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Commentary

"EXHIBITIONS galore" might well be the electronic industry's slogan for March and April. At the beginning of March the Television Society held their exhibition, this was followed at the end of March by the exhibitions organized by the Physical Society and by the Association of Supervising Electrical Engineers. In April there are three major events; the Radio and Electronic Component Manufacturers' Federation Exhibition, the Instruments, Electronics and Automation Exhibition and the London Audio Fair. In addition, there are several quite sizeable private exhibitions organized by individual firms. Altogether a total of six major exhibitions plus the two or three private ventures in a space of less than two months, and each and every one of them "bigger and better" than its predecessor.

Some of these exhibitions are not, of course, exclusively "electronic" in character but with the exception of the A.S.E.E. Exhibition, electronic engineering, in one form or another, plays the major role. There are a few firms who will be exhibiting at all of the public exhibitions while a considerable number of items of equipment will be exhibited at three or more of the exhibitions. This apparent duplication of effort is not necessarily a waste of endeavour but it does provide food for thought and makes one wonder whether a more beneficial arrangement for all concerned, both exhibitors and visitors, could be brought about by a measure of co-operation and co-ordination among the various organizing bodies.

This is, however, a subject on which much has been said and published in recent years and while not wishing, at the moment, to enter upon this controversy, a few words about the various exhibitions may not be out of place.

The exhibitions of greatest interest to the "non-domestic" side of the electronic industry are the Physical Society, the R.E.C.M.F. and the I.E.A. Exhibitions.

The Physical Society Exhibition was, as in the past few years, housed in the Old and New Halls of the Royal Horticultural Society. While this venue undoubtedly provides more room and makes for more comfortable viewing its departure from Imperial College at South Kensington has tended to rob it of its academic flavour. The individual and university research exhibits are very much in the minority, and it would seem to be in danger of becoming a commercial exhibition pure and simple—or maybe one should say not so pure and simple! In its earlier years it was an exhibition at which one rightly expected to see new techniques and theories demonstrated, but now, year by year, it becomes more than ever a display of commercially produced instruments. On looking through the catalogue the phrases "a production version of the prototype shown last year" and "a modified version of the equipment shown last year", become monotonous by their repetition; added to this, the modifications are sometimes so slight as to be hardly discernible. It may, of course, be in the general interest that this should be so, but it would, nevertheless, be refreshing to have one exhibition a year where all the exhibits had a genuine claim to originality.

The R.E.C.M.F. Exhibition this year is being split, as it was last year, between Grosvenor House and Park Lane House; an arrangement which for the visitors, and possibly the exhibitors, too, is not altogether satisfactory. The two venues are rather far apart and the splitting of the exhibition is very unequal; the latter is unavoidable since Park Lane House is in no way an ideal exhibition building. The solution here is not easy to see for, while the exhibition, as it now is, has obviously outgrown Grosvenor House, it has become so well established there that it could not be uprooted lightly, even if a satisfactory alternative venue could be found. A possible solution might be to limit this exhibition to components only and to exclude complete equipments such as test gear, which can be adequately exhibited elsewhere.

The Instruments, Electronics and Automation Exhibition is by far the youngest of the three. Its inception last year was, however, most auspicious and this year's exhibition promises to be even more successful. The five trade associations who sponsored last year's exhibition have been joined this year by the Electronic Engineering Association (formerly the Radio Communications and Electronic Engineering Association) and, in addition, the 250 British exhibitors will be joined by over 100 overseas exhibitors from ten different countries.

Whatever criticisms may be offered and whatever one's views may be about this galaxy of exhibitions, there is no doubt but that the range and quality of the products shown give ample testimony to the healthy state of Britain's electronic industry. To the engineer who has the time and opportunity to browse around them they will provide great interest and enlightenment—to do adequate justice to them he will, however, need to be superbly fit and well shod!

U.H.F.-S.H.F. TECHNIQUES

The May issue of ELECTRONIC ENGINEERING will be given over entirely to a survey of wide band multichannel communication systems in the u.h.f. and s.h.f. bands.

The survey will cover all aspects of British developments and practice and will take the form of contributed articles written by those in industry and research who are largely responsible for the present high standard of British techniques.

The price of the issue, which will be considerably enlarged, will remain unchanged at 3s.

Orders for additional copies should be placed, as early as possible, with the Circulation Manager, 28 Essex Street, Strand, London, W.C.2.

The Design and Manufacture of

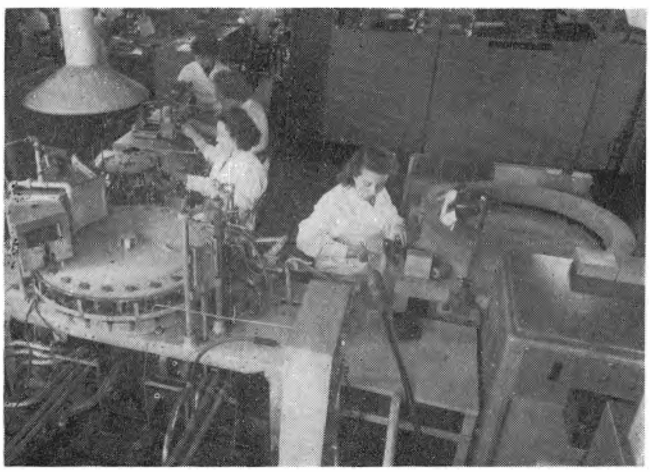
A Low-cost Cold-Cathode Trigger Tube

A. Turner*, B.Sc., A. Inst.P.

The long life, reliability and low power consumption of cold-cathode trigger tubes make them suitable for use in medium-speed computing and data processing applications. In the past the relatively high price of trigger tubes has limited their range of application, and the GTR120W tube has been designed to overcome this economic difficulty by the careful choice of materials and the development of processes which can easily be mechanized.

The cost of this tube has been reduced to only a fraction of that of other cold-cathode trigger tubes and it can therefore be used advantageously in applications where a large array of tubes is required.

COLD-CATHODE tubes possess properties which allow them to be used in certain applications with advantage over thermionic valves. They possess an inherently long life when in the standby condition and the problems associated with the supply and dissipation of heater energy do not arise. These properties make them suitable for use in computers where large numbers of tubes are required. However, they are slower in operation than hot-cathode valves, and so in complex computers, where a large number of successive operations are required and speed is essential, other devices will be preferred. In more specialized commercial computers where fewer sequential operations are performed, speed is less important than long life and low cost, but in the past the cost of cold-cathode trigger tubes has prevented them from being used economically in such equipment. The GTR120W tube has been developed to provide a tube at only a fraction of the cost of any other now available, while still maintaining the characteristics required for efficient operation. The cost of the tube can affect its range of application, since it is often possible to overcome limitations imposed by the operating speed and the wider characteristic tolerances by the use of a larger number of tubes than would be permissible with a higher priced tube. For example, in digital apparatus the parallel mode of operation uses more tubes but gives a higher effective operating speed than the serial mode.



A production unit in the valve factory

Mechanism of Cold-Cathode Discharges

As the voltage between an anode and a cathode in a gas at a pressure of a few millimetres of mercury is increased, an extremely small current will begin to flow between them. This current is caused by the formation of ionization in the gas by one of three alternative processes. These processes are

(a) Photo-emission from the cathode.

(b) Radioactive materials.

(c) A priming discharge.

Photo-emission from pure metals is very small but with coated-cathode surfaces in normal illumination an appreciable emission is obtained, which is proportional to the intensity of illumination, and so the process is only reliable where light conditions are adequate. The second and third processes, which are independent of external illumination, require either the addition of radioactive material or one or two auxiliary electrodes.

Electrons formed will be attracted towards the anode and, if the voltage between anode and cathode is sufficiently high, they will obtain enough energy to ionize gas molecules into electrons and positive ions. The electrons will be swept to the anode and the positive ions accelerated to the cathode where secondary electrons are produced by bombardment. Electrons and positive ions are continually lost to the discharge by recombination at the electrodes, and if they exceed the number formed the discharge will decay and only a minute current will flow due to the continual arrival of primary charged particles in the gap. As the anode-cathode voltage is increased, the ionization formed will steadily increase until each primary electron gives rise to more than one secondary electron by the processes of ionization and positive ion bombardment. At this voltage, V_s, termed the striking voltage of the gap, the discharge current increases until it is limited only by the resistance in the external circuit. At the same time the voltage across the gap decreases until it reaches a stable value, V_m termed the maintaining voltage of the gap. This is the minimum voltage which will maintain the discharge; if the applied voltage across the gap is decreased below this value the discharge will decay rapidly.

* *Ericsson Telephones Ltd.*

When a voltage greater than V_s is applied across the anode-cathode gap, there is a delay before the full current flows. This consists of (a) a 'statistical time lag' before a charged particle is drawn into the gap, and (b) a 'formative delay' in establishing the discharge. The former depends on the amount of inionization produced by the processes previously described and is normally so small that it can be neglected, while the latter decreases as the overvoltage applied to the anode above the value V_s is increased. The total delay is termed the ionization time of the gap.

A cold-cathode discharge can be extinguished only by decreasing the anode-cathode voltage to a value below the maintaining voltage until the ionization has decayed due to recombination of the charged particles. The time which must be allowed before the voltage across the gap is returned to any value V_a, such that $V_s > V_a > V_m$, is termed the deionization time of the gap, and it will depend on the value V_a and on the current which had previously been flowing across the gap.

Operation of Trigger Tubes

A cold-cathode trigger tube consists essentially of three electrodes, namely, an anode, a cathode and a trigger, in which the trigger-cathode gap is much less than the anode-cathode gap. It may be considered as two diodes with a common cathode, such that V_{st}, the striking voltage of the trigger-cathode gap, is much less than V_{sa}, the anode-cathode striking voltage.

When a voltage V_a, less than V_{sa}, is applied to the anode, the cathode being earthed, and V_t, less than V_{st}, is applied to the trigger no current will flow. If V_t is then increased above V_{st}, the trigger-cathode gap will strike and electrons from the trigger discharge will be attracted to the more positive anode. Provided that this priming is sufficient, the anode-cathode discharge will be initiated, and the anode 'takes over'.

Two important dynamic characteristics of trigger tubes are the ionization and deionization times; the former is the sum of the formative delay times of the trigger-cathode and anode-cathode gaps and it determines the delay between the application of a pulse to the trigger and the appearance of an output pulse across an external resistor. The repetition frequency with which a tube may be used is dependent on the sum of the ionization and deionization times.

Characteristics Required

Having described qualitatively the principles of operation, it is now possible to determine some of the desirable properties of cold-cathode trigger tubes from the examination of a few typical circuits.

If a tube is connected as in Fig. 1 when only the trigger gap is conducting, the current I_t which flows is given by

$$I_t = \frac{V_t - V_{mt}}{R_t + R_k}$$

where V_{mt} is the maintaining voltage of the trigger-cathode gap.

It is desirable that the voltage V_t applied to the trigger should be as small as possible, consistent with I_t being large enough for the anode to take over. In order to keep the input impedance high it is necessary to have as large a value of R_t as possible, provided that the reliable operation of the tube is not affected. Thus it is important that V_{mt} and the take-over current shall both be small.

The voltage output V_o across the external resistor R_k when the main gap conducts is given by

$$V_o = V_a - V_{ma(max)}$$

The anode supply voltage must be less than the anode striking voltage so that the tube will not strike in the absence of a trigger discharge, and therefore the requirements for a large output voltage are high V_{sa} and low $V_{ma(max)}$.

A circuit which is often used to couple two tubes so that a current flowing through the first causes the second tube to strike is shown in Fig. 2. The trigger of tube 2 is connected by a capacitor to the cathode of tube 1 and by a resistor to a bias voltage V_b which is, say, 10V less than $V_{st(min)}$, the minimum trigger striking voltage of the tube. When the anode-cathode gap of tube 1 passes current a pulse of amplitude V_o is applied to the trigger of tube 2,

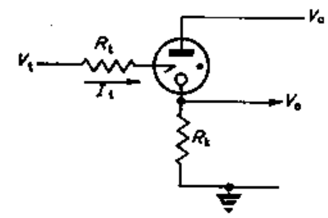


Fig. 1. Triggering circuit

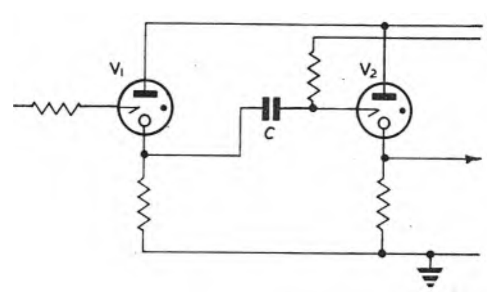


Fig. 2. Capacitor-coupling of two trigger tubes

and the condition that the trigger-cathode gap will strike is given by

$$V_b + V_{o(min)} > V_{st(max)}$$

but

$$V_b = V_{st(min)} - 10 \text{ and } V_{o(min)} = V_a - V_{ma(max)}$$

from which

$$V_{st(min)} - 10 + V_a - V_{ma(max)} > V_{st(max)}$$

and

$$V_{st(max)} - V_{st(min)} < V_a - V_{ma(max)} - 10$$

This shows that in this particular circuit the permissible variation in trigger striking voltage is dependent on the difference $V_a - V_{ma(max)}$ which may be obtained, and if the latter is large the tolerance on V_{st} is greater, giving the tube a longer life performance in this circuit.

If the ionization time is considered, the trigger formative delay may be reduced by increasing V_t and the anode delay reduced by increasing I_t and V_a. The deionization time is influenced more by the conditions of operation than by the static characteristics of the tube, being particularly dependent on the current prior to extinction and on the nature of the gas filling.

From the considerations discussed above, it is found desirable to have low values of V_{mt}, V_{ms} and take-over current and a high value of V_{sa}.

Design of Tube

In order to manufacture a low-cost tube it is necessary to use cheap materials, simple construction and to mechanize as many processes as possible.

Cathodes of pure metals may be used to provide very stable maintaining voltages but the values obtained are rather high. However, coated cathodes have lower maintaining voltages and, due to their efficient photo-emission, they may be used in normal light conditions without the need of additional sources of ionization, but they are normally expensive to prepare. An entirely different method of producing such cathodes which involves only easily obtainable material and elementary processing is used in the GTR120W. When a barium getter is fired in an inert gas, metallic barium is deposited over the getter frame, which, on combination with residual oxygen, forms a layer of barium oxide and free barium. In a suitable gas filling this surface gives an initial maintaining voltage of approximately 105V.

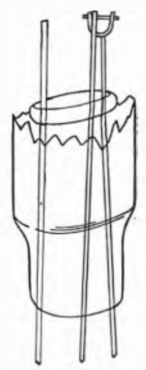
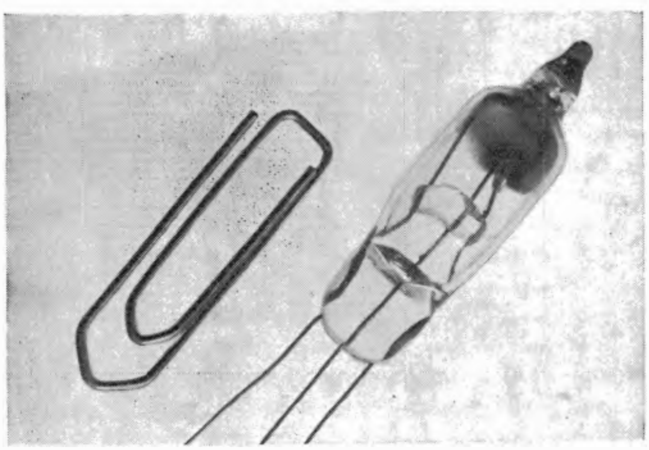


Fig. 3. Internal structure of the GTR120W tube



The GTR120W trigger tube

The nature of the gas filling is important, since sputtering due to positive ion bombardment causes a deterioration of the cathode surface. As light ions cause less sputtering than heavy ones, a pure gas of low atomic weight is required. In order that the copious emission of electrons from the cathode in high-intensity illumination should not cause the anode-cathode gap to break down in the absence of a trigger discharge, the gap must be made relatively insensitive to ionization. This will cause it to have a high take-over current and so a compromise must be made between the requirements of a low take-over current and a high anode striking voltage.

In its simplest form a trigger tube contains only three electrodes and since in normal illumination sufficient primary ionization will be produced by photo-emission from the coated cathode, no additional priming electrodes or material are added.

The internal structure of the GTR120W is shown in Fig. 3. The electrodes are composite wires of nickel and borated copper, the active surfaces being nickel, and the getter is welded to the end of the cathode support wire so that all are of the same length. The anode, trigger and cathode so formed are secured by a glass bead which prevents relative movement of the electrodes when this assembly is sealed into the glass bulb after the spacings have been adjusted. The borated copper leads provide vacuum-tight seals through the glass. The inertia of each electrode is small, thus giving good mechanical strength.

It is not necessary to pump the tubes to a very high vacuum, because a certain amount of residual oxygen is required in each tube, but the residual pressure before gas filling must be controlled within certain limits to ensure a standard product.

When a current is passed through cold-cathode tubes immediately after manufacture, large changes in electrical characteristics occur initially until a stable condition is reached. Before testing, the tubes are therefore subjected to an ageing schedule, which consists of passing through them a current which is higher than the normal operating current. The schedule for the GTR120W tube is short enough for it to be carried out on a rotary machine.

Tube Characteristics

The anode striking voltage is greater than 310V, and the recommended supply voltage is 180 to 310V. At 4·5mA the anode maintaining voltage is 95 to 140V, while with 310V on the anode, the maximum voltage at which trigger-cathode breakdown will not occur is +110V, and the minimum trigger voltage to cause breakdown is +170V. Thus it is possible to obtain an output voltage large enough to couple tubes in the circuit previously discussed.

A trigger bias voltage of 100V may be used and the minimum value of trigger coupling capacitor with trigger resistors exceeding 220kΩ is 150pF. Ambient illumination of 5 foot-candles is required for reliable trigger-striking and the minimum trigger-cathode current which must flow to cause the anode at 290V to take-over is 250μA; at 500μA the trigger-cathode running voltage is nominally 73V, and the trigger impedance in a conducting tube has a maximum value when the trigger-cathode voltage is 65V. While the static trigger striking voltage is 170V, an over-voltage is necessary to reduce the trigger formative delay; and with an anode voltage of 290V and the trigger pulsed to 200V the ionization time is less than 90μsec. The deionization time of the tube with standby voltages of 100 and 290V on the trigger and anode respectively is less than 3msec.

The maximum and minimum cathode currents are 9mA and 3mA respectively. When run at 4·5mA tubes may exhibit jumps of up to 10V in the maintaining voltage; these jumps will decrease as the current is increased and vice versa.

If 150 hours elapse without the tube passing current, self-ignition may occur with anode-cathode voltages of over 280V, but this may be prevented by passing a current of

3mA through the tube for at least one second. Thus satisfactory operation will be obtained by using anode voltages of less than 280V, or alternatively, if a larger voltage output is required necessitating an anode voltage of from 280 to 310V, the apparatus should be designed so that before normal operation all tubes are made to conduct for at least one second.

The failure of a tube is normally due to one or other of its characteristics passing outside the limits which may be tolerated in a particular circuit, and the circuit which allows larger tolerances will be more reliable and have a longer life.

In order to assist circuit design, the tube characteristics which have been quoted will be retained throughout the normal life of the tube. The life of a tube which passes

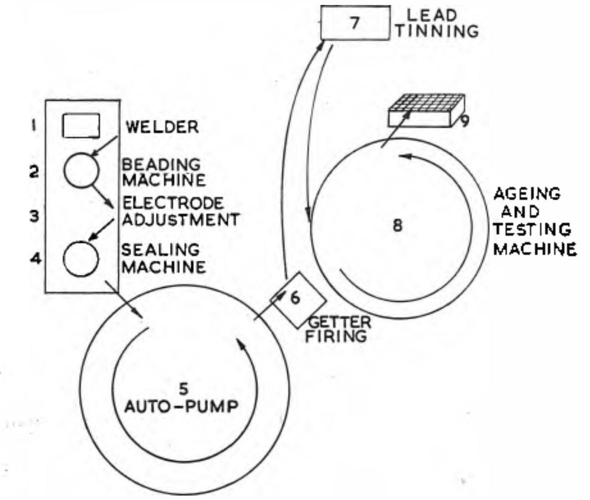


Fig. 4. Layout of a production unit

no current is over 10 000 hours, and the running life, when a current of 5·5mA is passed continuously, is over 2 000 hours. The latter condition is not normally used in practice, and lower duty cycle operations will give more than proportionately longer life, since the operating temperature of the tube will be reduced.

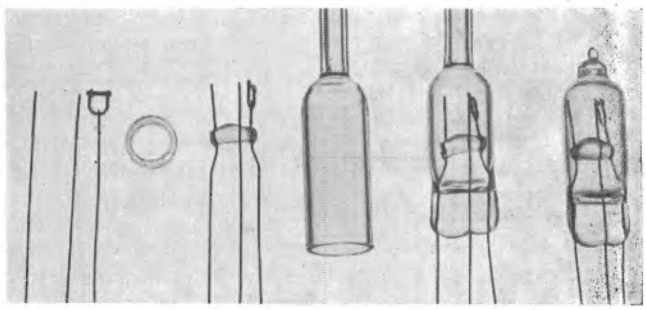
Manufacture

In order to keep the cost of the tube low it has been necessary to reduce the handling of components to a minimum and to mechanize as many processes as possible. Machines were made to manufacture the beaded assembly, and to seal this into the glass bulb, and as only elementary processing is required automatic pumping equipment is employed. A rotary ageing and testing machine was designed to eliminate expensive manual testing.

A diagrammatic layout of a production unit showing the flow of materials is shown in Fig. 4, and Fig. 5 shows the components used and the stages during manufacture. In position 1 (Fig. 4) the getter is welded to the cathode support wire, and the three electrode wires and a glass ring (Fig. 5(a)) are then loaded into a jig on the beading machine. As the machine indexes, sets of gas flames melt the ring into a bead which secures the three electrodes together in approximately their correct positions. On removal from the machine the assembly (Fig. 5(b)) is inspected, and the more critical trigger-cathode spacing is adjusted to within the tolerances required. The assembly is then placed into a jig on the sealing machine and the glass bulb placed over it. As the machine indexes, gas flames in each position melt the glass on to the borated copper wires and the seal is formed (Fig. 5(c)).

The tube is now inspected for cracks and other glass defects and placed in one of the positions on the automatic pumping machine (position 5) which passes it through the various stages of pumping and gas filling. The tube is first evacuated and leak-tested; if any defect is found the position is automatically isolated so that as the tube passes into the following stages it will not cause any deterioration of the pumping system. Tubes then pass into an oven which raises their temperature to release any gas absorbed on the internal surfaces. Following a period during which they are allowed to cool, the tubes are leak-tested again to check that no crack has occurred as a result of baking. In the next positions the tubes are filled with gas at the required pressure and sealed off. The tubes then leave the pump and the cathode surface is prepared by eddy-current heating the getter.

At this stage it is convenient to clean and tin the leads, which have become oxidized during the previous heating processes, so that good electrical connexions can be made.



(a) (b) (c) (d)

Fig. 5. Stages during manufacture

The tube is then secured by terminals on the ageing and testing machine, which passes it through several ageing positions and allows it to cool before reaching the testing positions. In each of these positions a certain test is performed, and when a tube fails it is released from the terminals and falls into a partitioned box under the machine which is periodically inspected to assess the causes of failure. Tubes which pass all tests are automatically painted for identification, removed from the machine and visually inspected for glass defects before being placed in a box for despatch.

The simplification of construction and processing, and the mechanization which it has been possible to achieve have all contributed to the low cost of the GTR120W tube.

Quality Control

The maintenance of a uniform quality of tube is ensured by the testing of incoming materials, check-testing after manufacture, the life-testing of samples and the testing of tubes immediately prior to despatch. The daily sample used for check-testing the tubes passed by the automatic tester is also tested manually for other tube parameters which require more intricate testing equipment.

When a tube is manufactured in such large numbers as the GTR120W it is possible to employ statistical sampling to avoid the use of an excessive amount of labour in performing these additional but very necessary tests. This involves the use of only a small sample from each batch and ensures that on average the specified level of quality is maintained.

Acknowledgments

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The Measurement of Spark Time Lags

By J. K. Wood*, M.Sc., Ph.D.

Two methods of measuring spark time lags are described. The apparatus is described and some results quoted to indicate some of the shortcomings of the apparatus and difficulties involved in making the desired measurements.

WHATEVER the subsequent mechanism of spark formation in air[1], the initial event is the production of an electron near the cathode of the spark gap by photo-ionization in the gas or by photo-emission from the cathode. The primary source of energy may be cosmic radiation, radioactive decay of a source placed near the gap or occurring naturally, or ultra-violet or X-radiation from a suitable generator.

If the electric field is sufficiently great the electron will ionize gas molecules in its path as it accelerates towards the anode. The number of free electrons in existence therefore increases exponentially with time and distance across the gap according to the relation.

$$n = n_0 e^{\alpha x} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

where n is the number of electrons a distance x across the gap from the point where n_0 electrons previously existed, or were created. α is the number of ion pairs produced by an electron in moving 1cm in the direction of the field. The process is known as an avalanche.

An exponential 'tail' of relatively immobile positive ions is left behind. Some electron-ion recombination occurs at the head of the avalanche, and the resultant electromagnetic radiation is responsible for photo-ionization in the gas and photo-emission from exposed surfaces as well as being the first visible phenomenon. Positive ions falling back to the cathode can cause electron emission from the cathode by impact. Some of these secondary electrons lead to further avalanches, and the gas continues to conduct. The gap current increases or decreases with time according to gap conditions.

If the current increases, it will rise rapidly until the external circuit is discharged or until it reaches a maximum value set by the operating conditions, as in a glow discharge tube. This is the basis of the Townsend mechanism of spark breakdown.

At higher pressures the space charge due to the positive ions left by the avalanche can become so high that secondary avalanches initiated by photo-ionization near the primary avalanche occur in the body of the gas, and by this means a densely ionized channel or 'streamer' propagates rapidly across the gap. This mode of breakdown is typical of the corona discharge and lighting, and is the basis of the streamer mechanism of the breakdown of a spark gap.

If breakdown by the Townsend mechanism is arbitrarily assumed to have taken place when the number of positive ions produced has risen to N, and if this takes s successive avalanches, then, neglecting the effect of space charge and assuming a single initiating electron,

$$N = e^{\alpha d} + \gamma e^{2\alpha d} + \gamma^2 e^{3\alpha d} + \ldots + \gamma^{s-1} e^{s\alpha d} \quad (2)$$

or

$$N = e^{\alpha d}\,[(\gamma e^{\alpha d})^s - 1]/[(\gamma e^{\alpha d}) - 1] \quad \ldots\ldots \quad (3)$$

where d is the gap length and γ is the number of secondary electrons produced per positive ion in the previous avalanche. Breakdown occurs when $\gamma e^{\alpha d} > 1$. α and γ can be determined from steady state measurements knowing the applied voltage, d and the gas pressure, for a given gas and electrode material and shape. Hence s can be determined for a given case in terms of N. A similar analysis has been applied to breakdown in nitrogen by Kachickas and Fisher[2].

Since the bulk of the positive ions and ionizing radiation are produced near the anode, the spark formation time t is given by

$$t = s(d/v) \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

where v is the velocity of the avalanche across the gap when the photo-electric secondary effect predominates ($\sim 10^7$cm/sec) or is the velocity of the positive ions ($\sim 10^5$cm/sec) if positive ion impact at the cathode is the predominant secondary process. If $\lambda e^{\alpha d}$ is only just greater than unity, s may be as high as 100 and t as great as 10^{-3}sec.

In cases where the streamer mechanism is held to apply, the space charge often becomes great enough to cause breakdown during the progress of the first avalanche, with the result that t may be of the order of 10^{-7}sec.

A large amount of work has been put into the determination of the limits of applicability of the theories, but the transition could not be accurately defined even by spark formation times which vary through several orders of magnitude. In the present work it was hoped to resolve the transition by means of a more refined experimental technique.

In common with previous work, an observation is commenced by superimposing a step voltage on a steady voltage across the gap. The total voltage exceeds the d.c. breakdown potential by a known percentage. It is necessary to eliminate any random delay which may occur between the commencement of the observation and the appearance of an electron leading to breakdown. This was previously achieved by irradiating the gap with an ionizing radiation to such an extent that the random delay, or 'statistical time lag' was small compared with the spark formation time or 'formative time lag'.

In the present instance it was felt that strong radiations would shorten the formative time lag, inducing space charge formation before the observation began. Thus a method was sought which would allow the measurement of formative time lags without the higher level of irradiation previously used. Although the overall space charges have been shown to be small by the workers concerned, it was felt that the microscopic space charges from avalanches in the approach voltage field would enhance emission from points on the cathode by increasing the field and by positive ion bombardment when the step voltage appeared across the gap.

Experimental Methods

Two methods were tried. Using radium sources, the total time lag was measured with various low intensities of irradiation, and the results extrapolated on the assumption that the average statistical time lag varied as the inverse of the intensity of irradiation to obtain the formative time lag. Secondly, using a controlled weak pulse of irradiation at a known time, the statistical time lag could be eliminated. By terminating the observation after a period just long

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enough to include the formative time lags, the spurious readings due to breakdown caused by extraneous sources of ionization can be made negligibly few.

Method Employing Radium Sources

It was felt that the reciprocity of mean statistical time lag and intensity of irradiation would be obeyed better by employing various sources at a fixed distance than by employing a given source at different distances, because the proximity of various materials would contribute to the total irradiation at the cathode by scattering. The method involved twenty measurements of a time lag with each of ten different radium sources, and repeating the experiment on ten different occasions. The sources were used in a different order on each occasion. Before each experiment the electrodes were repolished, so that it was possible to assess the effect of progressive oxidation of the electrodes. By this method the effect of differences in the surface finish of the electrodes and other random quantities were reduced.

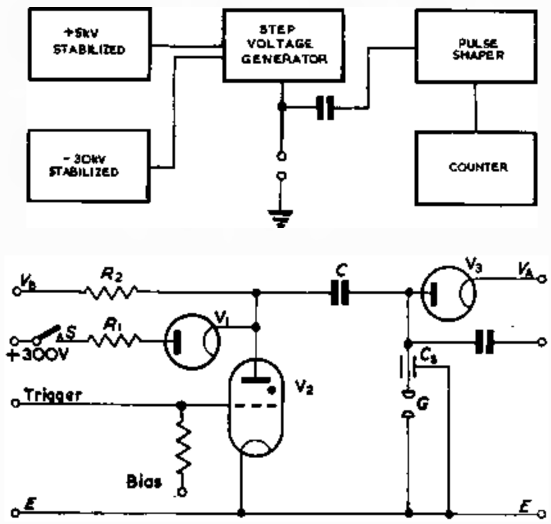


Fig. 1. (a) Apparatus employing radioactive sources (b) Step voltage generator

The slope of the line relating mean time lag to the inverse of irradiation intensity was determined by the method of least squares. The intercept corresponding to infinite irradiation was taken to be the formative time lag. The process was repeated with other gap conditions, each value of formative time lag depending on 2 000 individual measurements.

Established techniques of spark time lag measurement mainly cover small ranges and involve a certain amount of process work before the results are available. The analysis described above requires a wide range of measurement, and if the results are immediately available, a considerable amount of labour is saved. Consequently, a crystal controlled pulse generator and counter are used in the present case, which, with a suitable gate, measure a time in microseconds to the accuracy of the crystal, and $\pm 1\mu$sec.

A block diagram of the apparatus used is given in Fig. 1(a). Both e.h.t. units consist basically of oscillators, step-up transformers and rectifiers. The 30kV unit employs a Cockroft-Walton voltage multiplier with metal rectifiers. Stabilization is achieved in both cases by comparing a proportion of the output voltage with a reference battery and amplifying the difference to control the screen of the oscillator valve. The circuits are conventional, and similar to that given in Fig. 3. The high voltage feedback resistor is also used for voltage measurement.

The step voltage which is superimposed on the steady or approach voltage across the gap must rise rapidly, and remain constant at the new level for as long as any measurement is likely to last. This has previously been achieved by isolating the gap from the supplies with large resistances, giving a long time-constant with the capacitance discharged by the spark. This necessitates long charging periods between successive shots. In the present case, a high peak inverse voltage diode is substituted for the resistance, which cuts off when the step voltage is superimposed on the approach voltage and allows rapid recharging of the circuit.

Consider the circuit of Fig. 1(b). The capacitor C charges to a total voltage $V_a + V_b$, where V_a is the approach voltage and V_b the supply to the step voltage generator. The diode V_1 is cut off because V_b is greater than 300V. On receipt of a triggering pulse, the thyratron V_2 conducts, and its anode potential falls from V_b to V_c, the operating anode potential of the thyratron. Since V_c is less than 300V, the diode V_1 conducts and maintains the thyratron anode current, thus keeping the thyratron anode potential constant at V_c. Since high peak inverse voltage diodes often have an appreciable forward resistance, R_1 may in some cases be omitted, as its only purpose is to limit the current to a value within the rating of the components. The resistor R_2 limits the current from the e.h.t. source V_b to a safe value, which should not be large enough to operate the thyratron V_2.

When the thyratron anode potential falls, a redistribution of charge between C and the stray capacitance C_s occurs. The anode potential of diode V_3 is forced more negative and V_3 cuts off, thus isolating the spark-gap circuit. Equating charges before and after triggering thyratron V_2

$$C(V_a + V_b) + C_sV_a = C(V_g + V_c) + C_sV_g \quad \ldots . \; (5)$$

Hence

$$V_g = V_a + \frac{C}{C + C_s}(V_b - V_c) \quad \ldots\ldots\ldots \; (6)$$

where V_g is the voltage across the gap G.

A spark in the gap causes the potential across the gap to fall rapidly and V_3 conducts again. By under-running the filament of V_3 the conducted current is limited, so that the e.h.t. source V_a does not discharge through the spark. Since the time-constant of charge depends on R_2, this should be small; a time-constant of 0·1sec is easily obtained. At the termination of each observation the switch S is momentarily opened, and the thyratron is extinguished, allowing C to charge. This occurs whether or not a discharge has occurred in the gap. During the observation leakage currents cause the gap voltage to decrease with a time-constant of about three minutes.

A signal is taken from the high voltage electrode of the spark gap via C, which is a small electrode with a capacitance of about 3pF. This signal triggers a pulse shaper consisting of a bistable multivibrator and cathode-follower. The waveform is a square pulse lasting for the interval between the commencement of the observation and the appearance of a spark, and is applied directly to the counter as a gate pulse. Catching diodes ensure that the output voltage lies between the correct d.c. limits.

Analysis of readings revealed that the progressive oxidation of the electrodes did not affect the measured time lags, but in spite of the large number of readings taken, the error in the final value of formative time lag proved to be as large as the formative time lag itself. The main source of error was undoubtedly the large ratio of total time lag to the resulting formative time lag. This could only be reduced by increasing the level of irradiation, which would defeat the

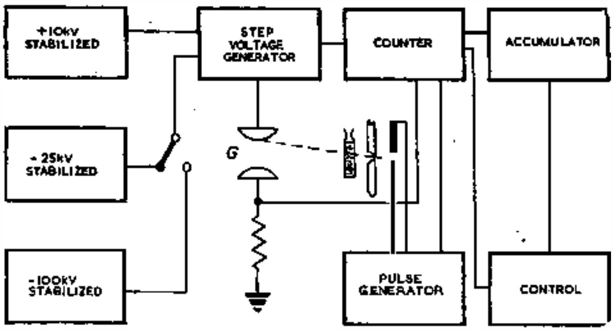


Fig. 2. Apparatus employing auxiliary irradiation

to allow measurements not only of formative time lag, but also of direct breakdown potentials, of currents in the gap immediately before and during breakdown, and of time lags with and without other sources of irradiation, and with the additional possibility of using gases other than air. A block diagram appears in Fig. 2.

Since a timing signal was required which would terminate the observation, it was felt that a binary counter would best fulfil the requirements. A series of logarithmically related timing pulses can be extracted at successive points in the circuit, and the observation terminated at 1, 2, 4, 8, 16, . . . µsec after start of the observation. This was ideal as the percentage error in formative time lag measurements was expected to be fairly constant, and the dis-

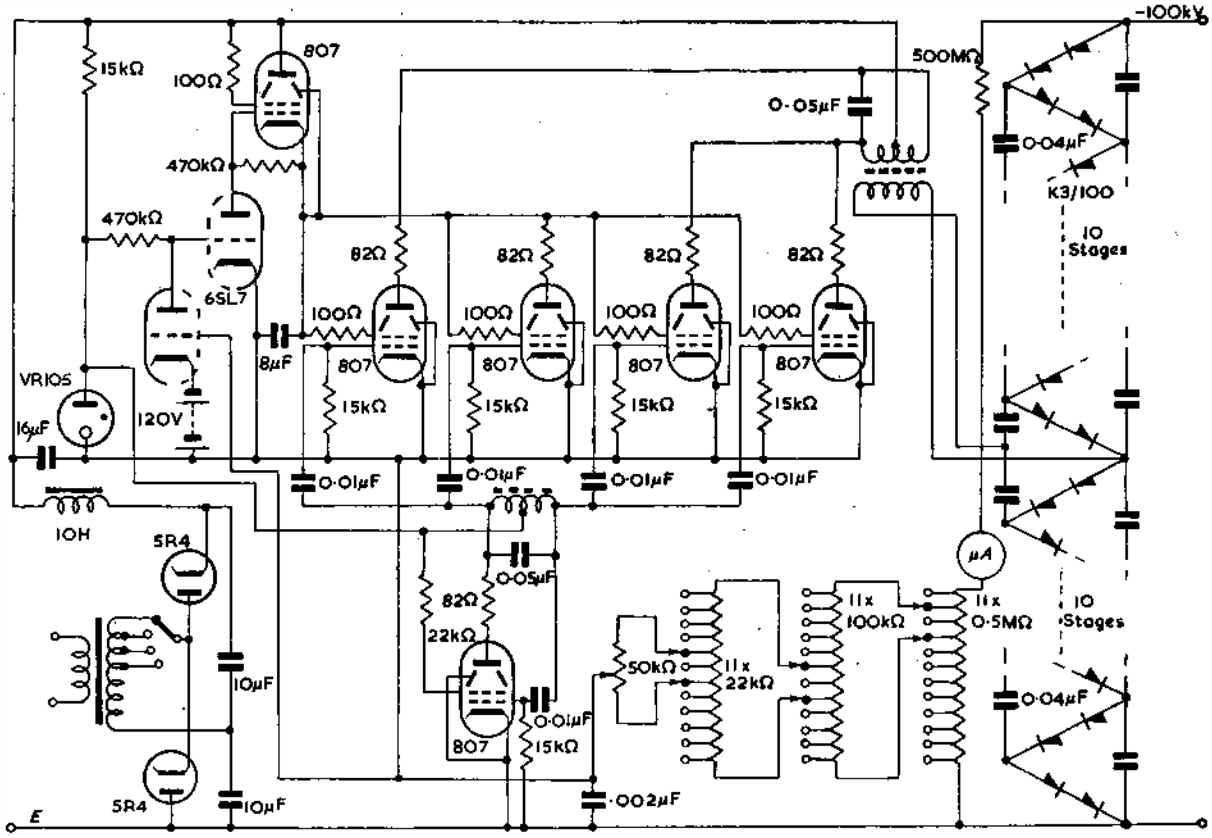


Fig. 3. 100kV generator

object of the experiment. It was for this reason that the second method of formative time lag measurement was attempted.

Method Employing Auxiliary Spark Irradiation

In this method no analysis is required to obtain the formative time lag, though the experimental technique involves a greater quantity of apparatus. Since a spark is an intense source of ultra-violet radiation, the beam, defined by a collimator, is attenuated by releasing only a small amount of energy, by omitting the usual quartz lens system, by placing the source some distance from the main gap, and by attenuation with organic liquids in a silica vessel in the path of the beam. The whole system was placed in a chamber allowing a pressure variation from one-tenth to about fourteen atmospheres. The earthy electrode of the main gap was connected to earth through a small resistance across which a signal was developed during breakdown. It was intended that the apparatus would be flexible enough

tribution of spurious time lags (sparks initiated by causes other than the auxiliary spark irradiation) exponential. However, a binary output involves conversion before the results are useful. Using components which were readily available, a simple relay operated binary accumulator was built[3] which would sum any given batch of results and thus save a considerable amount of labour in conversion. It was not considered worthwhile to build a binary to decimal convertor.

The variable e.h.t. sources were +10kV(max) from a conventional 50c/s transformer and rectifier, with series valve stabilization; and −25kV(max) and −100kV(max) from oscillator and voltage multiplier circuits, stabilized by controlling the oscillator amplitude. The circuit given in Fig. 3 is that of the −100kV unit.

The step voltage generator circuit is similar to that given in Fig. 2, with components rated to withstand higher working voltages than those met in the circuit described previously. Another thyratron is triggered simultaneously

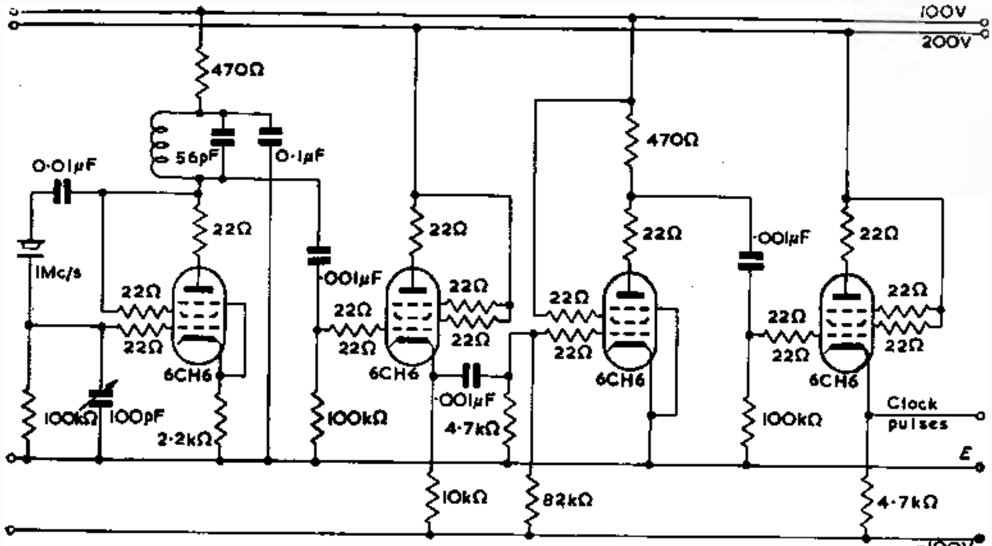


Fig. 4. Clock pulse generator

with that in the step voltage generator, delivering a 20kV(max) pulse to the auxiliary spark gap. This gap is about 1mm long, between 1mm diameter flat ended tungsten rods, and is severely over-volted to reduce the time lag to a value small compared with the lag in the main gap.

The main gap consists of solid brass electrodes machined and polished to a contour proposed by Farley and Hilditch[4], giving an area of uniform field in the centre. The fixed electrode is the high voltage electrode and the earthy electrode is mounted on a micrometer base allowing adjustment of the gap length up to 2cm without rotation of the electrode. The electric field pattern in the pressure chamber was checked in an electrolytic tank, using a wedge shaped model to simulate the cylindrical system. It was found that the electric field was nowhere higher than in the centre of the gap. The auxiliary spark-gap, collimator and silica container are housed in a long side arm to avoid distorting the field in the main chamber. A second shorter arm at right-angles to the first carries an access and inspection port with a window of thick glass allowing observation of the main spark-gap.

