inductance L

The resistance R may be given the value infinity, i.e., it may be omitted from the circuit, as a D.C. return path is provided by the discharge valve. A value of 100 may be obtained for A. Taking L = 100 Henries. Equivalent L = 10,000 Henries, and substituting in the formula $i = n^2Vt/4\pi^2L$, putting $n = 8$, $V = 200$, $t = 1/25$ seconds, we get:

$$i = \frac{64 \times 200 \times 10^3}{4 \times 10^4 \times 25} \text{ milliamps}$$

$$= 1.3 \text{ milliamps.}$$

This shows that a very slow time base may easily be constructed using this modified circuit. Unfortunately the voltage output is no longer equal to twice the H.T. voltage, but is now only two-thirds of the H.T.

Triggered and Strobe Time Bases

In certain applications such as radar, a time base of short duration recurring at comparatively long intervals may be required. This can also be produced by a simple modification of the above circuit, as shown in Fig. 12. It will be seen that $V1$ operates on zero grid bias, and a resistance r' is included in the grid circuit. The action when $r' = 0$ will be: In the steady state, since $V1$ is at zero bias, the cathode of $V1$ will approach H.T. and grid current will flow in L. After the first discharge of C through $V2$, the grid of $V1$ will be near earth potential, and the top of R will be a few volts higher. There is thus a small voltage across L, and current will grow in L, C will charge, the grid of $V1$ will rise in potential, the cathode will follow, being always a few volts higher than the grid. Thus the potential across L is maintained, and is always in the same direction, the cathode end of L being always positive with respect to the grid end. This process is maintained until the cathode of $V1$ approaches H.T. Grid current now flows, and maintains the current in L. Grid and cathode now remain at a voltage near H.T. + until the next discharge. The same process is repeated, except that the rise in voltage is now more rapid, due to the fact that current is already flowing in L. During the next rise, the current in the choke is again increased until the cathode nears H.T. +, when the choke current is again maintained by grid current.

During the grid current period there is some decay of current in L, on the time constant L/r'', where r'' is the resistance of the choke and the resistance of the cathode to grid path of the valve. The steady state is reached when the decay of current during the grid current period is equal to the rise during the charging period.

The resistance r' is included to control the decay during the grid current period. Values of r' are difficult to calculate, but it is a simple matter to find a suitable value for r experimentally. L and C are calculated as before. The wave form obtained is shown in Fig. 13.

Black Out Pulse for the Fly Back

Owing to the rapid fly back of the time base, for many purposes it is not necessary to black out the trace during the fly back. Should it be necessary, however, all that is required is to feed the positive pulses which are applied to the grid of $V2$ to the cathode of the C.R.T.

The Counted Down Time Base

If the time base is required for use in a general purpose oscilloscope, the circuit described above is not very suitable, as only one waveform will be displayed, and part of that

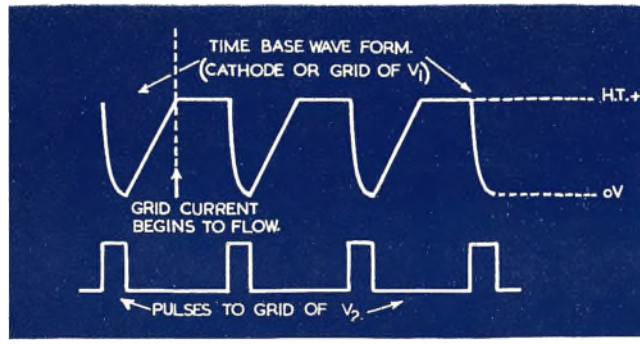


Fig. 13 (below). Wave form of strobe time base

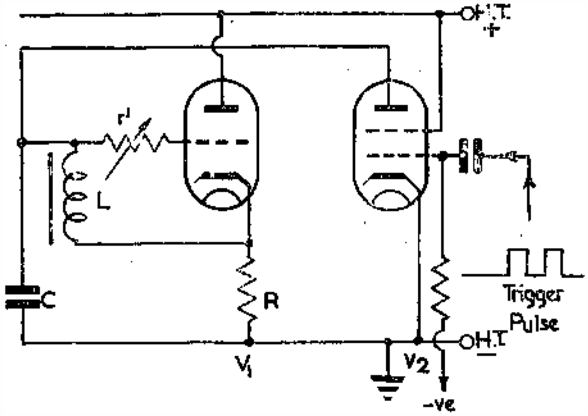


Fig. 12. Strobe time base. r′ controls the velocity

seen that two waveforms are displayed on the time base. While this is all that is required to examine the whole waveform, it has been found in practice that people who are used to the conventional time base circuits are apt to be unsatisfied by only two waveforms. It is purely a psychological effect, but in view of this, it may be found advisable to include a second counting stage (not shown in the diagram), so that four waveforms are displayed.

Alternatively, other methods of counting down independently of frequency may be employed; for example, methods in which the counting down ratio is determined by the ratio of the capacitances of two capacitors. This has the advantage that the number of waveforms can be conveniently altered from any value between one and ten or more.

will be obscured by the fly back. A circuit suitable for use in a general purpose oscilloscope is shown in Fig. 14. The action of the circuit is as follows.

V1. Double Triode Squarer

This is a directly coupled flip-flop circuit. The resistances are so designed that in the absence of a signal, the right half of the valve is conducting, this maintains the common cathode at a potential a few volts higher than the potential (80 V) to which the grid leak of the left half is returned. The left half of the valve is thus cut off. When a signal is applied to the grid of the left half of $V1$, the grid potential will rise, the left anode falls, this cuts off the right half, and the circuit tips over into its second stable condition due to positive feed back via $R6$. On the negative half of the signal the reverse process occurs. Thus any waveform fed into the left grid of $V1$ will result in square waves at the right anode.

This circuit will operate on almost any waveform. It will not operate, however, on short negative pips. To extend the scope of the circuit to cater for negative pips, a switch may be included to raise the potential to which $R1$ is returned. If this is made more positive, then the left half conducts in the steady state, and is cut off each time a negative pip occurs. Thus short positive pips are produced at the anode of $V1$. As an alternative to this circuit a straightforward overloaded amplifier may be substituted.

V2. Double Triode Counting Down Stage

This a directly coupled flip-flop of conventional design. The circuit has two stable conditions, with either the left half or the right half conducting. The positive square waves from the right anode of $V1$ are differentiated by $C2R13$, and the positive and negative pips so produced are fed to the common cathode of the two diodes. The positive pips have no effect as they do not pass through the diodes.

Suppose the left half of $V2$ is conducting. The grid of the left half of $V2$ is held at earth potential by grid current flowing in $R10$, $D1$ is conducting, $D2$ is cut off. When a negative pip is fed to the cathodes of the diodes, it passes via $D1$ to the grid of the left half of $V2$, and the circuit tips over. The next negative pip passes via $D2$ to the grid of the right half, and the circuit again tips over. This circuit therefore counts down by two independently of frequency.

The square waves at the right anode of $V2$ are differentiated by $C3R14$, and fed to the grid of the discharge valve. The waveforms are shown in Fig. 14. It will be

Indication of Frequency

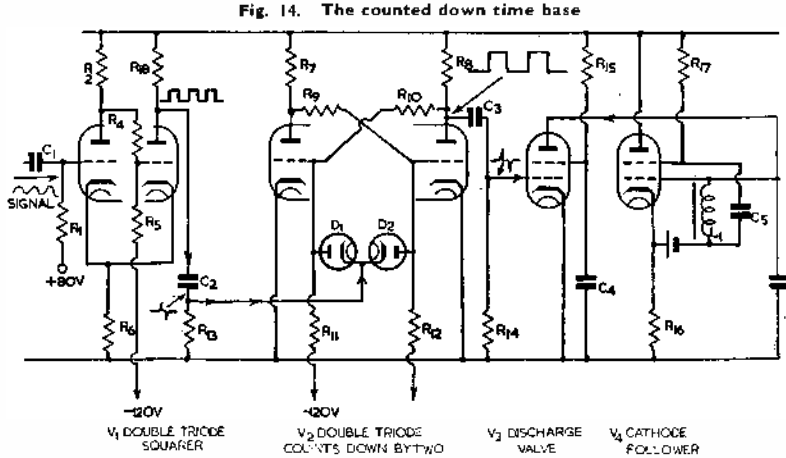
When using a conventional time base, the setting of the controls gives an approximate indication of frequency. In this time base there is only one coarse control to switch in values of L and C appropriate to the range of frequencies to be covered. Since this range is large, the operator has little idea of the frequency of the signal he is examining.

A D.C. milliammeter, suitably calibrated as a frequency meter may be included in the discharge circuit of the capacitor $C6$. The current in this circuit is directly proportional to the time base frequency.

References.

[1] Patent specifications 409, 571 ; 491, 934 ; 489, 275 458, 746 ; 471, 737.
[2] Patent Application No 15,836/47.

Fig. 14. The counted down time base



The Miller Circuit as a Low-Speed Precision Integrator

By I. A. D. LEWIS, B.A., A.Inst.P., Grad.I.E.E.*

THE theory of the Miller circuit as used for integration and the generation of various types of waveform at high repetition rates has been fully described in the literature.[1 ... 7] It has also been applied to low speed integration[8,9] and this discussion, in which sources of error are considered in some detail, may be of further interest. The analysis includes the effects of stray capacitances in the circuit and is therefore also applicable to high speed integrators.

Typical Arrangement

An instrument has been designed and constructed to integrate a varying voltage over a period of hours with an accuracy of within 0.2 per cent over most of the range. Results are obtained as counts on a mechanical counter and in this particular instrument a constant input potential of 10 V (the maximum value) corresponds to a counting rate of 3 per sec.

The circuit, shown schematically in Fig. 1, is fully described elsewhere.[10] In order to improve the linearity the effective gain of the Miller stage $V1$ is increased by the addition of a two valve directly coupled amplifier. The anode potential of $V1$ falls in the normal manner, on applying a positive input voltage. When a low value is reached a flip-flop circuit, which is directly-connected to $V1$, resets the integrator to the initial state. A pulse is also passed to the mechanical counter which registers one cycle of integration. The total number of counts registered in a given period of time is clearly proportional to the time integral of the input voltage. (It is possible to interpolate between whole numbers of counts by means of a voltmeter connected between the anode of $V1$ and earth.)

