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Commentary

THE construction of a 'round-the-world' Commonwealth telephone cable was considered at the recent Commonwealth Trade and Economic Conference and it was envisaged that its provision would be based on capital contributions from all Commonwealth countries and that the new cable would be operated on a self-supporting financial basis. The technical and financial issues arising in connexion with this proposal were examined at a conference of technical representatives from all Commonwealth countries which met in London earlier this year and which examined in detail a scheme for such a cable.

The first transoceanic telephone cable was opened in 1956 between the United Kingdom, Canada and the U.S.A., and the construction of a further cable between Canada and the United Kingdom is expected to be completed in 1961. This would form the first part of the Commonwealth link, the rest of the route being as follows:—The West Coast of Canada would be linked with New Zealand and Australia and from there the cable would run through the Indian Ocean connecting Malaya, India, Pakistan and Ceylon and from these countries by way of Kenya to South Africa; by providing a link between South Africa and the United Kingdom connexion could be made with the Transatlantic telephone cables.

The order of construction is, of course, a matter for further consideration by the Commonwealth Governments, but it may well be that the existing and proposed links from the United Kingdom to Canada would first be extended by a cable from the United Kingdom to South Africa followed by a cable across the Pacific to New Zealand and Australia; the links spanning the Indian Ocean, to be constructed later, would complete the girdle of the world. The total length of such a cable installation would be about 30 000 nautical miles and the estimated cost of construction some £88M.

Although the network is proposed as a Commonwealth project, individual governments will have, obviously, a special interest in different sections of it and it is likely that the various countries concerned will become partners in the ownership of particular sections as they are constructed.

In the 1920's there was considerable rivalry between the protagonists of cable and radio as a means of long distance telegraphic communication. However, the two systems gradually learned to live together in harmony and by the outbreak of war in 1939 they had reached an advanced stage of integration, and Commonwealth communications were secure against anything but a major catastrophe. But when Italy entered the war she immediately cut the Mediterranean cables between Malta and Gibraltar thus disrupting the main cable routes to the Middle East, India, the Far East and Australia. Later on the whole of the Far Eastern cable network was lost to the Japanese. To make good this very severe breach in world telegraphic communications the h.f. communication networks were expanded on an unprecedented scale and radio reigned supreme.

The above remarks apply to telegraphy only of course. As far as telephony is concerned there has been, until two or three years ago, no alternative to radio as a means of long distance transoceanic communication. But now it would seem that we must look to cables rather than to radio to satisfy the future requirements of the ever growing telephone and telex traffic between the Commonwealth countries; for although radio might suffice for a time on some routes where radio transmission is usually good, it does not satisfy even the immediate needs, in terms of quality and regularity of service, for some important Commonwealth services. There would appear to be little likelihood of any technical development in radio communication in the near future that will greatly enhance its performance; but with cable the position is very different for, not only is the design and construction of the cables themselves improving rapidly but also, and perhaps more important, the design of the necessary deep sea repeaters is improving steadily.

The present Transatlantic telephone cable is a twin-cable system, one cable carrying traffic in the westerly direction and the other cable carrying traffic in the easterly direction. Recent developments by the Post Office Research Laboratories in co-operation with the British communications industry have now made it possible to provide both-way connexion for telephone and telex services over a single transoceanic cable, thus conforming with British practice in shallower waters. This development has reached the stage where the engineering planning of an intra-Commonwealth scheme can proceed with this new technique and, indeed, it will be used for the first time in the cable to be laid between Canada and the United Kingdom in 1961. In addition considerable research is being directed towards the evolution of even lighter cables and repeaters using transistors instead of valves, which would be of great advantage from the points of view of size, reliability and, not least, the power supplies required.

The provision of an intra-Commonwealth telephone network of the type envisaged is no small undertaking but it is one which could play a leading part in bonding the member countries closer together and is, fortunately one in which Great Britain is technically equipped to play a leading rôle.

Some Novel Circuits Employing Cold-Cathode Tubes

(Part 1)

By R. S. Sidorowicz*, Dipl.Ing., D.I.C., A.M.I.E.E.

The circuits described are based on three types of cold-cathode tube: (1) a stable subminiature diode, type XC12, (2) a subminiature triode, type XC18, having a current rating of 1·0mA, and (3) a miniature 7·5mA triode, type XC23. The diode is used to stabilize the anode breakdown voltage of the triodes, and this technique leads to the development of a stable relaxation oscillator, a voltage discriminator or a delay circuit and two satisfactory monostable circuits. The 7·5mA tubes are employed in a special bistable circuit and in a univibrator. These circuits are capable of supplying currents of up to 8mA and are used in controlling of the 1·0mA triodes. This technique permits the development of a staircase waveform generator, a rectangular pulse generator, a fast decade counter, a six-tube decade counter and a five-tube decade.

THE aim of this article is to describe a number of novel circuits based on cold-cathode glow discharge tubes. The circuits are referred to as 'novel' not because they perform some novel functions, but on account of their special features. The circuits can be divided into two groups. The first group is characterized by the introduction of certain refinements into a number of standard, well-known circuits. These refinements result in an improvement of the stability and reliability of the circuits. The main feature of the second group is the combining of the tubes of two different sizes, i.e., the tubes whose current ratings differ by about one order of magnitude. The larger tubes in these circuits are employed to control the smaller tubes. The technique has led to the development of a number of novel circuits which have some useful properties.

It is not intended to discuss here the details of the design techniques of the circuits or to determine the optimum values of the circuit components. The reader who may be interested in the design aspects of cold-cathode tube circuits is referred to the papers of G. H. Hough and D. S. Ridler[1], R. Threadgold[2], J. E. Flood and J. B. Warman[3] and G. O. Crowther and K. F. Gimson[4]. Also, the physics of cold-cathode gas discharges will not be dealt with, since ample information on the subject can be found in a number of textbooks[5,6,7,8,9]. It may be useful, however, to consider some general properties of cold-cathode glow-discharge tubes.

Cold-cathode diodes and triodes can be divided into two classes. The tubes of the first class are characterized by breakdown voltages ranging from about 120 to about 350V and maintaining voltages of about 90 to 140V. The main physical feature of these tubes is that they are provided with pure-metal cathodes. The tubes of the second class are furnished with oxide-coated photo-sensitive cathodes, and have breakdown voltages ranging from about 75 to 250V and maintaining voltages of 55 to 80V.

The tubes of the first class have stable maintaining voltages and are normally employed as stabilizers and reference-voltage sources. However, their breakdown voltage is not very stable and they are also subject to statistically distributed ignition delays[1]. The tubes are therefore unsuitable for switching applications, where they have to be triggered by pulses, unless they are provided with an adequate source of priming electrons. One of the methods of supplying the required priming electrons is to furnish the tube with a quantity of radium or radioactive tritium. It appears however that the preferred method of priming consists of providing the tube with an additional discharge gap. The auxiliary discharge is kept on permanently and it results in a stable breakdown voltage of the tube. It should be clear that an auxiliary discharge gap makes the tube more complex; thus, in effect, a diode becomes a triode, and a triode becomes a tetrode. It would be wrong, however, to compare the auxiliary gap of a cold-cathode tube with the heater of a vacuum valve, since the gap consumes negligible power.

The priming electrons in the tubes of the second class are provided by the photo-sensitive cathode which has a comparatively low work function. However, the tubes must be provided with a transparent envelope and should be adequately illuminated. The normal ambient illumination is quite sufficient for a satisfactory operation of the tubes. It is necessary, however, to avoid direct sunlight, since this brings down the breakdown voltages; also, the tubes cannot be operated in complete darkness, since their breakdown voltages go up and become unstable (random in magnitude).