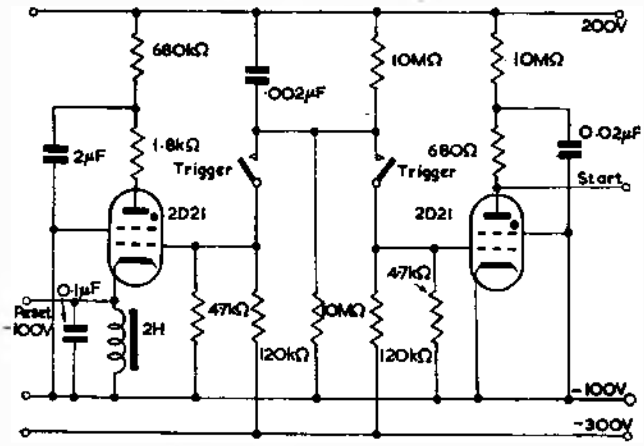
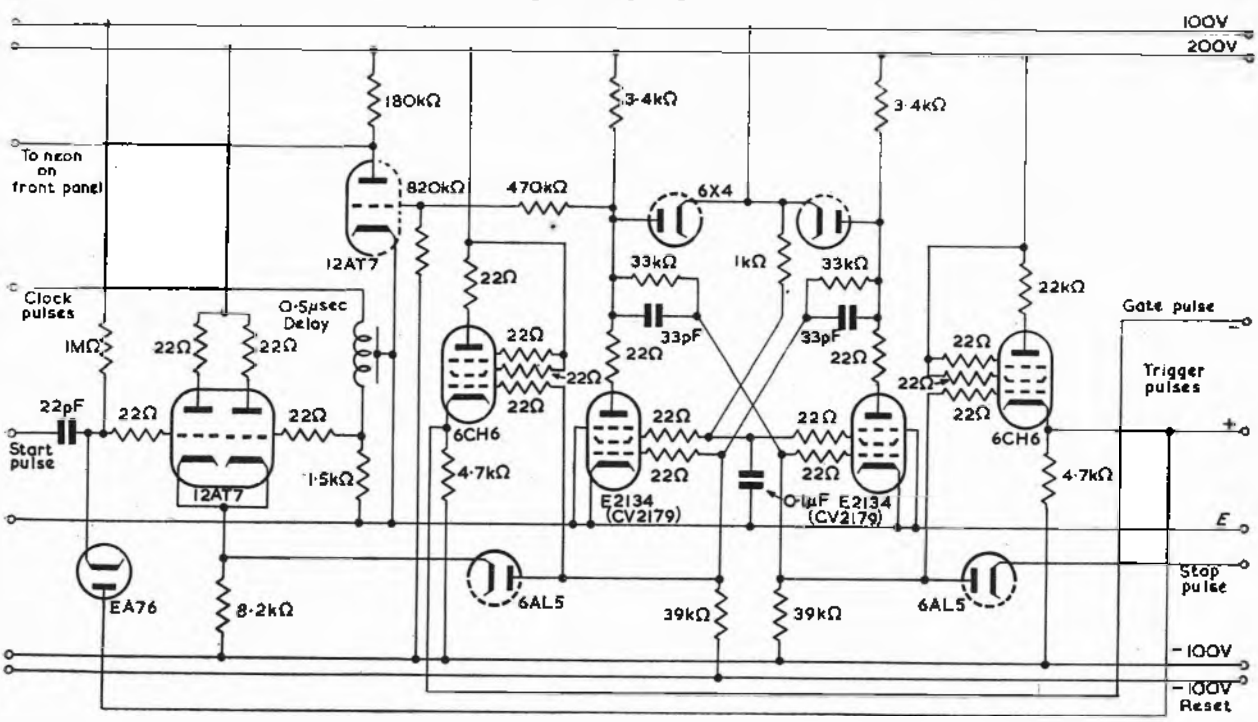


Fig. 5. Start and reset circuits

The counter consists of a crystal controlled oscillator, which delivers pulses at the rate of 10^6p/s to the counting circuits during the time interval to be measured (Fig. 4). The counter is reset by a thyratron circuit (Fig. 5) which injects a positive pulse into the negative line to which the

Fig. 6. Gate pulse generator



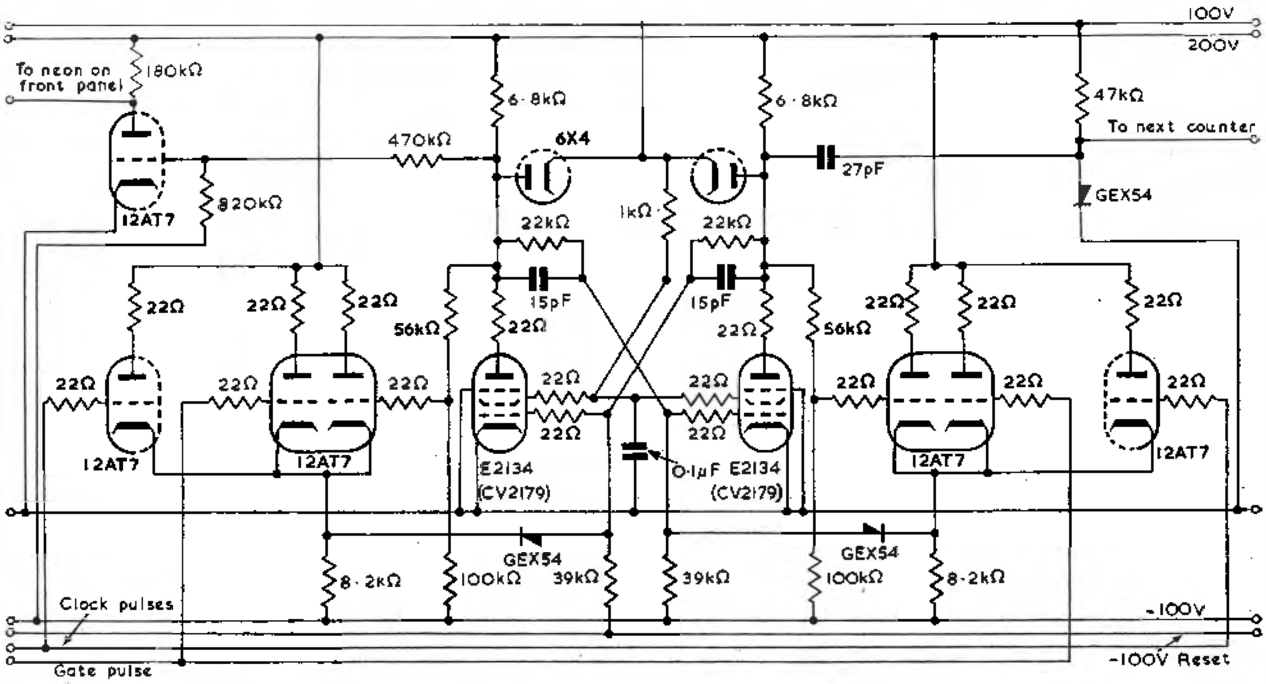


Fig. 7. Counters 1 to 4

appropriate grid leaks are returned. All the indicating neons are lit in the reset condition. The start pulse, also generated by a thyratron (Fig. 5) triggers a bistable multivibrator (Fig. 6) through a coincidence circuit which ensures that the output from the multivibrator, which admits pulses to the counting circuits, changes between two pulses. This waveform (gate pulse) is also used to trigger the main and auxiliary spark pulse circuits. The first pulse to the counters extinguishes the neons, so that the total is zero at the start of the observation; defined by the triggering of the spark pulse circuits. When breakdown occurs a pulse from the main spark-gap circuit resets the multivibrator and no more pulses reach the counter. Thus the neons indicate the number of microseconds to the next lowest whole number subject to the accuracy of the crystal.

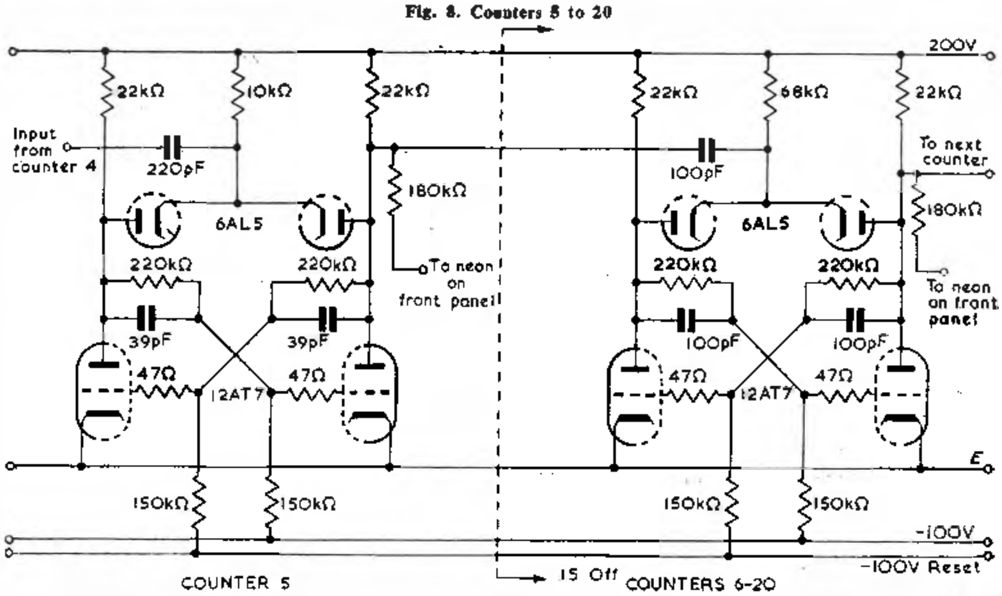
Thyratron circuits are used to eliminate 'double strobing' due to contact bounce, while coincidence circuits ensure a well-defined pulse for triggering purposes.

All the counters are bistable multivibrators; the first four employ high-slope pentodes (Fig. 7) and the remaining sixteen double triodes (Fig. 8). Only the first counter is controlled by the gate pulse. All the circuits are run well within the limit set by rise-times in order to ensure maximum reliability in operation. Since the output to the accumulator is taken from the anodes of the indicating neons they must all be run at the same cathode voltage. Consequently, those operated by the pentode counters require a buffer amplifier to adjust the d.c. level.

A further circuit triggers the accumulator when the counter stops. If no time lag is measured (the gap does not breakdown, or the counter fails to stop) then the counter is stopped by the reset action and no pulse is delivered to the accumulator. A short suppressor grid base pentode circuit achieves this, and is given with the neon indicator circuit in Fig. 9.

The output voltages from the anodes of the neon indicators are taken to the relay operated accumulator and the sum displayed on a bank of low voltage bulbs in the control panel. Also housed in the control panel is a clock unit which controls a uniselector. Pulses are delivered every half second and the 50 position uniselector completes its cycle every 25sec. The circuit is arranged to reset the counter, start the counter (which triggers the spark gap circuits), and terminate the observation after one second by interrupting the anode current of the thyratron in the step voltage generator. The effective length of the observation

Fig. 8. Counters 5 to 20



may be less than one second if the counter is arranged to ignore signals received after 2^n μsec as mentioned previously. The circuit is shown in Fig. 10, from which it will be seen that two cycles are available, 12½sec, and 25sec.

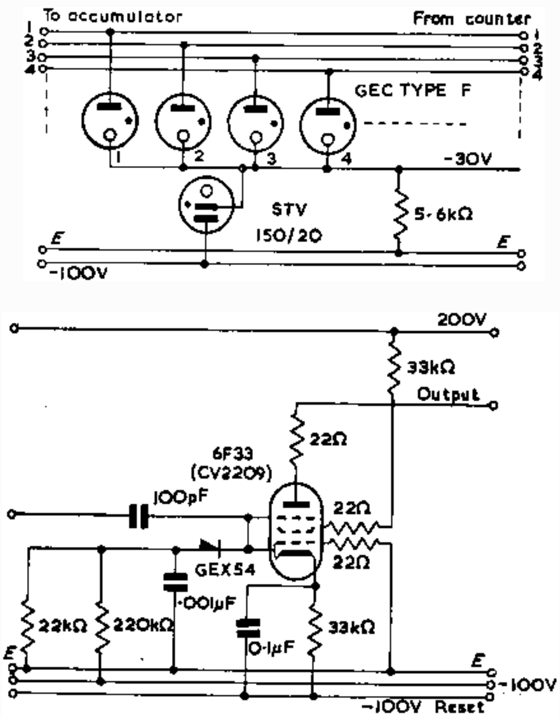


Fig. 9. Counter output circuits

It will also be seen that on switching off, the cycle is completed before the uniselector stops. The majority of the cycle is occupied by the simultaneously recharging of the high-voltage circuits and the operation of the accumulator.

Conclusion

Using radioactive sources a trend was observed in the variation of formative time lag with applied overvoltage which was of the shape observed by other workers, but with lags of the order of 1 000 times longer. While an increase in formative time lag was hoped for, an increase of this magnitude was more than could be explained on the basis of pre-breakdown space charges. Since the error was large (see Fig. 11 for a typical curve) no significance could be attributed to the measured values. It was therefore hoped that the second method, using pulsed irradiation,

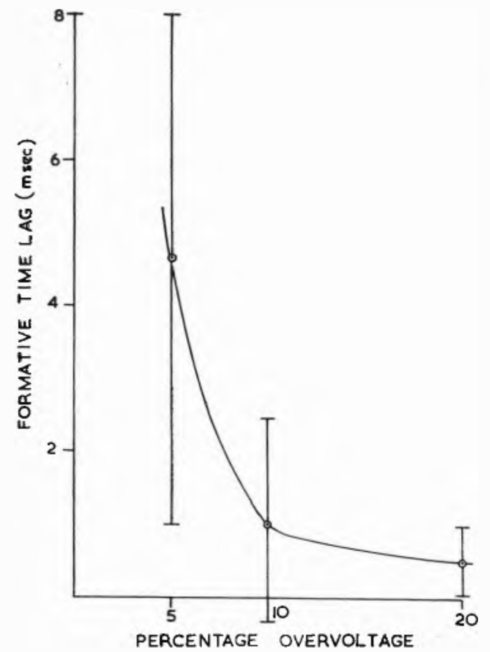


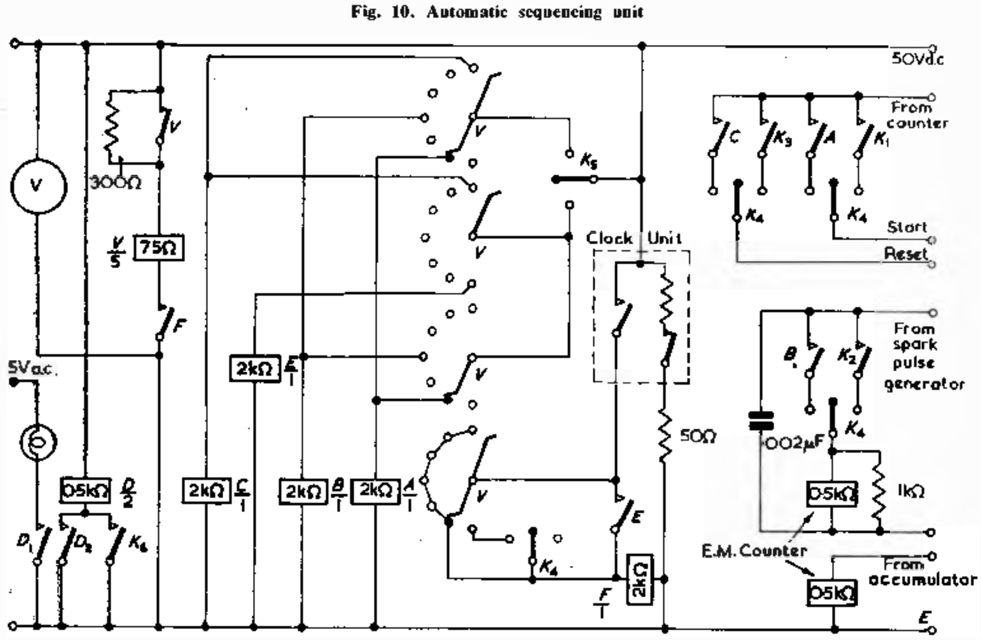
Fig. 11. Formative time lag versus percentage overvoltage

would be better in that the statistical time lag would be eliminated at source without introducing space charges known to be present in work involving continuous and more intense levels of irradiation.

A considerable amount of difficulty was experienced in attenuating the irradiation sufficiently; the fluids used appeared to be either transparent or opaque to the active wavelengths. Fig. 12 shows some results with various fluids, indicating their cut-off wavelengths. The average time lag is large in the opaque cases as the pressure chamber had a considerable shielding effect to natural radiations. The maximum possible time lag was a second in the arrangement used. It can be seen that for sparks in air at atmospheric pressure between brass electrodes the effective wavelengths are those below about 2 300Å. Since oxygen in the air begins to absorb radiations when the wavelength falls below 2 000Å only a narrow band is available. Although the curves imply a smooth transition from long to short statistical time lags. this was not the case; the intermediate values consist of an average of lags of the order of one second and lags of the order of one microsecond.

A typical curve of formative time lag against over-

Fig. 10. Automatic sequencing unit



voltage is given in Fig. 13. It will be seen that the lags are very similar in magnitude to those measured by Fisher and Bederson[5]. From this it is inferred that the lags measured by the first method are spurious, and that the effect of microscopic pre-breakdown space charges is negligible.

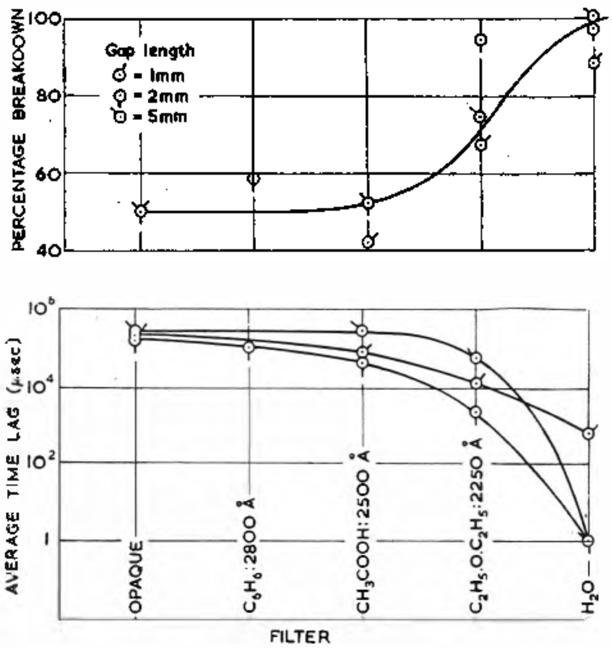


Fig. 12. Effect of irradiation wavelength on breakdown

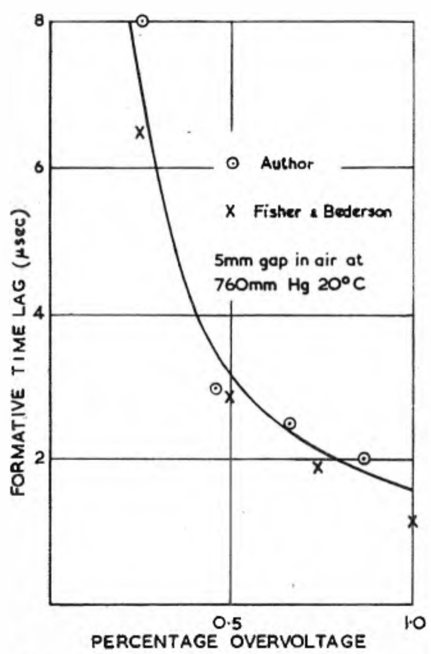


Fig. 13. Formative time lag versus overvoltage

Instead of an intense local emitting area on the cathode, resulting from a positive ion cloud left by an avalanche occurring before the observation began, causing a significant reduction in the formative time lag, the effect must be a slight reduction in the lag, the effective zero in time being approximately one avalanche transition time in advance of the measured zero. This difference would be so small as to be insignificant beside the variation between individual readings, as can be seen from the distribution shown in Fig. 14. It can therefore be stated that the fine structure of small space charges occurring during measurements of formative time lags does not measurably affect the results, and that the transition from Townsend to streamer breakdown is no more clearly defined than previously. Possibly the physical processes themselves merge in the transition region.

During the course of the experiments the apparatus functioned well, but considerable difficulty was experienced in keeping spurious pulses due to the sparks from affecting the operation of the electronic circuits. Shielding was introduced at sensitive points in the apparatus, and high impedance signal connexions were reduced to a minimum. This achieved the necessary freedom from stray pick-up, though the margin of safety was small. The apparatus was assembled in a rather small space, and if the electronic units

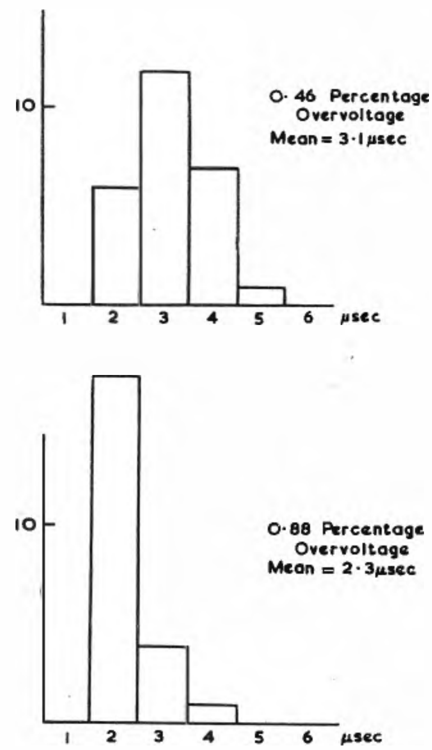


Fig. 14. Formative time lag distributions

could have been separated from the spark circuits by a reasonable distance and given a separate mains input and earth, then the stray pick-up would have probably been reduced to negligible proportions.

Acknowledgments

The author would like to thank Professor F. C. Williams for facilities provided, and the Department of Scientific and Industrial Research for a maintenance grant over a period during which this work was carried out. The author is grateful to Drs. D. R. Hardy and R. Cooper for their help and interest in the work and Drs. D. B. G. Edwards and G. E. Thomas who were largely responsible for the design of the counter described.

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A Crystal Controlled R.F. Induction Heater

By E. Cohen*, M.Sc., Grad.I.E.E., A.Inst.P.

The construction, operation and performance of a 2kW crystal controlled radio frequency induction heater is described. The several advantages associated with fixed frequency operation are discussed and it is shown that the greater complexity of the circuit is more than compensated by absence of radio frequency interference with essential services, greater efficiency conversion of d.c. input to r.f. output and ability to match an external load to the output impedance of the final power amplifying stage, by a novel construction of the output transformer. Experiments indicate that a graphite susceptor, if suitably thermally insulated, can be raised to a temperature in excess of 2 000° C in eight minutes.

The heater has been designed for the growth of single crystals of calcium oxide by the vapour phase transport method. The ability to produce such crystals is an essential prerequisite in studying their bulk semi-conducting properties and their role in the oxide coated cathode.

GENERATORS of r.f. power for industrial and medical purposes employ, almost universally, thermionic valve generators and in some cases, power amplifiers. Power is transferred from the generator to the load by a work circuit coupled to the oscillator or power amplifier tank circuit both of which produce radio frequency fields which may affect radio receivers in the vicinity. Further radio frequency currents may be injected into the power supply mains and cause interference by induction or re-radiation from the supply wiring.

The h.t. supply for the oscillator may be either unrectified a.c. or normally rectified single or three-phase a.c. The amount of smoothing is often small since usually a single reservoir capacitor and choke input filter is used, hence it is reasonable to suppose that amplitude and frequency modulation of the oscillator will occur due to the alternating component of anode voltage. Further the operating frequency may vary due to changes of the electrical characteristics of the load during the heating period so that interference may be caused over a band of frequencies about a mean value.

All commercial induction heaters at present available in this country consist of a self-excited oscillator with the work inserted either direct in the frequency determining inductive element or else coupled into the latter by means of a step down transformer to obtain a degree of impedance match between the work and the loaded tank circuit.

The generator will deliver full 'designed' power to the work circuit only when this presents the optimum load, and for heavier or lighter loads there will be a reduction in power. Since the optimum load will be presented during only part of the heating cycle for perhaps one of a range of applications, it is seen that the generator must have a considerable reserve of power in order to give a reasonably good performance over the whole range. Some types of equipment may be fitted with a variable device by which a degree of matching can be achieved for different applications but even when this is done there will be a change of load during the heating cycle.

These considerations have led to the construction of an r.f. induction heater employing a crystal oscillator followed by a buffer and a power amplifier stage conforming in the first place to the stringent requirements of the Atlantic City Agreement to which the United Kingdom is a signatory. This agreement stipulates the permitted working frequencies for industrial heating and diathermy applications for medical purposes. As the allowed bands are very narrow, 0·05 per cent wide at 13·66Mc/s, 27·32Mc/s, or 40·98Mc/s crystal drive is mandatory. This implies continuous tuning of the output circuit so that the power amplifier stage does not suffer serious damage by excessive and dangerous off-resonance currents. Class-C operation is used throughout in order to obtain the highest efficiency of conversion of d.c. into a.c. energy. Although the theoretical maximum for the conversion efficiency will be 100 per cent, in practice for a self-excited power oscillator the figure is 60 to 70 per cent, and 70 to 80 per cent for a class-C r.f. power amplifier. Hence with a given mains transformer a driven power amplifier is more efficient than a self-excited oscillator, assuming coupling conditions of the work remain the same. With any power amplifier or oscillator harmonic generation of the fundamental frequency is prolific owing to the nature of class-C operation. For good anode efficiency the voltage drop across the tank circuit (which determines the instantaneous anode voltage) should approach a sine wave characteristic. Although the anode current flowing through the tank is in the form of short pulses containing considerable harmonic energy, a resonant circuit discriminates against harmonic voltages across the circuit according to the Q of the circuit. If the Q is sufficiently high, the wave shape of the voltage drop across the tank circuit will be essentially sinusoidal. However, these harmonic voltages will radiate energy and may lead to parasitic oscillations if a reactive path is present. With self-excited oscillators that work directly coupled into the tank coil, little can be done to prevent dissemination of harmonic energy. With inductively coupled work, the valve and ancillary circuits are generally well screened, sometimes doubly screened, by heavy steel plates. It is possible, by appropriate circuits in the case of a power amplifier to by-pass harmonic and parasitic currents harmlessly to earth and copper mesh screening is adequate for all these cases. This latter case has been adopted in the r.f. heater described in this article and harmonic tests with a communication receiver and a sensitive grid dip oscillator show negligible harmonic energy radiation.

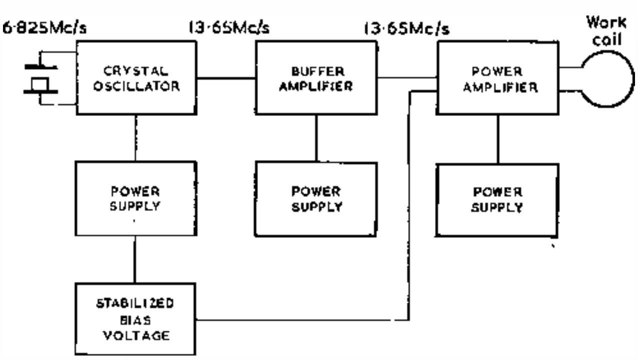

Fig. 1. Block diagram of crystal controlled r.f. induction heater

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Description of Circuit

A block diagram is shown in Fig. 1 and is self-explanatory. The entire equipment is contained within a two foot cube frame of slotted aluminium (Dexion) fitted with castors so that the heater is fully portable. The front panel carries various meters which monitor various voltages and currents and constitutes a means whereby the heater may be adjusted and tuned. Water cooling is essential to prevent the temperature of the work coil from rising excessively due to the close proximity of the heated work and also to reduce Joule heating (I^2R losses) in the power amplifier tank and work coil circuits owing to skin effect phenomena. Maximum power input to the final stage is 2kW (i.e. 4kV at 500mA) and at this d.c. input forced air cooling of the anode and grid terminals of the final amplifier valve (Type R.C.A. 833A) is advisable as the temperature difference of the metal to glass seals should not exceed 145°C. A small fan is used delivering approximately 40ft^3/min from a 2in diameter nozzle directed vertically on to the bulb between the anode and grid seals. R.F. power to the work coil is adjusted by means of a Variac in the mains side of the power transformer feeding the final stage.

The Oscillator and Buffer Amplifiers

Fig. 2 shows the circuit diagram of the complete r.f. heater. The oscillator is the well-known 'tritet' circuit and was chosen as it was of the electron-coupled type, the screen grid of the valve serving as the anode of a triode oscillator. A separate output tank circuit is used in the actual anode circuit. Because of the shielding effect of the screen grid, the coupling between the two circuits is comparatively small and exists principally through the common electron stream within the tube. Thus, when the load is coupled to the output circuit, its effect will be much less than if it were coupled directly to the frequency generating circuit. The anode circuit of V_1 is tuned to twice the crystal frequency and is coupled to V_3 and V_4 by π-section coupling (C_6, L_5, C_9, C_{10}). This type of circuit allows a relatively large fixed capacitance (C_9, C_{10}) connected directly from grid to ground in the buffer amplifier circuit and is highly advantageous in preventing v.h.f. harmonics generated in the grid circuit from developing an appreciable voltage between grid and ground. This not only prevents amplification of such harmonics in the anode circuit but also helps to keep harmonic currents from flowing in the d.c. grid return lead. C_9, C_{10} are also useful in stabilizing V_3, V_4 to prevent self-oscillation at the operating frequency. The larger the capacitance of C_9, C_{10} in comparison with the capacitances in use at C_6 the greater the impedance step-down between V_1 anode and the buffer amplifier grids, thus the crystal oscillator anode is reflected as a comparatively low resistance at the grid of the amplifier. This, together with the fact that any energy fed back from the buffer amplifier anode circuit through the valves' grid-anode capacitance cannot develop much feedback voltage across the large fixed capacitance between grid and cathode, effectively prevents self-oscillation and avoids the necessity for neutralization of the amplifier. L_4 is a small coil in the buffer grids to prevent v.h.f. parasitic oscillations in the buffer stage. Likewise R_9 and R_{10} serve the same purpose in the anode circuits of V_3, V_4.

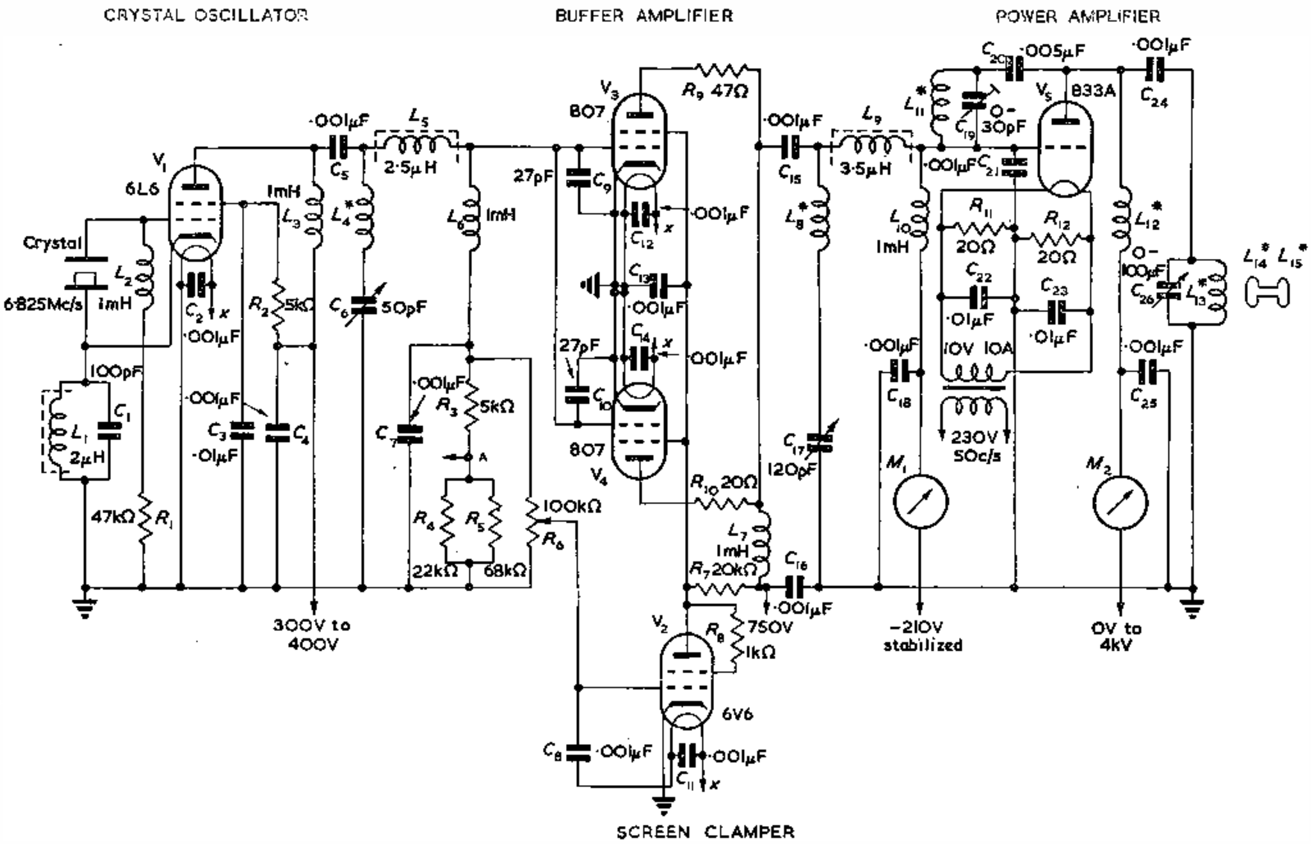

Fig. 2. Circuit diagram of crystal controlled r.f. induction heater

* L_4, L_8, *six turns, 14 s.w.g. on ½in former, parasitic chokes.* L_{11}, *high Q coil to resonate with* C_{10} *plus strays at operating frequency.* L_{12}, *80 turns, 20 s.w.g. enamelled wire on 1in o.d. Pyrex tube.* L_{13}, *six turns, ¾in o.d. copper tubing self-supporting coil of 3in i.d.* L_{14}, *single turn copper sheet around* L_{13} *and insulated from it by two p.t.f.e. sheets.* L_{15}, *single turn copper work coil of 1½in i.d.*

In the absence of any protective devices for the buffer stage, a failure of oscillations in V_1 (due, say, to catastrophic over-heating of the crystal, or other causes) would lead to the grids of V_3, V_4, being at earth potential owing to lack of flow of grid current in R_3. Such a course would lead to excessive currents being drawn with consequent overheating and failure of the valves. To prevent this, a small beam tetrode V_2, of the 6V6 type, is used as a screen clamper whose anode voltage is drawn from V_3, V_4 supply via R_7. The grid supply is d.c. derived from a 100kΩ potentiometer in parallel with the fairly low resistance of the grid leak R_3 (5kΩ) of V_3, V_4. With the slider of R_6 at the buffer amplifier grid end and under oscillating conditions, full negative d.c. grid drive is applied to V_2 cutting off V_2 anode current (or reducing it to a small value) and allowing the screen grids of V_3, V_4 to assume their rated values given and determined by R_7. (20kΩ – 30W). However, failure of V_1 to provide negative grid drive leads to V_2 consuming a large current depriving the screen grids of V_3, V_4 by a voltage given by the flow of V_2 current in R_7, i.e. V_2 has changed the screen voltage of the buffer amplifiers to a low value sufficient to prevent V_3, V_4 from passing excessive current.

The Power Amplifier Stage

The output of the buffer amplifier stage is π-section coupled to the grid of the final power amplifier. The remarks on π-section coupling units described earlier, apply here, but with one important difference. In the final stage neutralization is essential. This arises as follows: The input admittance of a triode is determined primarily by the capacitances existing between its electrodes. Since part of the current that flows to the anode depends upon the difference between the signal applied to the control grid and the amplified voltage developed in the anode circuit, the input admittance is affected by the magnitude and phase of the amplification. The exact relation[1] is

$$Y_I = \omega\, C_{ga}\,.\,A \sin\phi + j\omega[C_{gk} + C_{ga}(1 - A\cos\phi)]$$

where Y_I is the complex input admittance

C_{ga}, C_{gk} are the capacitances existing between grid, anode and cathode

ϕ is the phase displacement between the a.c. grid and anode voltage

A is the voltage amplification of valve.

Now when the output impedance of the final amplifier valve is inductive, the phase displacement angle ϕ is greater than 180° and less than 270°. As A, the voltage amplification of the valve is approximately 35 the input susceptance is very large and capacitive, i.e. for the 833A valve, C_{ga} = 6pF and C_{gk} = 12pF, the capacitive susceptance is of the order 200pF for $\phi = 7\pi/6$, say, and the input conductance $\omega C_{ga}\,.\,A\sin\phi$ is negative and equal to $-\ 1{\cdot}43 \times 10^{-3}\Omega^{-1}$. Hence, despite the fact that C_{21} places a 100pF capacitance between the grid and filament of V_5 an appreciable feedback voltage can exist across this capacitor. Therefore, with an inductive output impedance the input admittance is a negative conductance plus a positive susceptance. This means that some of the output power is fed back into the input circuit in such a phase relation that there is a component of input current π out of phase with the output voltage. Such a state leads to regeneration or oscillation and it is therefore essential to neutralize V_5 at the operating frequency. An effective method of neutralizing the C_{ag} of the 833A final amplifier is to make it part of a tuned circuit between anode and grid as exemplified by L_{11}, C_{19}. The coil L_{11} is a high Q component and with C_{19}, C_{ag} and distributed capacitance forms a high impedance rejector circuit effectively preventing the feedback of r.f. energy from V_5 anode to the grid. C_{19} is a pre-set neutralizing capacitor whose plates are of sufficient width apart to prevent r.f. flashover.

It is seen that despite the necessity for neutralizing, π-section coupling between the buffer and final amplifier has been retained owing to its valuable harmonic and parasitic suppression properties in consonance with remarks expressed earlier.

The filament circuit conforms to normal practice. R_{11}, R_{12} are low value resistors whose midpoint is earthed. The filament thus has a centre-tapped source with C_{22}, C_{23} acting as r.f. by-passing capacitors. A point worth mentioning is the construction of the r.f. choke L_{12}. As this component is connected directly across the high impedance of the anode circuit of the power amplifier, the r.f. voltage across the choke is, therefore, extremely high, and if it has sectional resonances in any waveband, severe loss of output will ensue, together with overheating or burning out of the choke. A low loss former is essential (in this case a hollow Pyrex tube), and winding should be single layer, the completed coil possessing an inductance about ten times greater than the inductance of L_{13}. A high grade low loss varnish is used to cement the turns in place.

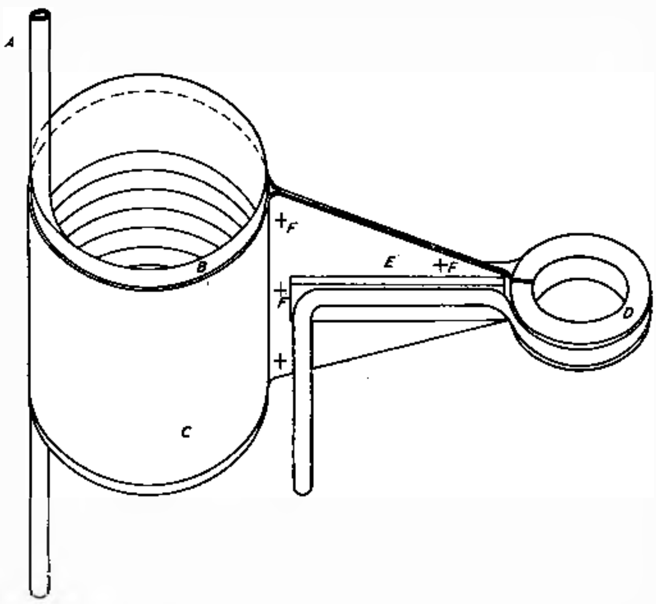

Fig. 3. Diagram of transformer coupled work coil

A—Primary of transformer. Six turns ⅜in copper tubing on 3in diameter circle. B—Two layers p.t.f. sheeting ·010in thick. C—Secondary of transformer single sheet copper with continuous coupling arms separated by one layer of ·010in p.t.f.e. sheet and bolted by ¼in nylon bolts F. D—Copper work coil 1¼in internal diameter, ½in wide with ⅛in copper tubing brazed on to it and to copper lugs. The latter being soft soldered to coupling arms E.

The Output Transformer and Work Coil

The primary of the output transformer, Fig. 3, consists of six turns ⅜in o.d. hollow copper tubing forming a self supporting coil of 3in internal diameter and rigidly attached to the variable tuning capacitor C_{26}. This is a 0 to 100pF component whose vane spacing is ½in, sufficient to prevent r.f. flashover at maximum voltage working. The rotor spindle arm of C_{26} is extended so as to protrude beyond the front instrument plate and may be grounded thus ensuring safety for the operator in manual operation. However, this implies that capacitor C_{24} be conservatively rated and in the present circuit is a T.C.C. transmitting type 0·001μF capacitor rated at 10kV working. The secondary of the transformer L_{13}-L_{14} consists of a single turn copper sheet wrapped around L_{13} as in Fig. 3 and insulated from it by two layers of 0·01in thick sheet polytetrafluoroethylene (p.t.f.e.). The form of this transformer construction possesses several advantages from the the point of view of efficient transfer of electrical power. It is shown in the appendix that for high energy transfer efficiency M/L should be large[3] where M is the mutual inductance existing between L_{13}, L_{14}. In addition $M = k\sqrt{(L_{13} . L_{14})}$ where k is the coupling factor. Hence not only should k be as near to unity as possible but also that $L_{13} > L_{14}$. For k to have a large value the spacing between L_{13}, L_{14} should be kept to a minimum consistent with voltage breakdown requirements. Such a secondary possesses a low inductance and resistance thus allowing large currents to flow without excessive voltage drop. To continue this state of affairs to the work coil D, the coupling arms are made integral with L_{14} and cut to the form shown in Fig. 3 to reduce the inductance between the work coil and the secondary of the output transformer. With regard to the value of a.c. resistance it may be noted that the current through the secondary coupling lugs and work coil will select its path in such a way that the magnetic energy of the circuit will be a minimum. In this way, the current encloses the smallest possible surface area. The

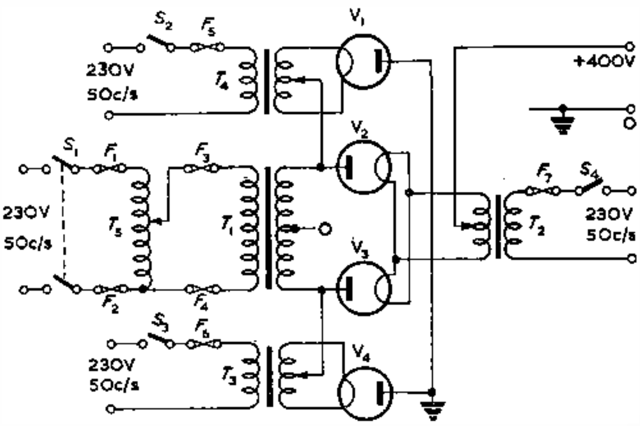

Fig. 4. Power supplies for oscillator bias and power amplifier

two coupling arms E are separated by a 0·010in thick sheet of p.t.f.e. with a similar air-spaced thickness in the work coil D. Water cooling is found to be adequate for all loadings, the temperature rise being of the order of a few degrees.

Setting Up Procedure and Performance

At point A, Fig. 2 a 0 to 1mA meter can be switched into circuit to measure the grid current of the buffer amplifiers, R_4, R_5 merely being shunting resistors to extend the range of the meter 0 to 10mA. The meter is additionally switched to measure anode, screen and grid voltages of V_1, V_3 and V_4 by appropriate series resistors (not shown on the diagram). After the crystal oscillator and buffer amplifiers have been adjusted for proper operating conditions, the r.f. heater is then tuned up by using only meters M_1 and M_2 of the final power amplifier.