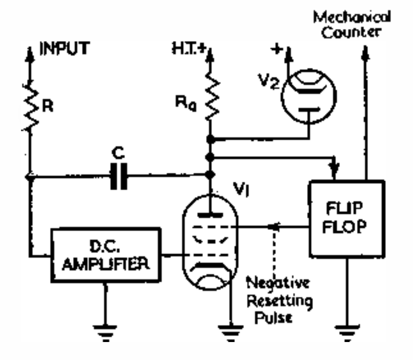


Fig. 1. Skeleton diagram of a repetitive integrator using the Miller principle

Such an instrument is subject to errors caused by (i) drifts (ii) non-linearity in time (assuming all the circuit elements to be linear) and (iii) the operation of resetting between cycles of integration. These will shortly be considered in turn.

Notation

For the purpose of analysis the circuit shown in Fig. 2 is used. The effect of a time delay, Ta, say, in the multi-stage amplifier (such as might be introduced to eliminate instability) is allowed for by including a factor $1/(1 + pTa)$ in the gain; p is the operator d/dt. C, R and Ra have the usual meanings and $R2$ represents any resistive load which may be connected across the output. The resistive component of the output impedance of the amplifier, R' say, is given by:

$$1/R' = 1/Ra + 1/R2 + 1/ra$$

where ra is the anode slope resistance of the valve. $C1$ and $C2$ are inter-electrode or other capacitances and R_L the leakage resistance of the capacitor C. Grid current is represented by the constant current generator I and the grid input resistance of the amplifier by $R1$. The lower ends of I and $R1$ may be imagined to be connected to a point at a potential, Vgo, say, equal to the working value of the grid voltage corresponding to some standard anode potential. The D.C. gain of the amplifier, with C and R_L disconnected, is denoted by G.

Drifts

We first consider single cycle, i.e., non-repetitive, integration and assume that $V1$ is conducting normally. The integrator can be caused to remain stationary, initially, by a suitable choice of the zero value of V determined by the condition that the algebraic sum of the currents flowing towards the lower plate of C be zero. Since the current I will not necessarily remain constant it should be small. An ordinary valve operating under special conditions may be employed[11] or, alternatively, an electrometer type used if a very long period of integration is desired. R_L may also change and in addition the leakage current through R_L will vary as Va changes; R_L should thus be large and the drift time cannot be made longer than the order of the natural leakage time of the feedback capacitor C. The effect of R_L is also included in the analysis below.

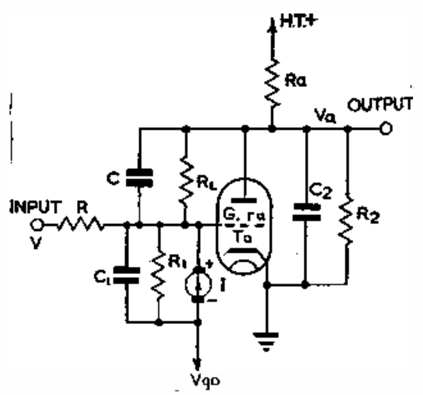


Fig. 2. Integrating circuit considered for analysis

The ordinary drift associated with D.C. amplifiers will occur and can be reduced by the use of a balanced circuit;[12] it is equivalent to a variation, D say, in the working value of the grid voltage (for some standard anode potential).

If $\triangle I$, $\triangle R_L$ are the expected variation in I and R_L, and K, $\bar{V}$ are the normal swing in anode voltage and the mean value of the input potential respectively, then the fractional error is of the order of:

$$\frac{1}{\bar{V}}\left[R(\triangle I + K/2R_L + Va\triangle R_L/R_L^2) + D\right] \quad \ldots\ldots (1)$$

Clearly $\bar{V}$ should be as large and R as small as possible; R is limited by the value of the output impedance of the source feeding the integrator and by consideration of the constancy of this impedance. A further apparent source of drift is mentioned below.

Non-Linearity

Let V be measured with respect to the zero determined above. If $V=0$,

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$Va = Vao$, say, at time $t < 0$ we require the form of the output voltage when a constant input V is applied at $t = 0$. It can be shown that the following relation holds just after $t = 0$:

$$Va(t) = Vao + \frac{1}{pR(C_1 + C_2 + C_1C_2/C)}.V \quad \text{.........} \ (2)$$

For later values of t we have:

$$Va(t) = Vao - \frac{1}{pT}.$$

$$[1 + a_2p^2 - bp - c - d/p + e_2/p^2] . V \quad (3)$$

Here $T = CR$ and the constants involve the circuit parameters. The following assumptions are made as to orders of magnitude:

$$R1\ R_1 \gg R;\ R' \ll R;$$
$$C1\ C2 \gg C;\ Ta \approx CR;\ G \gg 1$$

In deriving (2) no quantities are neglected, but third order small terms are omitted from (3). The latter is legitimate except in the case of positive powers of p for times just after $t = 0$. We write:

$b = b_1 + b_2$, $c = c_1 + c_2$, $d = d_1 + d_2$

where the suffixes 1 and 2 denote small quantities of the first and second orders respectively. The following values obtain:

$b_1 = Ta/G$
$c_1 = (1 + Ta/T)/G$
$d_1 = 1/GT + 1/CR_L$
$a_2 = b_1(b_1 - CR')$
$b_2 = CR'/G + (R'/R + C1/C - 2c_1)b_1$
$c_2 = (R'/R + C1/C)/G + (R/R_1 - 1/G - d_1T)b_1/T - c_1^2$
$d_2 = (R/R_1 - 1/G - d_1T)/GT - 2b_1d_1/T$
$e_2 = d_1^2$

If second order small terms are neglected (3) has the solution:

$$Va(t) = Vao - [-b + (1 - c)t - dt^2/2] . V/T \quad \text{..................} \ (4)$$

The output is not a linear function of time and the case for a rectangular shaped input is sketched (in exaggerated form) in Fig. 3. It is noticed that a "drift" occurs after the input falls back to zero due to the fact that since G is finite a change in output voltage implies a change in grid voltage from the initial value.

We also require the solution of (3) in the form in which t (or rather its reciprocal) appears explicitly. Putting $Vao - Va = K$ and again neglecting second order small terms we have:—

$$1/t = [1 - c - bV/KT - dKT/2V] + V/KT \quad \text{...........} \ (5)$$

Two ways of using an integrator are possible. We may measure either (i) the change in output voltage that occurs in a given time, or (ii) the time taken, t_1, say (or its reciprocal), for the output to change by a given amount K. An instrument of the latter type, like the one constructed, tends to be less accurate than one of the former, owing to the non-linearity in time. For small values of V, i.e., for large t_1 the error due to the d term increases. If t_1 is not too large, such that $dt_1^2/2 \ll 1$, equation (5) applies and the error in $1/t_1$ is simply a constant $-d/2$.

Arbitrary Input

If G and R_L were both infinite, all the error terms in (3) would disappear and we would then have the ideal result for an arbitrary input $V(t)$:—

$$\int_0^{t_1} V(t).dt = KT \quad \text{...........} \ (6)$$

When the errors are taken into account the output for an arbitrary input can be obtained from the solution for a step function input (4), by means of the Duhamel superposition theorem[13]. We then find:—

$$\int_0^{t_1} V(t).dt = \frac{KT}{1-c} + \frac{bV(t_1)}{1-c} + \frac{d}{1-c}\int_0^{t_1}\int^{t_1} V(t).dt.dt' \quad (7)$$

It is noted that no errors arise peculiar to any particular form which $V(t)$ may take.

Resetting

When the integrator is reset to the initial state between cycles an error arises due to the cessation of integration during the resetting periods. For the drift rate due to leakage and grid current to be small C should have as large a value as possible, but an upper limit is fixed by the maximum resetting time allowable.

The calibration of the instrument is determined to the first order by the quantities T and K only (6). The constancy of the latter depends on the stability of (i) $V2$ and its supplies, and (ii) the operating level of the flip-flop circuit; care must be taken to ensure that variations in these are negligible. In addition, there is a further source of variation in K. When, during the resetting periods, the anode potential of $V1$ has reached the value determined by $V2$, the grid potential falls to a value V_x, say. This is determined by the input current, which depends on V, and the grid voltage/grid current characteriste of the amplifier. If r_g is the slope of the grid characteristic curve in the region of grid currents of the order of V/R, then V_x is subject to an approximate variation of Vr_g/R (provided $V \gg V_x$). When the resetting pulse is over, anode current flows in $V1$ and the anode potential drops rapidly by a few volts in the usual manner. The magnitude of this drop is determined by the difference between V_x and the normal working value of the grid potential.* K is thus liable to a variation Vr_g/R.

Error Corrections

It is possible to correct for most of these errors provided they are

* In the stationary case the latter has the value V_{go} but when an input is applied it has a variation $R'V/CR$ due to the fact that $V1$ has to supply the current through C in addition to that through Ra. The change in the initial drop from this cause is negligible; the effect during the remainder of the cycle is allowed for in the previous analysis and, in any case, only the calibration is modified.