The circuits described below are based principally on two tubes with the oxide-coated cathodes and an auxiliary tube having a pure-metal cathode.

Characteristics of the Tubes Employed

The main electrical parameters of the tubes are listed in Table 1. Tube, type XC18, is a subminiature triode having a maximum continuous operating current of 1mA. Tube XC23 is a miniature triode with the maximum continuous current of 7·5mA. In many switching or pulse applications, these current ratings can be exceeded two-fold or five-fold, provided the tubes are given an adequate rest period between successive pulses. Both the tubes are provided with oxide-coated photo-sensitive cathodes and are filled with an argon-neon mixture. Envelopes of the tubes are transparent, so that their glow is clearly visible.

The meaning of most of the parameters in Table 1 should be fairly clear, but a word of explanation may be necessary with regard to the recovery time, t_x, and the dynamic transfer sensitivity, V_p. These parameters are measured in the circuit shown in Fig. 1. For the measurement of V_p the tube is given an anode potential of $V_b = 140$V and a rectangular pulse having a duration of 75μsec is applied to the trigger electrode via a ·001μF capacitor. The amplitude of the pulse is adjusted to a value V_p, such that the trigger-cathode gap of the tube will strike and the anode-cathode or main gap will become ignited. The recovery time, t_x, is measured in the same circuit except that the trigger is connected to earth via a 470kΩ resistor and the anode supply potential is in-

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creased to $V_b = 180$V. The tube is struck and the anode resistor R_a is adjusted so the tube carries the nominal current (1mA or 7·5mA). Negative rectangular pulses are then superimposed on the supply voltage; the amplitude of the pulses is such that the voltage varies from $V_b = 180$V to $V_{b1} = 50$V. The duration of the pulses is adjusted to a value t_x at which the tube is extinguished. More detailed information on the measurement of the transfer sensitivity and the recovery time can be found in several recently published papers[1,3,10].

The tube, type XC12, is a subminiature diode with a pure-metal cathode. This tube has a stable maintaining voltage and can be used as a reference voltage source or a stabilizer. It is subject however, to statistical ignition delays. Consequently, a tritiated version of the tube (type XC12T) is available, in which the ignition delays are reduced to a negligible value.

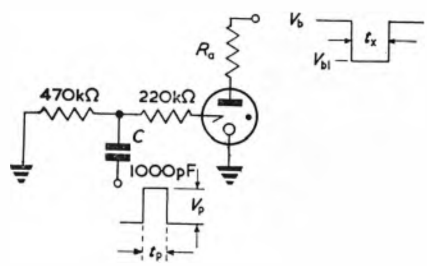


Fig. 1. Test circuit for the investigation of the dynamic characteristics of cold-cathode triodes

The Relaxation Oscillator

The basic circuit of a relaxation oscillator, based on a cold-cathode diode, is shown in Fig. 2. The circuit produces a positive pulse across its cathode resistor R_1 and a negative pulse across R_2. The pulses have an exponential tail with a time-constant $\tau = C(R_1 + R_2)$ and, provided the value of $R_1 + R_2$ is not too low, their combined amplitude is equal to $V_{max} = V_s - V_m$, where V_s is the breakdown potential and V_m is the maintaining voltage of the tube. The interval between two successive pulses is expressed by

$$T = RC \ln \frac{V_b - V_m}{V_b - V_s} \quad \ldots\ldots \quad (1)$$

TABLE 1

TUBE TYPES / PARAMETERS	XC18 (SUB-MINIATURE) (HIVAC)	XC23 (MINIATURE) (HIVAC)	XC12 OR XC12T (HIVAC)
Main gap (anode-cathode) breakdown voltage, V_{as}	210V min.	210V min.	104V (nominal)
Nominal main gap maintaining voltage, V_{am}	73V at 1mA (± 5V spread)	70V at 7·5mA	85V at 1mA
Nominal control gap (trigger-cathode) breakdown voltage, V_{ts}	68V (± 6V spread)	68V	
Nominal control gap maintaining voltage, V_{tm}	55V	55V	
Maximum continuous current I_k	1mA	7·5mA	1mA
Maximum transfer current (d.c.) I_t	30μA	90μA	
Recovery time, t_x	1msec max	1·8msec max	
Dynamic transfer sensitivity, V_p	85V max	85V max	
Dimensions	34×10·5mm	45×19·1mm	28×6·75mm

From equation (1) it is seen that, apart from the time-constant RC, the period of oscillation depends on three voltages: V_b, V_s and V_m. Assuming that V_b is stable, the frequency stability is dependent on V_s and V_m. In the tubes described above, the stability of V_m is quite satisfactory. On the other hand, the stability of V_s in an XC12 or even in an XC12T is inadequate, especially if it is required to operate the oscillator at frequencies of the order

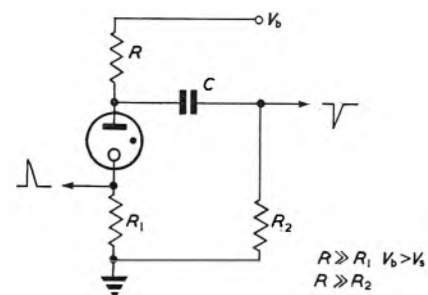


Fig. 2. Relaxation oscillator based on a cold-cathode diode

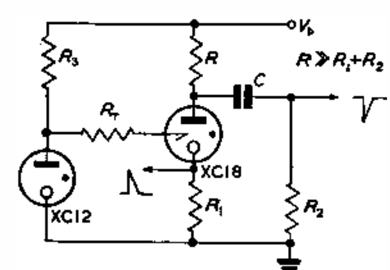


Fig. 3. Stabilized relaxation oscillator

of a few cycles/min or a few cycles/sec. At higher frequencies (over 50c/s) the stability becomes more satisfactory, since the tubes do not deionize completely between the successive discharges.

The triodes XC18 and XC23 can be used in the oscillators in two ways. The oscillatory circuit can be inserted in the trigger while the anode, with an RC network, acts as a pulse amplifier; a modification of this arrangement can be used as a delay circuit[11]. Alternatively, the trigger can be connected to earth via a high resistance and the oscillatory circuit inserted in the anode. Again, the stability of such oscillators is not satisfactory, especially at the low frequencies. The main reason for the instability is the dependence of V_s on ambient illumination. The situation can be remedied, if a constant priming current is maintained in the trigger-cathode gap of a triode, and its main gap is used as the oscillator. A circuit based on this principle is shown in Fig. 3.

It is seen that the trigger is biased by means of an XC12 or an XC12T. The maintaining potential of these tubes is higher than the trigger maintaining voltage of XC18 or XC23. Since V_m of the XC12 is very stable, a constant priming current can be maintained in the trigger gap of the triodes. This results in a stable breakdown voltage, V_{as}, of the main discharge gap.

In practical oscillator circuits, the resistor R_3 should be such that the XC12 passes about 0·5 to 0·75mA. V_b should be higher than 150V, since V_{as} of a triode primed by means of an XC12 is about 130 to 150V. If the oscillator employs an XC18, the circuit components should have the following values: $R \geqslant 2{\cdot}7\text{M}\Omega$, $R_1+R_2=6{\cdot}8$ to $15\text{k}\Omega$ and $R_T=1{\cdot}5$ to $0{\cdot}82\text{M}\Omega$. If the circuit is based on an XC23, $R \geqslant 1{\cdot}5\text{M}\Omega$, $R_1+R_2 = 33$ to $3{\cdot}3\text{k}\Omega$ and $R_T = 0{\cdot}82$ to $0{\cdot}47\text{M}\Omega$. The above values of $R_1 + R_2$ ensure that the combined amplitude of the output pulses is approximately equal to $(V_{as} - V_{am})$. R and C can be made very large, but it is important to use capacitors whose leakage is much higher than R. By a suitable choice of CR, it is possible to design satisfactory circuits with the oscillation periods as long as several minutes. On the other hand, CR should not be too low. This is undesirable for the following reason; if the oscillator operates at several kilocycles per second, the residual ionization between successive discharges tends to lower V_{as}; the output pulses are thereby considerably reduced in amplitude.