The power supplies of the crystal oscillator and buffer amplifiers are conventional, and r.f. filters are placed in all mains leads. The application of h.t. voltage leads to an immediate deflexion in the meter registering V_3, V_4 grid current. The anode circuit of V_1 is tuned to twice the crystal frequency by C_6 and is indicated by the maximum grid current (8mA). With an appropriate value of R_6 the anode circuit of V_3, V_4 is tuned for twice crystal frequency by C_{17}, resonance being indicated by a maximum value of power amplifier grid current. With a sensitive absorption wavemeter no harmonic or parasitic radiations were found for frequencies below 150Mc/s, higher frequencies being beyond the range of the instrument. Neutralization on the power amplifier is simply effected by adjusting C_{19} in small steps and with no driving voltage applied to the grid of V_5, turning the rotor of C_{17} until no perceptible dip in grid current is observed. The whole circuit may then be retuned to ensure that subsequent adjustments have little effect. In practice this was not found necessary. Grid drive and anode voltage can now be applied to the power amplifier such that the applied power does not exceed 2kW. The high voltage derived from the power supply Fig 4 is smoothed and filtered by two choke input filters (not shown

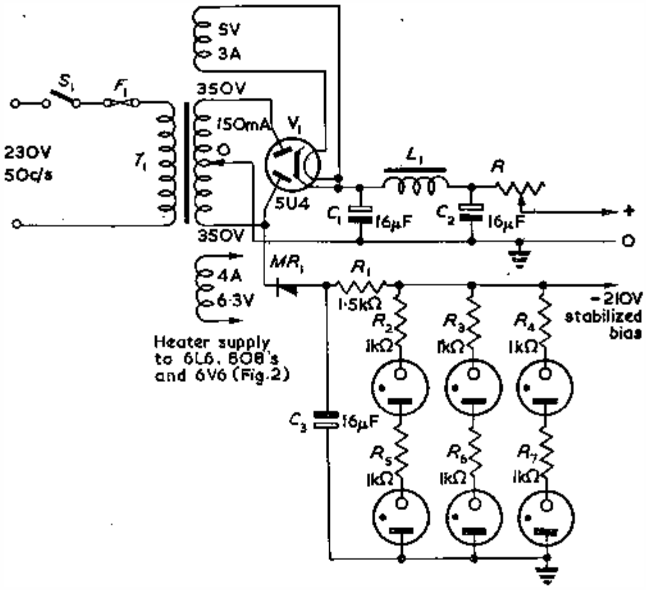

Fig. 5. Power supplies for oscillator and stabilized bias for power amplifier

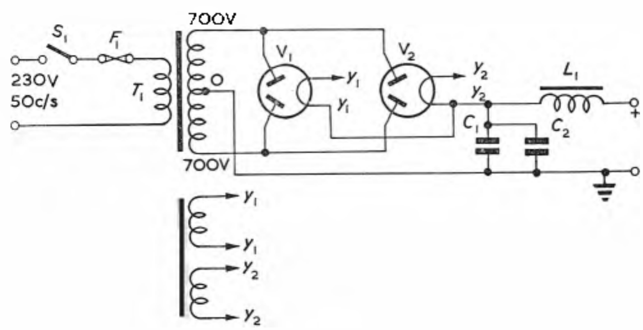

Fig. 6. Power supply for buffer amplifier

on diagram) and the voltage is variable from 0 to 4kV by the Variac T_5. Although gearing had been fixed on to an extension of the rotor of C_{26} for automatic tuning, the matter had not been proceeded with as it was found that the graphite susceptors used reflected only a small change in impedance into the tank circuit of V_5 and the retuning necessary was found to be small. A hollow cylindrical graphite crucible contained within a Syndanyo crucible for thermal insulation attained an uncorrected pyrometric temperature of 2 000°C eight minutes after insertion in the work coil. The true temperature was probably in the region of 2 400°C. At this temperature the graphite burns freely away in air and the Syndanyo melts and becomes exceptionally brittle. A further experiment was performed of suspending a graphite block in the work coil, the block being contained within a silica tube through which coal gas was circulating. A temperature of 1 450°C (uncorrected pyrometric) was reached and maintained for eight hours without requiring any adjustment of the tuning capacitor C_{26}. However, owing to the large heat losses it was difficult to exceed this temperature without exceeding the power ratings of the amplifier. It therefore appears that, owing to the low change of reflected impedance of a graphite susceptor over a wide temperature range, continuous tuning is not required, apart from the initial adjustments. However, for work of a ferromagnetic nature where the magnetic permeability changes rapidly with temperature automatic tuning is essential and would take the form of an automatic tuning control system[2]. From the final power amplifier tank circuit a signal can be derived which depends on the tuning condition of the output circuit, for example a signal the phase relation of which with respect to a reference signal depends on the tuning condition of the output load. This signal and the reference signal are fed to a measuring unit comprising a detector and a measuring element, the whole being able to ascertain the value of the controlled quantity. The output voltage of the measuring unit is applied to a controlling unit, governing a regulating unit, which may, for example, consist of a servo motor acting in such a way on the rotor of the tuning capacitor of the output circuit, that deviations of the controlled quantity are counteracted.

Acknowledgments

The author wishes to record his gratitude to Dr. R. W. Sillars and Mr. R. P. Chasmar of Metropolitan-Vickers Electrical Co. Ltd for their help and assistance in the preparation of this article, to Mr. C. M. Gateley of the Wigan and District Mining and Technical College for much advice and discussion and to London University for a research grant from their Central Research Fund.

Appendix

If in Fig. 3 L_{14} and L_{15} were dispensed with and the work of inductance L_W and resistance R_W were introduced directly into L_{13}, then one would simply have the case of a transformer with shorted secondary. Let i_1, i_2 represent the circulating currents in primary and secondary and replacing the value by a generator of e.m.f. e, one can write from ordinary coupled-circuit theory

$$i_1(R + j\omega L_{13}) + j\omega M i_2 = e$$

$$j\omega M i_2 + i\,(R_W + j\omega L_W) = 0$$

Hence the resistive component of the effective impedance of the work coil is R_1' where

$$R' = R + \frac{\omega^2 M^2 R_W}{R_W^2 + \omega^2 L_W^2}$$

Let $\omega L_W / R_W = Q_W$ then the reflected resistance becomes

$$\Delta R_1 = M^2/L_W^2 \,.\, R_W \,.\, \frac{Q_W^2}{1 + Q_W^2}$$

Let L_{14} and L_{15} now be included. The whole arrangement can now be regarded as two transformers in series and the main object of its use is to obtain a degree of impedance match between the work and loaded tank circuit. As L_{13} is greater than L_{15} and the coupling factor is high, the mutual inductance between them will be larger than L_{13}. In fact it will be larger than L_{14} and L_{15}. The circulating current in L_{14}, L_{15} is

$$i_{cc} = \frac{-j\omega M_{ps} i_1}{R_s + j\omega (L_{14} + L_{15})}$$

where i_{cc} = coupling circuit current

R_s = Resistance in secondary circuit

M_{ps} = mutual inductance of primary coil and secondary coil.

Hence $i_{cc} > i_1$ since $M_{ps} > (L_{14} + L_{15})$.

Now, the resistance reflected by the work in to the coupling coil circuit is given by ΔR_s where

$$\Delta R_s = M_w^2/L_w^2 \,.\, R_w \,.\, \frac{Q_w^2}{1 + Q_w^2}$$

where M_w is the mutual inductance between L_{15} and L_w of the work.

$$\therefore \Delta R_1 = \frac{M_{ps}^2}{(L_{14} + L_{15})^2} M_w^2/L_w^2 \,.\, R_w \,.\, \frac{Q_w^2}{1 + Q_w^2}$$

Since one has an enhanced reflected resistance and thus

$$\frac{M_{ps}}{L_{14} + L_{15}} \gg 1$$

loading as compared with a direct work circuit and hence a greater efficiency.

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Width Measurement in a Hot Strip Mill

The Evershed width gauge has been developed from a system initiated by the British Iron and Steel Research Association and is a servo driven optical arrangement for the monitoring of deviations in width in a continuous hot strip mill.

The system relies upon the natural radiation of infrared rays from the strip, and is sufficiently sensitive to respond to material that has cooled to a very dull red.

Two edge followers are used, each of which tracks the individual motions of the edges, the combined readings being summated electrically to give a signal proportional to the deviation in total width.

The edge followers are mounted upon a traverse which straddles the mill a few feet above the run of strip, and are motorized so that the two may be spaced apart to a distance equal to the nominal width of strip being produced.

The method of detecting and tracing the change in edge position reduces to the minimum variations that may arise in signal strength due to fluctuations in the rolling temperature. Radiation from the surface of the strip is projected through a lens and prism on to two photocells, placed side by side. One of the cells receives a constant amount of energy from a section of strip inboard of the edge, while the other is activated by the edge itself. Movement of the edge increases or decreases the radiation upon this cell dependent upon the direction in which the strip moves. A shutter is driven across the face of the other photocell reducing the energy upon it to a value equal to that obtained by its neighbour. This system of balancing radiation ensures freedom from temperature effects and compensates for changes in the cell characteristic due to ageing.

Each follower is self-contained inasmuch as it consists of an optical and photo-electric sensing device which presents a 'chopped' signal to a built-in 50c/s amplifier, the output of which drives the shutter mechanism by means of a small a.c. servo motor. Adjustable damping is provided by means of tacho-feedback derived from a small generator attached to the end of the motor. The unit operates on a 50V single-phase supply.

In a typical layout on a wide strip mill the followers are mounted on a special gantry above a footbridge. The traverse consists of two motor driven lead screws enclosed in an oil filled trunk upon which the two followers slide. The motor is at the left-hand end, and at the right is a shaft extension which is coupled by a system of bevel gearing to an indicator for the setting of nominal width. This indicator is purely a mechanical arrangement of two pointers and dials and is calibrated to read over the whole range of widths to the nearest 1/32in.

A small panel mounting roll chart recorder is used for the continuous recording of nominal width, and deviation up to plus or minus three inches. A remote indicator, also calibrated to plus and minus three inches, is mounted flush in the top of the finishing operator's control desk, which also contains the nest of push-buttons for control of the traverse motor.

Some Methods of Phase Measurement used in Transfer Function Analysis

By D. J. Collins*, B.Sc.(Eng.), A.M.I.E.E., and J. E. Smith*, B.Sc.

There are available, at the present time, a variety of instruments and techniques for the purpose of transfer function determination. The methods of phase measurement are diverse, and the article attempts to summarize the more conventional systems in use and to indicate their limitations.

In the development of closed loop systems, it is often necessary and always an advantage to know the transfer functions of the individual components of the loop. These transfer functions, although sometimes applying to purely mechanical devices, may be related to electrical parameters. This enables the components of a complex loop, incorporating both mechanical and electrical elements, to be referred to a common basis and facilitates the use of electrical test techniques.

For electrical measurement, the common analogy is an alternating voltage usually of sinusoidal form. The transfer function of a component is obtained by feeding it with a standard voltage and monitoring the output of the component. For convenience, the input voltage to the unit under test will be called the reference voltage V_R, and the output voltage will be called the signal voltage V_S. The transfer function is then defined as V_S/V_R.

The transfer function is usually mathematically complex and normally expressed in one of two forms:

(a) *Polar Form*

$$|V_S/V_R| \angle\phi$$

This implies that the component has an amplitude transmission of $|V_S/V_R|$ and a phase shift of ϕ.

(b) *Cartesian Form*

The information is contained in the two resolved components of $|V_S/V_R| \angle\phi$. The axes of resolution are the reference voltage axis and an orthogonal axis referred to as the quadrature axis. The reference or 'in-phase' component is then $|V_S/V_R| \cos\phi$ and the quadrature component is $|V_S/V_R| \sin\phi$.

Measurement of amplitude transmission of this type of voltage is relatively simple, and instrumentation is available of adequate accuracy for most purposes, over the relevant frequency spectrum. However, phase measurement seems to present some difficulty, and the difficulty is increased if the transfer process introduces distortion or noise content. There are a variety of methods of phase measurement, some of which are limited in frequency range and some of which have little use on distorted transfer voltages.

Some of the more common methods used in transfer function analysis will be discussed, and it will be readily apparent that, in some cases, the phase measurement process will also indicate the amplitude transmission.

(1) Methods Involving the Use of a Cathode-Ray Tube

In general, methods using cathode-ray tubes are capable of wide frequency range, but have poor accuracy and little harmonic rejection.

(a) Double Beam Oscilloscope

The use of a double beam oscilloscope will permit the direct measurement of phase shift by displaying reference voltage on one trace and signal voltage on the other.

By making simple length measurements, the relative phase angle can be estimated, as shown in Fig. 1.

* *Solartron Research and Development Ltd.*

It must be pointed out that the amplifiers associated with the two beams should have identical phase characteristics. In view of this requirement, a switched beam oscilloscope using a single amplifier will be more suitable than a split beam type with twin amplifiers. In a current type of switched beam oscilloscope, the phase error is less than 2° from d.c. to 1Mc/s. To obviate the need for direct-on measurement and to improve the accuracy, a calibrated phase-shifter could be incorporated in one channel and adjusted to superimpose the traces. The phase angle is then

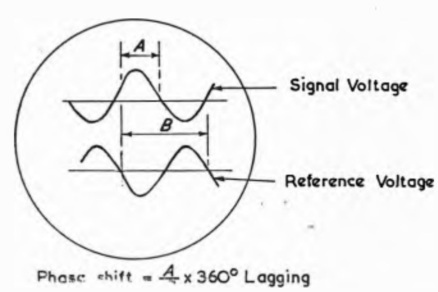

Fig. 1. Direct phase measurement on a double-beam oscilloscope

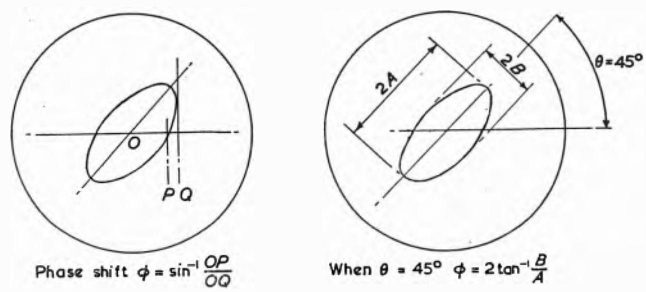

Fig. 2. Estimation of phase shift by Lissajous pattern

read directly from the calibration. A further improvement is to introduce a variable gain control in one channel, such that the superposition may approach perfection. Under these conditions, a certain amount of noise and harmonic content in the signal voltage can be tolerated.

(b) Single Beam Oscilloscope

The conventional method of applying reference voltage to the X plates and signal voltage to the Y plates, thus producing a 1 : 1 Lissajous pattern, can be extended in application by the use of X and Y amplifiers having similar phase characteristics.

In the Solartron CD 513 type of oscilloscope, the phase error between the X and Y amplifiers is less than 1° from d.c. to 100kc/s, when both amplifiers are used at a sensitivity of 10V/cm and 1V/cm.

The display is normally of elliptical form with the special cases of a circle (90° and 270° shift) or a straight line (0° and 180° shift), the phase shift for the ellipse is estimated as in Fig. 2.

As in method 1(a), this system can be improved by the use of a phase shifter adjusting for the straight line condition.

By use of gain calibrated amplifiers, the straight line can be adjusted to 45°, when the relative amplitude of signal to reference is unity.

(2) Wattmeter Methods

(a) Dynamometer Type

Consider a conventional coil type dynamometer wattmeter. If the voltage coil is fed with the reference voltage V_R and the current coil supplied with current in phase with the signal voltage, then the moving-coil will be subjected to a torque proportional to the product of the voltage and current in the appropriate coils.

The components are, in turn, proportional to the applied reference and signal voltages respectively, hence the ultimate indication of the wattmeter may be written as

$$\text{Indication} = K/T \int_0^T V_R V_S \, dt \quad \ldots\ldots\ldots\ldots (1)$$

where K is a constant and T is a time long compared with the periodic time.

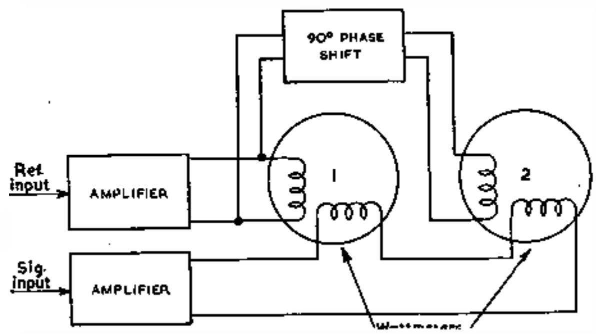

Fig. 3. Resolved components indicator using dynamometer wattmeters

If V_R is of the form sin ωt, the above integral will be zero for all forms of V_S other than $V_S = A \sin(\omega t + \phi)$. For $V_S = A \sin(\omega t + \phi)$, the indication is proportional to cos ϕ. The instrument can thus be used to evaluate ϕ. For a series of direct measurements of ϕ at varying frequencies, V_R and V_S must be maintained at constant amplitude. This can be a disadvantage. The disadvantage can be overcome in two ways:

(1) By using a second dynamometer whose voltage coil is fed with a voltage of equal amplitude but in quadrature with V_R and connecting its current coil in series with that of the first dynamometer. The indication of this second dynamometer for constant amplitude V_S input is proportional to sin ϕ. For a varying amplitude but constant phase signal voltage, it can be shown that the indication of wattmeter (1) is proportional to $A \cos\phi$, and that of wattmeter (2) to $A \sin\phi$ where A is the voltage amplitude. The two meters thus indicate the resolved components of the signal voltage vector. See Fig. 3.

(2) There are available specially constructed dynamometer type phase meters[1] giving a direct indication of ϕ.

(b) Thermocouple Type

Dynamometer type wattmeters are, however, somewhat restricted in frequency range, and the tendency when using this type of measurement has been to use bridge type thermocouple wattmeters. These have the advantage of wider frequency range and lower power consumption.

The thermocouple wattmeter arrangement comprises two fixed resistors, two vacuo-junction thermocouples, and a d.c. moving-coil meter. The fixed resistors and vacuo-junction thermocouples are housed in the meter case, and the overall assembly is provided with four input terminals. See Fig. 4.

The fixed resistors and thermocouple heaters are all of equal resistance, but the value of this resistance may vary between 90Ω and 120Ω.

The fixed resistors and thermocouple heaters are connected in Wheatstone bridge form, and the thermocouples are connected in series opposition to feed the moving-coil meter.

The response of the system will now be examined for the application of alternating currents.

It will be assumed that the thermocouple e.m.f. is proportional to the average heating effect, hence:

$$e = K_1 \int i^2 dt \quad \ldots\ldots\ldots\ldots\ldots\ldots (2)$$

where e = thermocouple e.m.f.

K_1 is a constant

i = instantaneous heater current.

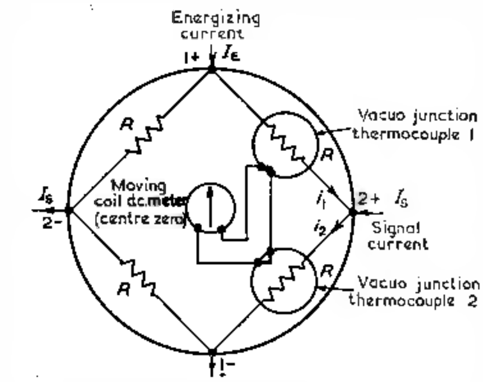

Fig. 4. Schematic of a thermocouple wattmeter

The meter indication which is proportional to the difference of the thermocouple e.m.f.'s may, therefore, be written:

$$\text{Indication} = K_2(e_2 - e_1)$$

where e_1 and e_2 are the e.m.f.'s of the corresponding thermocouples and K_2 is a constant.

Substituting from equation (2), this may now be expressed:

$$\text{Indication} = K_3\left(\int i_2^2 dt - \int i_1^2 dt\right) \quad \ldots\ldots\ldots (3)$$

where $K_3 = K_1 K_2$

From Fig. 4

$$i_2 = \tfrac{1}{2}(I_E + I_S)$$

$$i_2^2 = \tfrac{1}{4}(I_E^2 + I_S^2 + 2I_E I_S)$$

Similarly,

$$i_1^2 = \tfrac{1}{4}(I_E^2 + I_S^3 - 2I_E I_S)$$

Substituting in equation (3):

$$\text{Indication} = K_3/4\left(\int I_E^2 dt + \int I_S^2 dt + 2\int I_E I_S \, dt - \int I_E^2 dt - \int I_S^2 dt + 2\int I_E I_S dt\right)$$

$$= K_3 \int I_E I_S dt$$

This expression is the basic expression of response for a wattmeter. See equation (1).

The direct proportionality between thermocouple e.m.f. and heating effect, assumed in equation (2), is in practice good only to a first approximation. Further, the characteristics of the two vacuo-junctions employed are never absolutely identical, despite careful selection. The overall performance must, therefore, be taken as approximating to that of a conventional wattmeter. However, when a constant energization current is employed, the approximation is very close.

Instruments of this type have been used in phase sensitive voltmeters covering the frequency range 0·1c/s to 20kc/s.

The advantages of the thermocouple wattmeter as a phase measuring device are:

(1) Direct measurement of resolved components with good accuracy[2] (± 2 per cent of full scale).

(2) High degree of rejection of harmonics and unrelated frequencies. This is of great importance in the testing of servo mechanisms.

(3) Basic Electronic Methods

These include methods involving the direct use of known electronic techniques.

(a) Overlap Type Phasemeter

The operation of this type of phasemeter will be described with reference to Fig. 5. The reference and signal voltages are fully squared by limiting amplifiers, rectified and then applied to a gate circuit. This gate circuit allows constant current to flow through a d.c. meter, only when both inputs are simultaneously positive. The average current through the meter is thus proportional to the ratio overlap period/periodic time. Hence it indicates in linear terms of phase angle. By introducing known phase shifts into either

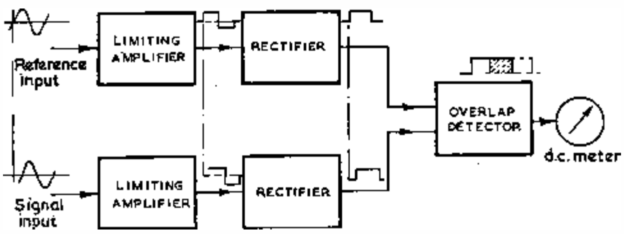

Fig. 5. The overlap type phasemeter

channel, the degrees of phase angle covered by the meter can be chosen at will.

The advantages of this phasemeter are:

(1) The indication of phase is independent of either input voltage over a wide range, due to the limiting action employed.

(2) The system in its ideal state is independent of frequency, and consequently phase changes can be observed by sweeping over a frequency band.

A grave disadvantage, however, is that the presence of harmonics in either channel will influence the instant at which the voltage is zero. The square wave action is operative on the zero voltage instant and, consequently, an error in indication results. The system is also known as the intercept type phasemeter.

(b) Phase Conscious Rectifier System

Phase conscious rectifiers can be designed using silicon, selenium, germanium or hard diode rectifiers.

The basic circuit of a simple phase conscious rectifier is indicated in Fig. 6.

The anodes of the diodes V_1 and V_2 are fed with a reference voltage from an isolating transformer T_2. As a result of this reference voltage, equal d.c. voltages are developed across resistors R_1 and R_2, due to the rectifying action of the diodes. The net d.c. output is thus zero. Consider the application of a signal voltage to the primary of T_1. Suppose that the voltage is in phase with the reference voltage, it will be apparent that on positive peaks of the signal voltage, one diode will be fed with a signal $E_1 + E_2$. where E_1 and E_2 are the peak amplitudes of the reference and signal voltages respectively. The other diode will, at the same instant, be subjected to a voltage $E_1 - E_2$. The diodes being peak detectors will then give a net output voltage of $(E_1 + E_2) - (E_1 - E_2)$, i.e. $2E_2$. It will be obvious that, if the signal voltage is reversed, then the output voltage will also be reversed. It can be shown that for quadrature signals there is a tendency for the net detected components of these signals to average to zero over a long period of time. Thus, the system is only sensitive to the in-phase components of the signal voltage, and gives an output proportional to that in-phase component. Obviously, harmonic distortion in the signal voltage will give an erroneous indication.

An extension of the diode phase conscious rectifier described above is illustrated in Fig. 7.

This circuit utilizes the rectifying properties of triodes. The reference input is applied to the anodes of the valves as a balanced voltage, such that the anode voltage applied to V_2 is 180° out of phase with the voltage applied to V_1. The triodes conduct by equal amounts on alternate half-

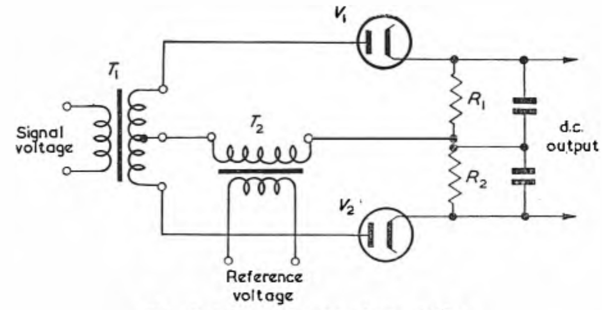

Fig. 6. Basic phase conscious rectifier

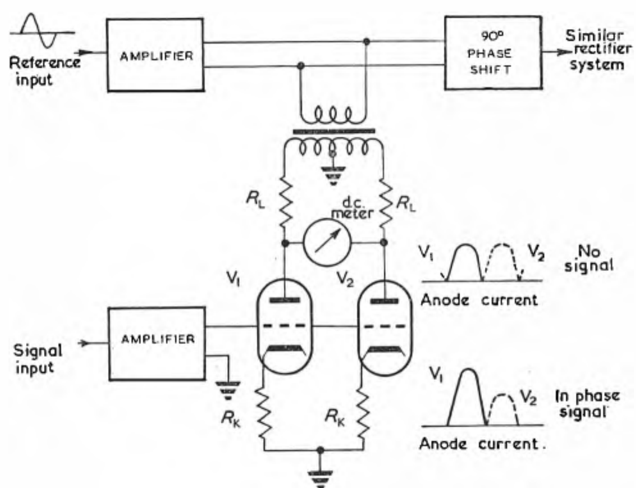

Fig. 7. Phase conscious rectifier type phasemeter

cycles. The d.c. meter connected between the anodes of the valves indicates the mean potential difference between the anodes which is obviously zero. Consider the application of an in-phase signal voltage to the grids of the triodes. When V_1 conducts, its control grid is being driven positive and when V_2 conducts, its control grid is being driven negative. The mean potential difference between the anodes of the valves will now assume some value dependent on the amplitude of the signal voltage. The reversal of the signal voltage will obviously give a reverse meter indication, and it can be shown that the meter indication for quadrature signal input voltages, and for harmonic components, will tend to zero. The simultaneous presence of quadrature and in-phase signal components affects the accuracy of indication of the in-phase component.

General disadvantages of phase conscious rectifiers are:

(1) Accuracy of indication is to a first order dependent on rectifier characteristics.

(2) The system is responsive to odd harmonics.

(3) Use of transformers limits frequency range.

(c) A Recent Development

This new device[3] is basically a resolved components indicator, displaying either component on a large scale meter, or indicating on a cathode-ray tube the vector condition.

The instrument utilizes a sampling technique whereby the instantaneous amplitude of the signal voltage is sampled at the instant when the reference voltage is maximum, thus giving a measure of the 'in phase' component. Similarly, when the reference voltage passes through zero, another sampling process provides the quadrature component. The two voltages are stored for several minutes and can be selected for display on the meter. If the voltages are used to deflect the beam of a cathode-ray tube in both axes, the spot position will indicate the limit of the vector. This facility enables a Nyquist plot to be made on the tube face.

The instrument covers a frequency range of 0·01c/s to 10c/s and has an accuracy of approximately 2 per cent.

(4) Computing Methods

These methods are characterized by the use of a trigonometrical identity in solving a vector computation.

(a) A Direct-Reading Phasemeter

This phasemeter[4] gives a direct indication of the phase angle and difference in level between two sinusoidal voltages. Its operation depends on the fact that the vector sum of two alternating voltages is a function of the phase angle between them and their individual amplitudes. By arranging that the two voltages are adjusted to the same predetermined level, the vector sum, which can be indicated on a meter, becomes a function of phase angle only.

The summation is carried out in an 'adder' stage which has three basic requirements:

(1) The voltages from the signal and reference inputs applied to the 'adder' stage must be equal. They are adjusted to the equality by means of the input attenuators.

(2) The phase angle between the voltages presented to the 'adder' stage must be between 90° and 270°. When the phase angle between the applied voltages is less than 90°, the phase of the reference signal may be reversed by means of a switch.

(3) The output meter must read full scale for 90° phase angle between the applied voltages.

With these requirements in mind, consider a typical condition (Fig. 8) on which OA is the reference channel voltage, OB is the signal channel voltage, both having been adjusted to the 'set level' amplitude. ϕ is the phase angle between OA and OB.

Thus OC, the voltage indicated on the meter, is the vector sum of OA and OB.

From Fig 8, it is apparent that

$$OD = DC = \tfrac{1}{2}\, OC \text{ and } \angle a = \angle b = \phi/2$$

Since ADO is right-angled

$$OD = OA \cos\phi/2 = \tfrac{1}{2}\, OC$$

$$OC - 2OA \cos(\phi/2)$$

The meter must give full scale when $\phi = 90°$

$$\text{i.e. when } OA = (1/\sqrt{2})\, OC$$

This provides the point for adjustment of 'set level'. For this example, the angle ϕ will be read on a scale ranging from 180° to 90°. The scale law conforms to the range 45° to 90° of the cosine law and is thus practically linear.

This instrument operates over the frequency range 2c/s to 100kc/s, and between 5c/s and 50kc/s has a phase measurement accuracy of ±1°. The instrument will not indicate correctly when the input voltages have distortion exceeding 5 per cent.

(b) The Solartron Process Response Analyser

The 'process response analyser' is intended to be used for the determination of transfer function elements used in process control. This implies that the frequency range must have a lower limit in the region of one cycle per hour.

It is possible to generate, electronically, sinusoidal reference voltages for testing at this frequency, but the measurement of phase represents a problem.

One solution is to record both reference and signal voltages simultaneously on a strip chart recorder, and to make an estimate of phase by direct comparison. This method is somewhat lengthy and not very accurate.

The method used in the indicator of the Solartron P.R.A. (due to Catherall[5]) gives direct indication of the 'in phase' and 'quadrature' components in the same manner as the thermocouple wattmeter. The multiplication process, however, is not carried out by thermocouples, but by analogue multipliers.

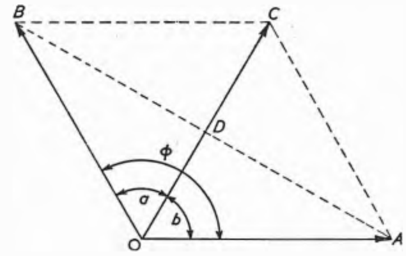

Fig. 8. Vector diagram for district reading phasemeter

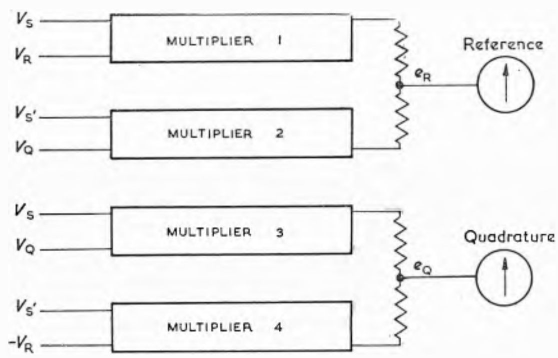

Fig. 9. Schematic of the indicator for the process response analyser

Consider Fig. 9 and assume we have four analogue multipliers fed with the voltages as shown, where:

$$V_R \equiv \text{Reference voltage} \equiv A \sin\omega t$$

$$-V_R \equiv -A \sin\omega t$$

$$V_Q \equiv \text{Quadrature reference voltage} \equiv A \sin(\omega t + 90) \equiv A \cos\omega t$$

$$V_S \equiv \text{Signal voltage} \equiv B \sin(\omega t + \phi)$$

$$V_S' \equiv \text{Quadrature signal voltage} - B \sin(\omega t + \phi + 90) = B \cos(\omega t + \phi)$$

The output of multiplier 1 is:

$$K \times A \sin\omega t \times B \sin(\omega t + \phi) = KAB(\sin^2\omega t \cos\phi + \sin\omega t \cos\omega t \sin\phi)$$

where K is a scale factor which is the same for all four multipliers.

The output of multiplier 2 is:

$$K \times A \cos\omega t \times B \cos(\omega t + \phi) = KAB(\cos^2\omega t \cos\phi - \sin\omega t \cos\omega t \sin\phi)$$

Summing the outputs of multipliers 1 and 2

$$e_R = (KA/2) \times B\cos\phi$$

e_R is thus a d.c. voltage whose amplitude is proportional to the 'in phase' component, $B\cos\phi$, of the signal since A is maintained at a constant level.

Similarly, the outputs of multipliers 3 and 4 can be summed to give

$$e_Q = (KA/2) \times B\sin\phi$$

$B\sin\phi$ is the quadrature component of the signal. e_R and

e_Q can thus be displayed simultaneously on two conventional d.c. voltmeters of the centre zero type.

It can be shown that signals with harmonic distortion give e_R and e_Q outputs having alternating components. These components integrate to zero over one cycle, but integration at low frequencies presents a problem. Despite this problem, the technique is extremely useful.

Conclusions

Methods of phase measurement at frequencies within the range from zero to 1Mc/s have been discussed. Although several of the methods have no inherent frequency limit, the problem of utilizing the method in a practical measurement or measuring instrument immediately imposes limitations.

The last method discussed seems to have the greatest possibilities, but the obvious limit is the upper frequency limit of the multipliers. However, there is a possibility that multipliers suitable for high frequency operation will result from investigations into the use of the Hall effect in certain semiconductors.

An important point which must be considered when choosing a measurement system is the question of distortion arising from the transfer process. Of the methods discussed, only the wattmeters have inherent harmonic and noise rejection. The other systems, of course, can be used, providing that they are preceded by some form of selective filter. Obviously, this filter will have to have a known transfer characteristic.

Without doubt, the instrument most suited and most widely used in transfer function analysis has to date been the thermocouple type phase-sensitive voltmeter.

Acknowledgments

The authors wish to thank the Directors of Solartron Research and Development Ltd for permission to publish this article.

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Impulse Voltage Wave Chopping Circuit for use with a Recurrent Surge Oscilloscope

By J. W. Armitage*, A.M.I.E.E.

A circuit is described which was designed to work in conjunction with an existing recurrent surge oscilloscope and was intended to provide better performance by the use of a hydrogen filled thyratron than had been possible with the mercury filled thyratron previously used. The chopping circuit is built up separately from the oscilloscope and is capable of chopping recurrent impulse waves up to 2·5kV at a selected instant over a wide range of times. The collapse of the impulse wave is very rapid and a biasing arangement has been introduced to eliminate the residual voltage drop in the thyratron from the chopped wave.

THE recurrent surge oscilloscope (henceforth abbreviated to r.s.o.) is commonly used to study the distribution of voltage in transformer windings subject to impulse voltage waves. The r.s.o. consists of a recurrent surge generator, the output of which is applied to the winding under investigation. It includes an oscilloscope together with a recurrent time sweep circuit, and the latter is synchronized with the surge generator. Thus the oscilloscope can display a standing picture of the voltage wave produced by the surge generator or of any selected voltage which results from the applied impulse. A comparatively low voltage may be used to simulate the conditions of full-scale impulse tests and the resulting voltages will bear a similar mutual relationship at any given time. Chopped wave applications to the winding are included in the high voltage test, so that the r.s.o. should be able to copy this condition. The chopping of the high voltage wave is effected by the flashover of a spark gap, which is connected externally between a transformer terminal and earth. The inductance and stray capacitance in the chopping circuit are such that the voltage at the terminal may collapse to zero, due to the chop, in a time as short as 0·1μsec, and this must be imitated in the low voltage test. In addition, there should be a wide range of control of the instant of time at which the applied wave from the r.s.o. is chopped.

An existing r.s.o. had been fitted with a chopping circuit which incorporated a type BT5 or 35 mercury filled thyratron as the chopping valve. However, this valve has a minimum ionization time of 0·35 to 0·6μsec, so that the rate of voltage collapse could not approach the value of 0·1μsec specified above. The earlier circuit also limited the range of voltages which could be chopped and the voltage did not collapse fully to zero. The present circuit was designed to overcome these defects. It will operate from any r.s.o. having a positive-going time-sweep voltage, and is capable of chopping a 2·5kV, 1/50μsec wave at any selected time after the wave has attained approximately 150V on the rising front. The wave is chopped to zero in about 0·1μsec. An additional electronic stage has been included to synchronize the chopping circuit with alternative synchronizing pulses. The existing r.s.o. generates 50 surges per second but the time sweep operates at 100c/s and alternative sweeps trace out a reference zero line or a time calibration.

Description of Equipment

The chopping equipment is separate from the r.s.o. and is built in an enclosed chassis 18in × 10in × 2in deep. A knob labelled 'chop position' controls the timing of the chop with respect to the impulse wave. A control labelled 'chop level' varies the bias voltage applied to the cathode of the chopping thyratron. This bias voltage is used to compensate for the arc drop in the thyratron so that an impulse wave can be chopped completely down to zero voltage. Two large terminals are provided on the front panel for connexion to the external points where chopping is required.

Two input sockets A and B (see Fig. 1) provide the alternative means of synchronizing the r.s.o. output wave and the operation of the chopping circuit. Socket A should be connected to the r.s.o. time-sweep circuit so that

* *Metropolitan-Vickers Electrical Co. Ltd.*

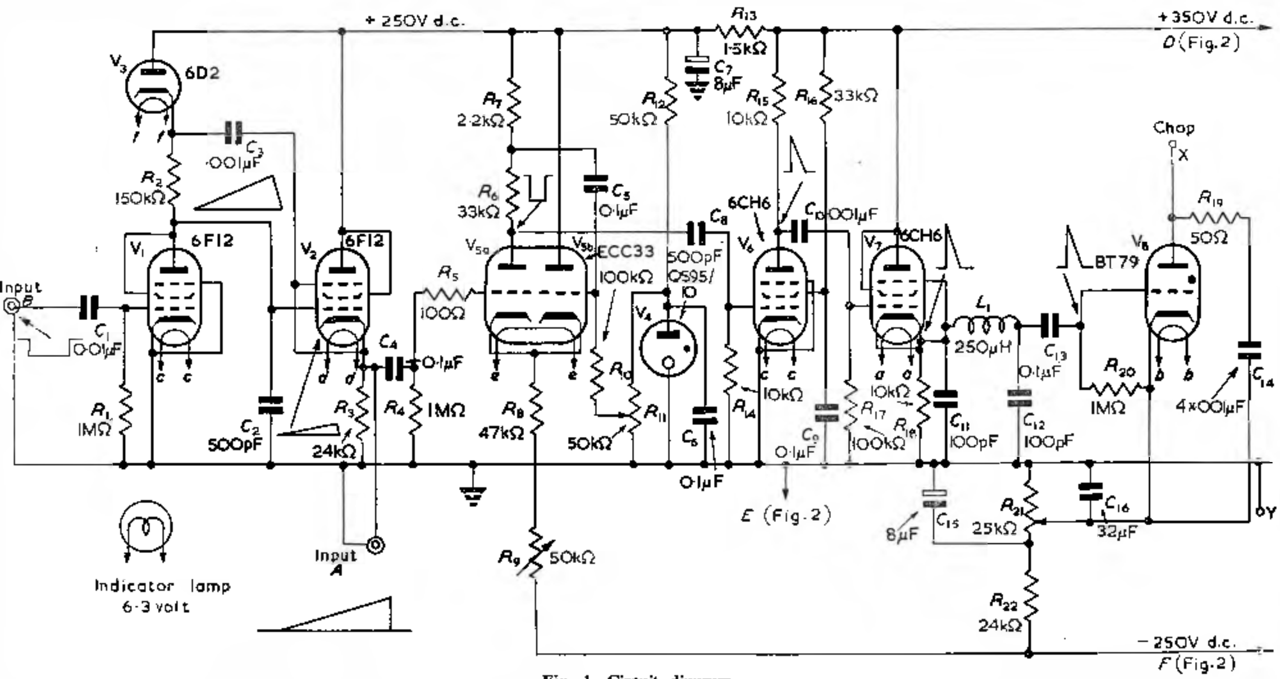

Fig. 1. Circuit diagram

a positive-going sawtooth voltage wave of about 100V is injected into the chopping circuit. This voltage waveform is available from the time-base thyratron of the existing r.s.o., and a suitable potential divider is incorporated in the r.s.o. to give the required voltage amplitude. (The pulse wave-shapes are shown at their related points in Fig. 1 for a recurrent frequency of 100c/s.) The slope of the sawtooth voltage rise is proportional to the time-sweep duration so that the time to chop, which is also proportional to the voltage slope, will change when the time-sweep is changed.

The alternative socket (*B*) requires a negative square pulse of about 20V amplitude which is obtained from the anode of the time-base triggering valve. In this case the time to chop is independent of the duration of the time-sweep.

Circuit Operation

The circuit diagram is shown in Fig. 1 and the power supply circuit in Fig. 2. A type BT79 hydrogen thyratron is used as the chopping valve. It has the required short ionization time and a peak current rating of 35A.

When the input *A* (Fig. 1) is used the sawtooth waveform rising to +100V is applied to the grid of valve V_{5a} via a coupling capacitor C_4 and grid resistor R_5. The cathode potential is set by the potentiometer R_{11} (labelled 'chop position') and the cathode-follower arrangement of V_{5b}. Thus V_{5a} will begin to conduct when the incoming grid signal approaches the cathode potential which is set by R_{11}. As soon as V_{5a} becomes conducting the anode potential falls, but some part of this negative change is applied to the grid of V_{5b} via the coupling capacitor C_5. This allows the cathode potential to fall and causes an increased anode current in V_{5a}. The effect is cumulative and gives a steep fronted negative pulse at the anode of V_{5a}.

The output pulse from the anode of V_{5a} is modified by the network C_8R_{14} and applied to the grid of the polarity reversing amplifier V_6. This valve, which is normally fully conducting, is then cut off by the negative pulse. Thus a positive pulse is generated at the anode of V_6 and fed through the capacitor C_{10} to the cathode-follower V_7. The low output impedance of the latter is necessary in order to fire the thyratron V_8. The filter $C_{11}L_1C_{12}$ prevents excessive feedback of the voltage pulse which often appears on the thyratron grid at the instant of firing.

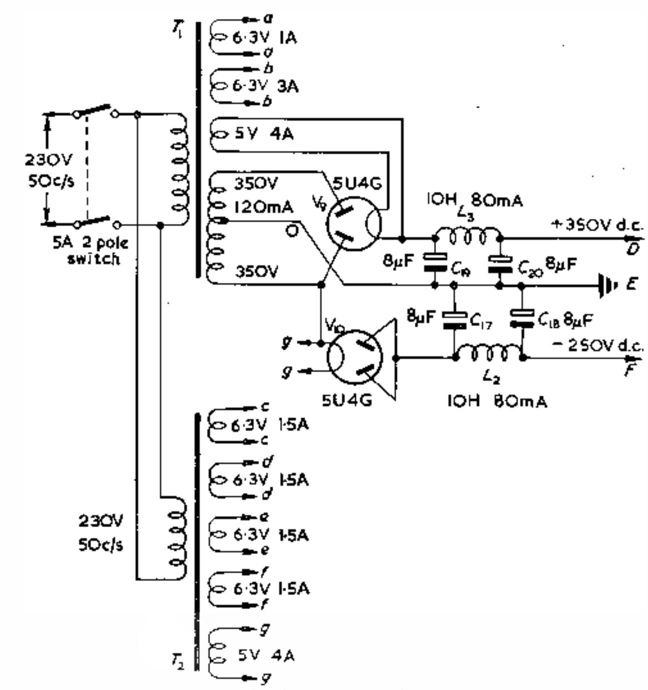

Fig. 2. Power supplies

Capacitor C_{14} and R_{19} are connected in series between the anode and the cathode of V_8. They help to increase the rate of rise of current in the thyratron and produce a voltage collapse in a time of less than 0·1μsec. The cathode of V_8 is biased to a negative potential the value of which is set by the potentiometer R_{21} (labelled 'chop level'). A large by-pass capacitor C_{16} is connected between the cathode and earth. The external terminals (*X* and *Y*) are connected by short leads to the anode of V_8 and earth respectively. The arrangement enables the voltage between *X* and earth (*Y*) to be chopped to zero without the arc drop of V_8 appearing as a residual voltage. Capacitor C_{16} is large so that the flow of the chopping current produces a negligible change in the voltage across it.