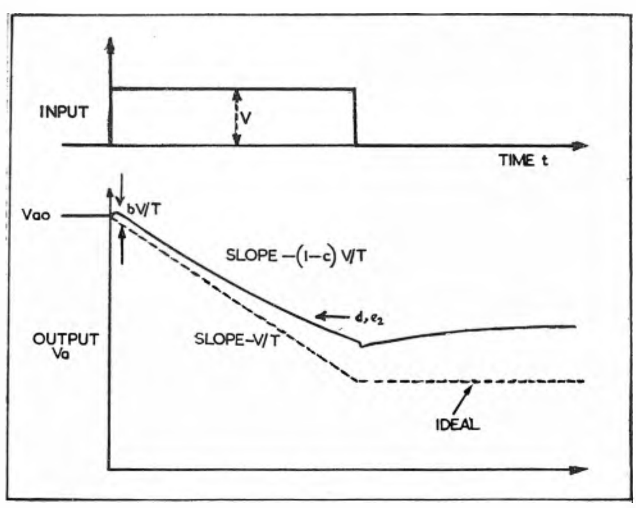

Fig. 3. Sketch illustrating how the response to a rectangular-shaped input waveform, of the circuit in Fig. 2 departs from that of a perfect integrator

Table of Design Procedure

Relevant Quantities Known	Quantity to be Determined	Relation Used
Source impedance and constancy, accuracy required	R_{min}	—
R	$R_1 min$	$R_1 \gg R$, expressions for c_2 and d_2
R, $\overline{V}_{min}$, accuracy required	$\triangle I_{max}$	1
$\overline{V}_{min}$, accuracy required	D_{max}	1
Accuracy required, constancy of K, convenience	K	
$\overline{V}_{max}$, $t_1 min$, K	T	6, assuming $c \ll 1$
T, R	C	$C = T/R$
$\overline{V}_{min}$	t_{max}	6
$t_1 max$, accuracy required	d_{max}	5, $dt^2_1 max/2 \ll 1$
R, K, d, } accuracy required	$R_L min$, $\triangle R_{L_1}$	1, $R_{L_1} \gg R$; $d_1 \approx d = 1/CR_L + 1/GT$ assuming $d_2 \ll d_1$
T, d	G_{min}	$d \approx d_1 = 1/GT + 1/CR_L$ assuming $d_2 \ll d_1$
V_{max}, K, T, $Max\|n-m\|$ } accuracy required	b_{max}	7; 8
G, b	$T_a max$	$T_a < T$, $b \approx b_1 = T_a/G$ assuming $b_2 \ll b_1$
$Max\|n-m\|$ accuracy required	$T_r max$	8
T_r, C, R_2 ... ... ; R	$R_3 max$; $R' max$	$1/R_3 + 1/R_4 > C/T_r$; $R' \ll R$, expressions for a_2, b_2 and c_2
C, $Max\|n-m\|$ accuracy required	$r_g max$	8
b_1, etc.	$C_1 max$	$C_1 \ll C$, expressions for b_2 and c_2
C	$C_2 max$	$C_2 \ll C$

small. Let N be the observed total number of counts, in a given time, and N_o the true value. It may then be assumed that, on the average, V has a value proportional to N_o during the resetting periods. It can be shown, partly with the aid of (5), that:—

$$N_o = N\,[1 - n_z/n + (n - m).(b + T_r - Cr_g)] \quad \ldots\ldots\ldots (8)$$

where n is the observed counting rate, n_z the "zero" counting rate and T_r the duration of the resetting period. The counting rate m corresponds to the value of the constant input at which the instrument is calibrated (say at about two-thirds maximum rate); the reading is taken as being exactly correct at this point, after allowing for n_z.

The zero counting rate is simply additive and need not necessarily be very small provided it is known; the error $-d/2$ in $1/t1$ (i.e., in n) obtained from (5) cancels when the zero rate is subtracted. In accurate work the integrator should be adjusted so that a definite drift rate exists. It is then possible to measure any change between the beginning and the end of an experiment; a change in the negative direction could not be measured if the drift rate were initially made zero.

The performance of the instrument can be assessed by reference to equations (1), (7) and (8).

Design Procedure

A guide to the order in which the various parameters should be determined is given in the table. Quantities assumed to be given are (i) the magnitude of V, (ii) the source impedance, and (iii) the overall accuracy required. The maximum counting rate, $1/t_{min}$, is determined by the mechanical counter to be used.

References

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[13] J. R. Carson: "Electric Circuit Theory and the Operational Calculus," McGraw-Hill, 1926, p. 16.

Electron-Optical Shadow Method

AS the result of a series of electron-microscope experiments at the National Bureau of Standards (U.S.A.), Dr. L. L. Marton, of the Bureau's electron physics laboratory has developed an electron-optical shadow technique* that provides a valuable tool for the quantitative study of electrostatic and magnetic fields of extremely small dimensions. The new method makes use of an electron lens system to produce a shadow image of a fine wire mesh placed in the path of the electron beam. From the distortion in the shadow network caused by deflexion of the electrons as they pass through the field under study, accurate values of field strength are computed. Thus it is possible to investigate quantitatively fields that have not been susceptible to other methods of investigation, for example, the fringe fields from the small domains of spontaneous magnetisation in ferromagnetic materials.

The new development, which is based on extensive theoretical analysis, should provide a powerful means for broadening present knowledge concerning space-charge fields, fields produced by contact potentials, patch fields in thermionic emission, charge distribution in a gaseous plasma, waveguide problems and the basic magnetic properties of metals. Though similar in some respects to the electron-optical Schlieren method† previously developed at the Bureau, the shadow method is much better adapted to

* For further technical details, see "Electron Optical Observation of Magnetic Fields," by L. Marton and S. H. Lachenbruch, *J. Research*, NBS 43 (Oct. 1949), RP2033. See also "Electron Optical Mapping of Electro-magnetic Fields," by L. Marton and S. H. Lachenbruch, published in *J. Applied Phys.* 20 (Nov. 1949).

† In the Schlieren method, a magnetic lens forms an image of a source of electrons on a small copper stop that intercepts all direct rays. If in the space between the electron source and the lens there is a variation of the index of refraction for electrons—in other words, a variation in electric or magnetic

precise determinations of field intensity.

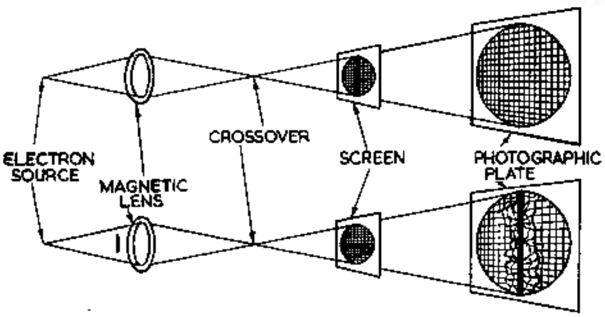
Principle

The principle of the shadow method was discovered in the course of a study of a recording wire magnetised in evenly spaced short pulses by means of a conventional magnetic recording head. In practice, the recording wire—or other object to be studied— is placed between an electron source and a system of electron lenses. The lens system focuses the electron beam to form an image of the wire on a fluorescent screen. By placing a wire mesh of known gauge just beyond the back focus of the lens system, a shadow image of the mesh is superposed on the image of the wire. This shadow image is formed by projection from the virtual source provided by the reduced image of the source of electrons. The portions of the shadow network adjacent on the screen to magnetised regions of the recording wire are then found to show considerable distortion.

A complete theoretical analysis of this effect has shown that the distortion of the shadow image is due to the deflexion of the electron beam by the field of the recording wire at each magnetised region. The result is a corresponding displacement of the reduced image of the electron source. This displaced image, acting as a virtual source, forms a shadow image, likewise displaced, of the network. Deflexion of the beam may also change the distance of the virtual source from the wire, in which case the magnification of the displaced image is affected.

Obviously, the displacement and change in size of the shadow image at any point depend on the strength of the field of the magnetised wire at a corresponding point.

Formulae have been derived by S. H. Lachenbruch, of the Bureau staff, that permit the calculation of consistent absolute values of field strength in magnetic or electric fields of various geometries from experimental measurements of the position of the wire mesh, the displacement and magnification of its shadow image, and the known constants of the apparatus. The patterns obtained also provide a qualitative visual representation of minute electrostatic magnetic fields. Although it is possible to compute field intensity from the intensity distribution of the pattern obtained by the Schlieren method, the shadow method is of far greater utility for quantitative work, since the image displacement and magnification can be measured much more accurately than can the intensity distribution across the Schlieren pattern on a photographic plate.



Principle of the electron - optical shadow method. The source of electrostatic or electromagnetic field is shown at the left of the lower diagram

Applications

Perhaps the greatest value of the electron-optical shadow method lies in its utility for exploring complex electric and magnetic fields of extremely small dimensions, or in which a probe of size greater than the electron would disturb the field under study. In the past, calculations of the field intensity at a point have been limited to those special cases in which the geometry of the field exhibits a high degree of symmetry. The shadow technique now provides data for accurate calculation of the absolute value of the intensity in the neighbourhood of a specimen of any size or shape without altering or disturbing the field.

The method is thus well adapted to investigation of the fundamental nature of ferromagnetism. Experiments now under way at the Bureau include a study of the behaviour of the fringe fields of the ferromagnetic domains; in this work a single crystal of cobalt having very large magnetic domains is being used. An extension to ferroelectric materials is also contemplated for the purpose of checking the domain theory of these substances; of particular interest will be a study of the polarisation of barium titanate and other high-dielectric materials, that are now being widely used in the production of small-sized capacitors for radio, radar and television.

In another application of the shadow method at the Bureau, space-charge fields in several types of apparatus employing electron beams are being investigated. In this connexion, use of the method with a pulsed electron source for the stroboscopic study of fields that vary with time is under study.

It has been suggested that the electron-optical shadow method may also be of value for the calculation of field intensities within a waveguide. Use of waveguides as conductors and circuit elements in ultra-high-frequency radar and communication often leads to arrangements whose geometry is too complicated for expression in any system of mathematical co-ordinates. Thus the electronics engineer, having in many cases only an intuitive picture of the field distribution at junctions and elbows of the guide, must rely on empirical methods in designing waveguide techniques and equipment. By the use of suitable auxiliary techniques, it is hoped that the shadow method may be adapted to the calculation of field intensities in regions of a guide that are not at present susceptible to analytical treatment.

The Bureau is also applying the principle of the shadow technique to the study of spherical aberration in electron lenses. When a fine wire mesh is placed in the focal region of a lens having spherical aberration, the shadow image of the network is enlarged either centrally or at the periphery, depending on the position of the mesh and the nature of the lens error. The resultant pattern may thus be interpreted to give information of value in correcting the lens.

—From the *National Bureau of Standards Technical News Bulletin*, **33**, 9 (September 1949).

field intensity—an image of that inhomogeneity will then be produced by means of the same lens in a conjugate plane beyond the stop. Thus a dark-field image of the magnetic or electric field is obtained on a fluorescent screen or on a photographic plane. See *Electronic Engineering*, 20, 330 (1948).