Apart from using the oscillator of Fig. 3 as a pulse generator, it is possible to modify it into a limit circuit or an amplitude discriminator[12]. If the overall amplitude of the input voltage is higher than V_{as}, the circuit can be in the form shown in Fig. 4. In this circuit, the d.c. component, V_1, of the input signal, $v_1(t)$, should be lower than V_{as}. The triode is therefore not discharging until the transient component, $u_1(t)$, lifts the voltage to V_{as}. This results in the appearance of a positive pulse across R_1. If after reaching V_{as}, the input voltage continues to rise, the circuit will produce further pulses. It should be realized that unless $u_1(t)$ is a comparatively slow waveform, there may be a considerable time lag between the input voltage and the potential appearing across C. If the d.c. component is disregarded, the relationship between $u_1(t)$ and the voltage across C can be expressed by the Duhamel integrals

$$v_c(t) = 1/RC \int_0^t e^{-\tau/RC}\, u_1(t-\tau)\, d\tau \quad \ldots\ldots \quad (2)$$

or

$$v_c(t) = 1/RC\; e^{-t/RC} \int_0^t e^{\tau/RC}\, u_1(\tau) d\tau \quad \ldots\ldots \quad (3)$$

For a step input transient having an amplitude U_1, the voltage across C is

$$v_c(t) = U_1(1 - e^{-t/RC}) \quad \ldots\ldots\ldots\ldots\ldots \quad (4)$$

If C was previously biased to U_o, the total voltage is

$$v_c(t) = U_o + U_1(1 - e^{-t/RC}) \quad \ldots\ldots\ldots\ldots \quad (5)$$

from which it follows that the tube will fire after a delay

$$t_1 = RC \ln \frac{U_1}{U_1 + U_o - V_{as}} \quad \ldots\ldots\ldots\ldots\ldots \quad (6)$$

It should be clear from the above that the arrangement of Fig. 4 can also be used as a delay circuit.

The input signals of a low amplitude can be applied to the modified limit circuit of Fig. 5, in which the capacitor is charged to a voltage $U_o = V_b R_6/(R_6 + R_5)$, such that $U_o < V_{as}$. The transient component appearing across C can be expressed by

$$v_c(t) = \frac{R_5}{[(R_5+R_6)R + R_5R_6]\, C} \int_0^t e^{-\tau/\beta}\, u_1(t-\tau)\, d\tau \quad \ldots . (7)$$

where

$$\beta = \left(R + \frac{R_5R_6}{R_5 + R_6}\right) C \quad \ldots\ldots\ldots\ldots \quad (7a)$$

or, if the input is a step having an amplitude U_o,

$$v_c(t) = \frac{U_o R_5}{R_5 + R_6}(1 - e^{-t/\beta}) \quad \ldots\ldots\ldots\ldots \quad (8)$$

The circuits of Figs. 4 and 5 can be used as upper-level voltage discriminators or as delay circuits. It is also possible to modify the relaxation oscillator of Fig. 3 into a circuit which will produce positive or negative pulses whenever the input voltage falls below a certain value. A lower-level voltage discriminator of this type[12] can be combined with the upper-level discriminator (plus a number of auxiliary circuits) to design a simple reversal or limit switch. In the following section, the upper-level discriminator will be used as a stable delay circuit.

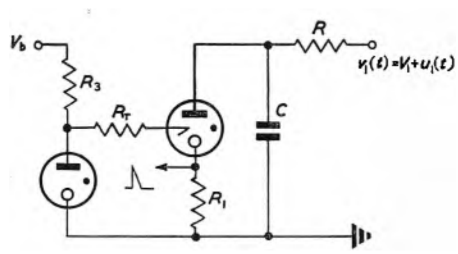


Fig. 4. An upper-level limit circuit for large voltages

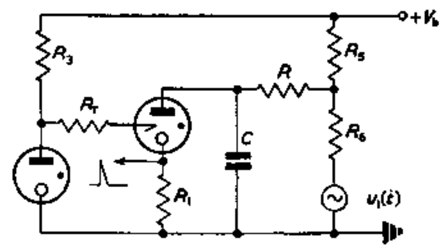


Fig. 5. An upper-level limit circuit for low voltages

Monostable Circuits

The monostable circuits (univibrators) based on cold-cathode tubes are derived from the two basic bistable circuits[1,3,4] shown in Figs. 6 and 7. Operation of the circuit of Fig. 6 is as follows. Initially, V_1 is conducting and V_2 is off; voltage at the anode of V_1 is V_m and the capacitor C is charged to a voltage $(V_b - V_m)$. A suitable pulse applied to the triggers of the tubes via the capacitors C_1 strikes V_2. This results in the anode voltage of V_2 being instantly lowered to V_m; the anode voltage of V_1 is therefore reduced to $(2V_m - V_b)$, which is lower than V_m. Provided the time-constant $T_1 = R_1C$ is sufficiently large, V_1 will be extinguished and C will be gradually charged to a voltage $-(V_b - V_m)$. The waveforms illustrating the operation of the circuit are also shown in Fig. 6.

In the circuit of Fig. 7, V_1 is initially conducting, while V_2 is off. The cathode of V_1 is at a potential $V_k = (V_b - V_m)R_k/(R_k + R_a)$, and the cathode of V_2 is at earth potential. The anodes are at a potential $V_a = V_m + V_k$. The application of a positive pulse to the triggers of the tubes results in the ignition of V_2. The anode voltage is thereby instantly lowered to V_m and, if the capacitance C is sufficiently large, V_1 will be extinguished. The cathode voltage of V_2 gradually rises to V_k, and that of V_1 drops to earth potential. The waveforms generated by the circuit and the time-constants governing them are given in Fig. 7.

The circuit of Fig. 6 can easily be modified into a monostable circuit. The resulting univibrator is shown in Fig. 8. In the stable state, V_2 is conducting while V_1

is off. The anode of V_1 has the full supply voltage V_b, while the anode of V_2 has a potential V_m. The trigger of V_2 is at a certain potential V_t which is dependent on the discharge conditions in the tube and is also a function of R_T and R. If V_1 is ignited by a positive pulse, its anode voltage drops to V_m and that of V_2 falls to $(2V_m - V_b)$. V_2 is thus extinguished and its anode voltage rises exponentially to V_b (see the waveforms of Fig. 8). Voltage of the trigger of V_2 will first drop to a value below V_t, but it will then gradually rise (depending on the time-constant $T = RC$), until it reaches the trigger strike voltage V_{ts}. At this point, V_2 will re-strike and V_1 will be extinguished. After a period depending on the time-constant $T_2 = R_2C_1$, the circuit is ready to accept another triggering pulse. It may be worth mentioning that the timing resistance R can also be connected directly to V_b or to any point on the anode resistor R_2.