When the chopping time is required to be independent of the time-sweep duration a negative square pulse is fed to socket B and thence to the grid of V_1 via the coupling capacitor C_1. The triode-connected pentode (V_1) is normally fully conducting, but is cut off by the negative pulse so that capacitor C_3 begins to charge up. This voltage change is applied to the grid of the cathode-follower V_2, the output of which is a voltage change of similar waveshape across R_3. Normally this change would be a positive-going exponential wave, but the addition of the feedback capacitor C_3 linearizes the change to give a linear sawtooth voltage wave across R_3. The duration of the rising voltage portion is governed by the duration of the negative input pulse at B. The diode V_3 is used as a convenient non-linear impedance.

The sawtooth wave is applied to V_5 which then operates in the manner described above.

Test Procedure and Results

The r.s.o. was modified by the addition of two external terminals (A', B') for the alternative connexions to the synchronizing terminals (A, B) of the chopping circuit. Input and output terminals A and A', when inter-connected, synchronize the chop at a point which is a given proportion of the time sweep from the start of the oscillograph trace, so that this connexion is suitable when only one sweep duration is required for a test series.

When the synchronizing terminals B and B' are inter-connected the time to chop is a fixed interval after the start of the time-sweep. Thus the time to chop with respect to the impulse wave is also fixed, irrespective of the sweep duration, provided that the starting time of the wave is not changed with respect to the start of the time-sweep.

Normally it is the output impulse waves of the r.s.o. which would be chopped, in which case the main terminals X and Y of the chopping circuit should be connected to the terminals of the r.s.o. (Y is the earth terminal). When both chopped and unchopped impuses are required during a test series the chopping circuit should be left connected but it should be switched off or the 'position' control should be turned to a long delay when chopping is not required. In this manner the loading effect on the r.s.o. impulse generator remains constant and the same wave front is retained for each test condition.

The chopping circuit could be used to chop to earth the voltage at a point other than the r.s.o. output terminal (e.g. the neutral point of a transformer with an isolated neutral).

Examples of chopped waves are shown in Fig. 3. Oscillograms (a) and (b) show unchopped waves applied to two transformers, and (c) and (d) show the same waves when chopped. In each case the collapse of voltage was largely completed within about 0·1μsec but there was a small oscillatory overshoot. The overshoot and oscillations were characteristics of the chopping and associated circuits, and it was probable that the time of voltage collapse was also more dependent on circuit characteristics than on the ionization time for the chopping thyratron. Oscillogram (f) shows a chop on a shorter time-base and here an oscillation was superimposed on the trace of the chopped wave. This was probably caused by a limited response to the collapse of the voltage across the thyratron, so that both the impulse wave terminals and the oscilloscope transient deflector plates were subjected to a voltage change which was slowed down by the series impedance and stray capacitance of the connecting leads. Furthermore the stray capacitance circuits were largely undamped. It is inferred from these comments that the actual and apparent rates of voltage collapse were limited by circuit characteristics.

Oscillograms (c) and (d) show that the voltage settled down to zero very soon after the chop, so that the thyratron arc drop was effectually compensated by the 'level' control.

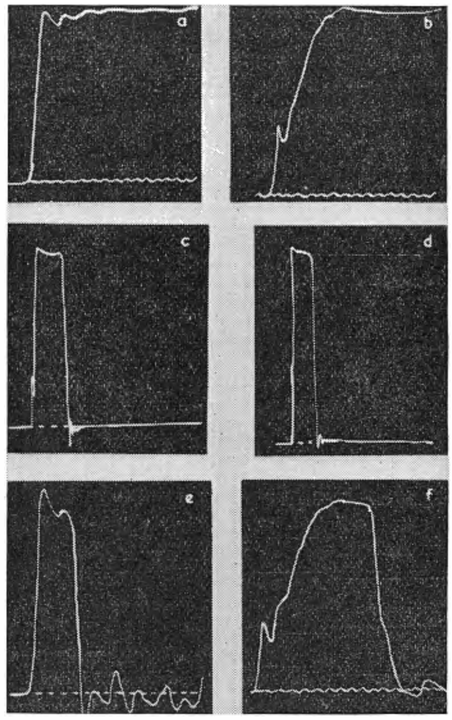

Fig. 3. Oscillograms

(a) Steep wavefront applied to a 110kV, 6 000kVA auto-transformer
(b) 0·5μsec wavefront applied to a 33kV, 2 000kVA transformer
(c) Wave (a) chopped after 4·5μsec
(d) Wave (b) chopped after 3μsec
(e) Voltage 5 per cent inside winding for wave (c)
(f) Wave (b) chopped after 0·9μsec
Period of timing oscillation or mark a, b and f = 0·1μsec c, d and e = 1·5μsec

Oscillogram (e) is representative of the voltage wave at a point 5 per cent within a winding which was subjected at its line terminal to the chopped wave shown in oscillogram (c).

Conclusion

The chopping equipment has proved satisfactory during about two years of use. Synchronization of the chop and the applied wave is effective with a negligible jitter, and impulse waves can be chopped over a wide range of time and of wave voltage. The hydrogen thyratron gives a rapid collapse of voltage at the instant of chop, and this time interval is about 0·1μsec. At this short interval the response of the test and measuring circuits restricts the rate of voltage collapse at points remote from the chopping thyratron.

The equipment was made up separately from the r.s.o. with which it was used, but there would be some advantage in combining both components in the same case.

Acknowledgments

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A Transistor Operated Self-Balancing Radiation Pyrometer

By D. W. Birnstingl*, B.Sc., A.C.G.I., A.M.I.E.E.

The article describes a radiation pyrometer capable of measuring surface temperatures down to 150°C from a distance, using transistor circuits and a lead-sulphide cell. A feedback technique is used to render the instrument insensitive to changes in the cell characteristics and the response time is less than a second.

The advent of lead-sulphide photo-conductive cells has enabled the range of high-speed radiation pyrometers to be extended to measure surface temperatures below 200° C from a distance. These cells respond to wavelengths between 0·3 and 3·0μ (i.e. from visible light into the infra-red) and the peak response occurs at about 2·5μ, but as the sensitivity varies widely with ambient temperature it is impractical to use the cell directly as a measuring device.

This difficulty is overcome in the pyrometer described in this article by using the cell as a null detector. The lead-sulphide cell is placed behind a rotating chopper disk (Fig. 1) in such a manner that it is irradiated alternately by the unknown source and by a tungsten filament comparison lamp. This produces an alternating voltage at about 1kc/s across the cell, which is dependent on the difference between the intensities of the two sources; and the control circuit is designed to adjust the comparison lamp current to keep the cell voltage continuously at a minimum. Thus the inherent drift in the cell sensitivity is unimportant and the radiation intensity is measured in terms of the comparison lamp current.

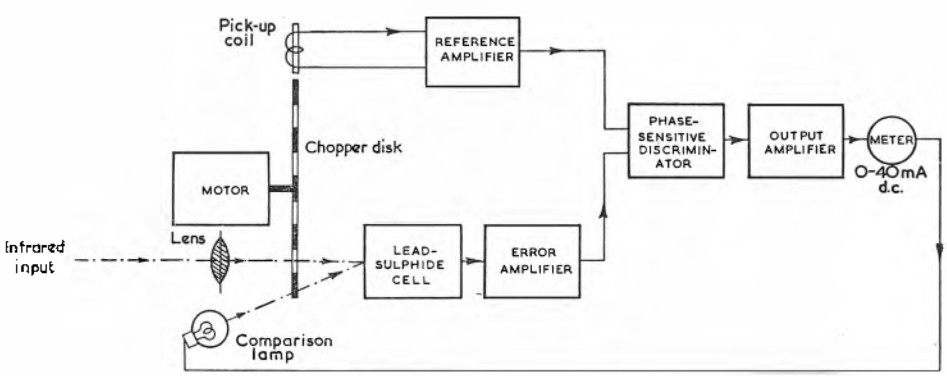


Fig. 1. General arrangement of the circuit

The pyrometer comprises a control box and a viewing head, and it is shown in Fig. 2.

Control Circuit

There are two inputs to the control circuit. The first is the error voltage from the lead-sulphide cell, which is proportional to the difference in the amounts of infra-red radiation coming from the external source and the comparison lamp, and is at a frequency of approximately 1kc/s.

The second is a phase reference voltage for determining which amount of the radiation is the larger, and it is obtained from a pick-up coil whose magnetic circuit is completed through the serrations of the chopper disk. This coil, T_7 (Fig. 3) is connected in series with the base lead of the transistor X_1, and the base bias current flowing through it provides the necessary excitation. The amplified voltage at the collector is applied through transformer T_5 to the base of a second transistor X_2, which is thereby driven alternately positive and negative, so that the transistor is alternately fully open or closed. Thus, a square wave of voltage is applied across the primary of the transformer T_6, connected between the negative line and the collector.

The transformer secondary has one terminal earthed and is shunted by a diode MR_2 so that the output swings from negative to zero in phase with the chopper disk serrations. This voltage is applied to the collector of the discriminator transistor X_5, which can conduct, therefore, only in alternative half cycles*.

At the instant when the driving transistor X_2 opens, considerable magnetic energy is stored in the transformer; excessive over-voltages are prevented by diodes MR_1 and MR_2 across the primary and secondary, which serve to dissipate this energy in the diode and winding resistances.

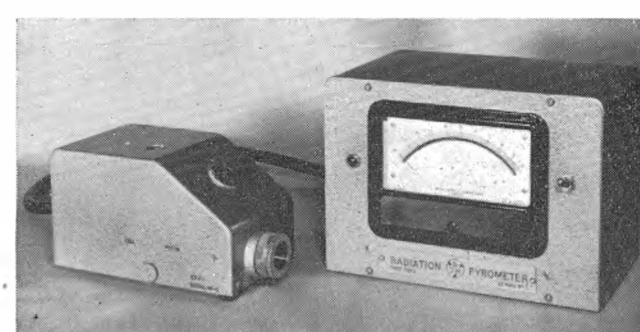
The signal from the lead-sulphide cell is fed through a 10 : 1 step-down matching transformer into a high-gain two-stage transformer-coupled transistor amplifier (X_3 and X_4). The output from this amplifier is applied to the base of the discriminator transistor X_5, which is thereby driven alternately positive and negative.

If the base is driven negative at the same time as the collector is negative, the transistor conducts, but if the signal is reversed the transistor is held open; thus, by adjusting the position of the reference coil, the discriminator can be made to conduct only when the external source of radiation is stronger than the comparison lamp.

When the discriminator transistor conducts the capacitor C_8 receives a negative charge, and the base of the output transistor X_4, which follows, is held negative.

This causes current to flow through the collector and

Fig. 2. The pyrometer control box and viewing head



* *The British Thomson-Houston Co. Ltd.*

* *Patent applied for.*

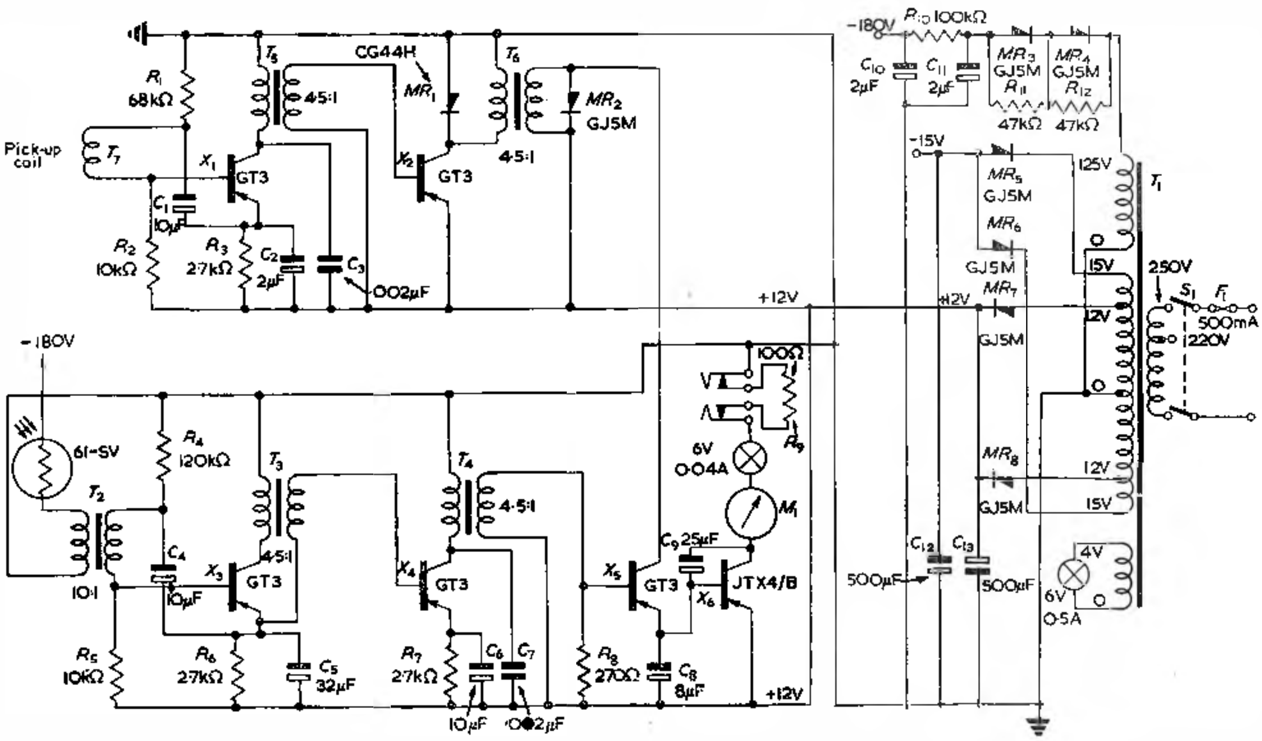


Fig. 3. The control circuit

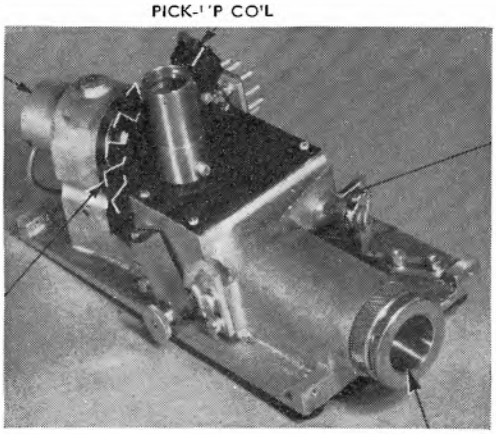


Fig. 4. Viewing head
(cover removed)

through the comparison lamp and milliammeter in series with it, and a continuous balance is reached in which radiation from the comparison lamp is just less than that from the external source. If it is greater, the discriminator cuts off and allows the lamp to cool down.

Viewing Head

The components of the viewing head are mounted in a jig-bored aluminium casting (Fig. 4), which holds the objective lens, the comparison-lamp focusing lens, the chopper-disk motor and the lead-sulphide cell in the correct geometrical relationship. The magnetic phase reference coil is mounted on a radius arm coaxial with the motor, and by rotating the arm the correct phase relationship of the output voltage with respect to the chopper-disk position can be obtained.

The objective lens is fixed at the inside end of a long tube, which acts as a lens-hood and can be pulled in or out to focus the image of the hot body on to the cell. For normal operations the tube is kept pushed right in, but when it is required to view a small object (down to 1·0cm diameter) the tube must be withdrawn.

As the instrument is responsive to the quantity of infra-red radiation falling on the cell, it is necessary when measuring higher temperatures to attenuate the light beam. For this purpose, two removable stops are provided which fit inside the lens tube. The first of these reduces the aperture area by ten, and the second by one hundred. (Further attenuation can be achieved by interposing heat absorbing glass between the hot body and the pyrometer.)

The temperature range covered on each stop depends on the characteristics of the comparison lamp: with a typical instrument the range on full aperture is 150 to 400°C; on medium aperture, 300 to 600°C; and on small aperture, 400 to 1 200°C.

A view-finding telescope is provided in the viewing head, in which a movable mirror behind the objective lens reflects the incoming rays on to a magnifying lens, and a graticule indicates the approximate area irradiating the photocell. Normally the mirror rests out of the way below the beam, and it is held either up or down by a permanent magnet.

Operation

The pyrometer does not measure temperature directly; it measures the amount of infra-red radiation falling on the cell, and the response depends on the emissivity of the surface, the size of aperture fitted, the extension of the lens tube and the distance to the pyrometer. Thus the calibration of each pyrometer depends on the conditions under which it is being used, and on the individual characteristics of the comparison lamp. The calibration curves for a particular instrument viewing a rough mild steel surface one foot away, through an aperture in the side of an oven are given on Fig. 5.

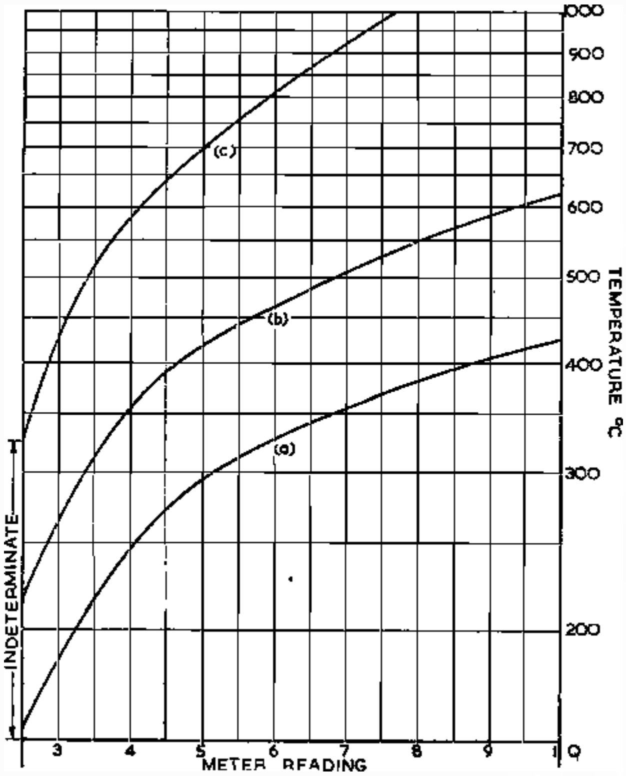


Fig. 5. Typical pyrometer calibration curve viewing hot plate at 12in
(a) full aperture; (b) medium aperture; (c) small aperture

When the temperature of the surface is low the signal-to-noise ratio of the cell is poor so the meter reading is erratic; it does not become steady until the radiation is sufficient to give a meter reading of about 25 per cent full scale deflexion. Thus, although the instrument may start to indicate at 80°C on full aperture, the reading is not steady until 150°C is reached.

When the lens tube is pushed right in, the pyrometer is focused at infinity and the meter reading is very nearly independent of distance, provided the surface being observed completely fills the field of view. Thus, with the instrument calibrated to measure at one foot, the temperature of a 400°C hot plate appears to 408°C at 3in and 392°C at 4ft—the error is only ± 2 per cent. Under these conditions the diameter D of the field of view at a distance A feet from the pyrometer is as follows:

With Full aperture: $D = 1 + 0{\cdot}8A$ inches

Medium aperture: $D = 0{\cdot}7 + 0{\cdot}7A$ inches

Small aperture: $D = 0{\cdot}2 + 0{\cdot}6A$ inches

Smaller surfaces than this can be viewed by pulling out the lens tube, but if this is done the calibration will vary widely according to the distance from the pyrometer.

A jack plug is provided on the control box to drive an external meter or controller. It is in series with the comparison lamp, and the full scale current is 40mA d.c. The load resistance should not exceed 220Ω.

Applications

The response time of the instrument is less than 1sec, and under identical conditions of surface emissivity and distance, the calibration is consistent to within 2 per cent. These factors make the pyrometer suitable for a wide variety of applications both in the factory controlling heat processes, and in the laboratory.

Acknowledgment

The author is indebted to his colleagues in the B.T.H Research Laboratory for encouragement and assistance in this work.

A Floating-Point Computer for Germany

Early in 1955 an enquiry about electronic computing equipment was received by Elliott Brothers (London) Limited from the Ernst Leitz Optical Works at Wetzlar in Germany. In the first place interest was expressed in computing equipment for both scientific and commercial applications, but due to the delay inevitably resulting from the analysis of the complex business procedures of Leitz, it was later decided to concentrate attention on a computer designed for scientific purposes only.

Elliott Brothers at that time had in production their 402 computers, and early in 1956 a comprehensive demonstration of one of these machines was given to two members of the Leitz staff, Dr. Franke and Dr. Marx.

Even at this stage Dr. Marx was becoming extremely interested in the use of floating-point methods of calculation, and towards the middle of 1956 a definite request was made that a floating-point extension should be developed for the 402, which would enable the computer to carry out its normal functions at normal speed in a floating-point mode, as well as the normal fixed-point mode.

Considerable discussion and design work followed, incorporating several suggestions made by the customer, and in August 1956 a contract for the supply of a new type of computer, designated the Elliott 402F, was agreed.

Design details were completed by February 1957, and the machine assembly completed in June 1957. Commissioning of the fixed-point section of the computer, essentially the same as the standard 402 machine, gave little trouble and was completed in three weeks. The floating-point extension was then commissioned and the computer, after passing a rigorous acceptance test laid down by the customer, has been used jointly by Elliotts and Leitz at Borehamwood since October last. The machine has recently been shipped to Germany where it is being used for the design of optical systems.

In the 402F floating-point mode of working, the normal 402 computer word of 32 binary digits is divided into two parts: (a) a mantissa of 26 binary digits representing a fractional quantity x lying in either the range $+\frac{1}{2} \leqslant x < +1$ or the range $-\frac{1}{2} > x \geqslant -1$, and (b) an exponent of 6 binary digits representing an integral quantity lying in the range $0 \leqslant y \leqslant 63$. The 402F then interprets the word for arithmetic purposes as the number $x \cdot 2^{y-31}$, unless $y=0$, in which case the word is interpreted as the number zero, or unless $y=63$, in which case the word is interpreted as an infinite quantity. The 402F stops if ordered to carry out an operation resulting in an indeterminate quantity, but otherwise performs all the normal arithmetic operations, correctly rounding-off the results.

The main advantages of the floating-point mode of working are first, that the numerical analysis and subsequent programming of a mathematical problem is very much simpler, and second, that the way is clear for the introduction of automatic programming. This is a method of using a floating-point computer in which details of a problem are prepared for the machine in a standard mathematical form by non-programming staff, followed by the subsequent translation of the mathematical symbols punched on paper tape into machine instructions.

1kc/s Transistor High-Gain Tuned Amplifier

(with provision for transistor noise figure measurements)

By R. A. Hall*, B.Sc.

By the use of junction transistors instead of thermionic valves an amplifier has been constructed which has the advantages of reduced power consumption, bulk and microphony. Particular attention has been paid to stability of gain and bandwidth against transistor variations; over a period of 12 months the maximum gain of about 125dB has changed by only 1·5dB with no measurable change in bandwidth.

The amplifier has been designed principally for use as a bridge detector and general-purpose amplifier, but provision has also been made for transistor noise figure measurements by the addition of a stage at the input in which the transistor under test provides sufficient gain to make the effective noise figure of the amplifier negligible for most purposes.

HIGH-GAIN thermionic valve amplifiers when used at low signal levels generally tend to be microphonic and extremely inefficient in terms of both power consumption and bulk. For the particular case of low frequencies it can be concluded that transistors offer considerable advantages over valves in this application.

The tuned transistor amplifier described below has been designed principally for use as a general-purpose amplifier and also as a detector for the Transistor T-Parameter Measurement Set[1]. But provision has also been made for transistor noise figure measurements by the addition of a stage at the amplifier input in which the transistor under test provides sufficient gain to make the effective noise figure of the amplifier negligible for most purposes.

Particular attention has been paid to stability of gain against transistor variations by the use of transformer-coupled common-base stages. In this type of circuit, provided for each transistor one has that generator impedance is much greater than transistor output impedance and load impedance is much less than transistor input impedance, then the circuit performance is dependent mainly on the coupling transformers since the effect of feedback within each transistor is made negligible and the working current gain of each transistor is approximately unity. However there is also the requirement of maximum gain per stage for which the conditions for each transistor are

Generator impedance = transistor input impedance

and

load impedance = transistor output impedance

so that in practice a compromise has to be found. Tuning of the amplifier is carried out by means of capacitance in shunt with the primary of each transformer but it will be seen later that this does not affect the argument so far but introduces the additional factor of bandwidth, which is required to be as narrow as possible and also stable against transistor variations.

Circuit Description

The amplifier uses STC junction transistors in a cascade arrangement of six common-base stages, the load in each case being a tuned circuit inductively coupled to the next stage, followed by an untuned common-base output stage. The input impedance is 600Ω, the primary of the input transformer being isolated, and the maximum power gain is about 125dB into a load of either 600Ω or 75Ω. The gain may be reduced to about 85dB by switching out two tuned stages and also by steps of 1dB using the 0 to 80dB variable attenuator which is incorporated. The amplifier is tuned to 1kc/s with a bandwidth of 80c/s between −3dB gain points when using six tuned stages and 100c/s when using four tuned stages.

The transistors used in the amplifier are STC type LS828 which is similar to the present Brimar TJ series but without selection according to current gain.

The transistor collector bias conditions are I_c = 1mA and V_c = 10V except for the final stage in which V_c = 30V. A meter is incorporated which, by means of switch S_2 (Fig. 1) may be used to check the 10V and 30V supplies and also the total collector current, including that of the transistor in the test stage, and resistor R_8 is a shunt to convert the full scale reading of the meter to 10mA in the 'Total I_c' position of the switch.

Noise measurements are made at a mean output level of 1mW and the final stage is biased so that negligible overloading occurs on noise peaks at this mean level. The resultant peak overload output level for a continuous sine wave signal is found to be 5mW.

The common supplies and meter circuit provide feedback paths so that very large capacitors C_6 and C_7 are necessary for decoupling because of the large gain involved. In order to avoid excessive charging currents, which may damage the meter and also shorten the battery life, a 'stand-by' position is included in the 'on-off' switch S_3 in which C_6 and C_7 are charged up via limiting resistors R_{10} and R_9 respectively.

A detailed description of the amplifier stages can be given with reference to the transformer data presented in Table 1 which includes the load resistance (assuming a transistor input resistance of about 80Ω) together with the required primary inductance and transformed load resistance for each transformer used to calculate the number of turns. The current gain for each transformer, calculated from the turns ratio, is also given.

Preliminary experiments showed that much higher values of Q factor can be obtained for an inductance of 1H using a Ferroxcube core than using 'Permalloy C' laminations, therefore in the tuned transformers Ferroxcube cores are used. But in a tuned stage at an output level of about 1mW the losses involved in a Ferroxcube core are found to cause considerable reduction of Q factor and hence transformer current gain which therefore becomes a function of power level. In the final stage an untuned transformer with a 'Permalloy C' laminated core is therefore used.

The primary winding of each tuned transformer with its secondary unloaded has, when on tune, a certain dynamic resistance R_d; considering one particular transformer such

* *Standard Telecommunication Laboratories Ltd.*

TABLE I

TRANSFORMER NO.	CORE	LOAD RESISTANCE (Ω)	TRANSFORMED LOAD RESISTANCE OHMS	PRIMARY TURNS	SECONDARY TURNS	PRIMARY INDUCTANCE (H)	CURRENT GAIN (dB)
T_1T_{10}	Fortiphone Transformer Type EX45 ('Permalloy C')	600	30kΩ	7·1/1 Ratio		15	17·0
T_2	'Permalloy C' ·015 in. 500 T Bobbin	900	600Ω	506 (38s.w.g.)	620 (38s.w.g.)	0·65	−1·75
T_3T_4 $T_7T_8T_9$	Ferroxcube 2 off FX1105/A4	300	40kΩ	632 (38s.w.g.)	55 (38s.w.g.)	1·0	21·2
T_5	Ferroxcube 2 off FX1105/A4	75	40kΩ	632 (38s.w.g.)	27 (38s.w.g.)	1·0	27·4
T_6	'Permalloy C' ·015 in. 500 T Bobbin	900	75Ω	179 (30s.w.g.)	620 (38s.w.g.)	0·3	−10·8
T_{11}	'Permalloy C' ·015 in. 500 T Bobbin	75	600Ω	620 (38s.w.g.)	219 (30s.w.g.)	1·0	9·0

as T_4 it can be seen that the collector load of transistor X_2 consists of R_d shunted by the primary transform of the 3rd stage input impedance and the generator impedance for the 3rd stage is formed by the secondary transform of the shunt combination of R_d and the output impedance of transistor X_2. It can be seen, therefore, that if one attempts to increase the gain per stage in the cascade amplifier by progressively reducing the number of turns in each transformer secondary then the load to each transistor is increased until eventually its input impedance begins to rise sharply due to internal feedback; at the same time the generator impedance to each transistor is reduced which causes a fall in its output impedance, again due to internal feedback. These two effects are additive so that, since each tuned circuit is shunted by the output of the preceding transistor and the primary transform of the input impedance of the next stage, it follows that a limit is set to the maximum gain and minimum bandwidth of the cascade amplifier. Also difficulty is caused in tuning and there is a reduction in stability of gain and bandwidth against transistor variations although these effects, particularly those related to tuning and bandwidth, are reduced by the insertion of series padding resistance in each emitter circuit.

Preliminary experiments on a single stage using a Ferroxcube tuned transformer with a primary inductance of 1H and an average transistor having $r_c \simeq 1\text{M}\Omega$ and $\alpha_{cb} \simeq 30$ showed that the transistor input impedance begins to increase when the transformed load resistance reaches about 40kΩ. The tuned transformers are therefore designed on this figure but it is necessary to add series resistance of 220Ω in the emitter circuit of all stages except the 1st and 4th in order to pad out the effect of variations in transistor input impedance, as mentioned above, so that the secondary load is taken to be 300Ω. The measured value of R_d for the primary winding of 1H when tuned is found to be about 400kΩ which leads to a calculated value of generator impedance for all stages except the 1st and 4th of 1·5kΩ, assuming a transistor output impedance of 400kΩ. In the 4th stage and also in the 1st stage, assuming an input termination of 600Ω, it is necessary to include a series resistance as large as 820Ω in the emitter circuit in order to obtain an adequate total generator impedance of 1·7kΩ; this has also the advantage of ensuring greater stability of the 600Ω input impedance of the amplifier and of the 75Ω load for the variable attenuator, although a considerable waste of power is involved in each case.

It is found that the most sensitive method of tuning the amplifier is to isolate completely each transformer in turn and tune the primary winding at 1kc/s with the aid of an oscillator feeding through a resistance of about 1MΩ, observing the voltage across the tuned circuit with an oscilloscope or valve-voltmeter. An alternative method is to tune each transformer in the amplifier as it stands, in which case it is advisable to begin with the 6th stage and work back to the 1st stage then repeat until no further improvement is obtained. The value of tuning capacitance of 0·02μF given in the circuit diagram is only approximate; tuning can be carried out satisfactorily by the addition of shunt capacitance in steps of 0·001μF and the total capacitance for each transformer is found to lie between about 0·02μF and 0·025μF.

The maximum gain of the amplifier assuming the transistors to be perfect (i.e. $\alpha = 1$ and $R_{out} = \infty$) may be calculated by addition of the transformer current gains in dB given above but there is a loss of 0·8dB in each tuned transformer due to finite value of R_d, so that one obtains a maximum gain of 135dB under such conditions. Agreement can be obtained between the measured maximum gain (= 126dB) and the computed maximum gain on the basis of each transistor having $\alpha = 0{\cdot}97$ and $R_{out} = 400\text{k}\Omega$.

It was found necessary to enclose the 1st stage of the amplifier in a 'Permalloy C' can to reduce pick-up in transformers T_2 and T_3 due to stray magnetic fields and also to screen the input leads and make all external leads generally as short as possible. The 1st stage transistor was specially selected for low noise figure so that, using a 600Ω load followed by an oscilloscope of sensitivity 100mV/cm, it is possible to increase the amplifier gain up to 84dB before noise becomes discernible in the oscilloscope trace. One therefore obtains that the maximum sensitivity of such a system is about 5μV/cm which is adequate for use as a bridge detector. If, however, the gain of the seven-stage amplifier is, for any reason, reduced by more than about 50dB using the variable attenuator it is possible for noise and pick-up in the stage immediately following the attenuator to reduce the output signal-to-noise ratio and hence also the maximum sensitivity of the system. Switch S_4 is therefore provided as an alternative means of reducing the amplifier gain by switching off and by-passing the 5th and 6th stages, thus reducing the maximum gain by about 40dB.

Noise Figure Measurements

For transistor noise figure measurements the complete seven-stage amplifier is used together with a thermo-junction output meter having a full scale deflexion of about 1mW in 75Ω. The transistor under test is placed in the test

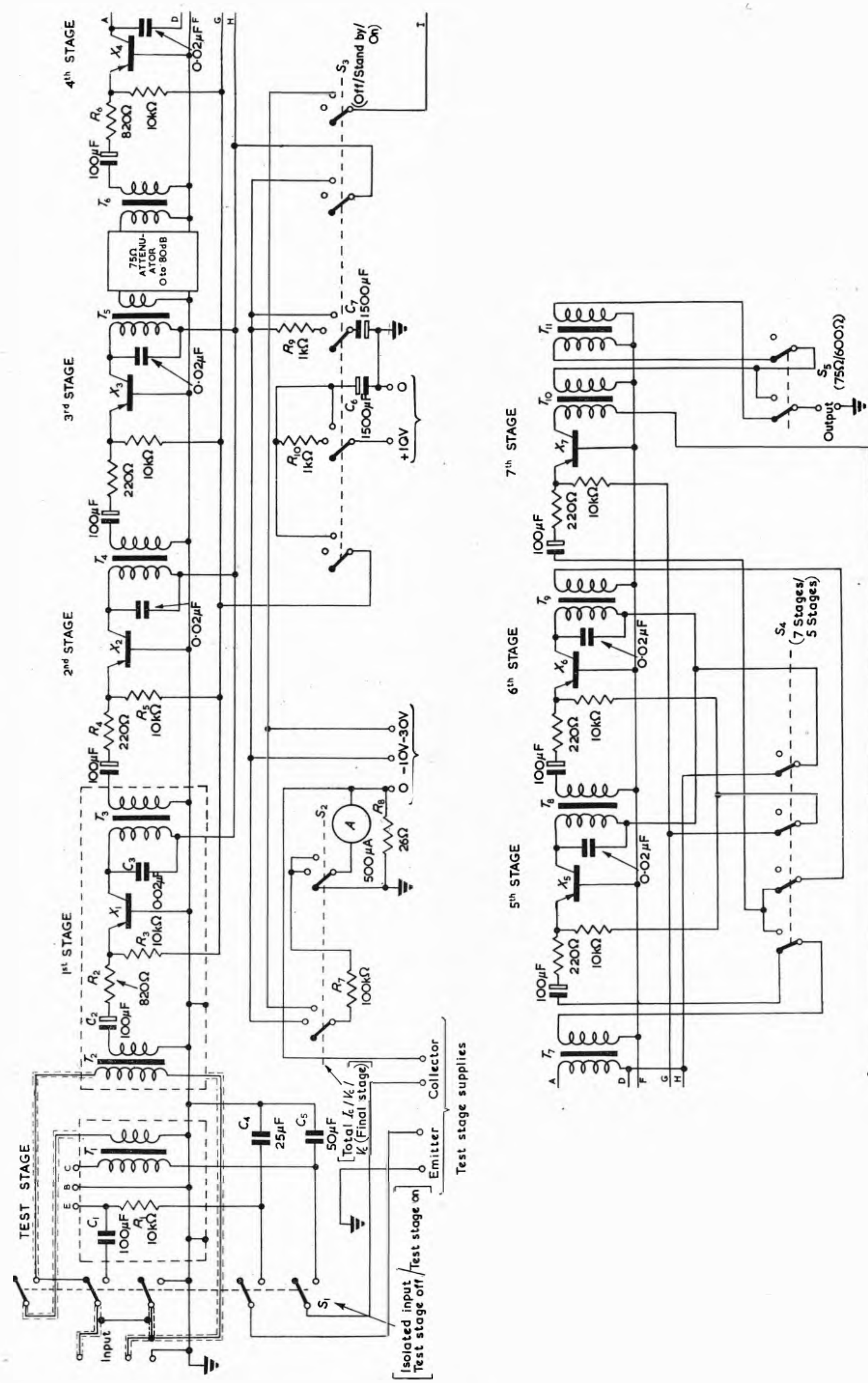


Fig. 1. The complete circuit

stage, shown in the circuit diagram (Fig. 1), which is a common-base circuit whose output is transformer coupled to the 600Ω input of the amplifier so that the stage has an insertion gain of about 22dB between a 600Ω generator and 600Ω load.

As in the 1st stage of the amplifier the test stage is screened and external leads are either screened or made as short as possible in order to reduce pick-up.

The method of measurement of noise figure can be described with reference to the simplified diagram of the test stage shown in Fig. 2. Connected to A (the input of the test stage) via a 600Ω variable attenuator is a 1kc/s oscillator with a 600Ω output impedance and the equivalent circuit of this combination as seen at A is a sine-wave generator of r.m.s. voltage e and internal resistance R (= 600Ω). In the circuit must also be included the thermal noise generator associated with R whose mean square voltage is given by

$$e'^2_N = 4kTbR \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

where k = Boltzmann's constant in Joules/°C

T = Absolute temperature

b = Bandwidth of tuned amplifier in c/s.

The method of procedure is to switch out the oscillator and hence reduce e to zero, then the amplifier gain is

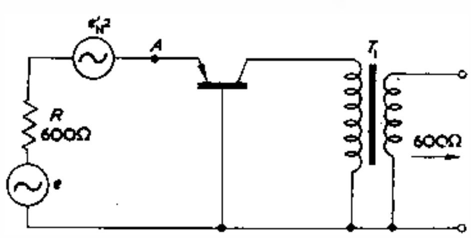


Fig. 2. Simplified diagram of test stage

adjusted until a certain reading of noise of approximately 1mW is obtained on the output meter. The amplifier gain is then reduced by 3dB and the oscillator output increased until the output meter reading reaches the same value as before. From the oscillator output voltage and the attenuator setting the value of e can be calculated, then the noise figure F_{12} of the transistor under test in cascade with the amplifier is given by

$$F_{12} = e^2/e'^2_N = e^2/4kTbR \quad \ldots\ldots\ldots\ldots \quad (2)$$

for an input termination of R (= 600Ω).
Now it can be shown that if

F_2 = noise figure of amplifier with input termination of 600Ω (measured by similar method)

F = noise figure of transistor under test with input termination of R (= 600Ω)

G = insertion gain of test stage between 600Ω generator and 600Ω load.

then

$$F = F_{12} - (F_2/G) \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

provided that $F_2 \gg 1$ and provided also that the noise output of the amplifier is not affected by replacing a 600Ω termination at the input of the amplifier by the output impedance of the transistor under test. By carefully selecting a transistor with a low noise figure for the 1st stage a value for F_2 of 16dB has been obtained and it is found experimentally that increase of input termination of the amplifier from 600Ω to infinity results in a reduction in noise output of less than 0·5dB. Hence the assumptions leading to equation (3) are considered to be justified and one obtains that

$$F_2/G = (16 - 22)\text{dB} = -6\text{dB}$$
$$= 0{\cdot}25 \text{ expressed as a power ratio.}$$

It can therefore be concluded that the correction term representing the noise contribution of the amplifier is negligible for most purposes so that, since there is no evidence to suggest a significant noise contribution from electrolytic capacitors C_1, C_4 and C_5, the noise figure of the transistor under test may be measured directly.

Summary of Data on Seven-Stage Amplifier

(1)	Input impedance	600Ω
(2)	Input capacitance (either side to earth)	400pF
(3)	Discrimination against in-phase voltages	−80dB
(4)	Load impedance	75Ω or 600Ω
(5)	Maximum gain	126dB
(6)	Peak overload output level	(Noise. About 1mW) (Sine wave. 5mW)
(7)	Bandwidth	80c/s centred at 1kc/s
(8)	Stability (over period of 12 months) Gain Bandwidth	 Reduction of 1·5dB No measurable change
(9)	Noise figure (600Ω input termination).	16dB
(10)	Microphony	Only effect due to magnetostriction in Ferroxcube transformer cores which is negligible under normal laboratory conditions.
(11)	Power consumption	160mW
(12)	Dimensions*	Front panel 19in × 8¾in Depth 6 in
(13)	Weight*	16lb.

Conclusions

It can be seen that by the use of junction transistors a high-gain tuned amplifier has been constructed which has the advantages of low power consumption, good gain and bandwidth stability and negligible microphony. As regards bulk it should be pointed out that the space on the front panel is occupied almost entirely by switches, terminals, meter, variable attenuator, decoupling capacitors, and screening boxes for test stage and 1st stage. The bulk due to transistors, transformers, etc., in the amplifying stages themselves is very small in comparison with the total and in fact it has been found possible to reduce the total bulk of four similar stages, each separately screened in a 'Permalloy C' box, to a size of 3in × 2in × 2in and a weight of 1½lb.

It is therefore concluded that, using junction transistors in this application instead of thermionic valves, the expected advantages of reduced power consumption, bulk and microphony can be obtained in practice.

Acknowledgements

The author is indebted to Messrs. Standard Telecommunication Laboratories Ltd for permission to publish this article and to Mr. T. H. Walker under whose supervision the work was carried out.

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* *Excluding batteries which are connected externally.*

The Application of Square Hysteresis Loop Materials in Digital Computer Circuits

By A. D. Holt*, B.Sc., M.Sc.

The properties and uses of magnetic materials, in particular those of ferrites, used in digital computers are described. Theories and the practical design of shifting registers are explained. Core matrix storage systems are reviewed.

A description is given of a store in which a shifting register is used to convert the parallel output into serial form; this register is also used during the writing process.

AN earlier article[1] stressed the need for reducing the total number of valves in a digital computer and showed how this could be achieved by using selenium rectifiers. In this article it is shown how further reductions become possible when square loop magnetic materials are used.

Square Loop Materials

These may be classified as ferrite, metal tape and evaporated metal film.

The ferrite core normally has higher coercive, and lower saturation values than the metal tape. To reduce eddy currents the latter, however, must be ultra-thin and so become fragile, the hysteresis loop then being easily distorted.

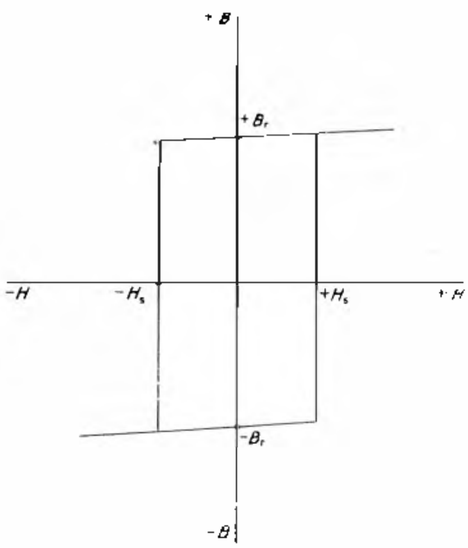

Fig. 1. Idealized hysteresis loop

It would appear that a suitable material, i.e. one having a short switching time, low coercive force and high saturation value can be obtained by using evaporated metal film techniques[2].