Measuring Steady Bioelectric Potential Differences

By W. T. CATTON, M.Sc. *

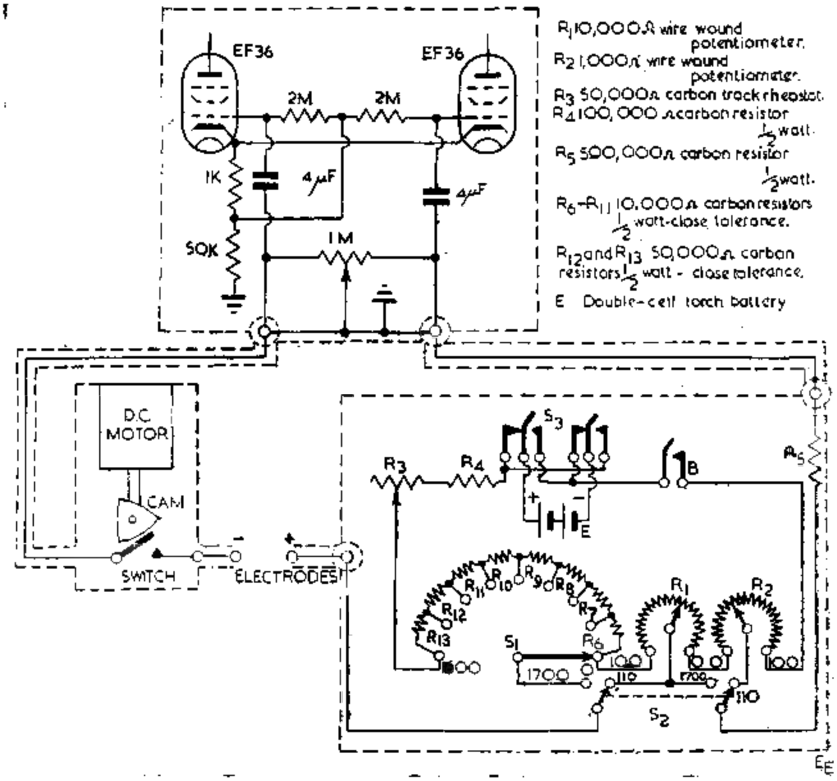


Fig. 1. Input circuit, switch and potentiometer

A SIMPLE "chopper" circuit and potentiometer can be used as an accessory to a biological amplifier for the measurement of steady bioelectric potential differences. The principle of the method is as follows. The input to the amplifier (a D.C. potential of small amplitude) is interrupted periodically at low frequency (about 25 times per second) by means of a mechanical switch introduced into the circuit. The output of the amplifier (rectangular pulses whose height depends upon the amplitude of the input potential difference) is presented on a cathode-ray oscilloscope, or alternatively a rectifying and smoothing circuit is added to the amplifier so that a steady D.C. potential is developed across a normal moving coil meter (Fig. 2). There is further introduced in series with the input circuit, a simple graduated potentiometer, able to provide a range of D.C. potentials from 0 to 1,700 mV. The potentiometer is connected in opposite polarity to the source (Fig. 1). In order to take a reading, the potentiometer is adjusted until the pulses disappear from the cathode-ray tube time-base, or until a zero or minimum reading is shown on the output meter; the potentiometer dials then give the value of the D.C. potential difference across the source.

A "backing-off" method was chosen in preference to that of making direct readings on the output meter or C.R.O., for the following reasons. It was found experimentally that the amplitude of the output pulses was proportional to a given input D.C. E.M.F. over only a limited range of inputs, since overloading of the later stages of the amplifier occurred at the higher input levels. In any case the output level was open to variations due to (a) alterations of amplifier gain from any cause, such as fluctuations of mains voltage, and (b) the variable amount of interference present at the input terminals, such interference reading as "noise" on the C.R.T. or as a constant reading on the meter in the absence of pulses. By using a "backing-off" method the following advantages are gained. (a) No current is drawn from the source, so that non-polarisable electrodes are not essential. (b) The system is largely independent of amplifier gain, spurious interference and variations in switching frequency or pulse duration within certain limits to be discussed later. (c) The ultimate accuracy of the readings depends upon the potentiometer, which may be made as accurate as is desired, or as is justified by the nature of the experimental application.

The mechanical switch employed is an Air Ministry surplus Type 156, Ref. 10FB/916, originally forming part of a radar equipment. The contacts are cam-operated, the cam having three "lifts" per revolution, and being driven by a 24 V motor. In the present application, one contact only is used (of four available) and the motor is supplied with 16 V D.C. from a rectifier, so that it runs at less than maximum speed. The actual speed of rotation is approximately 8 revolutions per second, giving a frequency of interruption of 24 per second. The shape of the cam and spacing of the contacts are such that the switch is open or closed for about equal periods of time, in this case about 20 milliseconds. It was found by experiment that small variations in motor speed did not affect the readings obtained, and in any case the motor maintains such a constant speed that the time-base of the C.R.O. remains in synchronisation for long periods with only a small amount of "*Sync*" control. It is a desirable feature of the system that both the pulse duration and the inter-pulse time should be considerably less than

* Medical School, King's College, Newcastle-upon-Tyne.

the time constant of the R.C. coupling circuits of the amplifier, so as to avoid rounding-off the pulses. In the amplifier employed, the coupling circuit time-constant is 100 milliseconds, the pulse duration 20 milliseconds, and the pulses show only a slightly sloping "top," with no rounding-off, as viewed without the loading of the rectifier and meter circuit applied to the output. With this latter circuit applied, some rounding of the pulses does inevitably occur, but since zero readings are being employed this is not important. In the practical arrangement, the meter and rectifier circuit are permanently mounted on the mains-driven section of the amplifier, and a switch ($S4$) is fitted, so that this circuit may be cut out when the amplifier is being used to study action potentials of nerve or muscle. There is no difficulty in using the meter separately from the amplifier in any convenient position.

A brief description of the amplifier employed follows. It is in two sections, the first section supplied by batteries only, the second entirely mains-driven. All stages employ two valves in push-pull. In the battery section the first stage makes use of two EF36 valves, with a common undecoupled cathode resistor, to increase discrimination against symmetrical interference potentials appearing at the input. The input impedance is 1 megohm grid to grid. In the second stage two EF50 valves are used, followed by a 6SN7 double-triode connected as a double cathode follower. The R.C. coupling in this section, as in the mains section, consists of 0.1 microfarad capacitors and 1 megohm grid leaks. The mains section, fed by screened cable from the battery section, employs two 6J7 pentodes in the first stage and two 807 valves in the output stage, with an H.T. line voltage of 450 volts. The overall gain has not been measured, but for the application described in this article, readings down to 1 mV are obtained with the gain control set only about 20 or 30 degrees from zero.

It will be noticed from the circuit diagram that the only earth point in the input circuit is the slider of the 1 megohm potentiometer, connected across the input terminals of the amplifier. It is therefore essential that all other components in the circuit, including the switch contacts, the potentiometer and the tissue under experiment should be insulated from earth. In addition, adequate screening is necessary, to reduce the level of picked-up interference to the lowest possible value. In the present arrangement the potentiometer is enclosed in a steel box connected to earth, the switch motor housing is earthed, the tissue or concentration cell used as a source of E.M.F. is mounted on rubber blocks, and all leads are screened. The presence of interference results in a "flat" minimum reading on the output meter and hence causes inaccuracy in reading. Readings from artificial sources have been made down to 1 millivolt, but below this level the amplitude of the interference prevents sharp minima from being obtained. Enclosure of the input circuit components, together with the battery amplifier, inside a screening cage would evidently reduce such external interference, and enable readings to be obtained at lower input levels.

The design of the potentiometer is based on similar devices used in ordinary pH meters, and little description is therefore necessary. E.M.F.s in steps of 100 millivolts up to 600 millivolts, and in two further steps of 500 millivolts up to a total of 1,600 millivolts, are available across the tapped switch $S1$. Wire-wound potentiometers $R1$ and $R2$ give ranges of 0-100 mV and 0-10 mV respectively. All resistors are chosen within close tolerances to rated values. Switch $S2$ is provided so that either $S1$ with $R1$ or $R1$ with $R2$ are brought into circuit, according to the range to be read. A ballast rheostat $R3$, together with fixed resistance $R4$ are included in the battery circuit, in order to adjust the E.M.F. across the potentiometer chain to the required value. R_3 is a panel control, used in conjunction with a standard cell in the external circuit, to provide calibration. A reversing switch $S3$ is provided so that the polarity of the battery may be reversed, as a check on the polarity of the source. A press-button B brings the battery into circuit only while taking readings. An additional fixed resistor $R5$ of $\frac{1}{2}$ megohm is included in the return lead to the amplifier, to raise yet further the impedance of the input circuit applied across the amplifier. Thus in series with the source there is a total impedance of not less than $1\frac{1}{2}$ megohms. Owing to the method of connexion, the 0-100 mV potentiometer must be double-scaled, since it reads in reverse when $S2$ is set to read 1,700 mV. The whole arrangement provides for the reading of E.M.F.s between 0 and 1,700 mV to the nearest mV, and from 0 to 110 mV to $\frac{1}{2}$ mV or less, according to the degree of interference present. Bioelectric potentials seldom exceed 100 mV, but the higher range is included so that standardisation against a Weston cell may be carried out, or for work on "model membranes" and concentration cells.

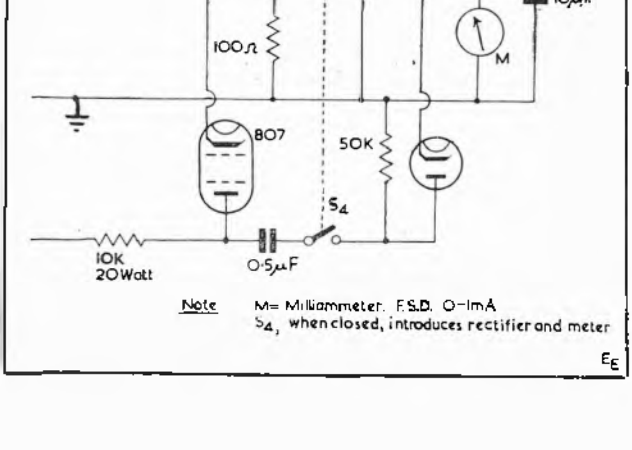
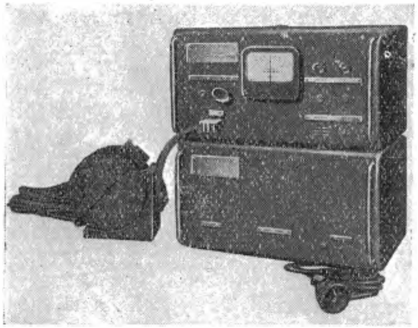


Fig. 2. Output circuit, rectifier and meter



Cable Eccentricity Gauge Type 155

THIS equipment serves for the non-destructive determination and control of the eccentricity of conductors relative to the surrounding dielectric in extruded cables and wires. It has been designed for continuous process control in the manufacture of conductors insulated with rubber or synthetic materials, as well as for laboratory and acceptance tests on the final product.