It should be clear that, if the circuit of Fig. 8 is based on a pair of XC18 or XC23, its stability will not be very satisfactory, especially when it will be required to generate

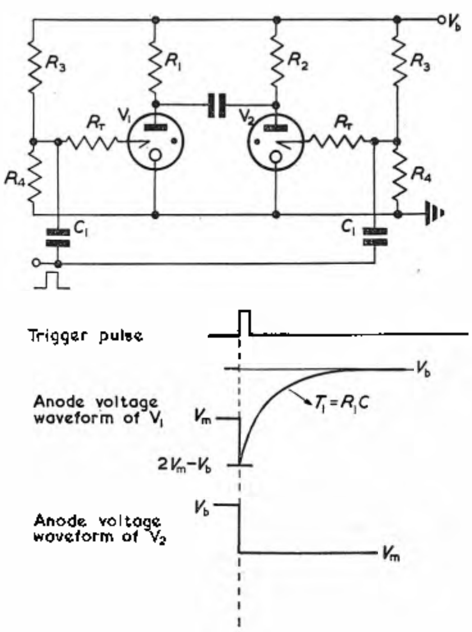


Fig. 6. A bistable pair with the quenching capacitor in the anodes

pulses of long duration (0·1sec to 1min). The situation can be remedied, if the univibrator employs a timing network based on the circuit of Fig. 4. The resulting monostable circuit is then in the form shown in Fig. 9.

The operation of this univibrator is essentially similar to that of Fig. 8, except that the duration of the monostable state is controlled by the limit circuit consisting of V_3 and V_4. The input signal to the RC network is taken from the anode of V_2, so that in the steady state the capacitor C is charged to the maintaining potential of V_2. The timing resistance R can also be connected to a tapping point on the resistor R_2, but the point should be chosen in such a manner that its steady state potential does not exceed the breakdown voltage, V_{as}, of V_3. When V_2 is extinguished, the potential across C rises gradually until it reaches the breakdown voltage of V_3. At this point, V_3 strikes and generates a positive pulse which re-strikes V_2. The cycle is thus completed.

It is possible to describe the operation of the univibrator of Fig. 9 analytically. Assuming that V_1 and V_2 are identical (i.e. $V_{m1} = V_{m2} = V_m$) and that R is connected

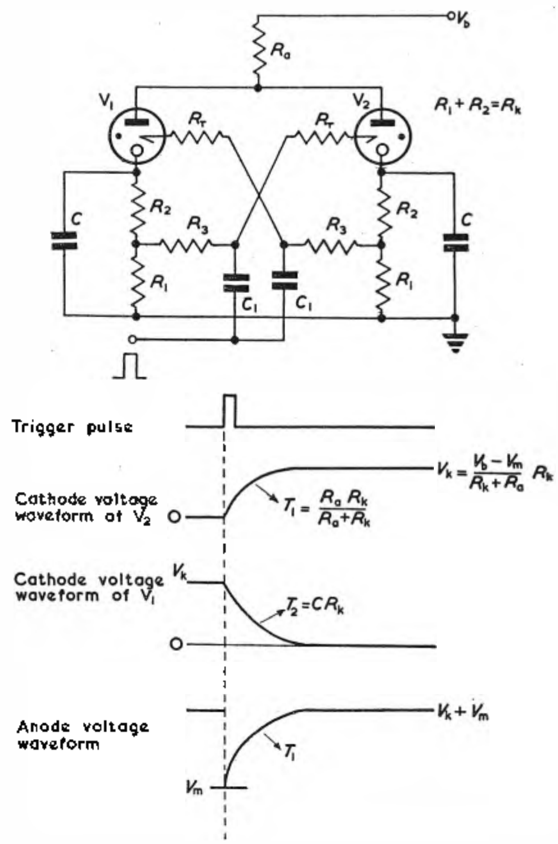
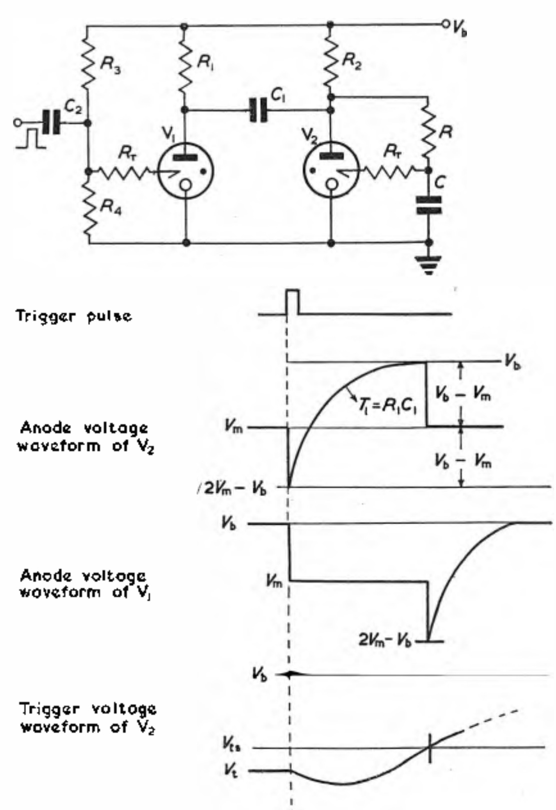


Fig. 7. A bistable pair with quenching capacitors in the cathodes

Fig. 8. Monostable circuit based on the binary pair of Fig. 6



to the anode of V_2, the waveform generated at the anode of V_2 when V_1 is ignited can be expressed by

$$v_a(t) = -(V_b - V_m) + 2(V_b - V_m)\,[1 - e^{-t/R_1C_1}] \quad \ldots\ldots (9)$$

where the first term represents a negative voltage step and the second term describes the exponentially rising portion of the waveform (see Fig. 8). In equation (9) it is also assumed that $v_a(t)$ is measured with respect to the steady-state component which, in this case, is equal to V_m. Voltage

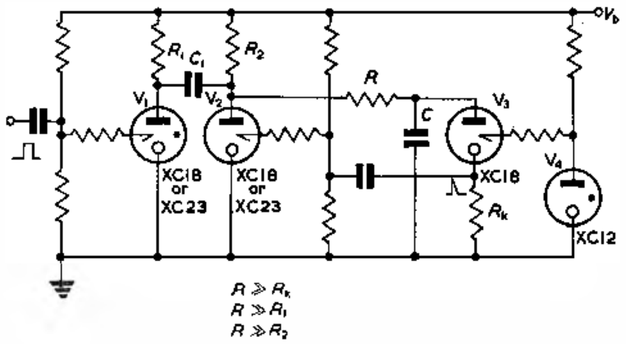


Fig. 9. Stabilized monostable circuit based on the binary pair of Fig. 6

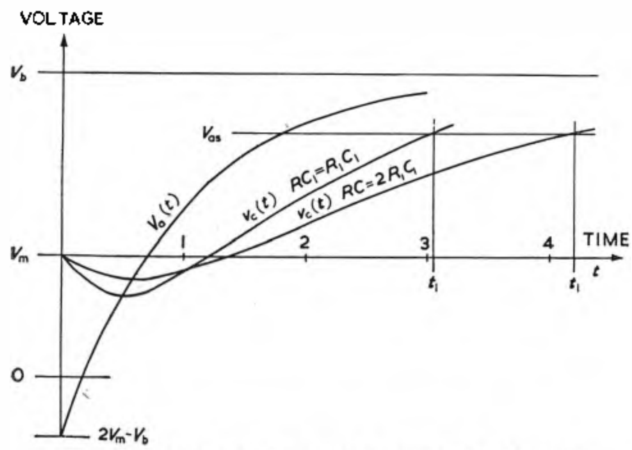


Fig. 10. Voltage waveforms of the univibrator of Fig. 9

across C is given by

$$v_c(t) = 1/RC \int_0^t e^{-\tau/RC}\, v_a(t-\tau)\,d\tau \quad \ldots\ldots (10)$$

or explicitly by

$$v_c(t) = (V_b - V_m)\,[1 - e^{-t/RC}] + \frac{2(V_b - V_m)}{RC/(R_1C_1) - 1}\,[e^{-t/R_1C_1} - e^{-t/RC}] \quad \ldots\ldots (11)$$

or if $RC = R_1C_1$,

$$v_c(t) = (V_b - V_m)\,[1 - e^{-t/RC}] - \frac{2(V_b - V_m)}{RC}\,t\,e^{-t/RC} \quad \ldots (12)$$