Fig. 1 shows an idealized hysteresis loop. The two stable states of magnetic remanance, $+B_r$ and $-B_r$, can be used to represent the binary numbers 1 and 0 respectively.

In many applications the properties of the cores must be maintained within close tolerances. As the cores are used under pulse conditions, they are tested by applying some repeated pulse train similar to that shown in Fig. 2 where $P_4P_5 \ldots P_{n+1}$ are half the current amplitude of P_1, P_2, P_3 and P_{n+2}.

From this test can be selected cores having the following properties.

(1) Similar outputs resulting from switching by P_2 and P_{n+2}.

(2) Small amplitude and duration of the outputs from $P_1, P_3 \ldots P_{n+1}$, compared to the full output; the former having decayed to negligible values when the latter are at peak amplitude.

Although for some uses the full test described above is not essential, if carried out, it does ensure that a uniform batch of cores is obtained.

Originally magnetic tape cores were used in computers, but in subsequent designs these have been replaced by the more robust ferrite cores. The latter are used in all the circuits described below. To maintain the reliability which these components are capable of achieving, semiconductor junction devices have been used, wherever possible, to replace point contact diodes and valves.

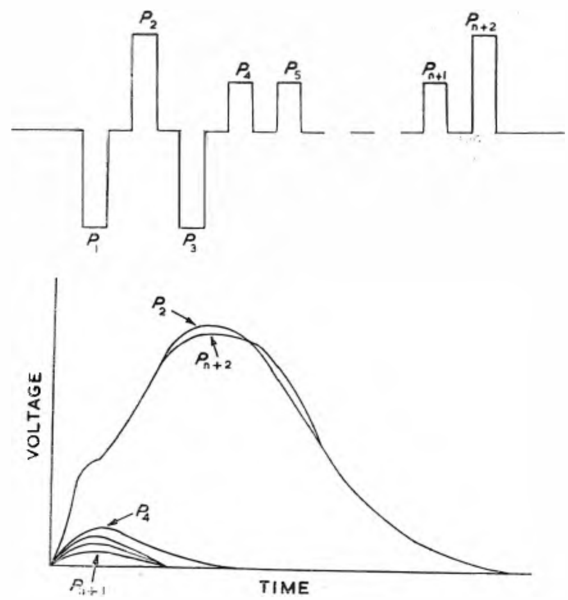

Fig. 2. (a) Pulse test waveform
(b) Output waveform

Shifting Registers

In the majority of computing machines the main number store has only one output. Since arithmetic processes such as addition require two numbers, a further digit store must be included from which a second number can be extracted when required. Such a unit can be a shifting register, the basic core circuit of which is shown in Fig. 3.

Let the current pulses A, B and C be applied sequentially to the cores so that the latter are driven to the $-B_r$ state. Assume that all the cores are in this state. When drive B is applied no flux change will occur in core 2, so that all

* *The Nelson Research Laboratories, The English Electric Co. Ltd.*

the cores will remain in the $-B_r$ state. Similarly, after the application of the remaining drive pulses the states of the cores will be unchanged. This action is the transfer of the digit 0.

Assume now that all the cores, except number 2, are in the $-B_r$ state and let drive pulse B be applied. A large flux change will then occur in this core as it switches from the $+B_r$ to the $-B_r$ state. This change of flux will induce a voltage across its output winding causing current to flow through the input winding of core 3. If this current is sufficiently large, then core 3 will be switched from the state $-B_r$ to $+B_r$ (i.e. transfer of the digit 1).

During this switching process a voltage will be induced across the output winding of core 3, and this must be isolated so that no other core is affected. A rectifier in series with the output winding of the core is used for this purpose.

When the voltage is induced on the output winding of core 2 a similar voltage is simultaneously developed across its input winding. The latter will cause a current to flow through the series rectifier and output winding of core 1 which would, if of sufficient amplitude, switch this core.

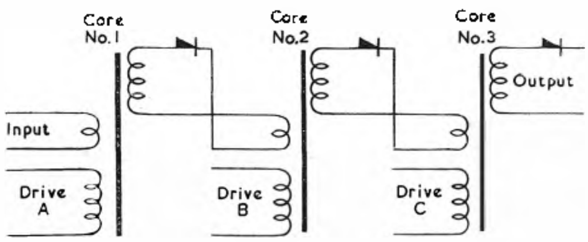

Fig. 3. Basic shifting register circuit

This effect may be overcome in the following ways[3].

(1) Drive pulses can be applied simultaneously to cores 1 and 2 (i.e. both driven to the $-B_r$ state).

(2) A shunt rectifier can be connected across the input winding of core 2. The equivalent (ideal) circuit is then unaltered for the input time, but during the output time the rectifier acts as a low resistance shunt across the output winding of the previous core and the series rectifier. It also loads core 2, and to reduce this effect a resistance is introduced between the junction of the two rectifiers and the input winding. This resistance is also used as a matching device.

(3) The ratio of the number of turns of the input and output windings can be so adjusted that the back flow of current (ampere-turns) is less than half that required for switching.

In the first system only, three cores are required for every digit stored, drive pulses being applied to two while the digit is being transferred into the third. Otherwise two cores per digit are needed since the rectifiers act as isolators

A circuit where each core can be used as a digit store is shown in Fig. 4. In this particular example only one winding is used for both output and drive. Assume that all the cores, except number 2, are in the $-B_r$ state; then when the drive pulse is applied little or no voltage will be developed and only a small current will flow in the output circuits of these cores. The winding of core 2, however, will present a high impedance and a large voltage will be developed across the drive/output winding which charges up the associated condenser. The latter discharges through the input winding of core 3 and, in doing so, switches it from the $-B_r$ to the $+B_r$ state at a time and rate determined by the properties of the core and the value of the series resistance.

The series rectifier in this circuit

(1) Prevents cores from loading previous ones.

(2) Allows the capacitors to discharge only through the input windings.

A backward flow of information is also prevented by a suitable choice of turns ratio.

Design Theories of a Two Core per Digit Register

Theories[4,5,6] published on the design of these units have assumed ideal components and square wave drive pulses. The core is treated as having an idealized rectangular hysteresis loop, while the rectifier resistances are assumed to have infinite reverse and constant low forward values. From these studies, the following four main results are derived when it is desired to operate the units under optimum energy conditions.

(1) The flux changes, both of the core driven and the one it switches, should occur in the same time and should be identical. If the former switches before the latter, incomplete transfer will result, while if the reverse occurs excess energy is dissipated in the resistor.

(2) The ratio of the output and input turns should be 2 : 1.

(3) The ampere-turns applied through the drive winding

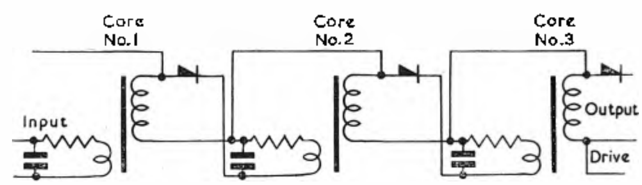

Fig. 4. Single core bit shifting register

of the core should be at least four times that required to switch an unloaded core.

(4) The sum of the forward resistance of the rectifier and that of the resistor must equal the equivalent resistance of the input winding.

Practical Design

Although the theoretical results may be used as a guide to design an experimental method is necessary, since the components used have not the assumed idealized properties. In addition the inductive load of the drive windings causes ringing if the drive pulse has both fast rise and decay times. This is particularly noticed during the latter, so that a slow decay time is necessary.

The pulse shape and current amplitude must be constant and independent of the load, the magnitude of which will depend upon how many cores in the register are in the '1' state at any particular time. If the cores are in the anode circuit of a valve driven from a cathode-follower, a feedback capacitor from anode to grid is often sufficient for both pulse shaping and stabilization. A small resistance, placed in series with the drive winding of the last core of each phase and the h.t. supply, will absorb, and so prevent reflections of the drive pulses.

The circuit used for the design is shown in Fig. 5 and simulates a section of a shifting register[7]. Two sets of drive pulses are alternately applied to the first core only. The tapped output winding from the first core is connected to the input of the second through a variable resistor and two rectifiers (one being in the reverse direction to the other) wired in parallel. Thus, when the first core is switched in either direction, current can flow through one or other of the rectifiers and switch the second core. The loading is now similar to that occurring in a shifting register when only the forward flow of information is considered. To simulate the loading in the reverse direction, a tapped input winding on the first core is short-circuited through a variable resistor and two further rectifiers connected as described above. During the experiment the values of the two resistors and the number of turns on the two input windings, although varied, are maintained equal. Pulses are applied and the flux or voltage outputs observed on a double beam oscilloscope. The number of turns used in the input and output windings is varied, and the two resistors adjusted, until the patterns are identical. A shifting register can now be constructed using the values found. If the additional load on the driver circuits alters the shape of the drive pulses, the flux patterns from the experimental circuit will no longer be identical and the two variable resistors will need a slight readjustment. When the register resistors are changed to this value, no further setting up procedure is required. To eliminate the need for this finalization a load, equivalent to that of a register, can be inserted in series with the drive windings of the experimental circuit.

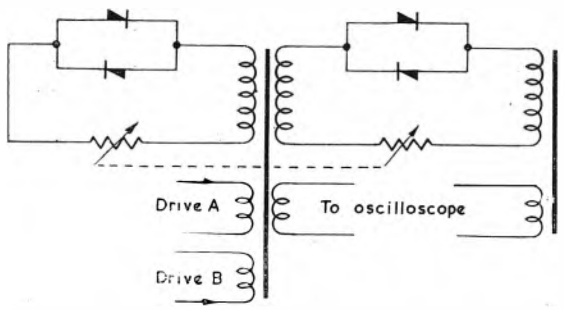

Fig. 5. Design circuit for shifting register

Using this method, shifting registers have been designed to work at digit shift rates of several hundred kc/s. Suitable cores are of 2mm o.d. and the rectifiers germanium junction.

Matrix Stores

Any core fulfilling the tests described above (see Square Loop Materials) can be used as an AND gate, i.e. its state of remanance will be unaffected by a current pulse producing a magnetization force equal to half its coercive force, but will be changed if two such current pulses are applied. The cores, therefore, can serve as a store and as a gate. These properties have been utilized in the majority of matrix stores constructed[8]. Other systems[9], not employing this dual facility, require at least two cores per digit stored so that construction costs are increased, although there are some advantages.

Fig. 6 illustrates the method used for the construction and wiring of a matrix plane. If half amplitude pulses are applied to one horizontal row and one vertical column, only the core situated at the intersection of these will be switched. The remaining cores on both the row and the column will, however, be disturbed and outputs similar to $P_4, P_5 \ldots P_{n+1}$ (Fig. 2) will be generated. Since the output wire threads the cores on the rows and columns in reverse directions, these disturbing signals are largely cancelled. Also these signals will have decayed to a small value at the time that the true output signal is a maximum, so that a strobing system can be used.

A single plane, similar to that shown in Fig. 6, could be used to form a serial store. Its mode of operation would be as previously described. An address system would be required to select the appropriate row and column. Pulses in the correct temporal sequence would be applied, first in one direction to 'read' the digit stored and then, if required, in the reverse direction to 'write' in a digit. The large amount of auxiliary address equipment needed would make such a store uneconomical. In practice it is usual to construct a parallel store where all the digits of one number are read from separate planes at the same time. These digits can then be serialized if required. The store is constructed by using n single planes where n is the number of digits in the word to be stored. The rows and columns of each plane are connected in series with the corresponding rows and columns of the other planes, so that when the pulses are applied only one core in each plane receives the full switch current. The writing process differs from that previously described. The two half write pulses are applied to all the cores of the number that is being written. Where the digit 0 is to be stored a bias pulse of half switch amplitude is applied to the appropriate plane.

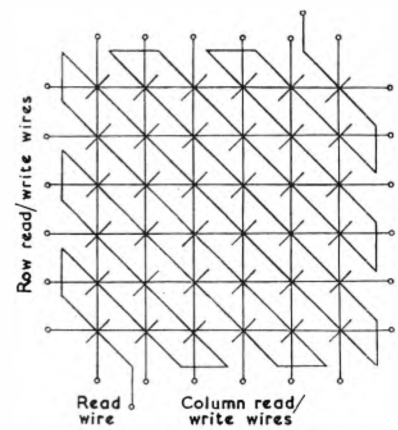

Fig. 6. Core matrix plane

A single plane can also be used to form a further type of parallel matrix store[10]. If all the digits of each word are stored on a row of cores and a full amplitude pulse is applied to this row, then outputs corresponding to the digits stored will be generated on the column wires.

As in the previous system, two half amplitude pulses are used for the writing process.

This system has the advantage that only cores being read receive the read pulse, so that no disturb signals are generated and strobing is generally unnecessary. In addition, if a pulse exceeding that required to switch the core is applied, the amplitude of the output signal is increased and its duration decreased. The disadvantage of this form of store is that the selection equipment is larger since each word must be selected.

Uses of Shifting Registers and Matrix Stores

Although the cores used in these units are capable of permanent storage, this property in general has not been fully exploited. For example, if a power failure occurs, or if the units are switched off, no attempt is made to retain the information they contain. Instead this, or new data, is inserted after the units have been brought back into operation.

Registers incorporated in computers have usually been made continuously recirculating, the input and output being gated. This is done due to the difficulty of gating the driver valves, especially when other cores are used as logical units. Here an economy results since more than one register can often be driven from one set of driver and pulse shaping valves. With this type of unit the main costs are the rectifiers, winding the cores and the wiring. The expense of the cores and driver units is only a small proportion of the total.

As semi-automatic methods are used for wiring the matrix planes their cost is only a small proportion of the total for a store of this type. In general, two valve amplifiers per plane are required, one to increase the size of the output signal and the other to generate the inhibit or write '0' pulse.

Selection equipment is required, but its cost decreases when the number of words from which selection has to be made is increased.

Thus, if only a small number of words are to be stored it is usual to use shifting registers; otherwise a matrix store is preferable.

Matrix Store for use in a Computer using Core Circuits

The store illustrated in Fig. 7 is suitable for a serial machine when a capacity of from ten to a hundred words is required. A parallel output is read from the matrix planes and is serialized in the register, the converse process being applied for writing.

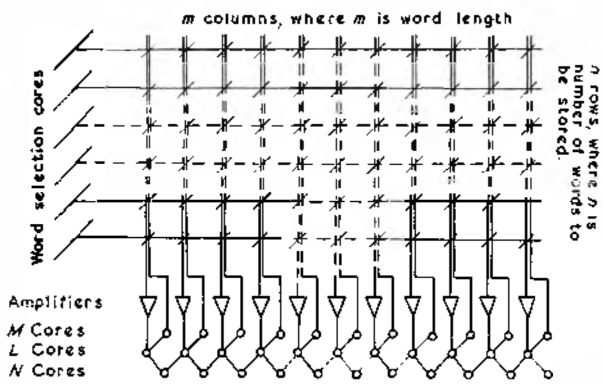

Fig. 7. Arrangement of matrix store

The pulses used for the reading process are obtained from cores. These are normally biased so that only one core can be switched during the reading time. The output, from many turns, will depend upon the flux change, the load and the amplitude of the drive. If the output is sufficiently large, the storage cores associated with it will all be left in the '0' state, those being switched inducing voltages in their output windings. The amplitude of these can be increased by using a winding of more than one turn but, since semi-automatic methods of winding are then impracticable, the expense of the matrix is increased if many turns are used.

These signals cannot be used to switch other cores directly, since the currents required would affect the matrix. Valves or transistors must be used to amplify the currents and switch the appropriate cores in L. These are connected by register type circuits to the cores in M and N, those in N being similarly connected to the ones in L. Thus a binary number read from the matrix store in parallel, can be obtained in serial form from the register.

The output windings of the M cores are connected to the corresponding matrix inputs. Their outputs are designed to be of half storage core switch current amplitude and so, by themselves, do not affect the store. Being driven at the same time the M and N cores are always in the same condition.

While the serialization is in process the next word, which has to be stored in the matrix, is inserted into the register. Should this be the original word the output of the register is fed back into the input as well as to the unit where it is required.

After the last digit of the stored word has entered this position, the next word will be contained in cores M.

The core used in the selection system is now set to its original state. The rate of change of flux is designed to produce a current flow of half switch amplitude in the storage cores with which it is associated. At the same time the M cores are driven, and the storage cores associated with those storing digits will receive two half current pulses and will themselves be switched.

Currents from the selection and M cores are stabilized by the use of series resistors. This is necessary due to the different impedances of a core when fully switched or only disturbed.

The register used is designed as previously described, except that the switched core must drive two others connected in parallel.

The drive current must be maintained constant for different numbers stored, otherwise the outputs from the cores would vary. This is achieved by using a large resistor in series with the drive windings and the h.t. supply.

The system has been described for use with a serial type store, but the same principles could be applied to a parallel machine. Its advantages are apparent when the remainder of the associated equipment consists mainly of circuits using cores.

Acknowledgments

The author wishes to express his thanks to Dr. A. D. Booth for the helpful discussions on the above work. Thanks are also due to Messrs. J. H. Millie and J. Dakin for their help in constructing some of the above circuits.

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Ultrasonic Metal Cleaning

In the past few years there has been a rapid growth in the development of practical uses of ultrasonics for cleaning, homogenization and sterilization, pigment dispersion and also for acceleration of fermentation of alcohols.

An application of ultrasonics to metal cleaning is now available combined with a vacuum technique from Electro-Chemical Engineering Co. Ltd., Woking, Surrey. Developed by Technochemie of Switzerland, EFCO this method makes a high standard of cleaning available to industry. It has already proved itself with nearly 100 production installations in Europe and America; most of these installations in Europe being by the Swiss watch industry. It was at first almost entirely devoted to the small intricate mechanisms of this type of industry, but now it is extended to very large installations which can deal with high production requirements and the cleaning of a considerable mass. The frequencies used are 400kc/s and 40kc/s and the transducers are piezo-electric and made of barium titanate. Frequencies lower than these can in some cases cause damage to fragile parts by vibrations, and could, if low enough, cause damage through erosion.

The advantages of the vacuum technique are considerable. With the removal of air from the solvent, cavitation is far more effective, air pockets are removed from any intricate component, thereby enabling the entire surface to have good solvent-surface contact which is essential for good cleaning. Impacted compound in small blind holes is also easily and readily removed with the use of this technique.

The cleaning sequence usually used is as follows: first there is a pre-cleaning operation with perchlorethylene or trichlorethylene with continuous filtration. Then follows the ultrasonic stage with or without vacuum, but with continuous filtration using the same solvent. There follows a continuously filtered rinse also using the same solvent, a vapour degrease process and finally a drying stage which may include the use of vibration, infrared and a stream of hot filtered air. All these processes may be incorporated in hand or fully automatic machines.

Controlled Saturation in Transistors and its Application in Trigger Circuit Design

(Part 2)

By N. F. Moody*, A.M.I.E.E.

A Complementary Pair Trigger Circuit Employing Controlled Saturation

The main circuit to be described has, as its design target, the specification of a similar binary element in which saturation is prevented. The linear circuit is described in a companion article[1], but its specification is repeated in Table 3 for ease of reference.

While this specification can be met, one item has been relaxed in order to reduce power consumption. It is marked by an arterisk in the specification for the saturated circuit, and may be varied at the designer's discretion. In the form here shown, the standby power consumption is minimized. It is about one half that of the linear circuit. The saturated circuit also provides a useful economy in the number of diodes required.

A simplified variant of the circuit is also mentioned, which can always be used when a longer resolving time is acceptable. In certain circumstances the simplified form will yield the same resolving time as the original circuit.

Detailed circuit discussions will be restricted to those aspects which involve saturation. The remaining principles are treated exhaustively in the companion article[1], to which the reader is referred for supplementary information.

The main circuit is shown in Fig. 5, and its description is divided into five sections. In the first section the output transistor X_2 is discussed separately. It is shown that its properties are similar to those of a linear emitter-follower. The design criteria for X_2 are then given, and the circuit is evaluated to meet the load called for in the specification.

The second section describes the conducting state of the complete trigger circuit; and shows that the conditions required for X_2, as outlined in the previous section, are met correctly. Then follows the third section, which concerns itself with the non-conducting state of the binary stage.

The fourth and fifth sections deal with the transient phenomena which occur during the changes of state.

The Saturated Emitter Follower

Reference to the circuit of Fig. 5 will show transistors X_1 and X_2 form the binary element proper. Transistor X_3 is an auxiliary, added to facilitate fast turn off. It remains non-conducting in both the 'on' and 'off' states of the trigger.

TABLE 3
Trigger Circuit Specifications

	A—LINEAR CIRCUIT	B—SATURATED CIRCUIT
Switching Time	<0·2μsec	<0·2μsec
Output Fall Time when Driving 5 Similar Stages	<0·2μsec	0·2 to 0·3μsec
Output Rise Time when Driving 5 Similar Stages	<0·5μsec	~1μsec*
Output Load Current (max.)	>30mA	>30mA
Output Impedance	<20Ω	~20Ω
Resolving Time	<1·5μsec	<1·5μsec

* Defence Research Board, Canada.

Transistor X_2 performs a dual role. Its collector-base circuit is a part of the trigger regenerative loop. The emitter-base circuit acts, semi-independently, to provide a low impedance high current source for the external load. It is this second, emitter-follower like action which is discussed in this section.

The circuit of X_2, when restricted to this one function, and when it is reduced to essentials, becomes that of Fig. 1. When saturated, the circuit in Fig. 1 continues to display properties which are analogous to those of a linear emitter follower, as will now be demonstrated.

The linear emitter-follower is characterized by a low impedance emitter output terminal, whose voltage approxi-

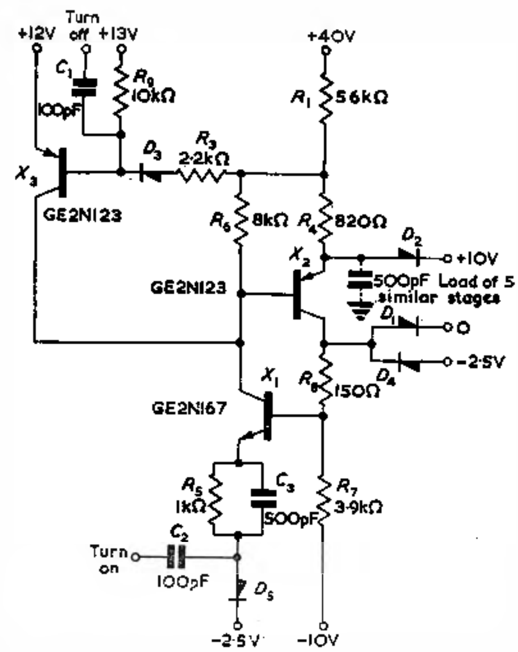

Fig. 5. The final trigger circuit

mates that of its base. Its saturated counterpart yields a low impedance emitter output terminal which, because the transistor is saturated, effectively takes up the collector potential. The only difference lies in the source of reference potential for the output electrode. Both circuits can handle emitter loads up to approximately $I_b\beta_n$. At this maximum load, similar values of β_n and T_n give identical storage in either circuit.

The design of a saturated emitter-follower differs little from its linear counterpart. First, the maximum emitter load must be known. A source of base current must now be provided, of magnitude sufficient for this load. Usually a range of β_n values will be associated with the type of transistor used: sufficient base current must be available to satisfy the transistor of lowest β_n.

For the circuit of Fig. 5, the maximum emitter load is 36mA (composed of 30mA external load as required by the specification, plus 6mA required by the trigger itself). If the minimum β_n is taken as 23, the available current must be 1·5mA.

According to the theory of Part 1, for minimum storage, the base current of a saturated transistor should be as small as possible, and so it should be set in accordance with the β_n of the sample. In the trigger circuit described here, no provision is made for adjustment of base current, whose value is set at a constant 1·5mA so as to satisfy the lowest β_n. Thus, when an unfavourable transistor is chosen, the maximum storage of the saturated circuit will exceed that of the linear form. The ratio is easily evaluated.

For a linear emitter-follower, at maximum design emitter load, the storage range has a ratio of $T_{n(max)}/T_{n(min)}$. For the saturated form under similar conditions the range is $(\beta_n T_n)_{max}/(\beta_n T_n)_{min}$.

Measurements on 50 transistors show the extreme ratio of storage in the linear circuit to be 3:1, and for the saturated circuit as 4:1 (see Fig. 6). It has been shown already that both circuit forms have the same minimum storage: thus the saturated circuit may store some 30 per cent more than the linear emitter-follower.

In practice this increased storage is not very important, for the total turn-off time of the complete trigger will prove to be appreciably longer than that calculated for T_s alone. This is shown in a later section.

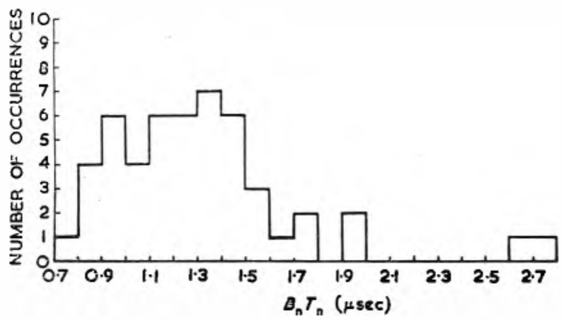

Fig. 6. Spread of the parameter

Test on 50 transistors type GE2N123 B_n and T_n at I_c = 25mA V_c = −4V. Values B_n are incremental (or a.c.) values samples have $B_n \geqslant$ 23.

Static Conditions when the Binary Stage is in the 'On' State

In the 'on' state, both transistors X_1 and X_2 are conducting. The former always operates under linear conditions, but the latter is in controlled saturation. It receives a fixed base current under all load conditions, whose value is set to 1·5mA for reasons given in the preceding section.

The base current of X_2 is set directly by the emitter current of X_1, as will be shown. This emitter current is itself defined in terms of R_5 and the base potential of X_1. A first step is to establish this base potential.

The base of X_1 is connected to the junction of R_7R_8 whose upper extremity is almost at earth potential because D_1 is conducting. If the network R_7R_8 were not loaded by X_1 base, the junction would be at −0·4V. In practice, this load is small enough to neglect (0·1mA). There is thus 2·1V between base and emitter supply of X_1. If the voltage drop from base to emitter, as well as that due to D_3 are ignored, R_5 sets the emitter current of X_1 at 2·1mA. In fact these drops nearly cancel a drop in D_1, so that this current is obtained in practice.

The base current of X_2 may now be evaluated: it is the difference between the collector current of X_1 and the current in R_6. The former is close to 2·0mA (since X_1 is not saturated) and the latter is approximately 0·5mA. Thus the base of X_2 is excited with a defined current of 1·5mA, which remains sensibly constant whatever the external load conditions.

It will now be shown that the diode D_1 is conducting (as was assumed) so that it defines the output terminal as near-ground potential at all loads. Thus the current in R_1R_4, plus the load current, is injected to the emitter of X_2. It appears at the collector diminished by the base current of 1·5mA. Since the current in the network R_7R_8 is only about 2·5mA, and further, since R_1R_4 injects 6mA, the diode D_1 is conductive at all external loads down to zero. It follows that the 'on' state is always stable. A major part of the load current leaves the structure via this diode; as is necessary, since it must not be allowed to drive X_1 into saturation. The regulation of the output voltage with load is determined almost entirely by the diode D_1, whose impedance is higher than the emitter-collector impedance of X_2.

Transistor X_1 is prevented from saturating by the voltage drop across R_8. It has been seen that the base of X_1 is at about −0·4V, while the collector is connected to the base of the saturated transistor X_2, which must me near zero voltage. Any voltage drop in the diode D_1 will displace the collector voltage from zero, but this does not disturb the voltage difference existing from collector to base of X_1. This difference (0·4V) is adequate to prevent saturation at the small current X_1 is required to handle.

Static Conditions in the 'Off' State

In the 'off' state, the output terminal is restrained at +10V, by some 4·5mA flowing from the network R_1 and R_4 into the diode D_2. The voltage drop across R_4, about 4V, applies cut-off bias to the base of X_2, while the bias of about 0·4V appearing across R_8 maintains cut-off on X_1.

The Transition from the 'On' to 'Off' States

The analysis of this transition is simplified where the trigger circuit feeds a capacitive load, and this situation will be assumed in the first place. Such a load is approximated when five similar circuits are being driven, when the capacitance is 500pF as shown in Fig. 5.

The turn-off transistor X_3 is normally cut off, because of bias applied via R_9. This resistance is made as high possible, consistent with maintaining X_3 cut off in the presence of its own I_{co} current together with the reverse current in D_3. When this is done, the presence of R_9 may be ignored during the turn-off action which follows.

The turn-off pulse is assumed to be a negative step waveform of from 6 to 10V in amplitude. This step causes charge to be withdrawn from X_3 base, of magnitude corresponding to the applied voltage step, less one volt of cut-off bias. The resultant collector current can be evaluated by the methods given in Part 1 where an identical circuit (Fig. 2) is analysed. It will range from 20 to 40mA (in correspondence with the amplitude of input pulse) and is associated with a decay time-constant 1 to 3μsec. This collector current is injected as reverse base current to X_2, and so initiates its cut-off which, in turn, causes the binary pair to switch to its non-conducting state.

The reverse current generated by X_3 closely approximates an idealized step during phase 1 of carrier clearance, and exceeds that which X_2 base can accept during phases 2 and 3. Thus the waveform satisfies the analysis of Part 1, which becomes applicable here. In that analysis, low impedance paths were required at both emitter and collector of the saturated transistor. Such an exit for carriers is provided at the collector by D_1; at the emitter the load capacitance of 500pF will serve the same purpose.

The conditions for carrier clearance in X_2 are now similar to those described in Part 1 with the exception that the network R_1R_4 may continue to inject carriers during the turn-off process. If the analytical methods of part 1 are applied to this case it will be found that the stored charge remains unmodified and that carrier clearance proceeds at an unchanged rate. Thus the earlier analysis remains valid, and equation (19) may be applied to determine the turn-off time T_s. An evaluated example based on sample 203 was given in Part 1. The value of T_s, 0·25μsec, which results, corresponds with minimum pulse applied to the trigger terminal. Because sample 203 has a time-constant $\beta_n T_n$ which lies at the upper boundary of the histogram in Fig. 6, this sample yields the longest turn-off time which is to be expected.

If the capacitive load on the output terminal is reduced, it ceases eventually to provide a satisfactory exit for carriers from the emitter of X_2. An extreme case occurs when the load capacitance is zero, for then no low impedance path can exist until the emitter has risen by 10V to catch on D_2. In this situation, as will be shown, the collector capacitances of all three transistors must receive charge also and the turn-off time is increased accordingly. Intermediate conditions of loading give times intermediate between these limit values.

When the load is purely resistive, the collector region is first cleared of sufficient carriers for the base to become unclamped from the collector. Reverse base current will now cause the base to become increasingly positive. During this phase the collector-base capacitances of all three transistors must be charged, until 10V of motion has occurred and the emitter is caught by D_2. A typical charge of 300pC is required for the purpose, and this must also be supplied by X_3. Once the emitter of X_2 is caught the elimination of carriers proceeds normally.

Thus, in the absence of an external capacitive load, X_3 must supply an added charge of about 300pC. This extends the clearing time of X_3 by 0·1 to 0·15μsec making a maximum clearing time of 0·4μsec.

The second transistor of the binary pair, X_1, commences to cut off during the time interval required to clear carriers from X_2. Once the collector current of X_2 falls below 2·5mA diode D_1 cuts off, and R_7 injects reverse base current into X_1. Since this reverse base current is of the same order as the emitter current of X_1, the collector current decays with a time-constant which approximates T_n. It is harmlessly small after a few time-constants, and is of little importance since other factors predominate in setting the overall turn-off time of the complete trigger.

The turn-off process is not complete until X_3 is itself free from carriers. It has been seen that the natural time-constant of decay of current in X_3 lies in the region 1 to 3μsec. A slow decay is desirable since it permits the initial large base current to be maintained throughout the turn-off of X_2. On the other hand, the continued ability of X_3 to deliver current when clearing of X_2 has been completed can, in unfavourable cases, limit the resolving time to 5μsec. Steps are taken, therefore, to inject reverse base currrent into X_3, at the appropriate time. The network D_3 and R_3 serves this purpose. The injection is made to occur when the emiter of X_2 is well recovered towards its final potential, for at this time its carriers are largely eliminated.

Thus, when the emitter of X_2 is within 2V of its final value, the diode D_3 is on the verge of conduction. Thereafter a current of increasing magnitude flows until, when the recovery is complete, R_3 can inject 0·7mA of reverse base current. The storage in X_3 is relatively small, and its carriers are cleared (from the emitter alone) in about 0·5μsec.

The time taken for the emitter of X_2 to rise to the critical value is determined by the nature of the load which it is driving. It has been seen that when this is resistive, the time will be rather less than 0·4μsec, so that the overall clearing time for the system is about 0·9μsec.

At the other extreme a load of 500pF is encountered. Then the time-constant due to this capacitance in conjunction with R_1R_4 becomes a predominant factor. It limits the emitter rise time to 0·8μsec*. The overall clearing time now slightly exceeds the sum of this time plus the clearing time of X_3: a figure of 1·5μsec is reasonable.

The resistance R_3 is included to limit the maximum rate of carrier clearance in X_3. The rate is limited purposely since conduction in X_3 must not taper off more rapidly than the residual phase 3 current in X_2.

Should the turn-off negative edge be followed, within approximately 1μsec, by a positive edge of equal amplitude, the latter itself will suffice to clear the carriers from X_3. The network D_3R_3 may then be eliminated without loss of resolving time. Alternatively, the network may be dispensed with when a resolving time of 5 to 6μsec is acceptable. This is the simplified circuit variant mentioned in the introduction to Part 2.

The Transition from 'Off' to 'On' States

This transition is easily understood qualitatively, and may be dismissed in a few words. The mechanism closely follows that of the linear form of the circuit, and therefore the reader is referred to the companion article[1] for accurate analytical methods.

The trigger pulse is applied via C_2, and is directed to the emitter by D_5 which serves this sole purpose. Capacitor C_3 passes the impulse to the emitter of X_1; the resultant collector current develops a voltage across R_6 which continues to increase until X_2 becomes conductive.

The resistor R_6 is chosen to meet three criteria: it must be sufficiently high to avoid loss of triggering sensitivity, and yet low enough to handle any small residual current from phase 3 of the turn-off action. (There may yet remain a small current when X_3 has become quiescent.) If it fulfils the latter requirement, it will fulfil the third: namely to absorb the I_{co} currents of X_1 and X_2 in the off state. The value of R_5 has been so chosen that it can inject a current of 0·5mA as X_2 comes into conduction. This current remains sensibly constant as the trigger relaxes into the 'on' state.

Once the process of extracting base current from X_2 has commenced, an amplified current rapidly appears in its collector. The latter overcomes the current in R_7R_8 (whose overall resistance is not critical) and so injects base current into X_1. At this stage the action becomes regenerative.

Capacitor C_3 now performs a second action: it allows the transient emitter current of X_1 to exceed its steady 'on' value of 2mA. Thereby X_2 receives an enhanced base drive during switch on, and so is able to deliver a much faster output waveform to capacitive loads.

General Remarks

The resolving time of the true trigger circuit has been shown to be in the range 0·9 to 1·5μsec, when performing two switching operations which commence and end with the 'on' states. It is considerably less for the alternative situation bounded by 'off' states. These results are substantially load independent, as are the triggering and re-triggering sensitivities.

The use of saturation yields output waveforms of nearly ideal shape, as can be seen from the unretouched photographs of Fig. 7. At the top are shown the trigger input waveforms, and it is the negative edge of each which actuates the circuit. Below are the output waveforms of the trigger at various loads. The lowest waveform simulates one trigger circuit driving five similar ones, according to the design target.

The circuits shown in this article, and the companion article[1], are members of a large group of complementary pair trigger circuits which can yield, with decreasing com-

* *A circuit addition, which involves one further transistor, allows the trigger circuit to recharge such capacitive loads as fast as it can discharge them. It may be employed in conjunction with either the saturated or non-saturated versions of the trigger circuit, and proves particularly useful when the capacitive load is required to exceed the values mentioned in this article. The reader is referred elsewhere[3] for further information.*

plexity, resolving times from 0·5μsec to 10μsec. Information on the various forms may be obtained elsewhere[3,4].

Conclusion

It has been shown that when saturation is employed under suitably controlled conditions, there need be little loss of speed as a result.

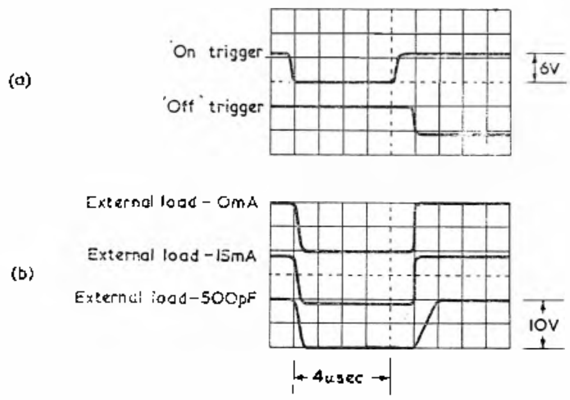

Fig 7. (a) Input waveform
(b) Output waveform

Furthermore, it is found that analytical methods yield simple first order solutions; so that the engineering assessment of saturated circuit performance can be accomplished as readily as for the non-saturated counterpart.

The practical embodiment of these conclusions has been demonstrated in a particular form of trigger circuit. It must be emphasized that the concepts are not restricted to such circuit forms, but are applicable to all switching problems.

The conclusions which relate to large signal conditions are necessarily based on a small number of transistor samples, supported by circuit tests which embrace perhaps 40 transistors. It is planned to build a computer which will use saturated trigger circuits, and it is hoped that this will be a source of statistical support to the conclusions.

Acknowledgements

The author wishes to record his indebtedness to Mr. C. D. Florida for many fruitful discussions, and to Messrs. D. Selwyn and B. Jones for experimental work and much patient measurement

Appendix 2

Output Current of the Turn-Off Transistor X_2

An accurate equivalent circuit of X_3 and its driving source is shown in Fig. A5. All potentials are shown with respect to ground, rather than to +12V as appears in the physical circuit.

This equivalence may be simplified with little loss of accuracy in order to ease the mathematical treatment. It will be recalled that the trigger circuit proper exhibits two main stages in its turn off. In phase 1, X_2 appears as a short-circuit, and the flow of current is set by X_3 itself. For this phase, therefore, the collector of X_3 may be short-circuited as shown in Fig. A6, and since the collector capacitance C_c is then unimportant it may be neglected. The duration of phase 1 is so short that loss of input charge due to R_9 is only one or two per cent. This resistance is therefore omitted, and the cut-off voltage it introduces is included by diminishing the input voltage from E to $(E-1)$ volts, which is written as e. The emitter resistance is properly shown as a diode in both circuits. Because of the rapid impedance fall with current that such an emitter diode exhibits, the form of the voltage drop across it approximates the shape of the input waveform. This small voltage is disregarded here, but it may be added to the cut-off bias if desired. Finally, the source impedance will be treated as a resistance, and its value added to the base resistance r_b' to form a net series resistance r. It will be found that this approximation is quite acceptable.

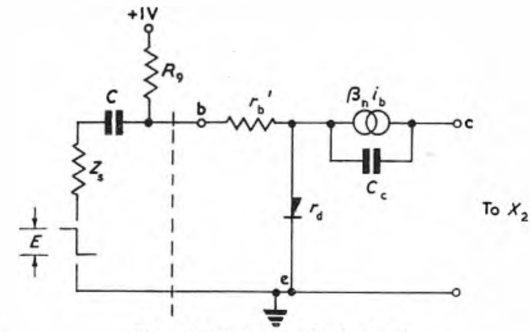

Fig. A5. Equivalent circuit for X_3

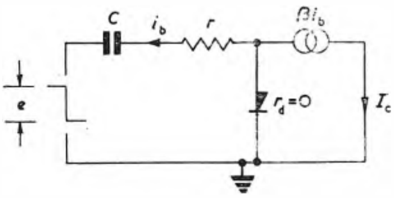

Fig. A6. Simplified equivalent circuit of X_3 for phase 1 of turn off

During the second phase of turn off the transistor X_2 itself limits the current it can accept, and the surplus delivered by X_3 charges the stray capacitances of all three transistors. The collector of X_3 then generates a voltage and C_c becomes significant. This action will be treated as an extension of the phase 1 analysis.

When the sum of the times of the two phases are added it is found that R_9 introduces a small charge loss. The analysis concludes by giving an approximate value for this quantity.

For the circuit shown in Fig. A6 the following relationships apply.

The base current i_b is

$$i_b = \frac{epC}{pCr + 1} \quad \ldots\ldots\ldots\ldots\ldots (10)$$

where $p = d/dt$

The collector current is

$$i_c = i_b \frac{\alpha_0}{1 - \alpha_0} \cdot \frac{1}{((pT/1) - \alpha_0) + 1}$$

$$i_c \sim i_b \frac{\beta_0}{p\beta_0 T + 1} \quad \ldots\ldots\ldots\ldots\ldots (11)$$

where T is $1/\omega_\alpha$ and α_0 is the low frequency value of α. From equations (10) and (11)

$$i_c = \frac{epC}{pCr + 1} \cdot \frac{\beta_0}{p\beta_0 T + 1} \quad \ldots\ldots (12)$$

The Laplace transform of (12), when e is a step function, and for zero initial conditions is

$$i_c = (e/rT) \cdot \frac{1}{(p + (1/Cr))(p + (1/\beta_0 T))}$$

and the solution is

$$i_c = \frac{q\beta_0}{rC - \beta_0 T} \left[e^{-t/Cr} - e^{-t/\beta_0 T}\right] \quad \ldots\ldots\ldots\ldots (13)$$

$$[\alpha_0 \to 1]$$

where $q = eC$

Typical numerical values for the transistor and circuit are

$r = 100\Omega$, $C = 100\text{pF}$, $Cr = 10^{-8}\text{sec}$,

$T = 2{\cdot}5 \times 10^{-8}$, $\beta_o = 70$, $\beta_o T = 1{\cdot}75 \times 10^{-6}\text{sec}$.

The time-constants which govern the exponential terms stand in ratio of 175:1, and so the rise-time (as set by the first exponential) is substantially unaffected by the decay time (set by the second member). It follows that the peak current is virtually uninfluenced by the actual time-constant values and closely approximates

$$i_{c(max)} = q/T \quad \ldots \quad (14)$$

$$[\alpha_o \to 1;\ \beta_o T/Cr \to \infty]$$

The independence of the factors which govern rise- and fall-times, coupled with the fact that the rise-time is set by factors extrinsic to the transistor (r_b' is here so considered), allows equation (13) to be modified for non-ideal step waveforms. Such waveforms occur when one binary stage is driving several others. Then, as seen in Fig. 7, a driven stage receives a degraded step rising in 0·1 to 0·3μsec. Since this rise is slow compared with the time-constant Cr, it will govern the rise of I_o, whose leading edge will then approximate the input voltage waveshape. Little loss of peak value will occur due to the slower drive waveform, since its rise-time is still much less than $\beta_o T$.

The total charge Q available from the collector of X_3 is given by integrating equation (13) and is:—

$$Q = q\,\beta_o \quad \ldots \quad (14)$$

irrespective of the ratio of the two time-constants.

During phase 2 of turn off, X_2 sets its own limit to the current it will accept. The surplus delivered by X_3 then charges the collector capacitances of all three transistors and C_c becomes significant. If the value of C_c is 10pF, then the 10V motion of the collector results in a charge of 100pC, and this is injected as reverse charge to the base of X_3. This charge is relatively small compared with that already existing in X_3; and furthermore, at the time when the full charge of C_c has been injected, it is required to commence the clearing of carriers from X_3. It may be shown by approximate methods that the average collector current during this rise is diminished by about 10 per cent, so that the influence of C_c is merely to slow the voltage rise slightly.