Type 155 equipment consists of three units: a gauge head, comprising a coil system housed in a split cast iron case, through which the cable under test passes; an oscillator unit, supplying a substantially constant current to the cable under test, and a control unit incorporating a detector amplifier which gives a direct reading of the eccentricity measured, and has facilities for the automatic control of the extruding machine and/or the operation of warning signals should the eccentricity exceed a specified tolerance. The equipment is A.C. mains operated.

One oscillator and control unit may allow the supervision of a number of extruders simultaneously, provided each channel is furnished with its own gauge head. A manually or motor-operated switch unit is employed in this case to transfer the oscillator output and amplifier input terminals, as well as any control lines, cyclicly from one measuring position to the next.

Addison Electric Co., Ltd.,
163, Holland Park Avenue,
Kensington, London, W.11.

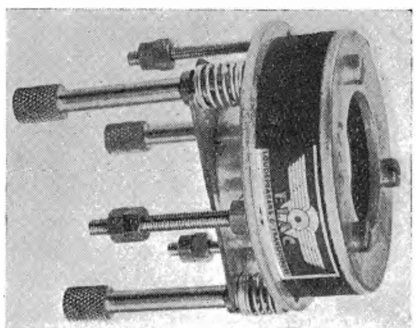
Permanent Magnet Focusing Units
(Illustrated below)

THE "Elac" Permanent Magnet Focusing Units have now been released for supply to the retail as well as the wholesale trade.

Owing to variations in the construction of cathode ray tubes, three models, having the same basic design but with different field strengths, are available. Type R17 has a low flux and is suitable for tetrode tubes. Type R20 will focus triode tubes operating on medium E.H.T. voltages, whilst Type R25 has a high field strength, and is suitable for tubes operating up to 15,000 volts and also some makes of triode tubes which are difficult to focus at lower voltages.

The unit itself is fixed rigidly to the receiver chassis, and all adjustments are made by three knurled headed screws which are readily accessible. Due to the small gap used in the "Elac" units, claim the manufacturers, a thin "lens" is produced which provides a more even focus over the whole of the tube face and also gives a smaller "spot" and therefore better definition.

Electro Acoustic Industries, Ltd.,
Stamford Works,
Broad Lane,
Tottenham, London, N.15.



Megatron Type C/I Cosine-Corrected Light Meter
(Illustrated left)

PHOTOMETERS with barrier-layer photo cells give correct readings up to angles of incidence not exceeding 60°-65°. For greater angles, although the amount of light reaching the photo cell remains proportional to the cosine of this angle, the current generated falls rapidly below the theoretical value. With an uncorrected light meter the error over an angle of incidence from 50° to 85° varies up to 40 per cent. With the Megatron Type C/I cosine-corrected Light Meter, however, the error is confined to angles of incidence 65°, and is only 12 per cent at its greatest.

This instrument has three ranges, a switch on the case selecting the range required. A separate switch marked "on-off" short circuits the moving-coil instrument in the "off" position, thus ensuring safety in transport. The Megatron type "B" photo cell is contained in an airtight mount covered by a perspex dome, and is connected to the meter by a 6 foot cable. Both the instrument and the photo cell are housed in a brown leather case.

Megatron, Ltd.
115a, Fonthill Road,
London, N.4.

Electronic

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

Mullard Valve Volt-Ohm Meter (Type E7555)
(Illustrated below)

THIS instrument, which has recently been introduced by the Electronic Equipment Division of Mullard Electronic Products, Ltd., has been specially designed to measure a wide range of A.C. and D.C. voltages and resistances. The input resistance and accuracy are extremely high, and a particularly important feature is the stable zero setting, which does not alter from range to range.

Voltage

D.C.—Pos. or Neg.	0-100V on four ranges
A.C.—Up to 10kc/s	0-1.0kV on four ranges
A.C.—Up to 30Mc/s	0-100V on four ranges

Resistance

Ohms	0-2M on six ranges

Frequency

Lower Limit	20c/s
Upper Limit (Input to panel up to 100V)	100kc/s
Upper Limit (Input to panel 300V and 1kV)	10kc/s
Upper Limit (Input to probe up to 100V)	30kc/s

Mullard Electronic Products, Ltd.,
Century House,
Shaftesbury Avenue,
London, W.C.2.

Equipment

tronic apparatus, components and supplied by the manufacturers.

Radiation Monitor

(Illustrated below)

THE instrument is a self-contained and portable ionisation chamber type radiation monitor, suitable for the measurements of Gamma-dosage, and for the qualitative detection of Beta-radiation.

For Beta-detection a thin window in the chamber end wall is exposed by rotating an external metal flap. All batteries are built-in, the filament cell being directly accessible from the top of the instrument, the remainder being located in the pistol grip.

Indication of dosage is made by a directly-calibrated meter, and a centre position of the *on/off* switch enables the filament cell to be tested, using the same meter. Provision is made for range calibration against an external source.

The range is 0-0.125 and 0-1.25 Rontgen per hour for Gamma radiation. The circuit comprises an E.1959 electrometer tube, with a backed-off meter which reads the change of anode current. The backing-off circuit contains the zero-setting potentiometer, whilst a variable meter shunt provides a calibration control.

The instrument weighs only 4 lb. and is run on one 9V and two 22½V miniature batteries, which last several months in normal usage.

E. K. Cole, Ltd,
Electronics Division, Ekco Works,
Malmesbury, Wilts.

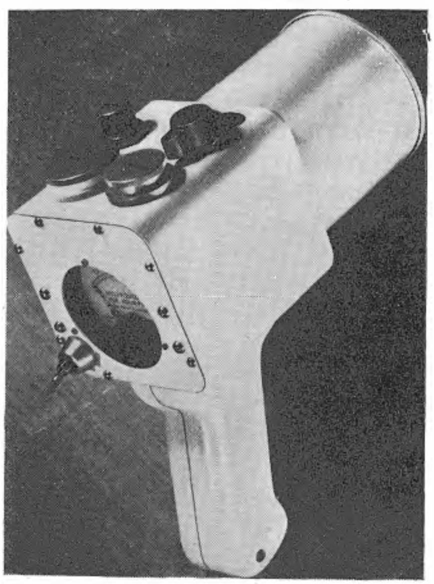


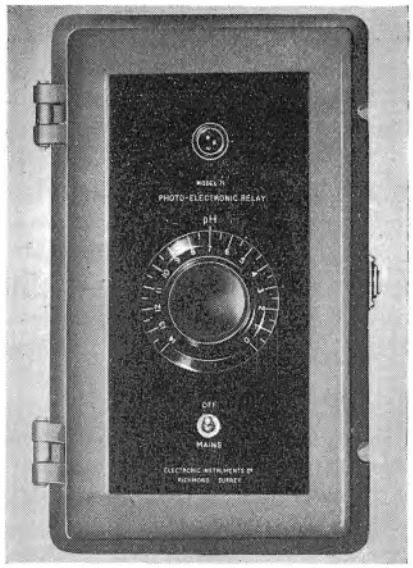
Photo Electronic Relay

(Illustrated below)

THE model 71 Photo Electronic Relay is fundamentally an ultra-sensitive current detector which, through the medium of a photocell and amplifier, operates a heavy duty relay in the output circuit. An important feature is that, unlike the conventional sensitive moving coil relay it is possible, by means of a single control, to set the operating point at any value between zero and 200 microamps. The sensitivity remains constant over the entire range and corresponds to 2 microamps or 1 per cent of the maximum input current, whichever is the greater. By means of appropriate shunts, the input current can be extended up to 2 amps.

The output circuit is completed or broken by change-over contacts handling 250 volts at 3 amps. A.C.

Electronic Instruments, Ltd.,
Paradise Road,
Richmond, Surrey.



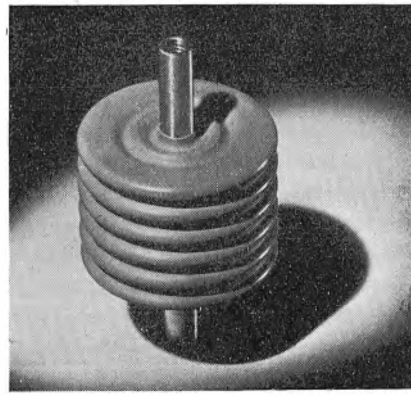
"Tom Thumb" Capacitors

THE "Tom Thumb" type of cubular dry electrolytic capacitor is now available in a range with greatly reduced dimensions, from 1¾ in. to 2⅜ in. length, ¾ in. to 1 in. in width. They have the added attraction of a coloured plastic covering to provide neat and clean appearance. The printing is permanent and cannot be erased by normal handling.

"Tom Thumbs" are suitable for smoothing applications, particularly where space is limited, and are available in six capacitances, as listed below :

List No.	Cap. uF	Max. Wkg. Volts
J650/NT	16	350
J651/NT	32	350
J652/NT	4	450
J653/NT	8	450
J654/NT	16	450
J655/NT	32	450

A. H. Hunt, Ltd.,
Wandsworth, London, S.W.18.



High Voltage Capacitor

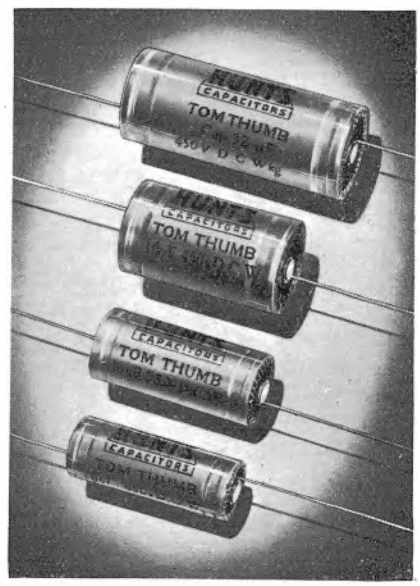
(Illustrated above)

THE Erie Type 410 high voltage Hi-K Double Cup Ceramicon is a capacitor designed with particular attention to the requirements of an R.F. oscillator or line time-base fly-back type of E.H.T. supply for the modern television receiver. The active centre web of the ceramicon is of adequate thickness for high voltage application, and the whole is encased in a low loss moulding, ribbed to ensure a long surface leakage path so desirable when the formation of dust or moisture film is probable.