The breakdown of V_3 occurs after a time t_1, when $v_c(t_1) = V_{as} - V_m$. Unfortunately it is impossible to evaluate t_1 explicitly, since equation (11) is transcendental. It is comparatively easy, however, to determine t_1 graphically. The procedure is illustrated in Fig. 10, where it is assumed that $V_b = 2{\cdot}5V_m$ and $V_{as} = 2V_m$; Fig. 10 shows $v_c(t)$ for $RC = R_1C_1$ and for $RC = 2R_1C_1$, and also a curve of the anode voltage of V_2.

It can be seen from the above, that the univibrator of Fig. 9 can be designed analytically without much difficulty. In general, all the necessary parameters (V_b, V_m and V_{as}) are either known or can easily be measured. On the other hand, a similar design procedure is not possible with the circuit of Fig. 8 where V_t is not known beforehand. Consequently, the univibrator of Fig. 9 compares favourably with the circuit of Fig. 8 in that its stability is more satisfactory and its design is simpler.

It has been mentioned that it is also possible to modify the circuit of Fig. 7 into a univibrator. This can be done by connecting a timing network to the trigger of one of the triodes. Since such a circuit would have all the disadvantages of the univibrator of Fig. 8, it should be preferable to employ a timing device similar to that of Fig. 4 or 5. The resulting univibrator is shown in Fig. 11. The functioning of the circuit is briefly as follows. In the steady state V_2 is conducting, so that its cathode is at

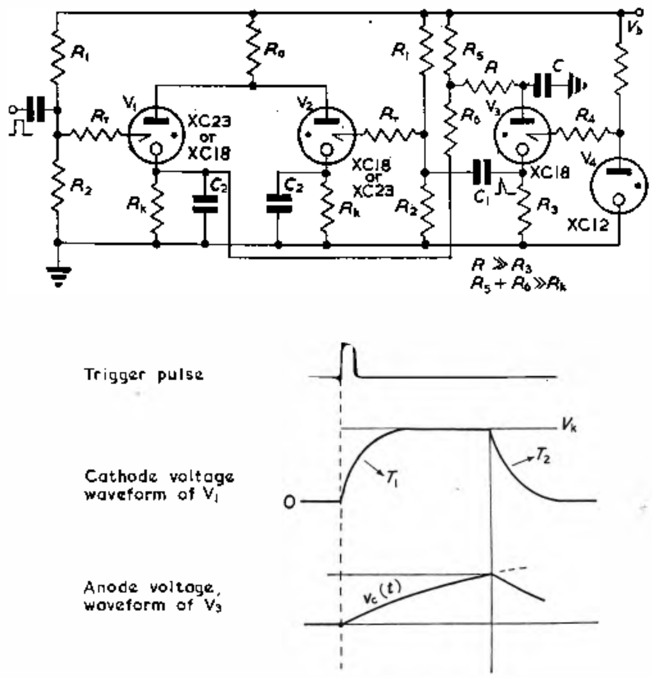


Fig. 11. Stabilized univibrator based on the binary pair of Fig. 7

a voltage $V_k = (V_b - V_m)R_k/(R_k + R_a)$; the cathode of V_1 is at earth potential and the anode of V_3 is biased through R_5 and R_6 to a voltage V_a, such that $V_a < V_{as}$. When V_1 is ignited by an external pulse, V_2 is extinguished. The cathode potential rises exponentially, as expressed by

$$v_k(t) = V_k\,(1 - e^{-t/T_1}) \quad \ldots\ldots\ldots\ldots (13)$$

where

$$T_1 = \left(\frac{R_aR_k}{R_k + R_a}\right) C_2 \quad \ldots\ldots\ldots\ldots (13a)$$

and the anode voltage of V_3 increases by an amount equal to

$$v_c(t) = \frac{R_5}{(R_5 + R_6)\beta} \int_0^t e^{-\tau/\beta}\, v_k(t-\tau)\,d\tau \quad \ldots\ldots (14)$$

where β is given by equation (7(a)). By substituting equation (13) into equation (14), the following expression for $v_c(t)$ is obtained

$$v_c(t) = \frac{R_5V_k}{R_5 + R_6}\,(1 - e^{-t/\beta}) + \frac{R_5V_k}{(R_5+R_6)(1-(\beta/T_1))}\,[e^{-t/\beta} - e^{-t/T_1}] \quad \ldots\ldots (15)$$

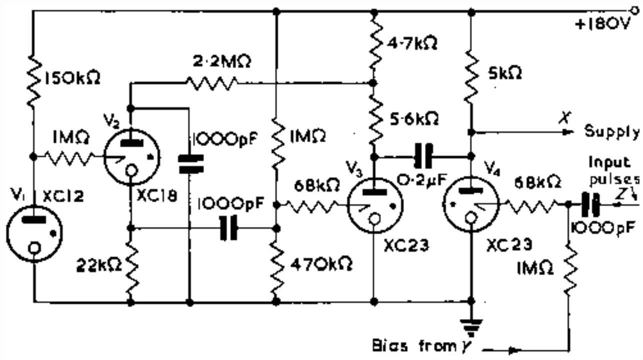


Fig. 12. A practical univibrator circuit

or, if $\beta = T_1$

$$v_e(t) = \frac{R_5 V_k}{R_5 + R_6}(1 - e^{-t/\beta}) + \frac{R_5 V_k}{R_5 + R_6} t\, e^{-t/\beta} \quad \ldots . (16)$$

The tube V_3 strikes after a time t_1, when $v_e(t_1) = V_{ss} - V_m$. The positive pulse obtained from the cathode of V_3 re-ignites V_2 and results in the extinction of V_1. The cycle is thus completed. As in the circuit of Fig. 9, the duration of the quasi-stable state, t_1, cannot be evaluated explicity from equations (15) or (16). Again, the equations can be solved graphically without much difficulty.

The univibrators of Figs. 9 and 11 can be used in various switching systems and timing devices. The decision as to which of the univibrators is preferable, will depend on a specific application. Where positive output pulses are required, the circuit of Fig. 11 will be useful. On the other hand, negative pulses having sharp leading edges can be obtained from the circuit of Fig. 9. An example of a practical univibrator is shown in Fig. 12. The circuit is based on two 'large' cold-cathode triodes (type XC23) and it is used to control a number of small tubes (type XC18). The duration of the quasi-stable state in the circuit is about 4msec.

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(To be continued)

Miniature Interdigital Magnetrons

At a recent exhibition, held by Mullard Ltd, of microwave and radar valves, a paper presented by N.N.G. King described a new miniature interdigital magnetron.

Magnetrons are normally considered as large expensive and high powered tubes, but some time ago it was decided to look into the possibilities of producing a valve that was small, light-weight and cheap, while still retaining the magnetron's high efficiency. A particular application might be in radar-sonde balloon equipments where a peak output of a few hundred watts is required and which use, at the moment, disk-seal triodes of much lower efficiency. The magnetron would need to be considerably more efficient than these while still being competitive in size, weight and cost. A target specification, therefore, was set to give a magnetron of greater than 200W peak output at 2850Mc/s with a pulse input of 2kV, $\frac{1}{2}$A.