Similarly the influence of R_9 is small. It injects about 0·1mA towards the base of X_3 throughout the turn-off operation. If the time of turn-off is 0·4μsec, a reverse charge of 40pC has accumulated at its termination, and this is negligible.

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Investigation of Magnetic Storms

A radio spectroscope, for investigating magnetic storms, has been built at Dapto near Sydney by the Radiophysics Division of the Industrial Research Organization. The information obtained is being forwarded to the International Geophysical Year centre in the U.S.A.

The background that inspired the creation of the radio spectroscope goes back to World War 2. On 27 February 1942, a source of noise intermittently blacked out all the radar screens in Britain. At sunset the interference vanished, but at sunrise next day it started again.

After examining all reports, Dr. J. S. Hey, of the British Operational Research Group, traced the direction of the jamming to the sun.

During that period an enormous sunspot, with an area greater than one five-hundredth of the hemisphere, was close to the centre of the disk. From midday until after 3 p.m. on 28 February, a tremendously brilliant flare was visible over the sunspot. The post-war publication of Dr. Hey's wartime report stimulated radio astronomers to start an intensive investigation of the sun.

Using sensitive radio receivers with special directional aerials, they followed it across the sky, recording its normal radio emissions and sudden bursts. Analysis of the recordings revealed the peculiarity that when a radio burst from the sun was recorded at two different frequencies. This suggested that the broadcasts from the burst must be coming from different parts of the sun's atmosphere. In other words, different radio frequencies had different penetrating power into the corona.

It seemed obvious that it would be possible to make a systematic exploration of all levels of the corona and to follow the development of a burst right from the depths. Knowledge of the electron population in the corona supported this belief.

At about 300 miles above the sun's surface there are 100 000 000 free electrons per cubic centimetre. Moving away from the surface a steady decrease takes place until, at about half-a-million miles, the number of electrons per cubic centimetre has fallen to 1 000 000.

Thus, higher radio frequencies could probe deeper into the corona before encountering electron densities great enough to stop them.

Paul Wild and L. L. McCready of Sydney, conceived the idea of exploring the sun's corona in depth with a radio spectroscope that could record emissions over a wide range of radio frequencies.

They set up an instrument covering the range from 70Mc/s to 130Mc/s at Penrith, near Sydney. This proved to be so successful that it was decided to build a bigger one covering a wider range.

The Dapto instrument was set up by Wild and J. D. Murray. It covers the range from 40Mc/s to 240Mc/s, and probes the corona from about 20 000 miles to about 200 000 miles above the sun's surface.

A broad-band antenna, consisting of three rhombic aerials in the order 40 to 70Mc/s, 70 to 140Mc/s and 140 to 240Mc/s, picks up signals from the sun and feeds them into a tunable receiver which is swept rapidly, by an electric motor, over the whole band, so that it records the signals received at all frequencies at the rate of twice per second. Each spectrum is displayed on a cathode-ray tube and automatically recorded on 35mm film.

Paul Wild found that there are two characteristic types of solar burst, each of which takes place at the time of a solar flare.

In one of the characteristic types found by Wild, the frequency drift indicates a source moving out through the corona at a speed of about 300 mile/sec.

Bursts of this type are rare. They occur only with very large flares. When they do take place they may cause magnetic storms on the earth one or two days later.

The other type of burst has an outward velocity from the sun of about 50 000 mile/sec—nearly one-third the speed of light. Suitably directed, these fast particles would reach the earth about a quarter of an hour after the flash of the flare. This is just the characteristic time delay between a great flare and an increase of cosmic rays on the earth.

The radio spectroscope records indicate that the entire flare phenomenon and its consequences is the result of a gigantic explosion near the surface of the sun. The causes of these explosions are not yet known, but they may well be natural 'hydrogen bombs' or vast electrical discharges. Whatever the cause, an explosion immediately ejects clouds of fast particles that pour into the corona at a velocity equal to about one-third the speed of light. At this high speed the atoms are stripped of their electrons, thus becoming invisible, and are directed along lines of magnetic force into jetlike streams which reach the outer layers of the corona within seconds, sending signals of their route by brief radio flashes.

When they escape from the sun's influence and reach the earth 15min after the explosion, they cause an increase in cosmic ray intensity.

For some time the radio spectroscope took the whole radiation from the sun. There was no attempt to locate the actual area of origin of the source. But Wild and his colleagues are now locating the centres of activity with an interferometer telescope operating on a single frequency of 100Mc/s.

Equivalent Circuits of Noisy Networks

By Leo Young*, M.A., M.S., A.M.I.E.E. Mem.A.I.E.E.

This article derives and presents new equivalent circuits useful for predicting the noise figure of a combination of amplifiers, attenuators, and terminations, which may all be at different temperatures.

A new method for the precision measurement of noise figure is formulated and described. Experimental results are given.

EQUIVALENT circuits have proved useful in representing complex or new circuits in terms of simpler or more familiar ones. Examples of circuits which are sometimes represented by equivalent circuits include thermionic valve circuits, involving amplification; and waveguide junctions, involving the reduction of distributed to lumped parameters. Similarly it will be shown that calculations involving noisy networks can be facilitated by first writing down the noise equivalent circuit.

Equivalent Circuits

An equivalent circuit is a combination of definable circuit elements consistent with the mathematical formulation. The circuit elements may be passive or active, but do not need to be physically realizable. They are described by parameters which are used to describe actual circuit elements; these parameters may be assigned in the equivalent circuit

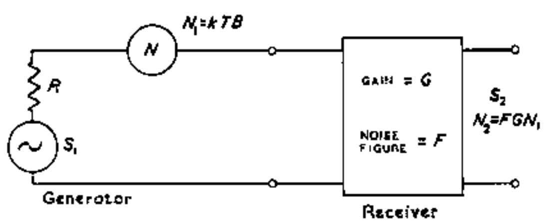

Fig. 1. Circuit representation of a noisy network

to elements which need not have a separate existence in nature, provided that the resulting behaviour is in accord with the equations. This may be stated otherwise by saying that the equivalent circuit postulates a mechanism which is more detailed than the equations warrant. For instance, a battery may be represented by a lossless generator and a resistance in series, though this separation is not feasible in practice.

Equivalent circuits play the role of a model and may be used to visualize the process described by the equations. They cannot, however, convey more information than is contained in the equations.

It is proposed to derive equivalent circuits of a resistor or waveguide termination, an attenuator, and an amplifier, when all have to be treated as sources of noise. It will be assumed that all networks are linear, and that all noise sources are uncorrelated, so that noise powers are additive.

Noise Figure

The noise figure[1,2,3,4] of a linear network, with a generator connected to its terminals, may be defined as the ratio of the available signal-to-noise power ratio at the signal generator terminals (weighted by the network bandwidth) to the available signal-to-noise power ratio at its output terminals. The temperature of the signal generator resistance is 290°K.

A suitable circuit representation is shown in Fig. 1, where

S_1 = available signal power at the terminals of the signal source,

S_2 = available signal power at the output terminals of the receiver,

N_1 = available thermal noise power at the terminals of the input termination at $T = 290$°K, in the receiver bandwidth B,

N_2 = available noise power at the output terminals of the receiver, when the input is thermal noise (N_1) at 290°K,

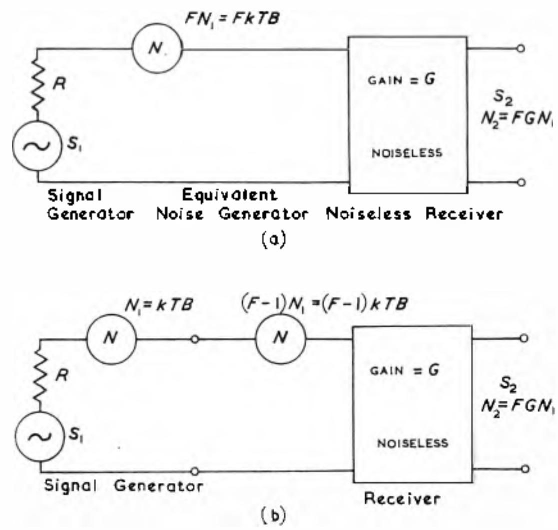

Fig. 2. Equivalent circuits viewed from output terminals

G = maximum available power gain of the receiver with respect to frequency,

B = effective bandwidth[1] of the receiver,

k = Boltzmann's constant = $1{\cdot}38 \times 10^{-23}$ J/°K/cycle.

Since

$$S_2 = GS_1 \qquad (1)$$

and

$$F = \frac{S_1/N_1}{S_2/N_2} \qquad (2)$$

$$= N_2/GN_1$$

therefore

$$N_2 = GFN_1 \qquad (3)$$

from which the equivalent circuit shown in Fig. 2 can be drawn. This equivalent circuit represents the actual circuit when viewed from the output terminals on the right.

An alternative definition[1,5] of noise figure can now be formulated:

The noise figure of a network may be defined as the ratio of the noise power output of that network to the noise power output which would exist if the network were noiseless. The temperature of the signal generator resistance is 290°K.

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This second definition is the more convenient for mathematical manipulation

The equivalent circuit of Fig. 2 applies to both an amplifying and an attenuating network. For an amplifier it cannot be further reduced; for an attenuator at room temperature the noise figure F is equal to its attenuation X, while for an attenuator having a noise temperature t, F is a function of X and t. In this case also another useful equivalent circuit will be derived.

Noise Figure and Equivalent Circuit of a Matched Attenuator

Let X (times) = attenuation of attenuator, where $X \geqslant 1$
t_x = noise temperature of attenuator.

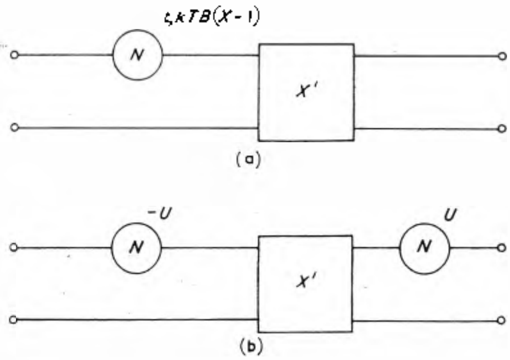


Fig. 3. Equivalent circuits to the right of an attenuator

Assume that the attenuator is placed in front of a matched termination also at a noise temperature t_x, and suppose that:

(a) The attenuator attenuates the available noise power from the matched load to the receiver, reducing it by a factor X.

(b) The attenuator contributes available noise power, P say, which is a function of X and t_x.

(c) The total available noise power appearing at the output terminals is independent of X for a matched attenuator and a matched load.

This gives

$$t_x(kTB/X) + P = t_x kTB \quad \dots\dots\dots (4)$$

Therefore

$$P = t_x \frac{kTB\,(X-1)}{X} \quad \dots\dots\dots (5)$$

This noise output would also result from a noise source of available noise power $t_xkTB(X-1)$ placed before the attenuator, as shown in Fig. 3(a), in which the symbol X' represents an attenuator in all respects similar to X except that it is noiseless (it may be thought of as being at a temperature of absolute zero). Comparing Fig. 3(a) with Fig. 2(b), it is seen that

$$F - 1 = t_x(X - 1)$$

or

$$F = 1 - t_x + Xt_x \quad \dots\dots\dots (6)$$

If $t_x = 1$ this reduces to $F = X$.

The equivalent circuit of Fig. 3(b) is also consistent with equation (5), which may be written

$$P = U - (U/X) \quad \dots\dots\dots (7)$$

where

$$U = t_xkTB \quad \dots\dots\dots (8)$$

The noise source $(-U)$ will be called a virtual or image source, by analogy with optics and electrostatics, since the noise situation to the right of an imaginary boundary between $(-U)$ and X' can be described by this equivalent circuit.

Application of Method

It is possible to rapidly calculate the noise figure of a receiving system by drawing its noise equivalent circuit. Suppose that a radio receiver has a noise figure F, that its antenna is matched into space and has a noise temperature t_a, and that the cable between them has attenuation X (times) and is at a noise temperature t_x. Such a situation is typical of radar receivers, where noise figure is an important consideration. Then the equivalent circuit is as shown in Fig. 4. The noise output from the whole receiving system is

$$\left\{\frac{(t_a - t_x)}{X} + t_x + F - 1\right\} GkTB \quad \dots\dots\dots (9)$$

The noise output without any noise sources except one of kTB at the input (into the antenna) would be

$$(G/X)\,kTB \quad \dots\dots\dots (10)$$

The noise figure F' of the entire receiving system is therefore given by the second definition of noise figure as the ratio of equation (9) to equation (10).

$$F' = t_a - t_x + X(t_x + F - 1) \quad \dots\dots\dots (11)$$

Measurement of Noise Figure

One of the most accurate methods of measuring noise figure is summarized in Fig. 5. The notation for noise figure F, attenuation X, and attenuator noise temperature t_x is as before.

Fig. 4. Equivalent circuit of a radio receiver

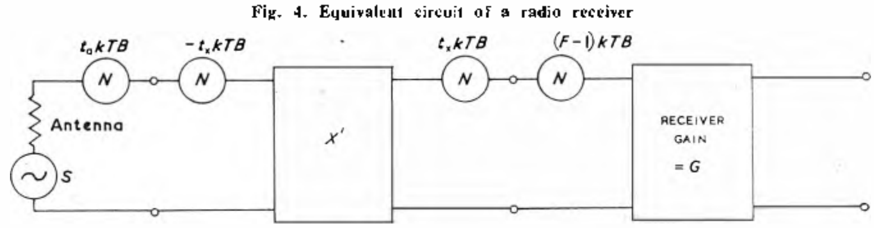
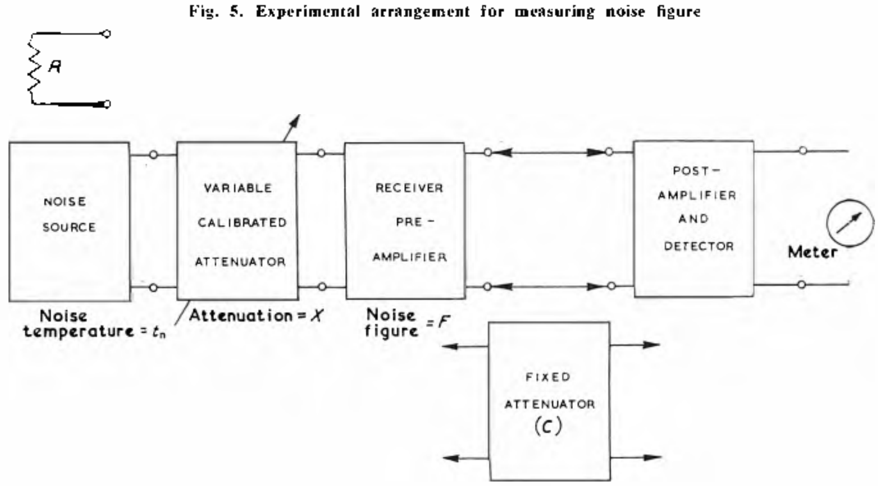


Fig. 5. Experimental arrangement for measuring noise figure



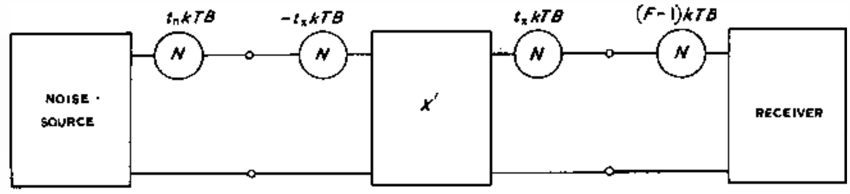


Fig. 6. Equivalent circuit for measuring noise figure

Let t_n = noise temperature of known noise source (a secondary standard),

C (times) = attenuation of fixed attenuator between the pre-amplifier of the receiver and the post-amplifier, (where $C \geqslant 1$).

A particularly simple measurement results if $t_x = 1$ and when $C = 2$. In this case, first take the fixed attenuator out of the circuit connecting pre-amplifier directly to post-amplifier; also connect the termination R instead of the noise source. R is made equal to the output impedance of the signal generator to be used, and is also the output impedance of the noise source. Set the variable attenuator to zero attenuation, $X = 1$, and note the meter reading. Then connect the noise source in place of R, and also switch the fixed attenuator into the circuit. Adjust X until the meter reading is as before. Then the excess noise $(t_n - 1)kTB/X$ must equal the receiver noise (assuming the noise contribution after the receiver pre-amplifier to be negligible).

$$FN_1 = FkTB = (t_n - 1)kTB/X$$

Therefore

$$F = \frac{t_n - 1}{X} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (12)$$

To work out the general case, when t_x is not unity and C is not equal to 2, resort is had to the equivalent circuit, which for this case is shown in Fig. 6. It is not necessary even to know the value of C. It will be shown that this method not only measures noise figure F, but also calibrates (or checks) the fixed attenuator C.

The noise source is connected to the variable calibrated attenuator followed by the receiver as shown in Fig. 5. The fixed attenuator is first switched out of the circuit as shown in the figure, and the variable attenuator setting is made $X = X_1$, say. The meter reading is noted. Next, the fixed attenuator is switched into the circuit, and some of the attenuation X taken out to bring the meter reading up to its previous value. Now $X = X_2$, say. Then since the two meter readings are the same,

$$\left(\frac{t_n - t_x}{X_1} + t_x - F - 1\right)kTB = \left(\frac{t_n - t_x}{X_2} + t_x + F - 1\right)(1/C)kTB \quad \ldots\ldots\ldots \quad (13)$$

Multiplying by $1/(t_n - t_x)$ and rearranging,

$$1/X_2 = (C/X_1) + \frac{(t_x + F - 1)(C - 1)}{t_n - t_x} \quad \ldots\ldots \quad (14)$$

which is of the linear form

$$v = Cu + b \quad \ldots\ldots\ldots\ldots\ldots \quad (15)$$

where

$$v = \frac{t_n - t_x}{X_2} \quad \ldots\ldots\ldots\ldots\ldots \quad (16)$$

and

$$u = \frac{t_n - t_x}{X_1} \quad \ldots\ldots\ldots\ldots\ldots \quad (17)$$

Therefore equation (14) or (15) is a straight line in the (u, v) plane having a slope C and an intercept on the v – axis of $b = (t_x + F - 1)(C - 1)$

When $v = u$, or $1/X_2 = 1/X_1$, $\quad \ldots\ldots\ldots \quad (18)$

then $u = \dfrac{t_n - t_x}{X_1} = -(t_x + F - 1)$

Several measurements, starting with various X_1, each yields a different X_2, and plotting $v = (t_n - t_x)/X_2$ against $u = (t_n - t_x)/X_1$ yields several points in the u, v plane. The best straight line joining them determines not only F but also C, when t_n and t_x are known.

An actual experimental plot is shown in Fig. 7.

These results were taken from measurements on a waveguide mixer and i.f. pre-amplifier combination, followed by a fixed i.f. attenuator of nominally 3dB ($C = 2$) which could be switched in and out of circuit. t_x was taken to be

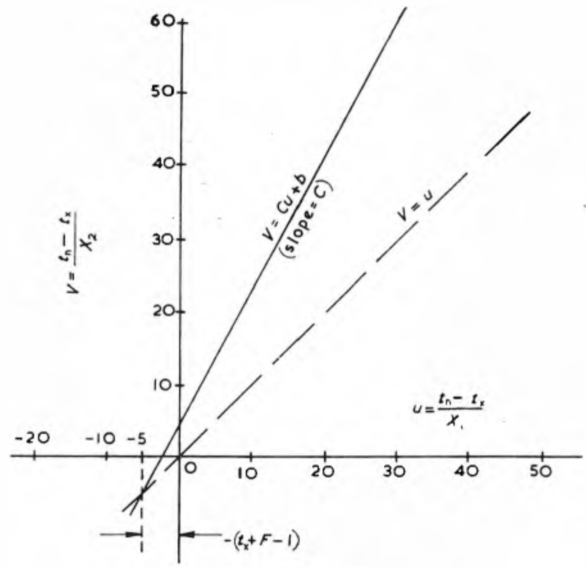


Fig. 7. Derivation of noise figure F and attenuation C

unity. A mercury gas discharge tube[6] was the noise source. The points lay along a straight line within experimental error. It was found that $C = 1{\cdot}86$ instead of 2 from the slope of the line. This results in a correction of 0·7dB to the noise figure, as compared to the first simple measurement procedure described, which assumed $C = 2$. A subsequent check on the attenuator made at i.f. confirmed $C = 1{\cdot}86$ very closely. For the receiver which gave Fig. 7, the noise figure F is 5, (or 7dB).

Conclusion

Several noise equivalent circuits for generators, amplifiers and attenuators have been derived. It has been shown how they may be applied to the calculation of noise figure for a given circuit; and how they may be used to determine an accurate experimental method for measuring noise figure.

Acknowledgments

The author wishes to thank Dr. William Altar and Mr. Henry De Francesco for some interesting discussions on this subject. This article was originally presented as a Conference paper at the A.I.E.E. Fall General Meeting in Chicago in October 1957.

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A Tunable Filter for Use in the Measurement of Excess Noise from Local Oscillators

By W. P. N. Court*, B.Sc., A.C.G.I., A.M.I.E.E.

A method of forming a filter from available microwave components is described. This filter is tunable over a wide-band and need not be removed physically from the circuit when filter action is not required.

WHEN a measurement of excess noise is made, it is customary to compare the noise observed, at a receiver output, when the input mixer is fed with a local oscillator under two conditions.

The noise is first removed from the oscillator by inserting a filter between it and the mixer. The filter is then removed and the change in noise output noted.

This filter usually consists of a fixed tuned band-pass filter or an H_{01} transmission cavity. In both cases physical insertion and removal of a filter is necessary.

This note is concerned with making use of a hybrid junction, together with an H_{013} cavity and a variable short-circuit, which are more generally available than either of the above devices.

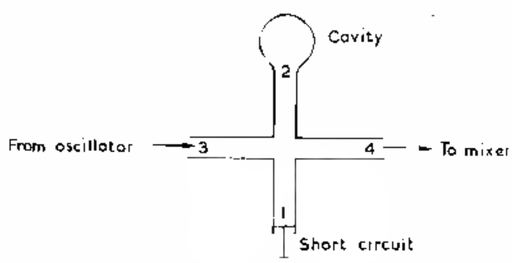


Fig. 1. Arrangement of filter components

The great advantage, however, of the proposed method lies in the fact that the physical removal of the filter becomes unnecessary. The change of filter condition is effected by simple tuning adjustments.

Method

Circuit Arrangement

Fig. 1 shows a magic-T junction (a typical hybrid junction) to one side-arm of which the H_{013} cavity is connected, while a variable position short-circuit is attached to the other.

Noiseless or Filter-in Condition

With the cavity off-tune, and hence appearing as a short-circuit at its coupling iris, the variable short-circuit is adjusted so that no output obtains from the junction.

The cavity is then tuned to the incoming frequency, thus disturbing the bridge-balance around its resonant frequency. The output is then maximal at this frequency and falls on either side of it at a rate dependent on the Q-factor of the cavity.

Since Q-factors of several thousand are readily obtainable over a wide-band, a narrow band-pass filter results.

Noisy or Filter-out Condition

For this condition the cavity is detuned and the variable short-circuit readjusted to give maximal output from the junction.

Insertion Loss and Bandwidth

It is shown in Appendix A that, for optimum performance, the cavity impedance at resonance should be matched to the waveguide.

Then, if the filter is connected between matched input and output impedances, the insertion loss for the filter-in condition is 6dB, while the bandwidth is that appropriate to a Q-factor half that for the unloaded cavity.

Similarly, for the filter-out condition, the insertion loss is shown to be zero.

Use as F.M.-A.M. Convertor

The device may also be used as a microwave f.m.-a.m. convertor, by suitably detuning the cavity so as to operate on the slope of the response curve.

The conditions for optimum conversion are given in Appendix B. It is shown that, using a matched cavity, an amplitude modulation percentage of $47Q_o/f_o$ per cycle deviation is obtained.

Conclusions

A filter has been made as described above and has been found to conform to theory.

The device has proved easier to use than a fixed filter and is valuable where measurements are to be made over a band of frequencies.

Acknowledgments

Acknowledgments are due to Mr. B. Davy, who assisted in the measurements which caused this development to take place.

Appendix A

Equivalent Circuit of Cavity

At the plane containing the coupling iris, the cavity may be represented by a simple parallel-tuned circuit, in the neighbourhood of a resonance. This is shown in Fig. 2, where Z_o is the wave-impedance of the associated waveguide, and n is a measure of the coupling. A matched cavity obtains when $n = 1$, and over- or under-coupling occurs when n is greater or less than unity.

The impedance Z_c, at frequency f, at this plane is therefore,

$$Z_c = \frac{1}{[(1/nZ_o) + j(\omega C - (1/\omega L))]} = \frac{nZ_o}{[1 + j\,Q_o(\omega/\omega_o - (\omega_o/\omega))]}$$

where $Q_o = (nZ_o/\omega_o L)$, $\omega = 2\pi f$, $\omega_o = (1/\sqrt{(LC)}) = 2\pi f_o$

$$\therefore Z_c \approx \frac{nZ_o}{[1 + j\,(2Q_o/f_o)\,.\,F]} = \frac{nZ_o}{(1 + j\,KF)} \quad \ldots\ldots\ldots\ldots (1)$$

* *Royal Radar Establishment.*

where, $F = (f \sim f_o)$, $K = (2Q_o/f_o)$, and $f_o \gg 0$.
For sufficiently large values of F, $Z_c \to 0$.

Filter-in Condition

In this case the cavity-iris and the short-circuit are equally distant, electrically, from the axis of symmetry of the T-junction.

If the source and load impedances of the T-junction are each equal to Z_o, an equivalent circuit may be drawn as in Fig. 3.

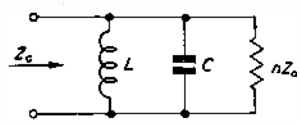


Fig. 2. Equivalent circuit of cavity at coupling iris

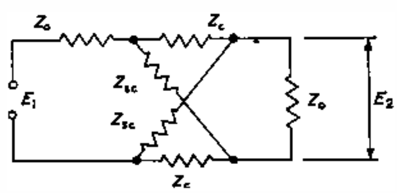


Fig. 3. Equivalent circuit of filter

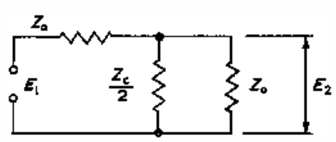


Fig. 4. Equivalent circuit of filter $Z_{sc} = 0$

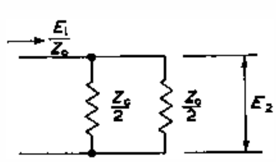


Fig. 5. Constant current equivalent of Fig. 4

In the present case $Z_{sc} = 0$ so that the circuit may be redrawn as in Fig. 4. This may be redrawn as in Fig. 5, from which it can be seen that

$$E_2/E_1 = \frac{Z_o Z_c}{2(Z_o + Z_c)} \cdot 1/Z_o$$

$$= \frac{1}{2(1 + (Z_o/Z_c))} \quad \text{.............} \quad (2)$$

$\therefore$ from equations (1) and (2):

$$E_2/E_1 = G = \frac{1}{2[1 + (1 + jKF)/n]}$$

and,

$$|G| = \frac{n}{2[(1+n)^2 + K^2F^2]^{\frac{1}{2}}} \quad \text{......} \quad (3)$$

$$\therefore \text{ at } F = 0,\ |G_o| = \frac{n}{2(1+n)} \quad \text{..........} \quad (4)$$

$$\therefore |G_o|/|G| = \left[1 + \frac{K^2F^2}{(1+n)^2}\right]^{\frac{1}{2}} \quad \text{..........} \quad (5)$$

The response is 3dB down when $|G_o|/|G_B| = \sqrt{2}$ and $(KF)_B = (1 + n)$.

Hence, bandwidth,

$$B = \frac{2(1+n)}{K} \quad \text{................} \quad (6)$$

$\therefore$ from equations (4) and (6):

$$|G_o|/B = \frac{Kn}{4(1+n)^2} \quad \text{..............} \quad (7)$$

This has a maximal value when $n = 1$. Hence the system is most effective as a filter when the cavity is matched to the guide.

Thus, with a matched cavity,

from equation (3),

$$|G|_{opt} = \frac{1}{2(4 + K^2F^2)^{\frac{1}{2}}} \quad \text{............} \quad (8)$$

and from equation (6),

$$B_{opt} = (4/K) = (2f_o/Q_o)$$

Hence the effective $Q_{opt} = (f_o/B) = (Q_o/2)$ .... (9)

At $F = 0$, $|G_o|_{opt} = \frac{1}{4}$, which represents an insertion loss of 6dB.

Filter-out Condition

If the cavity is now detuned and the short-circuit moved through $\lambda/4$, $Z_c = 0$ and $Z_{sc} = \infty$. It is then clear from Fig. 3 that this represents zero insertion loss and hence maximal output.

Calculation of Necessary Q_o

From equation (5), assuming $n = 1$,

$$|G_o|/|G| = (1 + (K^2F^2/4))^{\frac{1}{2}}$$

Suppose it is desired to reduce the noise power at $F = F_1$, where F_1 is the i.f., to one per cent.

$$\therefore (K^2F_1^2/4) = (Q_o^2F_1^2/f_o^2) \simeq 100$$

$$\therefore Q_o = 10(f_o/F_1) \quad \text{................} \quad (10)$$

Appendix B

Use as an F.M.-A.M. Convertor

A measure of the f.m.-a.m. conversion obtainable is derived from the slope of the curve of $|G|$ plotted against F. This slope may again be shown to have a maximal value when $n = 1$.

$$\therefore \text{ from equation (8) } (d|G|/dF) = S = \frac{-K^2F}{2(4 + K^2F^2)^{3/2}}$$

This may be shown to have a maximal value when $(KF)_{opt} = 2^{\frac{1}{2}}$.

$\therefore S_{max} = K/(2^{\frac{1}{2}} \cdot 6^{3/2})$ ignoring the negative sign. .. (11)

$$\text{Now, } S = (d|G|/dF) = \frac{d|E_2/E_1|}{dF}$$

$$\therefore (d|E_2|/dF) = |E_1| \cdot S$$

but, $|E_2| = |E_1|\ |G|$

$$\therefore \frac{d|E_2|/dF}{|E_2|} = (S/|G|)$$

$$= \frac{K}{2^{\frac{1}{2}} \cdot 6^{3/2}} \cdot \frac{2 \cdot 6^{\frac{1}{2}}}{1}, \text{ for optimum case}$$

$$= \frac{K}{3 \cdot 2^{\frac{1}{2}}} = 0{\cdot}235K\,\frac{\text{Volts/c/s}}{\text{Volt}} \quad \text{.....} \quad (12)$$

Hence the resulting percentage amplitude modulation of the output is 23·5K per cycle deviation.

LETTERS TO THE EDITOR

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

A Voltage Stabilizer Principle

DEAR SIR,—An interesting article written under the above heading by C. Billington and E. Chakanovskis and published in the August issue of your journal deals with a problem I have been occupied with for several years. I have been using this principle for various stabilization purposes since 1953, and I have come to some general conclusions which your readers may also find interesting.

The stabilizer of the above type can generally be represented by the circuit[1] shown in Fig. 1 and consisting of a main degenerative feedback loop modified by an auxiliary loop, which is regenerative and provides a gain approaching unity.

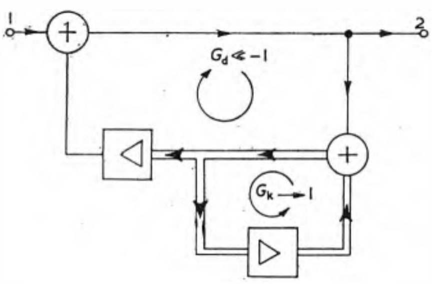


Fig. 1. Principle of the stabilizer

The transfer to the output point of a disturbing signal entering the single-line drawn region in an arbitrary point, is given by

$$G' = G\frac{1 - G_k - G_d}{1 - G_k} \quad \ldots\ldots\ldots\ldots (1)$$

where G is the transfer between the same points with the main loop disconnected.

The internal output resistance changes from its original value R_{i_2}, pertaining to a disconnected main loop, to

$$R_{i_2}' = R_{i_2}\frac{1 - G_k}{1 - G_k - G_d} \quad \ldots\ldots\ldots\ldots (2)$$

Both these values become zero for a perfectly compensated circuit, i.e. for $G_k = 1$, and their signs change to negative if $G_k > 1$.

The main loop keeps the entire circuit stable, as long as $G_k < 1 - G_d$ and the simple conditions governing the frequency response of the loops are fulfilled. These conditions may be derived simply from equation (1), presuming the transfer functions are of the simplest form as given by equation (3). For the case of complicated transfer functions the general derivation, however, is more difficult, but the possibility of securing stable circuit operation by simple means had been verified many times experimentally.

The specific qualities of the circuit described are as follows:

(1) Extreme stabilization (regulation) can be achieved with a moderate main loop gain[2,3].

(2) Values G' and R'_{i_2} may be varied over a considerable range on both sides from zero by a simple continuous control.
Negative output resistance can be reached (current-controlled type).

(3) A substantial part of the circuit—drawn by a single line in Fig. 1—is fully insensitive to disturbing signals[1].

(4) The circuit may be converted to a special d.c. amplifier, the gain of which is determined solely by the transmission properties of a certain resistance network and the output resistance equals zero[1].

Note: when correctly designed and properly placed in the main loop, the regenerative part of the circuit remains exactly adjusted even in the presence of a very large disturbing signal[4]. A voltage stabilizer was built delivering an output voltage of 200V constant within 0·0015 per cent with input voltage variations exceeding the limits 260 to 580V.

The operation of the circuit described gives an impression of paradoxy at first glance: when the stationary output deviation with an exactly compensated circuit becomes zero, the regulation signal seems to disappear and the regulator proper is no more called upon to intervene.

The compensating loop incorporates always at least one RC delay network having a transfer function of the operational form[1]

$$G(p) = \frac{1}{1 + p\tau} \quad \ldots\ldots\ldots\ldots\ldots (3)$$

where p is the Laplacean complex operator and τ is the time-constant of the delay network. The operational transfer of a signal passing through the portion of the compensating loop becomes

$$G'(p) = G\frac{1}{1 - G_k/(1 + p\tau)}$$

$$= G\frac{1 + p\tau}{1 - G_k + p\tau} \quad \ldots\ldots\ldots\ldots\ldots (4)$$

and for perfect compensation, i.e., for transfer constant adjusted to $G_k = 1$,

$$G'(p) = G(1 + (1/p\tau)) \quad \ldots\ldots (5)$$

Accordingly, this part of the circuit transforms the incoming signal to a value proportional to the sum of the original value and of its time integral. When the output deviation value transformed in this way is applied with suitable polarity to the control grid of the regulator tube, the regulator intervention will augment as long as the time integral of the output deviation increases, i.e., as long as the deviation becomes not zero again. The apparently paradoxical behaviour of the circuit described can be explained in this way.

Similar results will be obtained even with a more complicated transfer function of the compensating loop than that given by equation (3), presuming the circuit is still allowed to operate regularly. A more complete discussion of this subject is given in reference 1.

Obviously, this circuit is analogical to a first order servo-system, controlled by an error voltage and by its time integral. The perfect integrating effect which is accomplished by the regenerative feedback used in the described manner is not limited to electronic circuits only.

The present circuit was developed originally for a 2kV d.c. stabilizer, but later on it has been utilized successfully in solving various stabilization problems.

Sincerely yours,

M. PACAK,

Institute for Physical Chemistry of the Czechoslovak Academy of Science.

REFERENCES

1. PACAK, M. Dynamic Properties of a Degenerator. *Slaboproudy Obzor.* *18*, No. 9 (Sept., 1957).
2. MACKENZIE, R. B. A New D.C. Electronic D.C. Voltage Stabilizing Circuit. *Proc. Instn. Elect. Engrs.* *101*, Pt. 2, 59 (1954).
3. PACAK, M. Feedback Voltage-Stabilizer with the Possibility of Overcompensation. *Slaboproudy Obzor.* *16*, No. 2 (Feb., 1955).
4. PACAK, M. A Generalized Degenerative Stabilizer with a Possibility of Overcompensation. *Slaboproudy Obzor.* *16*, No. 2, 562 (1955).

The authors' reply:

DEAR SIR,—We were most interested to be informed of Mr. M. Pacak's work in the field of voltage stabilization. The Czechoslovakian journal which has published it has only recently become available in this country, and we regret that neither the journal nor abstracts from it were known to us at the time of our publication. The most recent of Mr. Pacak's articles (September 1957) has not yet been received here.

The first suggestion of using positive feedback in stabilizer circuits appears to have been made by Valley and Wallman[1] in 1948; and no doubt many workers have since used the idea for their own immediate needs. Our purpose in publishing was to record a fully practical development of this principle, embodying all the major requirements of a stabilizer with reasonable economy. Mr. Pacak has proceeded on slightly different lines and has produced a more specialized arrangement. While we would make several points in discussion, we readily concede that they may well not be new to Mr. Pacak.

Single stage amplifier

The first quoted of Mr. Pacak's articles (February 1955) uses an arrangement which is virtually identical to our Fig. 1, and obtains a comparable performance.

Two stage amplifier

The circuits of Mr. Pacak's second article (November 1955 Figs. 7, and 8, reproduced here as Figs. A and B) are comparable with the second circuit of our article (Fig. 2) since all these use one stage of amplification without positive feedback, and one stage (using two valves) with positive feedback. We must concede that Mr. Pacak's arrangements are superior to ours in that they do not raise the average potential of the signal line. Mr. Pacak demonstrates experimentally the superiority of using positive feedback before an ordinary amplifying stage (his Fig. 8) rather than afterwards (his Fig. 7, and our Fig. 2). This is because the dynamic range at input is much smaller than at output, and as we suggested at the conclusion of our article, this seems to be an essential condition for good operation of positive feedback.

As published, we feel there are three technical disadvantages to Mr. Pacak's circuits, viz:—

(1) They are as sensitive to heater voltage variations as are the prototype circuits by Hunt and Hickman[2].

(2) It would appear that their frequency response is very limited, because of severe capacitive shunting of the main amplifier.

(3) They do not appear to be capable of significant variation of output voltage without upsetting the positive feedback loop adjustment.

Criterion of Performance

We endeavoured to show that the real criterion for a stabilizer amplifier is the output dynamic range over which very high gain can be maintained; and that the function of the series valve should be assessed separately. This criterion is even more suitable for Mr. Pacak's wholly degenerative circuits than for ours which require the use of a cathode-follower whose anode supply has to come from the unstabilized side of the series valve. With 100V stabilized supply we obtained a 25V output swing which could be applied to the series valve grid. Mr. Pacak uses a 200V stabilized supply and a series valve with $\mu \sim 4$ so that the dynamic ranges of his input-compensated and output-compensated circuits are about 85V and 35V respectively.

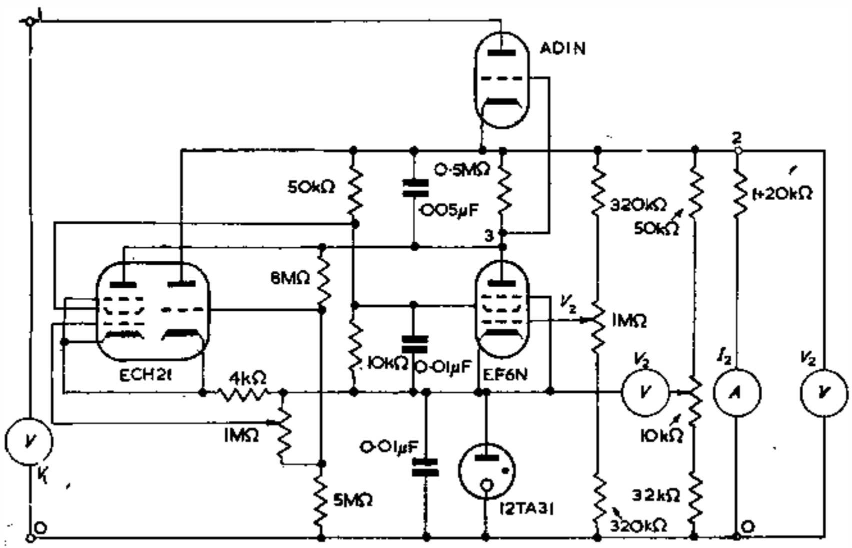
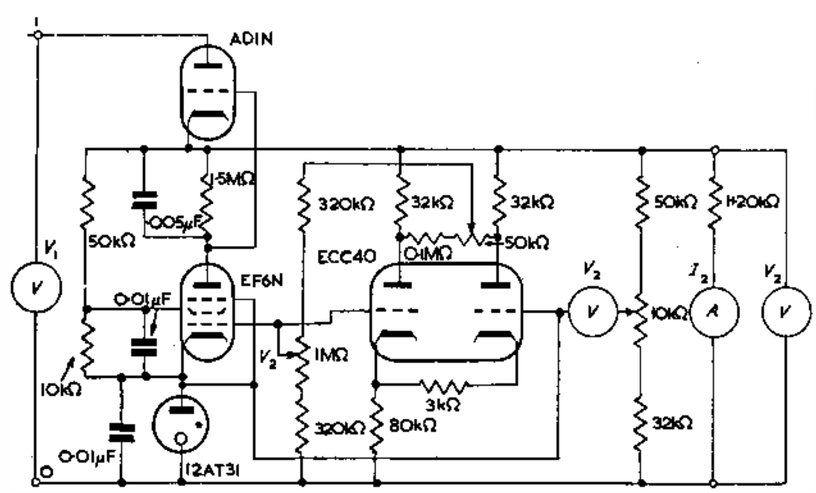


Fig. A. (above) M. Pacak's Fig. 7

Fig. B. (below) M. Pacak's Fig. 8



The other factors

It has been our experience that genuine stabilities better than one part in 10^4 are very hard to achieve without resorting to chopper techniques. Even though changes in unregulated input voltage may be rejected to a greater degree than this, the temperature of the reference valve has only to change by about 3° C, or carbon resistors differentially by 0·5 per cent to give a change of $1:10^4$. Again, even with carefully stabilized heater power, cathode conditions soon wander by this amount unless the whole equipment is in a constant temperature enclosure. We consider this complication to be unwarrantable within our terms of reference. We have already drawn attention to the importance of maintaining a good frequency response at least well into the audio range.

Mathematics

With regard to the mathematical treatment presented by Mr. Pacak, and left out of account by ourselves, we note that he appears to make no attempt to relate his mathematical findings with the performance of his experimental stabilizers. It is our view that the textbook equations that can be set up for these systems are derived from first-order physical considerations only, and that the arrangement is such that the first-order interactions are largely balanced. Thus the success of the treatment will be limited by the degree of appreciation of the higher-order interactions.

Yours faithfully,

C. BILLINGTON,
E. CHAKANOVSKIS,
Chemical Physics Section,
Division of Industrial Chemistry,
Commonwealth Scientific and Industrial Research Organization.
Melbourne, Australia.

REFERENCES

1. VALLEY, G. E., WALLMAN, H. Vacuum Tube Amplifiers. M.I.T. Radiation Laboratory Series, 18 (McGraw-Hill, 1948).
2. HUNT, F. V., HICKMAN, R. W. Electronic Voltage Stabilizers. *Rev. Sci. Instrum.* *10*, 6 (1939).

The correspondent replies:

DEAR SIR,—Mr. Billington and Mr. Chakanovskis have paid little attention to my first letter and my paper[1], and turned their attention to the circuits published in my early works[3,4]. These circuits were given as illustration of the principle rather than as elaborated stabilizers. Therefore, I would like to answer briefly their discussion points.

My early circuits were improved later in some respects, and they fulfilled their tasks well: The sensitivity to heating variations was substantially reduced by utilizing a symmetrical positive loop stage, and can be cancelled by well-known methods[5,6]. The unsymmetry of the main amplifier stage is practically irrelevant, see point three in my former letter. The frequency response, reduced by unfavourable location of a stabilizing capacitor in my early circuits, can easily be expanded over the desired range within the limits given by wideband technique. As my first instruments were utilized for constant current purposes only (electron-optic devices and electromagnets), the output impedance rising with frequency caused no harm. Adjustability of voltage was not required originally. Later, a circuit was developed permitting it over the range of several hundred volts without disturbing the positive loop adjustment.