Contact with each silver plate is made by a metal pressing and a 4 B.A. tapped pillar, which forms a convenient means of mounting and connexion. Full corona protection is afforded by the arrangement of silvering and the nature of the moulding.

The maximum working voltage is 15 kV at 85° C. (tested at 22.5 kV D.C.), and the capacitance is 500 pF. Body measurements are 1⅛ in. diameter by 4/5 in. (approx.) height, exclusive of mounting pillars.

Erie Resistors, Ltd.,
Carlisle Road,
The Hyde, Hendon, N.W.9.



BOOK

Historic Researches

Chapters in the History of Physical and Chemical Discovery. By T. W. Chalmers. 288 pp., with 85 illustrations. Morgan Brothers (Publishers), Ltd., London, 1949. Price 21s. net.

BETWEEN 1944 and 1948 a series of articles by Dr. Chalmers appeared in *The Engineer* dealing with some selected chapters of engineering, physics and chemistry from a historical point of view. It was a happy idea to collect these articles in a well printed and neatly bound volume and thus make them accessible to a wider public. The result is a delightful book which should appeal not only to the scientist and engineer but also to the general reader interested in scientific subjects.

The author rightly states that "the historical approach towards any branch of scientific study possesses not only intrinsic interest; as a practical educative method it undoubtedly has distinctive merits. It provides the easiest path towards the ready understanding of modern knowledge and opinion." And also that "physics of to-day is the engineering of to-morrow."

What makes the book specially valuable is the critical treatment of the experimental and theoretical results of the various investigators. Distinction is made throughout between the three stages of research—basic research, its interpretation and its practical application.

It is not possible to give, in a brief review, an indication of the vast amount of information contained in this book. Gaps in our knowledge are indicated throughout, but especially in the chapters on friction (I) and on specific heat (V). In chapter II, dealing with the mechanical equivalent of heat, the story of the replacement of the "caloric" theory by the dynamic theory makes fascinating reading. The student looking up in his handbook the figure for the heat equivalent has no idea what amount of experimental work had to be performed to arrive at that figure. The author rightly stresses "the importance which frequently attaches to the study of residuals," i.e., to the explanation of some minor deviations of a commonly accepted law.

In the chapter on electrodynamics (III) only early work is dealt with, and the story is told till Faraday's experiments of 1832 only, while in the fourth chapter dealing with ether drift quite recent experimental work is described, but no attempt is made to give the reader a guidance on "the long and arduous path leading to an understanding of the generalised theory of relativity." Of course, this would fall outside the scope of a book, the "object of which is to present a history of the subject and not to be a formal text book." The increasing complexity of theory is also stated in the chapter on specific heat (V), and the wisdom of Black who did not try to explain his experimental results but left this task to future physicists is mentioned. On the other hand Lavoisier, whose work is described in the chapter on elements and atoms (VI), was less an experimenter than an interpreter and thus became the founder of modern chemical doctrines. This chapter tells the story of the atom from the ancient Greeks up to Dalton and Avogadro. The later developments leading to Bohr-Rutherford's atom models and modern conceptions concerning valency are not dealt with.

The chapter on the classification of elements (VII) starts from Prout's hypothesis and its abandonment and deals mainly with Mendeleeff's and Lothar Meyer's, their precessors' and successors' work on the periodic system, the predictions and fulfilments based on it, and it points to the many open questions still to be answered. "Not Nature's but Dalton's atoms have been split," and "The real atoms are the electrons, protons, neutrons and so forth."

In the chapter on molecular physics (VIII) the principles of the kinetic gas theory are explained, starting from Robert Hooke's notion that gaseous pressure is produced by the impact of moving atoms on the wall of the container. Bernoulli's mathematical investigation of this phenomenon is given in brief and so are Joule's and Clausius' investigations on specific heat and Maxwell's distribution law of velocities. Particularly welcome is the description of modern experimental methods for determining the velocity distribution. Finally, the mean free path, the "radius of action," diffusion and viscosity of gases are described.

The chapter on the conduction of electricity through liquids (IX) is perhaps not quite as perspicuous as the rest. It tells the story from Grothuss' chain hypothesis, Faraday's two laws of electrolysis, the work of Ohm, Kohlrausch, Clausius, Maxwell to the more recent investigations of Arrhenius and to J. J. Thomson's discovery of the electron. The tenth chapter deals with electric conduction through gases, and

REVIEWS

is therefore of special interest to the electronic engineer. Davy's and Faraday's experiments are described and the work of Plücker, Hittorf and Crookes is critically surveyed. The discovery of Goldstein's "canal rays" and of Lenard's rays led to Perrin's experiment and its modification by J. J. Thomson, which proved that the rays were corpuscular and not undulatory. Thomson's, Wilson's and Millikan's work for measuring the ratio of charge and mass and their absolute values are described, and Stokes' formula regarding the fall of droplets through a viscous medium is simply derived in one of the mathematical appendices.

A comparably short chapter (XI) deals with X-rays, their discovery by Röntgen, the introduction of the hot cathode by Coolidge, the decision in favour of the undulatory nature of the rays by Laue's interference tests using crystals as a 3-dimensional diffraction grating, their elucidation by the Braggs, the explanation of ionisation through X-rays and Moseley's observation of the stepwise shift of the X-ray spectre of different elements.

The last chapter (XII) dealing with positive rays and isotopes and the explanation of the mass spectrograph and its theory is especially excellent. The work of Wien, J. J. Thomson, Aston and Dempster is described. It is stated that by the use of the mass spectrograph transient stages in chemical reactions not observable by other means may be investigated, and the discovery of the isotopes, especially neon, hydrogen and oxygen, has led to a revival and modification of Prout's hypothesis. Heavy water and its preparation are dealt with and the chapter closes with the discovery of the neutron.

It may be hoped that the author will be able to present his planned companion volume on nuclear physics in the not too distant future.

The present volume concludes with some brief biographical notes on 34 of the scientists mentioned in the book, with 8 concise mathematical notes and a combined name and subject index.

A few desiderata may be mentioned for a second edition of this volume. Besides the correction of a surprisingly small number of misprints (page 14 line 13, Fig. 5 not Fig. 4—page 213, line 6, insert "inverse of the" before "viscosity"—page 279, line 21, $y = Ae/mv^2$ instead of $Aemv^2$) the reviewer would suggest that a few more mathematical notes as excellent as those already presented would be desirable for explaining the expressions for the law of probability (page 151), the ratio of sound velocity and R.M.S. value of molecular velocity (page 154), the radius of the circle of reflexion of the beam in a magnetic field (page 208) and the viscosity of a gas (page 161). This last-named explanation would be specially desirable with regard to giving the dimension of viscosity used afterwards in mathematical note 7 and not easily found in usual handbooks.

More detailed study would be materially facilitated if a list of references to literature could be added. It appears dubious whether actually Geissler should be credited with the invention of the mercury vapour pump.

But these are minor blemishes which in no way impair the value of a book equally remarkable for the scholarly treatment of its subject, for the vast amount of work of compilation, for the critical attitude, and for the lucid and very readable style.

R. Neumann

Practical Electrician's Pocket Book, 1950

Published September 1949, Electrical & Radio Trading, 6 Catherine Street, Strand, London, W.C.2. 5s. net.

LATEST applications of electricity for increasing the output of factory and farm are explained, in practical detail, in the 1950 edition of the 52-year-old *Practical Electrician's Pocket Book.*

As regards factory, extensive chapters on motors, wiring and lighting are supplemented in this edition with a review of "Electronics in Industry" by R. C. Walker, B.Sc., A.M.I.E.E., A.M.I.Mech.E., and up-to-date information on "Ventilation" by N. F. Parry, A.M.I.H.V.E. Factory trucks, as well as road and other battery electric vehicles, are described by J. de Gruchy, M.Brit., I.R.E.

The many electrical aids to agriculture, horticulture and dairy farming are dealt with by J. Golding, of the Electrical Research Association, and practical information given on such matters as loadings and wiring. In fact, this chapter—like all sections of the "Pocket Book"—has been written to guide the practical electrician and inform the student apprentice.

Tables summarising the characteristics of instruments and giving preferred instruments for various measurements are now a feature of the Measurements Section. The

BOOK REVIEWS (Continued)

rewinding of small motors and the winding of resistances are explained by S. F. Philpott, A.M.I.E.E., and "Soldering" is another new heading.

The Mains Voltages Table, always a valued part of this book, now lists over 5,000 places, and gives, as well as voltage and frequency, the responsible Board and sub-area office.

Guide to Broadcasting Stations

Compiled by *Wireless World*. 88 pp. 5th Edition. Iliffe & Sons, Ltd. Price 1s. 6d.

ALMOST without exception Europe's four hundred long- and medium-wave broadcasting stations will be changing their wave-lengths in March when the Copenhagen Frequency Allocation Plan comes into operation. In order that this new edition should not become out of date within a few months, it has been enlarged and includes, in addition to the present operating details of Europe's stations, those which will come into force on March 15, 1950.

Operating details of nearly 1,300 short-wave stations of the world, which have been checked against the frequency measurements made at the B.B.C. receiving station at Tatsfield, are also given in tabular form both geographically and in order of frequency.

This booklet also includes details of Europe's television and E.H.F. broadcasting stations and special service stations—such as meteorological and standard frequency transmitters; world time constants; revised list of international call signs; and wavelength frequency conversion table.

International Radio Tube Encyclopaedia

Edited by B. Babini. 410 pp. 1st Edition, 1949. Bernards (Publishers), Ltd. Price 42s.

THIS book, as the publishers correctly state, is an important contribution to electronic engineering, containing operating characteristics and pin connexions of some 15,000 valves of all types manufactured throughout the world.

These 15,000 valves are classified into ten main categories, and once the reader has discovered the particular section to which a valve belongs, the search for its characteristics is generally an easy matter.

An important feature of the book is the inclusion of the types of valves used by the British, American and European services.

The sections containing the technical matter and instructions relating to the tables have been translated into fourteen languages including Hebrew and Turkish so that the book is truly international. Since the publishers propose to issue supplements annually on new valves there is no risk that the book will become obsolete.

H.G.F.

Notes on Soldering

W. R. Lewis. 88 pp., 47 photographs and diagrams. Tin Research Institute. Fraser Road, Greenford, Middx. Free of charge.

DURING recent years every detail of the soldering process has been scientifically investigated and a vast literature on the subject has accumulated. In a new edition of the Tin Research Institute's handbook "Notes on Soldering," W. R. Lewis, B.Sc., reviews these researches and presents a compilation of the more important facts which are likely to be of value to solder-users in a large variety of industries.