The design of this valve can be divided into two parts:

(a) The interaction-space, or electronic part.

(b) The resonant system, or r.f. circuit of the valve.

The interaction-space design, and in this is included magnetic field as well as anode and cathode geometries, was based on normal magnetron practice using an eight-segment anode excited in the π-mode, that is with alternate segments π radians out of phase. The requirements of minimum weight and highest efficiency are incompatible, since electronic efficiency rises with magnetic field, while the total valve weight is very largely that of its permanent magnet. Magnet weight was kept down, however, by using the smallest possible pole piece separation (magnet weight varies at the cube of this dimension) and this requirement was borne in mind when choosing the input conditions for the tube, since interaction-space axial length is inversely proportional to the input current, with the voltage and all other parameters constant.

For the anode system something very different to the conventional multi-cavity machined or brazed copper structure was needed, since this is both heavy and expensive to produce. Ideally one would like to avoid the use of any machined components in the tube and so a resonant system was designed which could be fabricated from blanked-out pieces of copper brazed together in a simple jig. The resonant element was a quarter-wavelength shorted lecher line, the required r.f. field pattern round the cathode being obtained by brazing the digits to alternate sides of the lecher line. Due to the singular resonant element this structure has fewer unwanted modes of oscillation than a multi-resonator system, and these modes have frequencies higher than and well separated from that of the desired mode.

There is, of course, one role which this anode structure does not fulfil and that is keeping out the air. However, from the start of this work it was thought that a glass envelope valve had much to recommend it, in ease of construction and cheapness although, since it is convenient to have the magnet pole-pieces outside the glass envelope it does lead to rather large pole-piece spacing and hence a large magnet. This was avoided in the present design by tailoring the envelope to fit the anode structure, and it is made from rectangular glass tube. The output is taken from the resonator by means of a loop and coaxial line of 70Ω impedance. A thin copper shield is placed round the centre part of the envelope to prevent radiation from the anode structure and also to serve as a mounting for the pole-pieces and magnet. The complete packaged valve weighs less than 4oz.

Results for this valve have been very satisfactory and when packaged with the small magnet a power in excess of 200W is achieved at the 2kV, $\frac{1}{2}$A point. With a larger magnet powers of up to 3kW are available. The duty cycle is governed by the anode dissipation limit of 4W. A typical value for the pulling figure is 20Mc/s for a 1·5:1 voltage standing wave ratio.

Work has now been started on an X-band interdigital magnetron along rather similar lines to the one described above and the target being aimed at is a valve suitable for very low-power marine radar applications where only a few kilowatts are needed. The major difference from the S-band valve is the use of a pill-box cavity resonator in place of the lecher system; by this means the unloaded Q is increased, giving better circuit efficiency at a moderate pulling figure. Another difference is in the output system, where a horn is used radiating through the glass envelope directly into X-band waveguide. It is not envisaged that this valve will be of the packaged type, since the magnet needed will be quite large, weighing several pounds, but it is hoped that the glass-envelope magnetron proper should be quite cheap to produce and be a serious competitor to more conventional valves.

The Differential Transformer as a Sensitive Measuring Device

By J. H. Heath*, M.Sc.Eng., A.M.I.E.E.

In the measurement of physical quantities such as strain acceleration etc, it is necessary to convert the mechanical movement into an electrical signal in order that use can be made of electronic amplifiers. This article is the result of a study made into a very sensitive measuring apparatus using a differential transducer as a sensing head and in which phase changes associated with the transducer output voltages have little effect upon the measuring system.

MOST high sensitivity measuring techniques involve the conversion of a mechanical movement (strain, acceleration, etc.) into an electrical signal in order that full use can be made of electronic amplifiers. From time to time various methods have been proposed[1] but among the very many systems that have been tried as transducers, three main methods are at present in general use:

(a) The change in resistance with strain.
(b) The change in inductance with the movement of some part of the transducer or its associated components.
(c) The change in capacitance with movement.

In order to detect the electrical signal produced in any of the above transducers the technique generally used comprises a balanced a.c. bridge network with the transducer as active arm(s) and resistors in the passive arms; by this means stability is obtained. When the system is in its passive state the bridge delivers zero output and in order to achieve this, both the resistive and reactive components must be equal[2] so that the process of balancing the bridge involves two distinct operations. Therefore, if a measuring system is designed in which the phenomenon is to be measured by means of a succession of highly sensitive ranges, each succeeding range must have resistive and reactive balance controls so that the zero of each range may coincide with the maximum of the previous one.

Greater flexibility in the design of transducer measuring circuits would be obtained if the reactive balance could be eliminated (or reduced), and this article analyses a method whereby this has been achieved.

The Differential Transformer[1]

This article is the result of a study made into a very sensitive measuring apparatus using a differential transducer as a sensing head. This inductive transducer consists of a primary winding coupled to two equal secondaries, the voltages in the two secondaries being produced by a movable dust core coupling both primary and secondaries forming a highly stable and sensitive instrument. The three windings can be connected in a variety of configurations, but only two will be examined here.

Consider Fig. 1(a), the primary L_p, R_p is coupled to the two secondaries S_1 and S_2, connected in opposition so that at one position of the dust core d, the output will be zero. If the two secondaries are not exactly equal there will be a residual voltage at E_{out} which cannot be eliminated.

By phase correcting the secondary voltages with r_1 balance conditions can be obtained. The zero output voltage is a function of the core position and as such limits the usefulness of the circuit; by injecting into the transformer secondaries a suitably phase corrected voltage, the electrical zero can be made independent of the core movement[3]. Neglecting the phase correcting potentiometer, the analysis is as follows:

By Kirchoff's Laws Fig. 1(a) yields:

$$\left.\begin{aligned} E\sin\omega t + I_p(R_p+DL_p)-(M_1+M_2)DI &= 0 \\ I_pD(M_2-M_1)+(2DL+2R+R_g)I &= 0 \end{aligned}\right\} \quad \ldots\ldots (1)$$

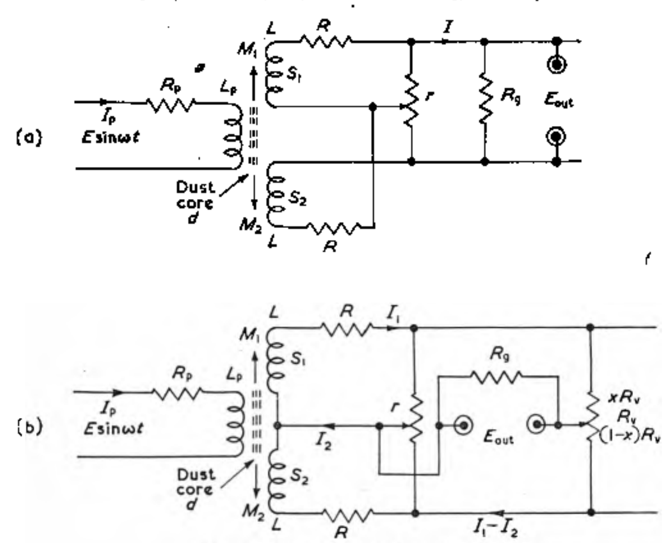

Fig. 1. The inductive transducer

From which,

$$I = \frac{(M_1-M_2)E\omega\cos\omega t}{(R_p+DL_p)(2DL+2R+R_g)+D^2(M_2-M_1)^2} \quad \ldots (2)$$