Criterion of performance, I believe, is given only by the dynamic range of the positive loop stage, presuming the gain in the main loop does not sink too much near zero. This can be deduced simply from the general conception of these circuits, and was demonstrated experimentally[4].

The other factors, such as component and reference valve instability and noise, reduce the performance of every stabilizer. But that fact does not limit the value of a circuit enabling one to cancel all the main (short-time) disturbing effects: the line voltage variations, the rectifier ripple, the output current variations by simple adjustment and in wide range.

The mathematical treatment of associated problems, given[1] and quoted in my first letter, though based on linear relation, was of great assistance to me during different realizations. I have, of course, no objections, if Mr. Billington and Mr. Chakanovskis prefer another way.

Yours sincerely,

M. PACAK,
Institute for Physical Chemistry of the Czechoslovak Academy of Science.

REFERENCES

1. PACAK, M. Dynamic Properties of a Degenerator. *Slaboproudy Obzor. 18*, No. 9 (Sept., 1957).
2. MACKENZIE, R. B. A New D.C. Electronic D.C. Voltage Stabilizing Circuit. *Proc. Instn. Elect. Engrs. 2*, 101, 59 (1954).
3. PACAK, M. Feedback Voltage-Stabilizer with the Possibility of Overcompensation. *Slaboproudy Obzor. 16*, No. 2 (Feb., 1955).
4. PACAK, M. A Generalized Degenerative Stabilizer with a Possibility of Overcompensation. *Slaboproudy Obzor. 16*, No. 2, 562 (1955).
5. MILLER, S. E. Electronics (Nov., 1941).
6. PACAK, M. A Balancing Heater Circuit for Improving the Stability of Direct-Current Amplifiers. *Slaboproudy Obzor. 15*, No. 12, 558 (1954).

"Dekatron" Tubes

DEAR SIR,—Ericssons have drawn our attention to the fact that the 10 way selector "Dekatron" tubes used in the versatile Pulse Pattern Generator described in the January issue (p. 39), were in fact experimental prototypes, and that in any future equipments to be built using the production type GS10D they should be run under the conditions recommended by Messrs. Ericsson in their Handbook "Cold Cathode Tubes."

Yours faithfully,

P. H. CUTLER,
L. R. PETERS,
Ministry of Supply
Signals Research and Development Establishment.

Filter Type Networks

DEAR SIR,—The recent publication of articles[1,2,3], giving design information for filter type networks using Insertion Parameter Methods and tabulating the element values of such networks against their insertion parameters enables time to be spent on the more sophisticated aspects of design. One such aspect is that of using a small number of components in the most efficient manner. The following design enables one inductor and four capacitors to be connected as either a high pass or a low pass filter with only one capacitor idle in each connexion. The resulting filters, although simple, have a useful insertion loss characteristic.

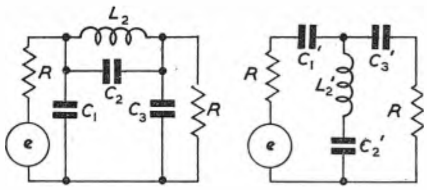


Fig. 1. The filter networks

The configurations, insertion loss characteristics, and relation of element values and design parameters are shown in Figs. 1, 2 and equations (1) and (2).

$$\left.\begin{aligned} C_1 = C_3 &= a_1 \cdot \frac{1}{2\pi f_1 R} \\ C_2 &= \frac{\Omega_{01}^2}{a_2} \cdot \frac{1}{2\pi f_1 R} \\ L_2 &= a_2 \cdot \frac{2\pi f_1}{R} \end{aligned}\right\} \quad \ldots(1)$$

$$\left.\begin{aligned} C' = C_3' &= \frac{1}{a_1} \cdot \frac{1}{2\pi f_2 R} \\ C_2' &= \frac{a_2}{\Omega_{01}^2} \cdot \frac{1}{2\pi f_2 R} \\ L_2' &= \frac{1}{a_2} \cdot \frac{R}{2\pi f_2 R} \end{aligned}\right\} \quad \ldots(2)$$

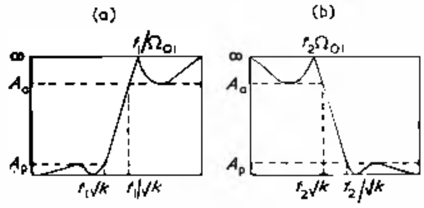


Fig. 2. The filter characteristics

Taking the same terminal impedances for the networks, and considering the design equations (1) and (2) one can see that to make $L_2 = L_2'$ one must choose f_1 and f_2 so that.

$$\frac{a_2}{f_1} = \frac{1}{a_2 f_2}$$

or $f_1 = a_2 f_0 \quad f_2 = f_0/a_2$
then simultaneously one can make
$C_1 = C_2'$, and $C_2 = C_1'$, if it is possible to make $\dfrac{a_1}{a_2} = \dfrac{a_2^2}{\Omega_{01}^2}$

If a particular value of A_p is taken, this relation can be satisfied by a corresponding unique value of k.

Using the design data of Skwirzinski and Zdunek[3] a graphical solution of the equation $a_1\Omega_{01}^2 = a_2^3$ gives the following parameters producing the desired filters.

A_p	=	3	1	0·5	0·2	0·1	dB
A_A	=	51	28	21	15·5	11·0	dB
K	=	0·347	0·597	0·675	0·714	0·717	
Ω_{01}	=	0·516	0·686	0·736	0·761	0·764	
a_1	=	5·48	2·285	1·570	1·07	0·835	
a_2	=	1·135	1·025	0·950	0·853	0·790	

As a particular example an inductor of 114·6mH was available for use. The case A_p = 1dB A_A = 28dB was considered the most useful, and a terminal impedance of 1Ω was chosen.

from the table; $k = 0{\cdot}597$, $\Omega_{01} = 0{\cdot}686$, $a_1 = 2{\cdot}285$, $a_2 = 1{\cdot}025$ for the low pass filter, $f_1 = \dfrac{1000 \times 1{\cdot}025}{2\pi \times 0{\cdot}1146} = 1422\text{c/s}$

$$C_1 = \frac{2{\cdot}285}{2 \times 1422 \times 1000} = 0{\cdot}2655\mu\text{F}$$

$$C_2 = \frac{(0{\cdot}686)^2}{1{\cdot}025 \times 2\pi \times 1422 \times 1000} = 0{\cdot}0513\mu\text{F}$$

for the high pass filter

$$f_2 = \frac{1000}{1{\cdot}025 \times 2\pi \times 0{\cdot}1146} = 1354\text{c/s}$$

$$C_1' = \frac{1}{2{\cdot}285 \times 2\pi \times 1354 \times 1000} = 0{\cdot}0514\mu\text{F}$$

$$C_2' = \frac{1{\cdot}025}{(0.686)^2 \times 2\pi 1354 \times 1000} = 0{\cdot}2655\mu\text{F}$$

These filters will have insertion losses as follows

Low Pass

$\leqslant$1dB below $1422 \,.\, \sqrt{k}$ = 1098c/s
$\geqslant$28dB above $1422/\sqrt{k}$ = 1840c/s
∞ at $1422/\Omega_{01}$ = 2072c/s

High Pass

∞ at $1354 \,.\, \Omega_{01}$ = 930c/s
$\geqslant$28dB below $1354 \,.\, \sqrt{k}$ = 1054c/s
$\leqslant$1dB above $1354/\sqrt{k}$ = 1753c/s

Filters constructed to the above design gave almost perfect agreement with the predicted responses.

The filters can be used in noise and distortion measurements giving a separation of at least 27dB between the wanted and unwanted bands of 0 to 1040c/s and 1840 to ∞c/s. In particular the high pass filter was found useful in checking the distortion of a tape-recording system at 930c/s, suppressing the fundamental to −46dB while attenuating the 2nd and 3rd harmonics by less than 1dB. By switching from the high pass to the low pass condition the ratio of fundamental to harmonics could be determined almost instantaneously for any chosen level of signal.

The basic design work for this filter was done while the writer was working at the Radio Section of the New Zealand Post Office.

M. R. NICHOLLS,
Decca Radar Ltd.

REFERENCES

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2. WEINBERG, L. Additional Tables for Design of Optimum Ladder Networks. *J. Franklin Inst. 264*, No. 1, 7 (July, 1957) No. 2, 127 (August, 1957).
3. SKWIRZYNSKI, J. K., ZDUNEK, J. Design Data for Symmetrical Darlington Filters. *Proc. Instn. Elect. Engrs.* Monogr.227R (March, 1957).

Short News Items

Pye equipment was used in Antarctica by Dr. Vivian Fuchs for the measurement of the thickness of the ice at intervals along the route. The seismic method used for these measurements involves the detonating of a charge of explosive and measuring the time that it takes for the pulse to travel down to the interface between the bottom of the ice and the underlying formations, and after reflection to travel back to the surface. The sensitive seismic detectors (geophones), which are placed out along the surface to detect the return of the pulse, and also the 25 trace high speed galvanometer recording oscillograph, which makes the record, were built by W. G. Pye & Co Ltd. The equipment was supplied through a Pye associate company, Seismic Instruments Ltd, to British Petroleum Co Ltd, who not only supplied the apparatus, but also seconded a geophysicist to Dr. Fuch's party to conduct the work of the survey.

The only wireless communication sets used by Dr. Fuchs were two Army type C12 sets designed and manufactured by Pye Telecommunications Ltd.

The Annual Dinner of The Institution of Electrical Engineers was held at Grosvenor House, London, on February 27, when the President, Mr. T. E. Goldup, C.B.E., was in the Chair. The principal guest was The Rt. Hon. Lord Adrian, O.M., M.D., F.R.S., Master of Trinity College, and Vice-Chancellor, University of Cambridge, who proposed the toast of "The Institution", to which the President replied. The toast of "Our Guests" was proposed by Sir Gordon Radley, K.C.B., C.B.E., Ph.D. (Eng.), (Immediate Past-President), and the reply was given by Sir George Edwards, C.B.E., B.Sc. (President, The Royal Aeronautical Society).

A Commonwealth Engineering Conference was held in Australia in March and was attended by the Presidents and Secretaries of the Institution of Civil Engineers, the Institution of Mechanical Engineers and the Institution of Electrical Engineers. From the Australian meetings at Sydney, Melbourne and Canberra, the party will fly to New Zealand for meetings at Christchurch, Wellington and Auckland and then across the Pacific, via Honolulu, to San Francisco and Vancouver for meetings in various Canadian cities. By the end of April they will reach New York for a meeting, the first to be held in America, of the Conference of Engineering Societies of Western Europe and the U.S.A. (EUSEC), where they will be joined by Sir George Nelson, Bt., President of the Mechanicals and Past President of the Electricals. After the conference, some members of the party, among them Air Marshal Sir Owen Jones, K.B.E., C.B., Vice-President of the Mechanicals, will visit the Joint Oversea Groups of the three Institutions in Trinidad and also the Argentine. Others will attend the Annual Meeting of the Engineering Institute of Canada in Quebec. This will be the first time the representatives of the three Institutions have made a complete circuit of the globe, and this fact emphasizes the importance now given to the exchange of technological information between professional and scientific bodies.

The Electronic Engineering Association is the new name for the Radio Communication and Electronic Engineering Association. This change was decided upon after a recent meeting of the Council. Mr. F. S. Mockford (Marconi's Wireless Telegraph Co Ltd) was re-elected chairman for the ensuing year and Mr. V. M. Roberts, O.B.E. (British Thomson-Houston Co. Ltd), was elected vice-chairman. The Council of the E.E.A. for 1958 consists of The British Thomson-Houston Co Ltd, A. C. Cossor Ltd, Decca Radar Ltd, Electric and Musical Industries Ltd, Ferranti Ltd, Kelvin and Hughes Ltd, Marconi's Wireless Telegraph Co Ltd, Metropolitan-Vickers Electrical Co Ltd, Mullard Ltd, Murphy Radio Ltd, the Plessey Co Ltd, Standard Telephones and Cables Ltd.

Decca Radar Ltd and the Decca Navigator Co Ltd have moved their head offices to Decca House, Albert Embankment, London, S.E.11.

The Shell Petroleum Co Ltd have ordered a high-speed electronic computing system to solve complex mathematical problems which arise in the planning of economic activity. It will be installed in the London office and used for special calculations covering commercial aspects of all activities from exploration to marketing. Based on the Ferranti Mercury machine with additional punched card, magnetic tape and high-speed printing units, the computing system will be one of the largest in the world to be used mainly for operational analysis. It will also be one of the fastest systems in Europe for electronic processing of data. The equipment will be able to store the equivalent of more than 60 000 000 decimal digits at a time, and the computer itself will perform up to 16 000 individual operations per second. The present order has been largely influenced by Shell's successful experimental work over the last three years on the Ferranti electronic computer in the Amsterdam laboratory. The new computing system is expected to be in full operation early in 1959.

The Wilmot Breeden Group of Companies, whose manufacturing interests include motor vehicle and gas turbine components, hydraulics and electronics, are sponsoring two Fellowships, each worth £1 000 per annum, one at the University of Birmingham and the other at the College of Technology. An unusual feature of these Fellowships is that the successful candidates will divide their time between the University or College and the Company. The terms of the Fellowship awards are so arranged that at any one time there will be a Wilmot Breeden Fellow working in association with the University and another with the College. The Fellowships will be advertised every year in March, in 'odd' years in association with the University and in 'even' years in association with the College. Each Fellowship will normally be held for a period of two years. A candidate for a Fellowship should normally have had two or three years research or industrial experience. He must be acceptable to the academic authority and would be expected to have an honours degree in a University in the British Commonwealth, a Diploma in Technology, or an equivalent qualification. An application for a Fellowship should outline a two-year investigation or project on which the candidate seeks to work. Further information may be obtained from the Secretary, Wilmot Breeden (Holdings) Ltd, Amington Road, Birmingham 25.

An International Convention on Transistors and Associated Semiconductor Devices is being organized by the Institution of Electrical Engineers in London from May 25-29 1959.

Servomex Controls Ltd announce that Dr. John Douce of Manchester University has been retained as consultant to advise on theoretical problems arising in the field of non-linear control systems, and systems subject to random fluctuations.

Errata. The following amendments should be made in the article "ZETA (The Main Recording and Monitoring Equipment)", which appeared in the March issue. Page 118, paragraph commencing "The resulting fall in voltage at the cathode of V_5 . . ." for "V_5" read "V_{5a}". Page 119, for "100k/s" read "100kc/s." Page 119, the opening paragraph under "Stabilized Power Supplies" should read as follows. "Some points of interest will be described. Four standard Cawkell PE31 Model A supplies are fitted (Fig. 9(a)). These embody a printed plug-in electronic control circuit with interlocking plug, whereby certain control resistors and the tapping on the transformer h.t. secondary are changed. . . ."

BOOK REVIEWS

Passive Network Synthesis

By J. E. Storer. 319 pp. 211 figs. Medium 8vo. McGraw-Hill Book Co., New York and London. 1957. Price 64s.

THE three fundamental concepts of network theory are the network, the excitation, and the response. The two fundamental problems of network theory are analysis and synthesis. Network analysis is the determination of the response when the network and excitation are prescribed; network synthesis is the determination of the network when the excitation and response are prescribed.

Modern network synthesis has its origin in the reactance theorem enunciated by Campbell in 1922 and adapted by Foster in 1924 to the design of passive non-dissipative 1-ports. With the extension of Foster's work by Cauer in 1927, the theory of passive two-element 1-port synthesis was complete. In 1931, Brune showed that the impedance function of a passive 1-port is a positive real function. Conversely he showed that any such function can be realized by a passive 1-port and gave a procedure for the synthesis of 1-ports containing three kinds of passive elements. Cauer, Darlington, and Cocci demonstrated that any positive real function can be synthesized as a reactive 1-port containing a single resistance but, as in Brune's method, the synthesis yielded ideal transformers. Bott and Duffin have since shown how the Brune synthesis can be modified to avoid ideal transformers. The problem of passive 2-port synthesis was solved by Gewertz in 1933. In 1939, Darlington introduced a reactive 2-port insertion-loss synthesis employing a frequency transformation previously introduced by Bode, to account for the dissipation of the reactive elements. Cauer and Guillemin also made significant contributions to the theory of reactive 2-port synthesis but the problem of passive n-port synthesis remained unsolved until eight years ago when Oono proposed a procedure yielding a $2n$-port with n of the ports closed with resistors. The n-port problem was also solved independently by Bayard and Leroy. In 1951, Belevitch introduced the scattering matrix into n-port synthesis and Tellegen, in 1953, extended Brune's technique to the synthesis of n-ports and thus arrived at realizations with the minimum number of elements. Because of the frequent appearance of coupled inductors and ideal transformers in many synthesis solutions, considerable research has recently been directed to the problem of realizing networks devoid of mutual-inductance. Notable advances have been made in this connexion by Guillemin, Dasher, Fiallow, Gerst, Ozaki, Lucal, and Adams. The advent of the transistor has now opened the way to the synthesis of active networks.

The literature of modern network synthesis is extensive but the number of books devoted to the subject is at present very small. This new book by Dr. J. E. Storer is the first of its kind and has as its purpose a concise survey of the field of synthesis. The book is divided into four parts the first of which contains nine chapters and deals with the realization of passive 1-ports. Beginning with an examination of the properties a function must possess to be realizable as a driving-point immittance, the first part continues with an exposition of the Foster-Cauer approach to reactance synthesis. Brune's procedure for designing any 1-port is then introduced and is followed by a discussion of the Darlington and Bott-Duffin methods. The first part ends with a brief indication of methods leading to realizations with dissipative reactors.

The remainder of the book deals with the synthesis of 2-ports. Eight chapters constitute the second part and are devoted to a review of the early Campbell-Zobel image parameter methods. The topics discussed include m-derived filters and delay lines, constant resistance lattices, and maximally flat delay lines. The third part introduces the modern insertion-loss technique originated by Darlington. Eight chapters are devoted to this topic. Two-element syntheses are treated in detail and the discussion concludes with a survey of techniques using active elements and a description of an iterative procedure due to Scott and Blanchard.

Network synthesis involves the solution of two distinct problems: the approximation problem and the realization problem. The latter is the problem of realizing the network according to a prescribed rational function or functions of complex frequency. The approximation problem is that of finding the physically realizable rational functions which satisfactorily approximate the desired network behaviour. The last part of the book contains five chapters dealing with techniques relating to the approximation problem. First to be discussed is the method of finding a polynomial representation of the required function using orthogonal polynomials and the subsequent conversion to a rational fraction using Padé approximants. The use of the potential analogy is also considered and the final chapter deals with the problem of finding a realizable rational function approximation when the required time response of the network is prescribed.

It seems reasonable to suppose that future developments in network technique will depend a great deal upon current research in network synthesis. This belief has stimulated considerable activity in recent years but many problems remain unsolved. Not the least of these is the presentation of much of the work in a form suitable for ready application by the engineer. Dr. Storer's book is addressed to the reader with a thorough background in modern network analysis. It does not pursue the subject beyond the synthesis of 2-ports but the information contained will be welcomed by those who recognize the importance of preparing the way for network engineering in the future.

S. R. DEARDS

Microwave Measurements

By Edward L. Ginzton. 515 pp. 90 figs. Demy 8vo. McGraw Hill Book Co., New York and London. 1957. Price 90s.

THIS new book is the product of many years of work in the field by the author, who is Professor of Applied Physics and Electrical Engineering, and the late W. W. Hansen.

The scope and limitations of the book are very clearly defined in the Preface. The measurements of Power, Impedance, Wavelength, Frequency, 'Q' and Attenuation are covered in exhaustive detail; no attempt is made to cover specialized activities such as aerial and propagation measurements.

Despite the fact that the preparation of this work is implied by the author to have been spread over more than ten years, it is remarkably up-to-date. A welcome feature is the adequate treatment of British contributions to the science.

The book appears to have been written for readers well versed in communication engineering but not necessarily with previously acquired knowledge of microwave methods. Consequently, the reader who has access to the M.I.T. series of volumes may find that much space is devoted to background information that is obtainable eleswhere. Some idea of the completeness of the treatment is given by the fact that the first chapter, "Generation of Laboratory Signals," extends to 107 pages, and not only gives a critical comparison of the many types of oscillator available but describes their principles of operation in some detail.

Although the emphasis of the book is naturally on the electromagnetic principles of the techniques described, the mechanical design requirements of such instruments as standing wave detectors receive adequate attention.

The bibliography, appearing as footnotes to the text, is very comprehensive, reference being made to the appropriate chapters of textbooks as well as to original papers.

The standard of production of this book, which will inevitably become an important standard work, is of the quality that one has over many years learned to associate with the publishers.

S. S. D. JONES.

High Quality Sound Reproduction

By J. Moir. 591 pp. 120 figs. Demy 8vo. Chapman & Hall Ltd. 1958. Price 70s.

High Fidelity Sound Reproducti n

Edited by E. Molloy. 200 pp. 60 figs. Demy 8vo. George Newnes Ltd. 1958. Price 20s.

IT is unfortunate that two books dealing with exactly the same subject matter and having substantially identical titles should appear at the same time. However, there need be no confusion, because there is little comparison between the treatment of the contents.

Mr. Moir's monumental work bears the stamp of the impartial research worker, underlining the fact that an independent investigator can always present the subject matter in a more reasoned, comprehensive and less biased manner than the expert associated with a particular trade product.

It is stated that the contents are written for both the professional engineer and the amateur; but this is hardly possible with equal emphasis, therefore we find the best parts really appealing to the knowledgeable enthusiast. Although the book is by far the most comprehensive text so far available on sound reproduction, the sections on amplifiers, valves and output transformers are somewhat condensed and few examples have component values appended. On the other hand, matters which are more associated with Mr. Moir's interests, such as hearing, loudspeakers, room acoustics, pickups and stereophonic reproduction are dealt with in a masterly fashion. Since these represent the least understood factors in reproduction, the reader will be amply repaid by chapters I-IV, VII and VIII, and XIII-IXX.

Few criticisms can be offered. The references are excellent but far from complete. The peak powers for a concert or cathedral organ would be about 10dB up on the values given for a cinema organ in the region 60-20c/s. Mr. Moir could perhaps have devoted some attention to the psychology of listening, the change in sensitivity of the hearing system after the first stimulus, and the unsuitability of the Fletcher-Munson curves for measurements on complex tones. In these matters the work of Prof. Seashore at the University of Iowa is illuminating.

Although it can be argued that some of the subject matter is more fully covered in specialized texts, on the whole the balance of material is deftly handled and the book is strongly recommended; one might go so far as to say that it renders obsolete any previous work on reproduction.

One conclusion (common to both books) is worth emphasizing; many impartial surveys show an overwhelming preference for reproducing equipment which cuts off at 8 to 10kc/s, and no one enjoys a system extending to 15 or 20kc/s for either domestic or professional reproduction of music. This applies to listeners of every class, nationality and age group.

Turning to the second book, this must suffer by comparison though it stands reasonably well on its own. The team of expert contributors have condensed the various sections too much, though not all to an equal extent. This may be because of the estimated selling price of the work, but it is not a good policy because it is aimed at the "engineer, dealer and the keen amateur". The different requirements of these groups cannot all be met in a survey of 200 pages. To take an example, the tuned pipe is put forward as a simple alternative to the vented loudspeaker enclosure; but nothing is said about the time lag due to the inertia of the large mass of air in the pipe, which in fact renders such pipes unsuitable for programme material. In other instances, more detailed information would have cleared up a number of obscure points, but this is the price to be paid for condensing so much into so small a space.

On the other hand, many complete circuits with component values are included, although no help is given with the design of output transformers. The section on electrostatic loudspeakers is particularly good, written as it is by the leading expert in this technique. A chapter deals with v.h.f. radio receivers and some good circuits are included, with constructional details. A number of useful dimensioned designs for loudspeaker enclosures should satisfy any discriminating reader.

The book is very well produced and represents good value at the price. The enthusiast will find much of practical value though in places it is a little advanced for the beginner or the musician; but if one requires an authoritative summary of the art the book is recommended; and if the reader wishes to delve further into the finer points of sound reproduction, he can graduate to Mr. Moir.

ALAN DOUGLAS

Radio Technology

By E. J. Voght. 556 pp. 325 figs. Royal 8vo. Sir Isaac Pitman & Sons Ltd. 1958. Price 45s.

IN this book, the author has included the necessary d.c. and a.c. theory to prepare the student for a proper concept of the basic radio phenomena that are presented as the text progresses. The earlier chapters present nomenclature, axiometric definitions, and the essential mathematics of radio.

It was the author's purpose to make available, within one volume, the necessary material for a complete understanding of the radio phenomena with which the practical radio engineer must work.

Basic Physics

By Alexander Efron. 692 pp. 800 figs. Royal 8vo. John F. Rider Publisher Inc., New York. 1957. Price $8.95.

THIS book comprises two volumes in a single binding. The author is Chairman of the Physics Department of the Stuyvesant High School, New York. Basic demonstration experiments are described and model problems of graduated difficulty are to be found at the end of each chapter. The volumes are simply explained and would appeal to the non-mathematical student as well as those with scientific knowledge.

NEW EQUIPMENT AT THE EXHIBITIONS

A description, compiled from information supplied by the manufacturers, of a few of the exhibits at the Physical Society Exhibition, the Radio and Electronic Component Manufacturers' Federation Exhibition and the Instruments, Electronics and Automation Exhibition.

(Letters in parentheses indicate the exhibitions at which the manufacturers will be represented)

AERO RESEARCH LTD. (R).

Duxford, Cambridge.

RESINS

Aero Research will be showing a recent development known as 'Araldite' dipping resin. This is a two component formulation, providing a thixotropic mixture with a long potlife at room temperature. The mixture will give a substantial coating to components immersed in it. Flexible resins possessing various degrees of elasticity are readily obtainable by the addition of a special plasticizer. Coloured resins may also be obtained by the use of specially developed 'Araldite' colouring pastes.

'Araldite' resins are being increasingly used in connexion with printed circuits and a new adhesive has been developed for bonding the copper foil to the conventional paper based laminate.

ADVANCE COMPONENTS LTD. (I).

Roebuck Road, Hainault, Ilford, Essex.

L.F. SIGNAL GENERATOR

(Illustrated below)

The type 81 l.f. signal generator has a frequency range of 15c/s to 200kc/s. covered in four bands.

The accuracy of the frequency calibration is ±1 per cent +1c/s up to 50kc/s and ±2 per cent from 50kc/s to 200kc/s.

The frequency stability is better than 0·1 per cent at 1kc/s after a warm up period of one hour. With mains voltage, variation of ± 10 per cent drift is less than ± 0·04 per cent at 1kc/s.

The total harmonic and hum content compared with fundamental above 100c/s is better than 34dB down 2 per cent at full output and 40dB down 1 per cent at 1mW. There is a slight increase in distortion below 100c/s.

The hum and noise content is less than 0·25 per cent of maximum output.

The output is controlled by units and decade attenuators.

The output impedance is 600Ω centre tapped, balanced or unbalanced terminations with respect to earth or 300Ω unbalanced.

The attenuator accuracy is ± 1·5 per cent of attenuator reading and frequency errors are negligible.

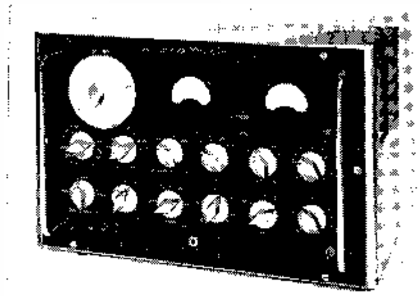
AIRMEC LTD. (P.I.).

High Wycombe, Buckinghamshire.

TRANSISTOR TEST SET

(Illustrated below)

The transistor tester type 236 is a general purpose tester for pnp junction and point type transistors. It is entirely mains operated and enables static tests to be carried out up to a current of 40mA and a maximum collector voltage of 80V.



In addition certain small signal parameters are measured directly by a bridge method at a frequency of 420c/s.

From these measurements simple calculation will give the remainder of the usual transistor parameters, thus determining

(a) the equivalent 'T' resistances r_e, r_b, r_c, r_m, α
(b) the hybrid 'h' parameters h_{11}, h_{12}, h_{22}, h_{21} and h'_{11},
(c) the mullard parameters r_{in}, r_{11}, r_{out}, r_{22}, α and r'_{in}, r'_{11}, r_{out}, r'_{22}, α'

This instrument contains its own bridge oscillator and detector, and facilities are provided for the connexion of an external oscilloscope to check for non-linearity of the small signals measurements, or telephone for aural bridge balance detection.

V.H.F. WAVE ANALYSER

The v.h.f. wave analyser type 248 is a selective measuring set suitable for use in the frequency range 5Mc/s to 300Mc/s.

It consists essentially of a selective amplifier working on the heterodyne principle and incorporating built-in l.f. attenuators and externally connected h.f. attenuators. Indication of the output, which is in the form of a supersonic signal, is provided by means of an average reading meter.

Briefly, the circuit consists of a grounded grid r.f. amplifier feeding a diode mixer operating on the fundamental of the injected local oscillator signal. Wave band switching is accomplished by a special v.h.f. coil turret, individual gain adjustment being provided on each band. The resulting i.f. signal is applied to a two-stage negative feedback amplifier, using a low noise input valve, followed by switched low-pass filters giving a choice of three cut-off frequencies. The i.f. signal then passes through precision coarse and fine attenuators before entering the final three-stage negative feedback amplifier which operates the meter circuit. Germanium crystals are used for rectification, thus dispensing with the usual set zero controls. A further amplifier is provided for modulation monitoring. It acts as a linear amplifier for amplitude modulation monitoring and as limiter amplifier for frequency modulation monitoring. After demodulation, either a.m. or f.m., the audio signal is amplified before being passed to the phones jack. A high degree of stability is achieved by feeding the r.f. amplifier and oscillator from a stabilized h.t. supply.

EXHIBITION DETAILS

The Physical Society Exhibition was held from 24 to 27 March in the Halls of the Royal Horticultural Society, London.

* * *

The R.E.C.M.F. Exhibition will be held from 14 to 17 April in Grosvenor House and Park Lane House, London.

* * *

The I.E.A. Exhibition will be held from 16 to 25 April at Olympia, London.

* * *

ELECTRONIC ENGINEERING

will occupy Stand Number 154 at the R.E.C.M.F. Exhibition (Park Lane House) and Stand Number 714 at the I.E.A. Exhibition and visitors will be welcome.

ANALYTICAL MEASUREMENTS LTD. (I).

Dome Buildings, The Quadrant, Richmond, Surrey.

POCKET pH METER

(Illustrated above right)

Completely self-contained with batteries, in a bakelite case, this instrument is furnished, camera fashion, in an ever-ready case with novel plastic tubes of buffer KCl solutions. The meter is scaled from 2 to 12pH for easy reading, and an adjustment gives readings from 0 to 14. Accuracy of 0·1pH is obtainable. Hearing-aid type batteries provide up to 1 300 hours of operation. The electrometer tube, switch and input connector are sealed in a single unit to ensure freedom from high humidity difficulties. The

one-knob control and continuous reading features of this instrument simplify operation for untrained personnel. Grounded samples can be directly measured because of no external power connexions. The instrument and electrode system are completely shielded.

AVO LTD. (P.I.).

Avocet House, 92-96 Vauxhall Bridge Road, London, S.W.1.

MAGNETIC MEMORY SCALING UNIT

This is a small light-weight scaling unit of extremely low power consumption. The unit employs minute ferrite magnetic cores as bistable elements in a scale of ten circuit, together with a new form of magnetic count indicator in which the currents flowing in the scaling circuit magnetize thin strips of nickel iron mounted around a balanced pair of compass needles.

The current pulses to operate the indicator are obtained from a circuit employing transistors. The power required to operate a complete scaling unit is little more than the power required for the first scale of ten; this is of the order of 3mW for a count rate of 1 000/sec. The resolving time of the circuit is approximately 3μsec, and the maximum mean count rate fixed by the collector current rating of the transistor is 250kc/s.

BRAYHEAD (ASCOT) LTD. (R).

Fall View Works, Kennel Ride, Ascot, Berkshire.

VALVEHOLDERS

A range of B9A valveholders has been designed with the particular needs of printed circuits in mind. The main features of these valveholders are, free-seating contacts (preventing damage to valves); removable and replaceable contacts (no chance of solder running into contacts); contact soldering to chassis if required; wedge and spring locking action of can to saddle (no 'rock' or earthing noise). The bodies are moulded in ceramic or phenolic material with extremely low-loss properties and screening cans in all standard sizes and finishes are available.

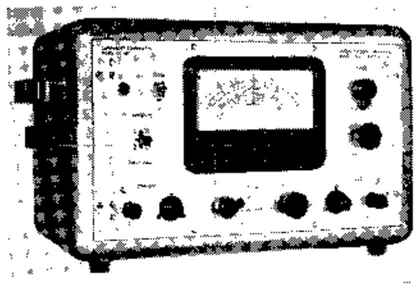
BRITISH PHYSICAL LABORATORIES (R.P.I.)

Houseboat Works, Radlett, Hertfordshire.

COMPONENT COMPARATOR

(Illustrated above right)

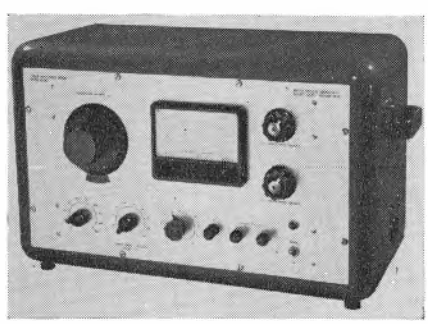
The component computor type C7457 has been designed for comparing inductive, capacitive or resistive components rapidly against the user's standard. It will deal with components having ranges from 10Ω to 1MΩ, 100pF to 10μF, and 5mH to 5H. Division limits are ±2 per cent ±10 per cent and +25 per cent to −35 per cent. These limits are indicated on a 5in centre zero meter. The bridge frequency is 1kc/s.



1Mc/s CAPACITANCE BRIDGE

(Illustrated below)

The capacitance bridge type CB.181 is another recent development and enables the precision measurement of dielectric loss and capacitance at 1Mc/s within the range 0 to 100pF with an accuracy of ± 0·1pF. Another model is available having a range of 0 to 1 000pF with an accuracy of ± 1pF. The value of capacitance is directly indicated on an engraved dial and the value of dielectric loss is directly expressed in terms of conductance from 0 to 1 000 × 10^{-8}mhos. In order to obtain the value of dielectric loss expressed as percentage power factor, it is only necessary to divide the indicated value of conductance by the indicated value of capacitance and multiply by a factor of 0·159. The bridge consists of a special substitution circuit patented by A. C. Lynch, of the Post Office Research Station.



CAMBRIDGE INSTRUMENT CO. LTD. (P.I.)

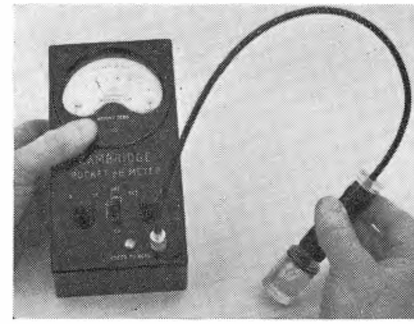
13 Grosvenor Place, London, S.W.1.

POCKET pH METER

(Illustrated above right)

This self-contained, pocket sized, pH meter of high stability and accuracy is useful in the works, laboratory or field, where spot checks are required.

The scale, 0 to 14pH units, is subdivided into 0·1pH units and the instrument has a high overall accuracy. To save space and to simplify operation the measuring and reference electrodes have been combined into a single unit. The electrode system is fitted into a plastic container so that small quantities of liquid



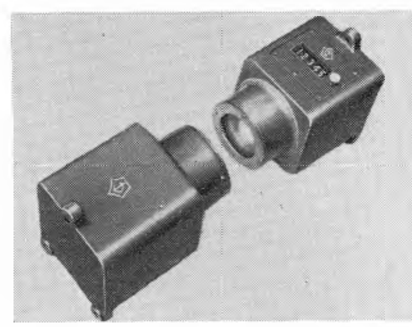
can be tested without need of beaker or stand. The instrument, electrode system, and buffer solutions can be contained in a small carrying bag with a shoulder strap. A dry cell electrical supply is contained within the instrument case and, when necessary, can be easily replaced. The instrument is simple to use and will withstand rough handling.

COUNTING INSTRUMENTS LTD. (I).

5 Elstree Way, Boreham Wood, Hertfordshire.

PHOTO-TRANSISTOR COUNTERS

(Illustrated below)



This photo-transistor counter has been designed as a packaged unit, comprising light source with a built-in 5-figure resettable counter and transistor photocell.

Both units are sealed and therefore can be used in damp or dirty conditions. The whole unit is self-contained and merely requires to be plugged into a suitable power supply.

HIGH SPEED ELECTRICAL COUNTER

This counter has been developed to meet the current trend for counters having facilities for either push-button or electric reset, combined with high speed, long life and large figures.

If required, this counter can be operated on a make and break pulse of 20msec duration.

DYNATRON RADIO LTD. (P.I.)

Maidenhead, Berkshire.

COINCIDENCE UNIT

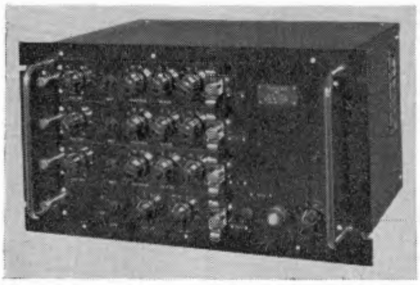
(Illustrated next column)

The Dynatron coincidence unit type 1036C is now available commercially and contains a three-channel coincidence circuit which may be used in any one of the following arrangements:—

(1) Channels 1 and 2 in coincidence.

(2) Channels 1 and 2 in coincidence, the output being in coincidence with channel 3.

(3) Channels 1 and 2 in coincidence,



the output being in anti-coincidence with channel 3.

(4) Channel 3 in coincidence with channels 1 or 2.

(5) Channel 3 in anti-coincidence with channels 1 or 2.

Facilities are provided within the unit for amplitude discrimination variable paralysis time and variable delay time.

ELLIOTT BROS. (LONDON) LTD. (P.I.)
Century Works, London, S.E.13.
TAPE CONTROLLED PRINTER

This equipment, specially developed for printing output information from digital computer and data-processing systems, is entirely automatic in operation. The output information is accepted as punched paper tape which is fed, via a high-speed tape reader, to the control unit. The control circuits convert the information into a form suitable for operating the Comuprinter which will print letters, figures and a selection of punctuation marks, etc., at the rate of twenty per second.

The character spacing is twelve to the inch. The paper is fed by sprockets at either side, the maximum width of paper being eighteen inches. The paper feed provides a line spacing of six lines per inch.

E.M.I. ELECTRONICS LTD. (P.R.I.)
Hayes, Middlesex.
ANALOGUE COMPUTING
(Illustrated above right)

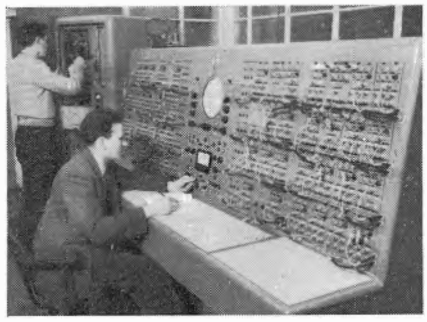
A demonstration will be given of a complete electronic analogue of a complex chemical plant. This affords a detailed study of plant operating characteristics and thus a new approach to chemical plant design.

The plant which is being simulated forms a closed loop and is made up of three kinds of units—storage tanks, operating columns and connecting pipes with their associated pumps.

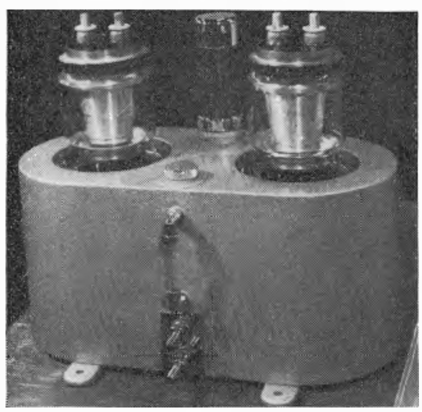
Individual characteristics of each of these units are simulated on the computer: for example, in the case of an extraction column, damping effect of the plates and the dead time or transport lag while the liquid passes from one plate to the next has to be taken into account; while in the case of a gas absorber, compressibility must be allowed for and the volume of liquid or gas must be a function of pressure.

Similarly, the pumps and valves associated with the connecting pipes have special features to be allowed for in the simulation Each valve has an associated level controller or flow controller, the valves and controllers have physical limits on their speeds and positions and as the level controllers send a hydraulic signal to the valves there is dead time and simple lag.

By means of this analogue of the complete plant it is therefore possible to assess the following important characteristics: sensitivity of control and stability margin, whether simple proportional control is sufficient, and the rate of propagation of disturbance (e.g. how soon the failure of any pump will cause trouble in the rest of the plant).



In addition, it is possible to simulate the starting up of the plant and so arrive at the most satisfactory starting routine; to check whether the pumping capacity and characteristics are adequate and further, to obtain a guide as to the influence of plant controllability on plant economics.



ENGLISH ELECTRIC VALVE CO. LTD. (P.R.I.)
Chelmsford, Essex.
VAPOUR COOLED VALVES
(Illustrated above)

Conditions sometimes exist both in industry and in the transmitting field, where neither forced-air cooling nor water cooling of high power valves is entirely satisfactory. Air may be so dirt-laden that the cleaning of filters and radiators calls for undue effort and cost. At the same time water may be expensive, scarce, or unsuitable for other reasons.

In these conditions vapour cooled valves have the advantages of low water costs, built-in protection against failure of water supply, quiet operation and greater overload capacity.

In the vapour cooled valve a closed system, called the boiler, contains distilled or otherwise purified water in contact with the specially designed valve anode. When the valve is operating, heat from the anode rapidly converts some of the water into steam which passes via a condenser system back to the reservoir of water at the bottom of the boiler. Efficient circulation is brought about by the provision of specially designed circulation holes in the walls of the anode.

Two different condenser systems are employed and the choice between them is largely dependent upon local conditions.

In the external condenser system the steam is led outside the boiler to the condenser, where cooling may take place by natural convection or by any other convenient means.

In the boiler-condenser system the steam is condensed by being directed over the top turns of a cooling coil inside the boiler. A supply of cooling water is necessary for the coil, but the consumption is only a fraction of that for water cooled valves of similar power capabilities.

The E.E.V. boiler-condenser is fitted with a protective device for shutting off the power supply in the event of water failure or of serious overloading of the anode. Visual indication of water level is also provided.

A number of valves in the E.E.V. range of forced as or water cooled valves are now available as vapour cooled versions.

ERICSSON TELEPHONES LTD. (P.I.)
Beeston, Nottingham.
BATCH COUNTER
(Illustrated below)

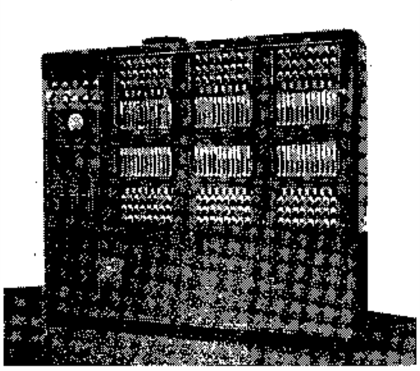
The batch counter type 108C is designed for counting a wide variety of objects of varying size and shape into batches of any required number up to 10^4. The count is displayed on Dekatron tubes, and the composition of a batch can be preset by rotary switches. At the end of each batch count, the instrument automatically resets to zero prior to commencing the next batch. The instrument operates from input pulses provided by mechanical, photo-electric, or electromagnetic devices, and provides output facilities suitable for control of chute flaps. Additional equipment can be incorporated in the instrument to meet specific requirements. These consist of an electro-mechanical register for recording the number of batches, a warning signal unit for operating a warning device 10 or 100 counts before the end of a batch, and a control unit which provides alternate

output pulses suitable for operating a pair of solenoids to control chute mechanisms at the end of each batch. Spurious pulses which are liable to occur when counting irregularly shaped objects are overcome by seven switch selected paralysis times. A d.c. input unit can be incorporated when it is desired to count large slow moving objects whose transit time is greater than the longest available paralysis time.