Advances in soldering technique are discussed and the various forms of solder and methods of applying heat to the joints are described in outline. The importance of design for soldering is emphasised.

The fundamental principles underlying the production of "wiped joints" in lead pipes are fully covered and alternative types of joint in lead pipe are described and illustrated. Aluminium, stainless steel, cast iron and other "difficult-to-solder" alloys can be soldered by the special methods suggested. The behaviour of solders at various temperatures, under tensile and shear stresses, and under creep conditions is important to the designer and to the production engineer, and data on these properties, together with notes on the metallurgical constitution of the tin-lead solders, is included.

"Notes on Soldering" also contains an adequate bibliography and a detailed scheme for the chemical analysis of tin-lead solders devised by Dr. J. W. Price.

Readers in the United States of America are requested to apply to Dr. Bruce Gonser, The Battelle Memorial Institute, 505 King Avenue, Columbus, 1, Ohio, U.S.A.

LETTERS TO THE EDITOR

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

The Design and Limitations of D.C. Amplifiers

DEAR SIR,—Your correspondent, Mr. Vernon, has criticised Messrs. Harris and Bishop for recommending the 6J6 (Mullard equivalent ECC 91) as a commendable double-triode for use in the early stages of a D.C. amplifier, quoting the Radiation Laboratory's experience as recorded in the McGraw-Hill series. From personal experience I support this view, but both parties are guilty of misconceptions which cloud the basic issue.

It is generally recognised, as the authors state on page 335 (September 1949 issue), that the best performance (i.e., minimum output voltage drift) is provided by those valves which remain most nearly balanced when the heater current varies, so that twin triodes in one envelope and with a truly common cathode, shared between the assemblies, should be used. The cathodes are then most nearly balanced, both in thermal and emissive properties. Furthermore, the two grid and anode assemblies must be matched to each other and to the cathode and this would be expected to be more difficult in miniature rather than standard-size assemblies.

Mr. Vernon's conclusion that "the best performance is apparently obtained from those twin triodes formed of two separate triodes in one envelope" is clearly untenable, all the more so when it is realised that the Radiation Laboratory's considerations exclude, for some strange reason, one very important standard-size type, the 6SC7 and 12SC7, although recommending the separate-cathode near-equivalent 6SL7.

Messrs. Harris and Bishop (page 335) agree that the 12SC7 is frequently specified for direct-coupled circuits, but say "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." They appear to have overlooked the point that, although the grid-anode assemblies are physically separated vertically, the heater-cathode assembly rises with a single cathode sleeve through both sections and is therefore very nearly balanced as compared with two independent cathode sleeves heated by separate filaments. One further theoretical improvement, as in the 6J6, would be to mount the two sections on opposite sides of the same cathode sleeve, rather than with longitudinal displacement, but in the 6J6 the miniature construction is probably responsible for the bad matching.

In operation it is also desirable to balance the anode circuits and currents whenever possible, as discussed on pages 355-6. If, however, the two halves of the tube are not perfectly balanced, but a truly common cathode is used, it can be reasonably assumed that the net effects of heater variations will produce proportional changes, upon which effect Miller's compensating circuit (page 356) depends for perfect adjustment to balance.

Miller's amplifier, which appears to have been the first cathode-coupled and highly-stable D.C. amplifier, incorporating highly-stabilised D.C. heater supplies and considerable negative feedback, employed the 12SC7 and it may be of interest to note that the 1634, which the authors also recommend (page 433) is merely a 12SC7 selected "for applications critical as to matching of the two triode units" (most of the RCA 1600 series are similarly selected tubes of various types).

In my own experience of various D.C. amplifiers for measuring purposes, the 6SC7 or 12SC7, even when unselected, is far superior in drift stability to any other type of either American or British origin, when used under the best conditions. Unfortunately, supplies generally depend on War Surplus disposal, as these types are made only in the U.S.A. Johnston, in his excellent articles on an electroencephalograph amplifier (author's ref. 29), has used the 6SC7, but mentions the possibility of using the 6J6, not then available (1947), as another particularly suitable type. The authors (ref. 30) subsequently gave a suggested modification of Johnston's preamplifier, with circuit values for 6J6 input, but no drift data. Such data is also conspicuously sparse in their review, which began promisingly enough, but is disappointing in containing no complete practical circuits together with performance figures. My own experience of the 6J6 type is confined to one sample ECC91 tube, used under balanced and stabilised conditions, but so bad in regard to drift that I did not consider it applicable to high-gain D.C. amplifiers. In view of the conflicting views on tube selection, it is clearly desirable in the general interest that representative tests be carried out on an adequate number of samples of the various type by a disinterested organisation with the necessary equipment and personnel; e.g., the R.A.E., N.P.L. or T.R.E.. Some criterion for tube matching selection, if found necessary, must then be established.

One factor which clearly arises from the controversy is that there is a definite need for a British double-triode high-mu voltage amplifier with a drift performance at least as good as that of the 6SC7 type, particularly in view of the increasing interest in high-gain D.C. amplifiers for industrial and scientific purposes. I hope that those who build and use such apparatus will lend their support to prove the demand to our tube manufacturers and to stimulate them into doing some useful work in

LETTERS TO THE EDITOR (Continued)

this direction. In addition to the fundamental question of the best form of cathode and general electrode assembly, it is evident that a practical compromise will have to be reached between the less critical tolerances and easier matching of a standard-size electrode assembly and the more rapidly attained and evenly distributed thermal stability, minimum pick-up and compactness of a miniature type.

Yours faithfully,

JOHN C. FINLAY,
Welwyn Electrical Laboratories, Limited.
Bedlington Station,
Northumberland.

The Authors reply:

"We should like to record our support of Mr. Finlay's plea for a British equivalent of the 6 and 12SC7, preferably as a miniature type, adding the suggestion that attention be paid to low grid current, which can be troublesome in the SC7 valves.

With regard to his experience of ECC91's, we can quite appreciate his poor opinion if gained on a single sample. We once tested a group of 37 mixed 6J6 and ECC91's, kindly loaned to us by Mr. Johnston. Only about 30 per cent were acceptable on the score of balance and freedom from microphony. Probably inadequate care in selection accounts for the adverse criticism expressed in the Radiation Lab. Handbook, although, naturally, it would save time and expense if more stringent production specifications had been applied by the makers, as is the case for 1634's and 6SU7's.

There is some difficulty in understanding why two halves of a thin cathode assembly will necessarily be better balanced with respect to heater supply changes when vertically above one another than when separated and side by side. Surely the temperature is determined by the balance between energy dissipation and radiation loss and the thermal conductivity along the cathode assembly is unlikely to provide a very effective means of keeping top and bottom at the same temperature. It was for this reason that we originally tried 6J6's, which have a small compact cathode with a much better chance of thermal gradients between the two halves (here the two sides) being minimised. Clearly any advantage accruing from this factor will only be apparent if other obscuring factors of greater magnitude have been obviated by selection. A residual instability is bound to remain in any case on account of statistical fluctuations in the ion/electron density in the semiconducting coating, which we suppose is the origin of flicker noise. A short but useful treatment of the influence of temperature on cathode emission will be found in Arguimbau's book "Vacuum Tube Circuits."

Mr. Finlay will find some of the information which he considers to be missing from our review in the reference (11), which gives a figure for the residual instability of a particularly good 6J6 stage. Provided sufficiently stable supplies are used, the instability will be determined by the valve used in the first stage, and not by the circuit arrangement (unless subsequent stages are worse by a factor approaching the total gain up to that stage). The various circuit devices are all used with a view to minimising the effects of supply variations, and their performance in this respect can be calculated, given the values of the valve parameters and components. As we were trying to survey the design of direct coupled amplifiers it would have been inappropriate to have included a detailed description of an apparatus made to fulfil a particular function. We have, in fact, submitted for publication a full description of a 4-channel alternatively direct or capacitor-coupled apparatus with high impedance cathode follower input stage, and provision for oscilloscope or pen recorder display."

Review of Recent Technical Films

Marine Radar Film

"In All Weathers," a film produced by C.O.I. Films for the Ministry of Transport, tells of the development and use of British mercantile marine radar.

The opening sequences deal with a cargo ship held up by bad weather without a radar system. Then scientific investigation, early experiments and war-time sets are portrayed, followed by the post-war developments such as construction of the prototype set for use in merchant ships, and the actual experiments with this set. The remainder of the film describes the passage of a merchant vessel, fitted with a British radar set, through a channel to a Canadian port in bad weather.

This film is a clear, concise and interesting account of how radar can help commercial shipping, and should be a valuable contribution to selling British radar equipment overseas. It is, however, considered a pity that "In All Weathers" has not been thought suitable for general release to the public.

E.D.A. Educational Films

The British Electrical Development Association has produced three more films in their third series.

"Electricity and Heat" illustrates the different effects resulting from the flow of electric current through a good conductor, such as copper, and a higher resistance conductor, such as nickel alloy producing heat. It is demonstrated how the amount of heat generated can be controlled by the thickness and length of the conductor.

"High Frequency Heating" should always be shown after "Electricity and Heat," for it deals with dielectric and induction heating, and describes actual processes.

In the film "Electro-Chemistry" the principle of electro-chemistry is shown as being a migration of ions between two electrodes. Industrial applications of the principle are shown—in the refining of copper and the reduction of alumina to aluminium.

In the same way electro-chemical action is shown in the production of an electric current from which a simple primary cell is built up.

These three films are an excellent addition to the 12 educational films already made for the Electrical Development Association, and have the same high standard of production.

All the E.D.A. educational films should have a wide field of interest, especially in schools and colleges.

NOTES FROM THE INDUSTRY

Television Exhibition for Birmingham

The Radio Industry Council announces that a radio exhibition, with television as the main feature, is to be held at Castle Bromwich, Birmingham, in September next—the exact date to be announced later. As previously stated, there is to be no National Radio Exhibition in London in 1950.

Convention in Radio-physics Measurements

A two-day convention on radio-physics measurements, with special reference to local conditions, will be held at the University of Malaya, Singapore, on February 17-18, 1950.