and E_{out} =

$$\frac{ER_g\omega(M_1-M_2)\sin(\omega t+\phi)}{\sqrt{[(R_p(2R+R_g)-2\omega^2LL_p-\omega^2(M_2-M_1)^2)^2+\omega^2(2L_p+L_p(2R+R_g))^2]}} \quad \ldots\ldots (3)$$

where

$$\phi = \tan^{-1}\frac{\omega(2LR_p+L_p(2R+R_g))}{R_p(2R+R_g)-2\omega^2LL_p+\omega^2(M_1-M_2)^2} \quad \ldots\ldots (4)$$

when R_g approaches ∞,

$$E_{out} = \frac{E\omega(M_2-M_1)\sin(\omega t+\phi)}{\sqrt{(R_p^2+\omega^2L_p^2)}} \quad \ldots\ldots (5)$$

and

$$\phi = \tan^{-1}(\omega L_p/Rp) \quad \ldots\ldots\ldots\ldots (6)$$

In Fig. 1(b) the two secondaries are connected in series so that in effect the whole secondary is centre-tapped.

* *U.K.A.E.A., formerly Admiralty.*

Again:

$$E \sin \omega t + I_p(R_p + L_pD) - M_1DI_1 - M_2(I_1 - I_2) = 0 \quad \ldots\ldots (7)$$

$$-M_1DI_p + LDI_1 + (R + xR_v)I_1 + I_2R_g = 0 \quad \ldots\ldots (8)$$

$$-M_2DI_p + LD(I_1 - I_2) + (R + (1 - x)R_v)(I_1 - I_2) - I_2R_g = 0 \quad \ldots\ldots (9)$$

The phase balance potentiometer r is neglected as its inclusion would only complicate the analysis and would not contribute to the argument.

$$E_{out} \text{ will be } I_2R_g = \frac{\Delta I_2 \,.\, R_g}{\Delta} \quad \ldots\ldots (10)$$

which can be shown to be:

$$\frac{\left[E\omega R_g \cos \omega t\left\{(R + LD)(M_1 - M_2) + R_v(M_1 - x(M_1 + M_2))\right\}\right]}{\left[(LD + R + xR_v)\left\{-M_2^2D^2 + (R_p + L_pD)(LD + R + (1-x)R_v + R_g)\right\} + \left\{LD + R + (1-x)R_v\right\}\left\{R_g(L_pD + R_p) + M_1M_2D^2\right\} - D^2(M_1 + M_2)\left\{M_2R_g + M_1(LD + R + R_g + (1-x)R_v)\right\}\right]} \quad \ldots (11)$$

If no load is drawn from the bridge $R_g \to \infty$ and

$$E_{out} = \frac{E\omega \cos \omega t\left\{(R + LD)(M_1 - M_2) + R_v(M_1 - x(M_1 + M_2))\right\}}{(R_p + L_pD)(2LD + 2R_v) - D^2(M_1 + M_2)^2} \quad \ldots\ldots (12)$$

Now $(R_p + L_pD)(2LD + 2R + R_v) \gg -D^2(M_1 + M_2)^2$

Thus $E_{out} \simeq$

$$\frac{E\omega \cos \omega t\left\{(R + LD)(M_1 - M_2) + R_v(M_1 - x(M_1 + M_2))\right\}}{(R_p + L_pD)(2LD + 2R + R_v)} \quad \ldots\ldots (13)$$

$$\simeq E\omega \sin \omega t \times \sqrt{\left[\frac{\left\{(M_1 - M_2)R + R_v(M_1(1 - x) - xM_2)\right\}^2 + \omega^2L^2(M_1 - M_2)^2}{(R_p^2 + L_p^2\omega^2)([2R + R_v]^2 + 4\omega^2L^2)}\right]} \times e^{j\phi} \quad \ldots\ldots (14)$$

where

$$\phi = (3\pi/2) + \tan^{-1}(\omega L_p/R_p) + \tan^{-1}\frac{2\omega L}{2R + R_v} - \tan^{-1}\frac{\omega L(M_1 - M_2)}{R(M_1 - M_2) + R_v(M_1 - x(M_1 + M_2))} \quad \ldots\ldots (15)$$

when $x = \frac{1}{2}$.

$$E_{out} \simeq \frac{E_\omega(M_2 - M_1)\sin \omega t}{2\sqrt{(R_p^2 + \omega^2L_p^2)}} \,.\, e^{j\phi} \quad \ldots\ldots (16)$$

and

$$\phi = (3\pi/2) + \tan^{-1}(\omega L_p/R_p) \quad \ldots\ldots (17)$$

Thus when the slider on R_v is in the centre position E_{out} can be reduced to zero by moving the core to a point where M_1 equals M_2; if, however, this is undesirable, x can be moved a distance such that from equation (13):

$$x \simeq \frac{1}{(M_1 + M_2)R_v}\left\{R(M_1 - M_2) + M_1R_v \pm \sqrt{[M_1RR_v(M_2 - M_1)]}\right\} \quad \ldots\ldots (18)$$

and ϕ will be given by equation (15) when x is substituted according to equation (18). Thus E_{out} can always be reduced to zero by an adjustment to the voltage potentiometer R_v combined with the phase control. r.

Use of Differential Transformer for the Measurement of Small Increments in Displacement

When large displacements have to be measured to a high degree of accuracy as in the case of investigations into the creep testing of metals, it is necessary to divide the total movement into a succession of highly sensitive ranges: when the differential transformer dust core has moved a distance equivalent to the first range, the output voltage must be reduced to zero at the commencement of the second range.

The sensitivity obtained in each range is only limited by the amplification available, the heating limits of the differential transformer and the stability of the measuring circuits. Equation (5) shows that the series opposition method is twice as sensitive as the series connexion (equation (16)).

Most of the early measuring techniques using differential transformers are based upon the series opposition method but the analysis in this article shows that E_{out} is only zero for one position of the dust core showing that it is not easy to design a measuring system with a succession of ranges, although a method is shown by Heath[3] whereby a suitably connected anti-phase voltage is injected into the differential transformer secondary at the commencement of each range.

The series connexion has been shown as suitable for the sub-division of the transducer linear range into a succession of highly sensitive sections, the analysis has proved that a balance can always be achieved at any part of the range. Equations (14), (15) and (18) show that the zero of each range must be achieved by both a voltage and phase adjustment; the equations also show that balance conditions will be true for only one set of positions of the dust core. This severely restricts the practical use of the transducer as it means that mechanically the core must be capable of being moved to the zero position of the first range. In effect, apart from its ability to obtain a balance for any position of the dust core, the series connexion in the particular application discussed in this article is restricted just as much as the series opposition connexion in as much as the dust core must be movable in the initial setting up of any apparatus in which the transducer is used.

The Elimination of Phase Shift

Consider equations (14) and (15) when R_v approaches ∞:

$$E_{out} = \frac{E\omega(M_1(1 - x) - xM_2)\sin \omega t \,.\, e^{j\phi}}{\sqrt{(R_p^2 + \omega^2L_p^2)}} \quad \ldots\ldots (19)$$

where

$$\phi = 3\pi/2 + \tan^{-1}(\omega L_p/R_p) \quad \ldots\ldots (20)$$

and x the movement of the slider on R_v for a balance when the core is moved:

$$\frac{M_1}{M_1 + M_2} \quad \ldots\ldots (21)$$

Let M be the mutual inductance of each half of the transducer when the core is in the electrical centre, and let the core be moved a distance y from the centre. It is assumed that the mutual inductance changes linearly with the core movement.

$$E_{out} = \frac{E\omega M(1 + y - 2x)\sin \omega t \,.\, e^{j\phi}}{\sqrt{R_p^2 + \omega t L_p^2}} \quad \ldots\ldots (22)$$

and hence to reduce E_{out} to zero on the slider on R_v must be moved a distance:

$$x = \frac{1 + y}{2} \quad \ldots\ldots (23)$$

Thus the output voltage is no longer phase conscious with the core movement and once a given set of ranges have been decided upon, these are no longer dependent upon the core position.