FAIREY AVIATION CO. LTD. (I)
24 Bruton Street, London, W.1.
ANALOGUE COMPUTER
(Illustrated below)

This analogue computer is designed on a unitary basis to fit standard 19in × 6ft racks. A standard range of items is available enabling a computer of any desired size to be built up with maximum flexibility in arrangement and usage.

Special attention has been paid to the design from the operational standpoint; the knowledge required by the operator is solely that needed to apply analogue



techniques to the computer, while the layout ensures a minimum loss of computer time in servicing.

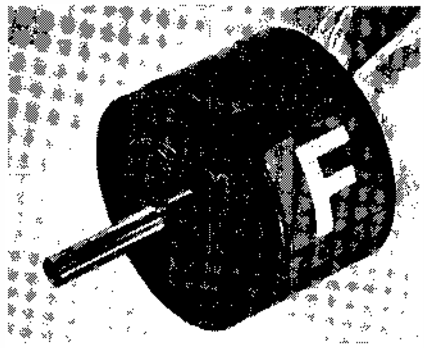
The basic arrangement comprises a control rack, housing control, distribution, and checking equipment for up to 96 operational amplifiers. Control racks may be inter-connected for larger computers. The basis of the computer is the computing panel which houses 12 plug-in drift corrected operational d.c. amplifiers, together with plug-in function units which enable the computer to perform all the standard linear and non linear functions.

Other units such as servo multipliers, angle resolvers etc., are also available, these are plug-in units housed in a separate standard servo function panel containing servo amplifier and buffer amplifiers.

Solutions may be displayed and recorded by means of a multichannel cathode ray oscillograph recorder or a quick response pen recorder.

FERRANTI LTD. (R.P.)
Hollinwood, Lancashire.
PRECISION POTENTIOMETERS
(Illustrated above right)

The Ferranti range of precision potentiometers is designed to provide analogue conversion from mechanical rotation to



an electrical signal. Linear and sine/cosine laws are available and other non linear functions can be manufactured. Four standard ranges of potentiometers are in production and are available in single and multi gang units. These potentiometers have extremely low starting torques, good resolution and linearity and still retain small dimensions. They are capable of reliable operation under conditions of severe shock.

FIELDEN ELECTRONICS LTD. (I).
Wythenshawe, Manchester.
LEVEL CONTROLLERS

Fielden Electronics are exhibiting two new industrial level controllers, the Tektors Major and Minor. These instruments replace the Tektor J's because, although the latter were reliable and stable instruments, it was felt that their sensitivity and general performance could not be further improved within the limits of their design. The Tektors Major and Minor operate on the well established and patented capacitance method.

These instruments incorporate printed circuits and have a control knob on the front of the chassis which enables the set-point to be easily adjusted on site. Other features include triggered positive relay action, quick release chassis, sensitivity with stability, rugged terminal blocks and industrial weatherproof cases. Greatly extended cable lengths are available which is a considerable advantage in that the electronic unit may now be installed some distance away from the probe. With the Major, for instance, a maximum of thirty-five feet of cable may be used for high level and twenty feet for low. The performance of the instruments is independent of cable length up to the stated maximum.

The Tektor Minor, while incorporating many of the features of the Major, has been designed to fill the need for a cheaper yet still reliable level controller limited to 110 or 240V operation, 10ft of coaxial cable and 250V 5A a.c. switching capacity, single pole changeover contacts.

THE GENERAL ELECTRIC CO. LTD. (P.I.)
Magnet House, Kingsway, London, W.C.2.
PACKAGED TRANSISTOR OSCILLATORS
(Illustrated above right)

To assist electronic equipment designers who require an accurate frequency standard, G.E.C. quartz-crystal transistor oscillators are now available in packaged form. Two self-contained units will be shown.

For demonstration purposes the two frequencies chosen are 1kc/s and 100kc/s. Each unit has a low-power sine-wave output with a total harmonic distortion of less than 1 per cent.

The 1kc/s oscillator circuit, which uses two GET3 transistors and three silicon diodes, is in 'potted' form on an international-octal valve base. The oscillator and the quartz-crystal units are plugged into a die-cast box containing connexions for a 12V d.c. supply and the coaxial output terminals.

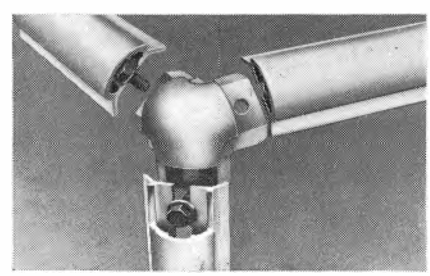
The 100kc/s unit is in tubular form and uses one GET3 transistor in a series-

resonant circuit. It operates from a self-contained 4V mercury cell which will give about 1 500 hours satisfactory operation.

HALLAM, SLEIGH & CHESTON LTD. (R.I.).
Oldfield Road, Maidenhead, Berkshire.
CABINETS
(Illustrated below)

The Widney Dorlec cabinet system has been internally modified so as to considerably reduce the assembly time hitherto necessary. The modified system replaces the 12 screws previously used in each complete corner assembly by 3 bolts.

Thus, all that is now needed for



assembly of main frame Widney Dorlec "Quick-Fit" components, as the new parts are called, is one bolt concealed in each section passing through the main hole in the corresponding corner flange.

HILGER & WATTS LTD. (P.I.)
98 St. Pancras Way, Camden Road, London, N.W.1.
DIGITIZING EQUIPMENT

Of particular interest among the latest Hilger and Watts developments in digitizing equipment is an optical tachometer. It consists of a glass disk, rotating between a lamp and a photo-transistor, and divided into 2000 alternately opaque and transparent sectors. The photo-transistor detects the pulses of light and responds with electrical pulses, which are counted by conventional means. The duration of count is so adjusted, and the total of one count stored while the next is being made, that the speed of rotation can be displayed on in-line indicators. The advantages of the optical tachometer are low torque and long life, the lamp only needing occasional replacement.

Examples of optical digitizers, which have binary coded disks in place of the disk in the optical tachometer, will be shown. Their detectors may be phototransistors, Schwarz cells, or photomultipliers, all of which have advantages for some applications.

Binary coded straight-line scales, with fine cursor readings when necessary, will be shown.

ALFRED IMHOF LTD. (R.I.)

112-116 New Oxford Street, London, W.C.1.

FORCED VENTILATION RACK

This latest addition to the Imhof range of standard racks is a specially designed version of an enclosed rack which gives unusually effective forced ventilation. The most efficient cooling commensurate with economy is ensured by the incorporation of a powerful blower unit and by the semi-sealing of all doors; the side doors, too, are fitted with three-point locking controlled by a single handle on each door. Forced ventilation racks are available as standard in heights of 4ft, 5ft, 6ft and 7ft with panel widths of 19in and 22½in and frames 19in and 24in deep in each height.

The blower unit consists of an axial flow fan, mounted on a sturdy chassis. The large inlet is fitted with a glass wool filter; this can be replaced as necessary by simply undoing the panel fixing screws and withdrawing the chassis about 4in. On the outlet side, the deflecting chamber is fitted with an all-round flange, so that trunking can be connected if localized cooling is required.

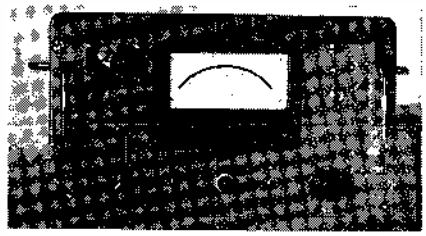
THE INFRA RED DEVELOPMENT CO. LTD. (P.I.)

40 Tewin Road, Welwyn Garden City, Hertfordshire.

MICROMANOMETER

(Illustrated above right)

This is a sensitive pressure measuring device capable of measuring differential pressures as low as ±·005in water gauge for full scale deflexion. Its speed of response is such that it can follow sinusoidal pressure variations up to 200c/s. The instrument consists of two cavities to which the pressures to be measured are applied, between which is mounted a thin diaphragm. The movement of the diaphragm is measured as a capacitance change which detunes two identical tuned circuits coupled to a stable oscillator circuit operating at about 2Mc/s. The instrument is of particular use in measuring small draughts such as are experienced in chimneys, adjacent to open fires etc, and may also be used



for the measurement of low gas velocities.

MARCONI INSTRUMENTS LTD. (I.)

Longacres, St. Albans, Hertfordshire.

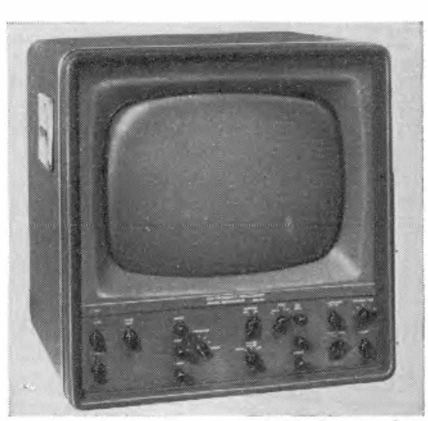
L.F. OSCILLOSCOPE

(Illustrated below)

This large screen, high-resolution oscilloscope type TF1159 permits the detailed investigation of voltage waveforms in the frequency range 15c/s to 20kc/s; it is particularly suitable for use in conjunction with sweep generators, such as the Marconi types TF1099 and TF1104, for the presentation of frequency response displays. The size of screen also makes it very effective for lecture-room demonstration purposes.

Displays of up to 24 × 32cm can be accommodated on the 17in rectangular-faced c.r.t. The Y amplifier has a high order of linearity and gives faithful reproduction; its sensitivity is such that a display height of 10cm can be obtained with signal levels less than 50mV. Maximum adaptability in presenting displays is ensured by the fact that the time-base, synchronizing and blanking signals can be applied either internally or externally. For frequency-sweep displays, externally-derived marker signals can be added for accurate frequency identification; for evaluation of relative response amplitudes, the combination of fine trace delineation and large screen size, made possible by the electromagnetic deflexion system, gives high discrimination with speed and certainty of measurement.

The vertical deflexion sensitivity of the Y amplifier system is variable from 4mV/cm to 12·5V/cm by means of a fine gain control and an attenuator with six 10dB steps; the attenuator calibration can be standardized against an internally-generated reference voltage. The frequency range of the Y input circuits is 15c/s to 10ks/s and can be extended to 20kc/s at reduced deflexion.



A linear sawtooth time-base can be generated internally by a Miller-transitron oscillator whose repetition frequency is continuously variable from 15c/s to 5kc/s in four ranges; alternatively, an external time-base drive can be applied, which enables the display to be synchronized with a sweep generator.

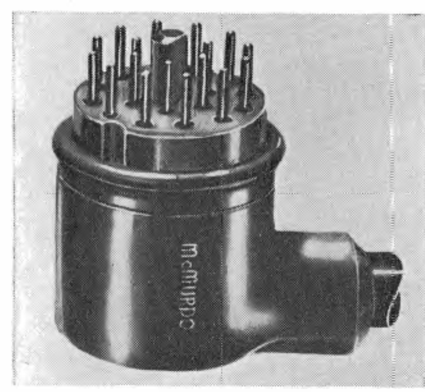
The internal time-base can be synchronized with either the Y input signal or external pulses; flyback blanking can also be applied internally or externally. An external marker signal, in the form of positive or negative pulses of known p.r.f., can be used to intensity modulate the display, giving bright or dark frequency-marker points on the trace.

MCMURDO INSTRUMENT CO. LTD. (R.)

Victoria Works, Ashtead, Surrey.

POTTED PLUGS AND SOCKETS

A range of miniature plugs and sockets, known as 'Pokonnectors' which



have potted cable connexions, is now in production. The system is that the makers take the customers' cables, connect to the plug or socket, and pot the assembly in Polythene. It is claimed that the potted connexions are far less likely to give trouble than those enclosed in the conventional metal cover with cable clamp. Also, although not a hermetic seal, the Polythene potting affords considerable protection from dirt, grease and moisture.

MEASURING INSTRUMENTS (PULLIN) LTD. (R.I.)

Electrin Works, Winchester Street, Acton, London, W.3.

TEST-SET

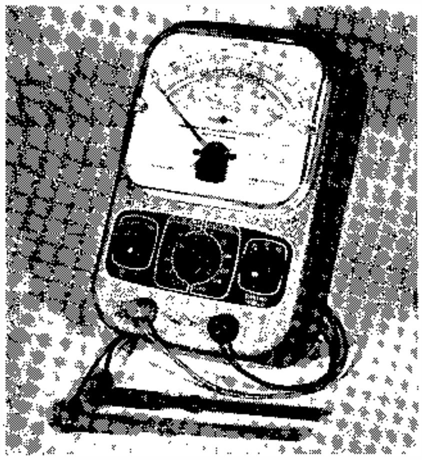
(Illustrated above right)

The present series 100 test set has been completely re-styled to provide a more robust instrument, together with the following additional features.

A new type Diakon plastic cover is fitted to the meter, which gives a very wide angle of vision, and good illumination of the dial surface.

An increased scale length is possible due to the use of the new cover.

The test set is fitted with a stand which makes it possible to use an instrument in a horizontal, vertical or an inclined position. The case has been re-styled in aluminium alloy and a finger recess for carrying is fitted.



The meter has a $3\frac{1}{4}$in mean scale length, rectangular pattern, with knife edge pointer, and a sensitivity of 100μA for full scale deflexion.

The test leads have been improved and they are fitted with retractable probes and a swivelling mechanism which enables the leads to be used in any position without risk of damage.

A printed circuit has been designed for the majority of the wiring, which gives a much more rugged construction, facilitates service, and enables all components to be accessible when the back cover is removed.

METROPOLITAN-VICKERS ELECTRICAL CO. LTD. (P.I.)

Trafford Park, Manchester, 17.

DIGITAL COMPUTING ELEMENT

A single basic transistor circuit has been designed from which a complete digital computing system may be constructed. The technique provides systems which are simple and economical to design and maintain, and which should prove more reliable than more complicated existing circuits. Although the present circuit is slower than some of these it will find applications in the fields of process control and data processing, where speed is often far less important than reliability and ease of maintenance.

A binary/decimal convertor constructed entirely from these basic units will be shown. This machine will accept any three-figure decimal number which may be set up manually, convert it to the corresponding binary number, and display the result on indicator lights. Further numbers may be set up and converted, and each is added to the total already stored, which may be reconverted and displayed in its decimal form at will. The basic frequency of the system is 10kc/s and notable features are its compact construction and low power consumption.

MUIRHEAD & CO. LTD. (P.I.)

Beckenham, Kent

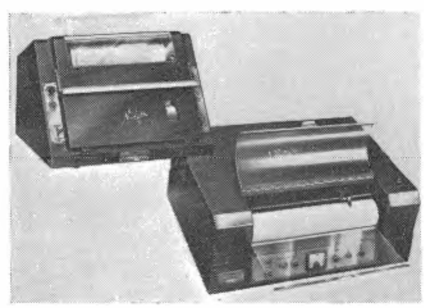
AUTOMATIC FACSIMILE SYSTEM

(Illustrated above right)

The new D-900 and D-901 Mufaxequipment will provide an inter-office or inter factory automatic facsimile system over ordinary telephone lines. Teleprinter copy, consignment notes, bank statements or memoranda—anything written, typed or drawn may be transmitted without error.

The D-901 transmitter accepts copy of any size up to 14in $\times$ $8\frac{1}{2}$in by means of a spring loaded transparent blind which locates the message round the drum. Transmission is commenced by pressing the start button; the distant receiver is alerted and transmission commences. Documents of practically any texture or colour may be transmitted; white, off-white or lightly coloured backgrounds are reproduced perfectly white. A warning lamp indicates the end of message and successful reception. For full size copy the transmission time is $6\frac{1}{2}$min or $4\frac{1}{2}$min depending on type of equipment used.

The D-900 receiver functions entirely automatically. Received copy is on white translucent Mufax paper feeding continuously during transmission from a roll 200ft in length—sufficient for 16 hours of continuous message reception. No attention is necessary other than removing the messages for distribution. The Mufax record is instantly visible—no processing being required.



For inter-office use one transmitter can feed a number of receivers located in various departments. Additional receivers may be in other parts of the country. The telephone lines may be local wires or long distance trunk circuits.

Where common power supplies are available the synchronization between transmitter and receiver is by means of the local a.c. supplies. For non-synchronous power supplies precision tuning forks are fitted to each instrument to ensure that the transmitter and receiver are synchronized.

PAINTON & CO. LTD. (R.I.)

Kingsthorpe, Northampton.

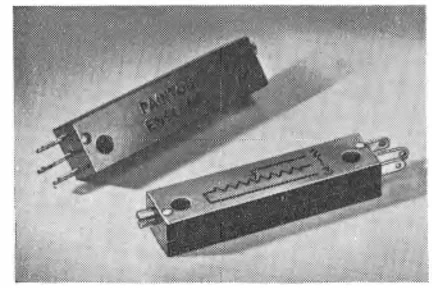
MINIATURE FLAT POTENTIOMETERS

(Illustrated above right)

An addition to the Painton range this year is the 'Flatpot'—a miniature preset wirewound potentiometer. The flat form of the potentiometer, which allows stacking and facilitates mounting, is also suitable for use with printed circuits and for this purpose a version is available with right angle tags. All tags and mounting holes are on a 0·1in module. Dimensions are only $1\frac{1}{4}$in $\times$ 5/16in $\times$ 7/32in, the resistance range is 10Ω to 10kΩ.

PRINTED CIRCUIT CONNECTORS

A range of printed circuit connectors is now available which in addition to the 10 way Microcon connector includes a more robust 15 way plug and socket fitted with steel dowels. This latter connector with contacts on 0·2in connectors is especially



suitable for use with printed circuit boards on plug-in chassis, when the dowels eliminate the need for expensive jigging. In the final stage of development is a minature 20 way printed circuit plug and socket employing a new design contact with 0·1in centres; this new connector incorporates 20 contacts in line in a total length of only $2\frac{3}{4}$in. A further development is an edge connector having 60 ways comprising 3 groups of 20 on 0·1in centres. This connector also employs a similar design of contact to that used on the 20 way but on a large scale and tests have shown that this form of contact greatly prolongs the life of the edge of the board. Other sizes are being designed.

THE PLESSEY CO. LTD. (R.P.I.)

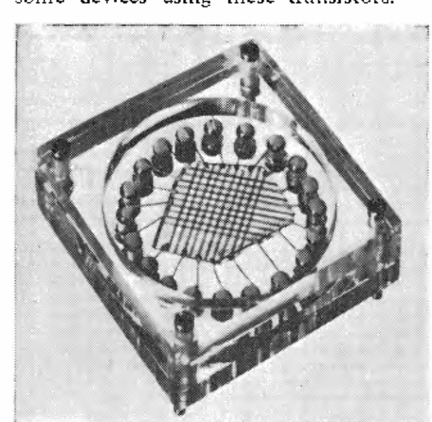
Vicarage Lane, Ilford, Essex.

SEMI-CONDUCTORS

(Illustrated below)

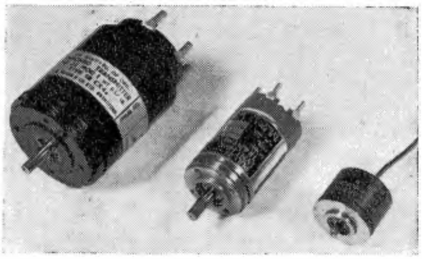
Significant advances have been made in the development of intermetallic semiconductors and these will be illustrated by a display of semi-conducting indium antimonide and indium arsenide crystals. Both these materials have applications as infrared detectors and as Hall and magnetoresistance probes. Typical devices such as the infra-red cell type PC 21, which utilizes the photo-electromagnetic effect in a single crystal filament of indium antimonide, will be shown and demonstrated in operation.

Other new heat sensitive components to be shown will include an extensive range of thermistors, suitable for applications such as temperature measurement and voltage regulation. These are ceramic components having a negative temperature co-efficient of resistance at 25°C of about 6 per cent/°C. A typical unit gives 500Ω at 25°C and 12 to 15Ω at 100°C. Germanium and silicon high-frequency transistors made by electro-chemical techniques will also be shown together with some devices using these transistors.



Progress in solar generation will be demonstrated on a pn junction which is formed within 0·0005in of the surface in a slice of low resistivity single crystal silicon. Under the influence of light, minority carriers diffuse to the junction and so generate a current. The e.m.f. is approximately 0·5V and the efficiency 5 to 10 per cent. (See Illustration).

Another new and interesting development on view will be a single crystal memory store. This compact information store is produced by evaporating or printing electrode systems on to single crystals of barium titanate. Stores, such as the one just described, with access time of 10μsec, are slightly slower than analogous magnetic devices but can operate at the low powers and voltages that can be obtained from transistor sources.



R. B. PULLIN & CO. LTD. (I.)
Phœnix Works, Great West Road, Brentford, Middlesex.
SERVO COMPONENTS AND SYSTEMS
(Illustrated above)

A servo system consists essentially of error signal elements, an amplifier, and a motor which drives the load through gearing, in order to reduce the error to a minimum.

The design of gear box to be associated with the load is essentially dependent on each application, and, therefore, it is not generally economic to attempt to provide standard gear systems as ratio, size and configuration can vary so much. The rest of the system, however, consists of:—

(1) A pair of control synchros, (2) Two-phase a.c. servomotor, and (3) A compact amplifier
all of which can be standardized with advantage to the user of servo systems.

In the past much time has been spent by each instrument engineer in the designing and manufacturing of his own servo system, particularly the amplifier. R. B. Pullin & Co. now hope that by providing compact amplifiers to be associated with the a.c. servo-components they produce, they will solve this problem for the majority of instrument engineers.

The transistor amplifier, which will be designed to operate with the size 10, size 11 or size 15, 400c/s a.c. servomotors, will be extremely compact and totally immersed in resin. Means will be provided so that the amplifier characteristics can be controlled by the user to suit a broad range of servo system applications, and match each particular system requirement.

The Kearfott Division of R. B. Pullin & Co. will manufacture size 11 synchros and a range of small motors and motor tachos which may be used in association with packaged amplifiers.

Illustrated is a size 18 C × 4a synchro, size 11 C × 4 synchro and the size 10 servomotor.

RECORDER CHARTS LTD. (I.)
Clyde Vale, Dartmouth Road, London, S.E.23.
RECORDING CHARTS

A selection of charts will be displayed representing the wide range which is produced.

The exhibit will indicate the specialized service which is offered to manufacturers of recording instruments, who require real precision charts, whether their requirements are for rolls, circles or sheets. Special high stability papers, which ensure minimum expansion or contraction through varying degrees of humidity, are used for charts for conventional ink recording. Also available is a range of inkless papers requiring heat, pressure or electric current as the recording medium, and these can be supplied in any chart form.

W. H. SANDERS (ELECTRONICS) LTD. (P.I.)
Gunnels Wood Road, Stevenage, Hertfordshire.
TRANSDUCTORS
(Illustrated below)

A range of encapsulated transductors is available consisting of five frame sizes giving outputs of 0·045, 0·09, 1·15, 5·5 and 11·5W d.c. with a supply frequency of 50c/s and 1·1, 6·8, 25, 70 and 145W d.c. with 400c/s, using the auto self-excited connexion. The high grade nickel iron laminations are inserted and sealed after the windings have been encapsulated in an epoxy resin, thus avoiding damage to the strain-sensitive core material.

The transductors are used in a standard range of magnetic amplifiers which are available either as separate units, or as complete automatic control systems, suitable for reliable control of motor speed, generator voltage and frequency etc.

SERVOMEX CONTROLS LTD. (P.I.)
Crowborough Hill, Jarvis Brook, Sussex.
D.C. SERVO SYSTEM COMPONENTS
(Illustrated above right)

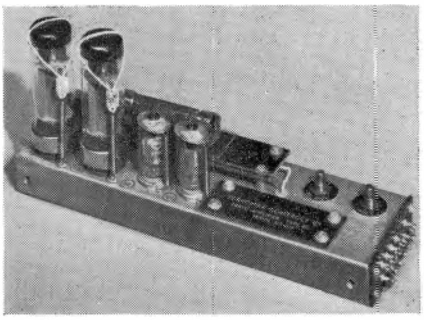
Units for the construction of d.c. servo systems for industrial and educational purposes will be shown. These are as follows:—

Servo amplifier SA.61; this is designed to give a full range of field control on standard split field servo motors requiring 80mA for full torque. Maximum output is obtained for 65mV d.c. input.

Armature supply AS.69: this unit is intended for the 'constant current' armature supply for small d.c. servo motors. Two currents are available from the standard unit, 0·35A and 0·48A.

HT supply DC.68: this unit is particularly suitable for feeding the servo amplifier, but also has many other applications in experimental work. It provides three 6·3V unregulated a.c. heater supplies and a stabilized h.t. supply. The h.t. potential can be set in the range 250 to 350V at currents ranging from 0 to 150mA.

These three units are supplied as unprotected chassis mounted assemblies which reduces the cost and permits the individual chassis to be incorporated easily in larger systems. The three units together are ideally suited for driving standard d.c. servo motors.



These sub-assemblies are used in standard Servomex control systems and thus they can offer technical colleges and others equipment of the highest quality at a price made possible by quantity production.

At the Exhibition, these units will be shown built into a self-balancing Wheatstone bridge system. This will be shown as an example of how a simple servo system can be used for the rapid sorting of resistive elements. The system will also be subjected to test signals generated by the low frequency function generator type LF.51.

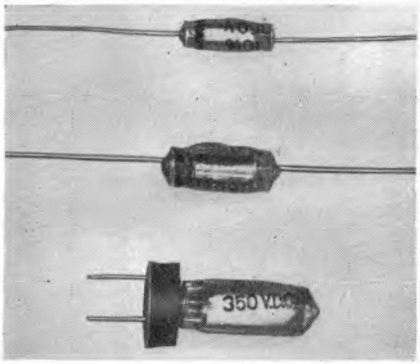
SOUTHERN INSTRUMENTS LTD. (P.I.)
Frimley Road, Camberley, Surrey.
PORTABLE OSCILLOSCOPE

The oscilloscope type M971 is a compact general purpose oscilloscope measuring 10·5in × 13in × 17in overall and containing a 3in flat-faced cathode-ray tube. The viewing hood and graticule are detachable and may be replaced by a recording camera, using the same fixing holes.

The Y amplifier has a maximum sensitivity of 0·2V/cm and a bandwidth from d.c. to 80kc/s. Input voltage range is 2, 6, 20, 60, 200 and 600V with voltage calibration to an accuracy of ±5 per cent. An a.c. pre-amplifier with a gain of 100 gives an overall bandwidth from 3c/s to 40kc/s.

The time-base has three modes of operation: triggered, single-stroke and beam trigger. Sweep durations are from 0·3 to 1000msec, calibrated in time.

The X amplifier has a bandwidth from d.c. to 80kc/s and maximum sensitivity 0·6V/cm. Expansion is continuously variable in the ratio 5:1.



SUFLEX LTD. (R.)
35 Baker Street, London, W.1.
PRINTED CIRCUIT CAPACITORS
(Illustrated above)

Suflex are making three distinct approaches to the problem of using capacitors in printed circuits. These are as follows:—

(1) End Mounting. The terminations of this component come from one end of the body, and are brought through a rubber or Neoprene bung. The latter ensures a constant lead spacing, and also gives a flat base for stable mounting prior to clinching.

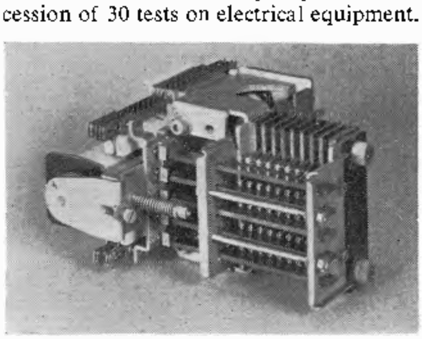
(2) Thicker Wires, Double-ended components are offered with thicker wires, i.e., 23 gauge, which makes them more suitable for insertion into printed boards.

(3) Bronze Wire Terminations. It would not be possible to use 21 gauge copper wire with small Polystyrene capacitors, and in order to obtain similar mechanical strength, 25 gauge tinned bronze wire terminations are available.

In addition to the above, Suflex are prepared to supply sample quantities of capacitors belted for automatic insertion for use with the Dynasert system.

TELEPHONE MANUFACTURING CO. LTD. (R.I.)
Martell Road, West Dulwich, London, S.E.21.
CROSSBAR BRIDGE
(Illustrated below)

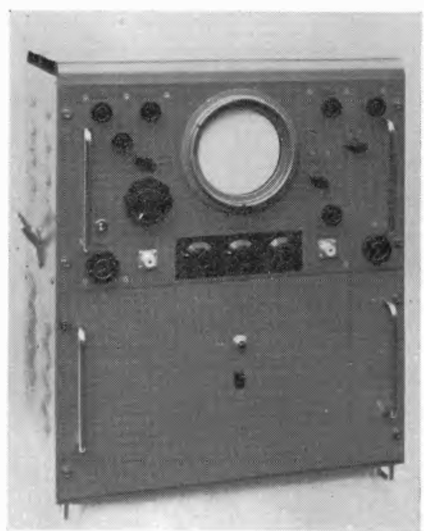
An unusual exhibit is the TMC crossbar bridge, a compact multi-point switching unit for use in circuits where it is required to select automatically a number of different switch points, to count impulses or hunt over a series of outlets. Two or more bridges, interacting electrically can provide complex counting and stepping facilities. The TMC crossbar bridge is finding increasing use in telemetering, process control and serial testing switching systems. It will be demonstrated in use in an automatic circuit tester which automatically imposes a succession of 30 tests on electrical equipment.



ULTRASONIC CO. (LONDON) LTD. (P.I.)
Sudbourne Road, Brixton Hill, London, S.W.2.
CONSOLE FLAW DETECTOR
(Illustrated below)

The Ultrasonic type B.1 (available in bench model or console form) is suitable for the research laboratory or in production where portability is not vital, i.e., for large immersion test rigs. Its size allows control and variation of a greater range of variables than in the standard portable flaw detection apparatus. Features include rectified or unrectified signal display, multi-channel bandwidth selector, frequencies to customer's requirements, single or double probe working and very high time - base speeds for examination of pulse shapes.

The flexibility resulting from these and other facilities makes the equipment suitable for application research, the immersion testing technique generally and for pure research and measurement.

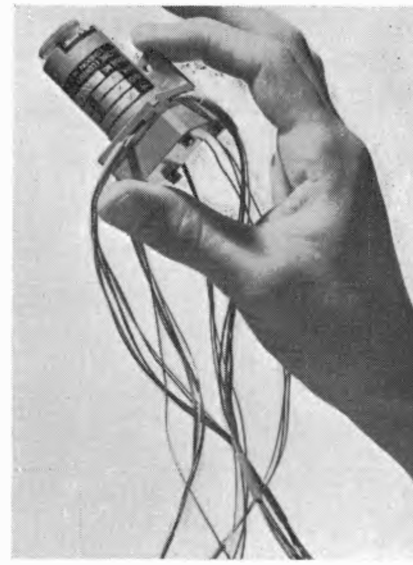


VACTRIC (CONTROL EQUIPMENT) LTD. (R.I.)
23 Mossop Street, London, S.W.3.
HIGH SPEED ROTARY SWITCH
(Illustrated above right)

The Vactric high speed rotary switch has been specially designed to fulfil many of the requirements now demanded by remotely operated systems.

Its construction permits it to be one of the smallest of its kind and particular care has been taken to extend the life of the switch without maintenance, and without the use of oil baths, etc. Because the manufacture of the contact plate is made basically through a photographic process, repetition of other units with exactly the same specification is no problem. Wiper pressure is low, compatible with contact wiper bounce, and because of this, several banks may be ganged together on a common shaft, utilizing the 1in diameter motor shown in the illustration.

In high frequency systems where intercoupling between channels is difficult to eradicate, this type of switch provides the minimum possible coupling between segments. Segments may be inter-connected



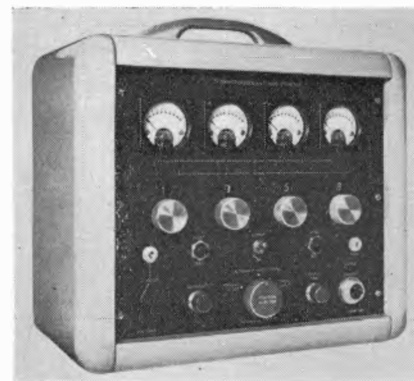
to provide a large variety of mark to space ratio, or if desired, special track configurations will be considered. Leads are supplied plain or can be screened and in either case they are left free for attachment to the user's terminal block.

The life of the switch is in excess of 100 hours at 6 000rev/min and approximately 1 000 hours at 600rev/min.

The number of poles is either 24 or 48, and can be break before make or make before break.

VENNER ELECTRONICS LTD. (P.I.)
Kingston By-Pass, New Malden, Surrey.
BATCHING COUNTER
(Illustrated below)

The batching counter type TBC is an all transistor equipment containing four decades, each of which is controlled by a 10 position switch calibrated from 0 to 9. A number is set up on the counter and input pulse rates varying from 1p/s to 10kp/s are injected. The counter stops after the required batch has been reached. Apart from straight-forward industrial applications where objects are to be counted, the instrument is also extremely convenient for providing precise delays. If, for example, the injected frequency is 1kp/s a batched number of 1234 pulses may be set up when an interval of 1·234 seconds will elapse prior to an output pulse being generated. This device will be demonstrated fed from inputs varying in frequency from 1c/s to 10kc/s.



Meetings this Month

THE BRITISH COMPUTER SOCIETY

Date: 21 April. Time: 6.15 p.m.
Held at: York Hall, Caxton Hall, London, S.W.1.
Lecture: The Use of an Electronic Computer in Research Statistics.
By: F. Yates.

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 23 April. Time: 6.30 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Lecture: Measurement of the Frequency Response of a Nuclear Reactor.
By: R. J. Cox and R. B. Stevens.

South Midlands Branch

Date: 25 April. Time: 7 p.m.
Held at: The North Gloucestershire Technical College, Cheltenham.
Lecture: The Automatic Factory.
By: J. A. Sargrove.

North Eastern Section

Date: 9 April. Time: 6 p.m.
Held at: The Institution of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle upon Tyne.
Annual General Meeting followed by a programme of technical films.

Scottish Section

Date: 17 April. Time. 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow.
Annual General Meeting followed by film evening.

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings, unless otherwise stated, will be held at the Institution, commencing at 5.30 p.m.

Radio and Telecommunication Section

Date: 9 April.
Informal evening: U.H.F. Test Transmissions.

Ordinary Meeting

Date: 10 April.
Lecture: Radio Observations on Artificial Satellites.
By: J. A. Ratcliffe.

Informal Meeting

Date: 14 April.
Discussion: The Future Trend of Installation Inspection.
Opened by: Forbes Jackson.

Measurement and Utilization Sections

Date: 15 April.
Lectures: A Train Performance Computer.
By: E. Bradshaw, M. Wagstaff and F. Cooke.
The Simulation of Distributed-Parameter Systems with particular reference to Process Control Problems.
By: J. F. Meredith and E. A. Freeman.
A Magnetic-Drum Store for Analogue Computing.
By: J. L. Douce and J. C. West.

Supply Section

Date: 16 April.
Lecture: Operational Experience—Calder Hall.
By: K. L. Stretch.

Radio and Telecommunication Section

Date: 23 April.
Lecture: Survey of Performance Criteria and Design Considerations in High-Quality Monitoring Loudspeakers.
By: D. E. L. Shorter.
Date: 28 April.
Informal Evening: Economic Usage of Broad-Band Transmission Systems.
Talk by: R. J. Halsey.

East Midland Centre

Date: 15 April. Time: 6.30 p.m.
Held at: Loughborough College.
Discussion: The Application of Digital Techniques to Control.

Cambridge Radio and Telecommunication Group

Date: 22 April. Time: 8 p.m.
Held at: The Cavendish Laboratory, Free School Lane, Cambridge.
Lecture: Underwater Television.
By: D. Allanson.

North-Eastern Centre

Date: 14 April. Time: 6.15 p.m.
Held at: The Royal Station Hotel, Newcastle upon Tyne.
Annual General Meeting and Conversazione.

Tees-Side Sub-Centre

Date: 2 April. Time: 6.30 p.m.
Held at: The Cleveland Scientific and Technical Institute, Corporation Road, Middlesbrough.
Lecture: Some Transistor Input Stages for High-Gain Chopper-Type D.C. Amplifiers.
By: G. B. B. Chaplin and A. R. Owens.

North Midland Centre

Date: 1 April. Time: 7 p.m.
Held at: The Royal Station Hotel, York.
Lecture: The Control and Instrumentation of a Nuclear Reactor.
By: A. B. Gillespie.

South-West Scotland Sub-Centre

Date: 2 April. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow, C.2.
Lecture: Ship Stabilization: Automatic Controls Computed and in Practice.
By: J. Bell.
Date: 8 April. Time 7 p.m.
Held at: The Royal College of Science and Technology, George Street, Glasgow, C.1.
Lecture: Storage and Manipulation of Information in the Brain.
By: R. L. Beurle.
Date: 22 April. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow, C.2.
Lecture: Operational Experience—Calder Hall.
By: K. L. Stretch.

South Midland Centre

Date: 14 April. Time: 7.30 p.m.
Held at: The Winter Gardens, Malvern.
Lecture: Developments in Semi-Conductors.
By: E. G. James.

South Midland Radio and Measurements Group

Date: 28 April. Time: 6 p.m.
Held at: The James Watt Memorial Institute, Great Charles Street, Birmingham.
Lecture: Servo-Operated Recording Instruments, a Review of Progress.
By: A. J. Maddock.

North Staffordshire Sub-Centre

Date: 14 April. Time: 7 p.m.
Held at: The Duncan Hall, Stone.
Informal evening on The Importance of Research in Hearing and Seeing to the future of Telecommunication Engineering.
Talk by: E. C. Cherry. (Joint meeting with Stone-Stoke Centre of The Institution of Post Office Electrical Engineers.)

Southern Centre

Date: 2 April. Time: 6.30 p.m.
Held at: The C.E.G.B. Offices, Portsmouth.
Lecture: Junction Transistors.
By: E. Wolfendale.

West Wales (Swansea) Sub-Centre

Date: 10 April. Time: 6 p.m.
Held at: The Conference Rooms, South Wales Electricity Board Showrooms, The Kingsway, Swansea.
Lecture: Nuclear Engineering.
By: D. R. Smith.

THE INSTITUTION OF ELECTRONICS

Date: 24 April. Time: 7 p.m.
Held at: The Assembly Hall, The University of London Institute of Education, Malet Street, London, W.C.1.
Lecture: Radio Astronomy and the Sputniks.
By: F. G. Smith.

PUBLICATIONS RECEIVED

THE IMPERIAL COLLEGE OF SCIENCE AND TECHNOLOGY, University of London, has issued a book on postgraduate courses for 1958-59. This session about a third of the 860 full-time postgraduate students are taking such courses. Forty-one separate courses leading to a higher degree and/or to the Diploma of the Imperial College are available, of which thirty-four have already been accepted by the Department of Scientific and Industrial Research for the tenure of their Advanced Course Studentships. Prominent among the courses are those in branches of engineering, for example in automatic control systems, nuclear power and gas turbine technology, but the range of courses is wide and includes subjects as far apart as applied entomology and pure mathematics. The courses provide the opportunity for students, including many who have already spent a period in industry, to further their knowledge in a particular specialized field, and at the same time to learn from experts their experiences of the application of this knowledge in industry. 'Postgraduate Courses 1958-59' is obtainable, price 1s., post free, from the Registrar, Imperial College of Science and Technology, Prince Consort Road, London, S.W.7.

COLLOQUE INTERNATIONAL PROBLEMES PHYSIQUES DE LA TELEVISION EN COULEURS. The proceedings of this International Symposium on Physical Problems of Colour Television held in July 1957, have been published by Acta Electronica, 23 Rue du Retrait, Paris XXe. The text appears in both French and English.

PORTABLE TRANSISTOR RECEIVER, Sensitive Superheterodyne Circuit Incorporating R.F. Transistors, by S. W. Amos has recently been published for "Wireless World" by Iliffe & Sons Ltd, Dorset House, Stamford Street, S.E.1. Price 2s. 6d. It describes the theoretical and practical design of a battery-operated, self-contained, portable set using a total of seven junction transistors, four of which are the a.f. type and three are the latest r.f. type.

PROBLEMS IN RADIO ENGINEERING by E. T. A. Rapson. In the 7th edition of this book the text has been completely reset and revised. Some earlier questions have been deleted and many numerical ones added. Material has been introduced on feedback amplifiers, valve reactors, and frequency discriminators and the m.k.s. system of units has been used throughout. Sir Isaac Pitman & Sons Ltd, 39 Parker Street, Kingsway, London, W.C.2. Price 15s.

G-V THERMAL RELAYS is the title of a folder produced by Mercia Enterprises Ltd, Codiva House, Allesley Old Road, Coventry. G-V relays represent an entirely new departure in the field of thermal timing and sensing devices and offer many advantages and possibilities that have been unobtainable heretofore. Their exceptional environmental characteristics and small weight and size make them particularly useful in the automation, aircraft and missile fields. Mercia Enterprises Ltd are pleased to announce that the conclusion of recent licensing agreement permits them to offer this component to the British electronics industry.

REPAIRING HI-FI SYSTEMS by David Fidelman describes how to find troubles and repair faults in high fidelity equipment with no test equipment, with simple equipment or with elaborate equipment. It discusses servicing of amplifiers, preamplifiers, tuners, tape recorders, etc., as well as the installation of these units. John F. Rider Publisher Inc., 116 West 14th Street, New York 11, N.Y., U.S.A. Price $3·90.

HIGH FIDELITY SIMPLIFIED, by H. D. Weiler, 3rd edition, has recently been published by John F. Rider Publisher Inc., 116 West 14th Street, New York 11, N.Y., U.S.A. Price $2·50. A substantial portion of the book has been revised.

VALVES FOR R.F. HEATING is the subject of a comprehensive catalogue produced by the English Electric Valve Co. Ltd, Chelmsford, Essex.

CONFERENCE TERMINOLOGY, a manual for conference members and interpreters in English, French, Spanish, Russian, Italian and German, by Jean Herbert, has been published by the Elsevier Publishing Company, Amsterdam/New York/Princeton. Cleaver-Hume Press Ltd, 31 Wright's Lane, London, W.8. Price 12s. 6d.

TUBI ELETTRONICI by A. Pinciroli deals with questions which were the object of several years' teaching at Turin University. Scienza Nuova, Via Andrea Doria 39, Milan. Price L1.800.

ATOMIC POWER FOR PEACE VOLUME 3 RADIOISOTOPES AND TRACERS IN AGRICULTURE METEOROLOGY AND MEDICINE by W. J. Vogt, 2nd edition, has recently been published by Atomic Power Review, 414/416 Elizabeth House, 18 Pritchard Street, Johannesburg.

CLOSED CIRCUIT TELEVISION SYSTEM PLANNING by M. A. Mayers and R. D. Chipp. This book is a guide for those who have the responsibility of planning and evaluating closed circuit television systems. It answers questions related to the organization of closed circuit television systems such as space requirements, cost of equipment and its installation, types of equipment available and manpower needed to operate and maintain the equipment. John F. Rider Publisher Inc., 116 West 14th Street, New York 11, N.Y., U.S.A. Price $10.