A programme of lectures, discussions and demonstrations will be arranged, under the following general headings:—

(a) Basic measurement techniques.
(b) Measurements under tropical climatic conditions.
(c) Standards and sub-standards available in Malaya, and co-operation between interested bodies.

Thermionic Valve Data

In order to simplify problems in the use of the Thermionic Valve in scientific instruments from the point of view of design, servicing and export conditions for members of the Scientific Instrument Manufacturers Association, their Electronic Section prepared a list of the preferred types. The list endeavours to provide a working collection of a limited number of valve types on which new designs of instruments may be based.

The advantages of limiting the number of types of valves used from the point of view of productivity is obvious, and with the increasing number of British electronic instruments being shipped overseas, the reduction of the number of types required for servicing should be of great value to the export drive.

The list was originally prepared for circulation among the members, but it was felt that other industries might be made aware of such a step with advantage, and accordingly the list was brought up-to-date, and information on the characteristics of the valves concerned was included. Copies of this list can be obtained from the Scientific Instrument Manufacturers Association, 17 Princes Gate, London, S.W.7, price 2s. 6d., post free.

1950 New Year Honours List

Mr. J. B. Stevenson, a director of E.M.I. Factories, Ltd.—a subsidiary company of Electric and Musical Industries, Ltd., has been awarded the M.B.E. for his services to the industry.

Mr. J. A. Hunt, general manager of the Hymatic Engineering Co., Redditch, Worcs., and chairman of the Birmingham Advisory Committee, Midland Regional Board for Industry, has been awarded the M.B.E. for his work in the field of industrial relations.

Bellahouston Park

On December 12 a new transmitting station in Bellahouston Park started radiating the Third Programme for Glasgow listeners. It took over the service from the transmitter that was installed in Broadcasting House, Glasgow, as an emergency measure in 1940.

The new station, which is on a $2\frac{1}{2}$-acre site, has two transmitters, each capable of 2 kilowatts output power, and a T aerial supported by a pair of tubular masts, 126 ft. high. By having duplicate transmitters, a very reliable service is ensured.

The twofold increase in power and the more efficient aerial system at the new station give listeners to the Third Programme a stronger signal in most parts of Glasgow.

British Standard for Soft Solders (B.S. 219 : 1949)

The need to economise in raw materials during the war led to the issue of a war emergency edition of B.S.219, Soft Solders, restricting the number of grades to five.

It has now been found possible to revise this standard to include five antimonial and five non-antimonial solders, and it is considered that these will cover all common requirements. Slight alterations have been made to the composition of some of the solders that were covered in the earlier edition.

Copies of this standard may be obtained from the British Standards Institution, Sales Department, 24 Victoria Street, London, S.W.1, price 2s., post free.

New Telephone Number

E. K. Cole, Ltd. announce that, as from January 9, 1950, the telephone number of the London offices and showrooms of E. K. Cole, Ltd., and Ekco-Ensign Electric, Ltd., at 5 Vigo Street, became Regent 7030/9.

Publications Received

Industrial Photocells.

This publication, MV1823, is intended to serve as a guide to the use of the Mullard range of emission photocells which have been specifically developed for industrial applications. *Mullard Electronic Products, Ltd., Century House, Shaftesbury Avenue, London, W.C.2.*

Hazlehurst Designs.

This collection of data sheets gives a table of the standard and non-standard types of radio frequency high voltage units made by *Hazlehurst Designs, Ltd.,* 186 *Brompton Road, London, S.W.*3.

Glass-to-Metal Seals.

This publication gives details of the Nilo series of alloys, and may be obtained from *Henry Wiggin and Co., Ltd., Wiggin Street, Birmingham* 16.

Mallory Resistance Welding.

A publication issued by *Johnson Matthey and Co., Ltd.,* 73-83 *Hatton Garden, London, E.C.*1, dealing with their Mallory products for resistance welding.

Radio and Electronic Instruments and Components.

This catalogue describes the many types of radio and electronic equipment retailed by *Berry's (Short Wave), Ltd.,* 25 *High Holborn, W.C.*1.

R.S.G.B. Publications

Amateur Radio Receivers by S. K. Lewer describes the various types of communications receivers of interest to amateurs, and includes their construction, power supplies, fault finding, adjustments and calibration. Price 3s. 6d.

Amateur Radio Simple Transmitting Equipment by W. H. Allen, M.B.E., and J. W. Mathews. This booklet discusses transmitters for amateur use, covering aerials, fundamentals, a simple V.F.O. unit and crystal controlled frequency sub-standard. Price 2s.

Amateur Radio—The Transmitting Licence. This publication presents the essential information required by those who wish to obtain an Amateur Transmitting Licence. Price 9d.

These booklets may be obtained from the *Radio Society of Great Britain, New Ruskin House, Little Russell Street, W.C.*1, and 3d. postage on each book must be added.

MEETINGS THIS MONTH

Institution of Electrical Engineers

Unless otherwise stated, all meetings are held at the Institution of Electrical Engineers, Savoy Place, W.C.2.

Measurements Section

Date: February 7. Time: 5.30 p.m.
Lecture: A New Precision A.C. Voltage Stabiliser.
By: G. N. Patchett, B.Sc., Ph.D.
Date: February 21. Time: 5.30 p.m.
Discussion: Electrical Methods of pH Measurement.
Opened by: G. I. Hitchcox and G. R. Taylor.

Radio Section

Date: February 15. Time: 5.30 p.m.
Lecture: Ground-Wave Propagation over an Inhomogeneous Smooth Earth — Experimental Evidence and Practical Implications.
By: G. Millington, M.A., B.Sc., and G. A. Isted.
Date: February 27. Time: 5.30 p.m.
Discussion: Mobile Radio Power-Packs.
Opened by: Air-Commodore R. L. Philips, C.B.E.

Informal Meetings

Date: February 13. Time: 5.30 p.m.
Discussion: Interference Suppression.
Opened by: E. M. Lee, B.Sc.

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

Cambridge Radio Group

Date: February 7. Time: 8.15 p.m.
Held at: The Cavendish Laboratory, Cambridge.
Lecture: " Some Considerations in the Design of Negative-Feedback Amplifiers. (Joint Meeting with the Cambridge University Radio Society).
By: W. T. Duerdoth, B.Sc.(Eng).
Date: February 28. Time 6 p.m.
Held at: Cambs. Technical College, Cambridge.
Lecture: A Survey of the Possible Applications of Ferrites.
By: K. E. Latimer, Ph.D. and H. R. MacDonald, B.E.

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

North-Eastern Radio and Measurements Group

Date: February 6. Time: 6.15 p.m.
Held at: King's College, Newcastle-on-Tyne.
Discussion: Control and Measuring Equipment: Electronic versus Non-Electronic Apparatus.
Date: February 20. Time: 6.15 p.m.
Held at: King's College, Newcastle-on-Tyne.
Lecture: Theory and Design of Magnetic Amplifiers.
By: H. M. Gale, B.Sc.(Eng.), and P. D. Atkinson, M.A.

Asst. Secretary: G. A. Kysh, Carliol House, Newcastle-on-Tyne, 1.

South Midland Radio Group

Date: February 27. Time: 6 p.m.
Held at: James Watt Memorial Institute, Great Charles Street, Birmingham.
Lecture: Magnetic Amplifiers.
By: A. G. Milnes, M.Sc. (Eng.).

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

The Television Society

Constructors' Group

Date: February 9. Time: 7 p.m.
Held at: Cinema Exhibitors' Association, 164 Shaftesbury Avenue, W.C.2.
Lecture: Mullard Projection Receiver.
By: Emlyn Jones.

Main Society

Date: February 24. Time: 7 p.m.
Held at: Cinema Exhibitors' Association, 164 Shaftesbury Avenue, W.C.2.
Demonstration and Discussion on Television Aids—polaroid lenses, filters, lighting, special aerials, interference suppressors.

Secretary: T. M. Lance, 35 Albemarle Road, Beckenham, Kent.

Midlands Centre

Date: February 6. Time: 7 p.m.
Held at: Lecture Hall (1st Floor), The Crown Restaurant, Corporation Street, Birmingham.
Lecture: The V.H.F. Link.
By: A. H. Mumford, M.I.E.E.

Secretary: W. Summer, 31 Beech Road, Bournville, Birmingham 30.

British Sound Recording Association

London Meeting

Date: February 24. Time: 7 p.m.
Held at: Royal Society of Arts, John Adam Street, Adelphi, W.C.2.
Lecture: High Quality Reproduction—How to Achieve it.
By: H. J. Leak.

Hon. Gen. Secretary: R. W. Lowden, " Wayford," Napoleon Avenue, Farnborough, Hants.

Inc. Radio Society of Great Britain

Date: February 24. Time: 6.30 p.m.
Held at: Institution of Electrical Engineers, Savoy Place, W.C.2.
Lecture: Panoramic Reception
By: J. Neale, B.Sc.(Eng.), A.M.I.E.E.

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

Institution of Post Office Engineers

Date: February 22. Time: 5 p.m.
Held at: Institution of Electrical Engineers, Savoy Place, W.C.2.
Lecture: Relay Manufacture.
By: R. Moxham.

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

British Institution of Radio Engineers

London Section

Date: February 23. Time: 6.30 p.m.
Held at: London School of Hygiene and Tropical Medicine, Keppel Street, W.C.1.
Lecture: Travelling Wave Tubes.
By: R. L. Kompfner.

General Secretary: 9 Bedford Square, W.C.1.

West Midlands Section

Date: February 22. Time: 7 p.m.
Held at: Wolverhampton and Staffs. Technical College, Wulfruna Street, Wolverhampton.
Lecture: Electronics and the Brain.
By: H. W. Shipton.

Hon. Secretary: R. A. Lampitt, 20 Northfield Grove, Merry Hill, Wolverhampton.

North Eastern Section

Date: February 15. Time: 6 p.m.
Held at: Neville Hall, Westgate Road, Newcastle-upon-Tyne.
Lecture: Single Sideband Systems Applied to Long Range Wireless.
By: Major S. R. Richman.

Hon. Secretary: C. Stokes, 6 Esterton Close, Coventry.

Scottish Section

Date: February 2. Time: 6.45 p.m.
Held at: Institution of Engineers and Shipbuilders, Glasgow.
Lecture: The Performance and Stability of Permanent Magnets.
By: A. J. Tyrrell.

Hon. Secretary: A. A. M. Turnbull, 68 Launderdale Gardens, Glasgow, W.2.