The effect of making $R_v \to \infty$ can be achieved by operating the transducer into a valve measuring circuit as shown in Fig. 2.

Equations (19) and (20) show that neither the secondary self inductance L nor the resistance R influence the output voltage; in fact, provided the mutual inductance M_1 and M_2 change linearly with movement of the core, they need not have the same intrinsic value and E_{out} will always change linearly with core movement. Another important result of the analysis is that the transducer winding specification can be relaxed considerably: there is now no need to closely match the two halves of the secondary, to take a specific example they could be wound in different gauges of wire and E_{out} would still be linear with core movement.

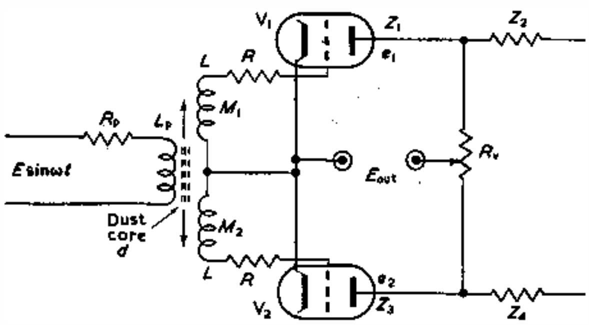

Fig. 2. Operation of transducer into valve measuring circuit

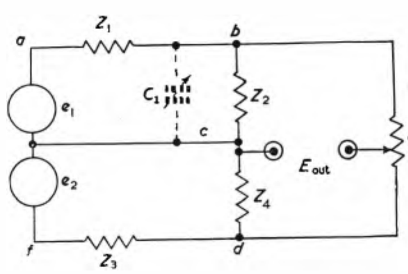

Fig. 3. Circuit derived from Fig. 2

Vectorial Analysis of Measuring Circuit

Consider Fig. 3 which shows the circuit derived from Fig. 2 where e_1 and e_2 are the transducer secondary voltages after amplification by V_1 and V_2. Z_1 and Z_3 are anode impedances of V_1 and V_2 and Z_2 and Z_4 the anode load impedances.

If e_1 and e_2 have a relative phase shift $\theta°$ and $Z_1 - Z_4$ are complex, the vector diagram will be as shown in Fig. 4(a). It will be seen that ce, the output voltage, cannot be reduced to zero unless there is both a voltage and phase adjustment in the circuit. If vector cb be corrected by a capacitor C_1 and e moved along bd, E_{out} can be reduced to zero as shown in Fig. 4(b). If e_1 and e_2 change linearly with core movement and θ remains unaltered, the relative displacement of bd remains unchanged. Consequently E_{out} is either in or out of phase travelling linearly with core movement along vector bd, and its value depends upon the position of e on R_v and at one position will be zero.

Thus for all positions of the dust core within the transducer range a balance can be achieved by adjustment to R_v only.

The effect of the impedance changing network V_1 and V_2 is to make R_v in Fig. 1(b) much greater than any other parameter in the circuit, and for all practical purposes the desired conditions in equations (19) and (20) have been reached. The vector diagram also shows that E_{out} is directly proportional to the change in e_1 and e_2, i.e. the circuit output is linear with core movement. In practice there is no relative phase shift between e_1 and e_2 as no current is drawn from the transducer.

Now R_v in Fig. 2 will have distributed inductance and capacitance so that as the slider is moved there will not be a purely resistive change in the network. The result is that vector ce in Fig. 4(a) cannot move along the straight line bd but follows a shallow hyperbola[2]; an additional amount of capacitance will therefore be required in C_1 to correct the slight change in vector configuration when e is moved. This means that the system is balanced for one position of the core and if it is moved from this position the point c moves along bd; when e is moved along R_v to reduce ce to zero again a small correction must be made to C_1. It should be noted that this capacitance change is due purely to the imperfection in the potentiometer R_v and, provided that the transducer output is linear with core movement, for a given core movement anywhere within the transducer linear range, the change in C_1 will always be the same.

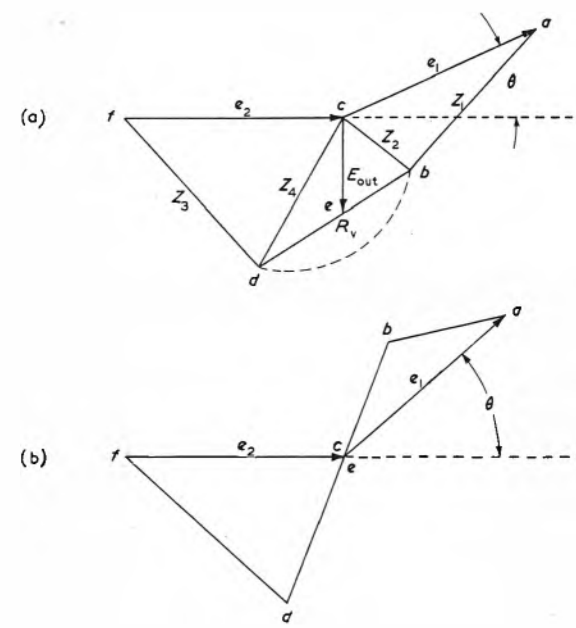

Fig. 4. Vector analysis of Fig. 3

This property leads to an important application of the circuit: once a given set of core movements has been decided upon (by a series of positions of e on R_v), it will only be necessary to rebalance the first range, in order to commence at any part in the transducer linear range, the other ranges will always be balanced as the phase change in each is due only to the distributed reactance of R_v.

Practical Application

V_1 and V_2 in Fig. 2 are in effect a differential amplifier stage subtracting the amplified signals from the two halves of the transducer secondary. Fig. 5 shows a practical circuit developed from Fig. 2.

The transducer B has both a centre-tapped primary and secondary, the primary being fed from a push-pull oscillator A. The output voltages M_1 and M_2 are fed into the measuring stages V_3 and V_4 via cathode coupled matching stages V_1 and V_2. At the anodes of V_3 and V_4 the amplified signals M_1 and M_2 are subtracted and E_{out} obtained from the switching network S_1, S_2 and S_3. On range 1 with the core of the transducer B at some point in its linear range E_{out} is reduced to zero by means of equal ganged potentiometers VR_1 and the differential capacitor C_1. When the limit of the first range is reached S_1, S_2 and S_3 are switched to the second range which is balanced by VR_2 and C_2. As many ranges as desired can be accommodated by increasing the number of switch points S_1-S_3. It will be recalled that the capacitor C_1 corrects the phase change due to the alteration in distributed reactance when the potentiometers VR_2, VR_3, etc., are moved.

Once the system has been set up, the ranges set by VR_2, VR_3, etc., remain approximately fixed for any given starting point in the transducer linear range, thus if it is desired to commence the ranges at some other point the first range is balanced by VR_1 and C_1, the remaining ranges being automatically balanced. It will be seen that the change required in C_1 to rebalance range 1 is due to reactance alterations in the network when VR_1 is moved. A coarse/fine ratio control for balancing can be achieved by a suitable choice for VR_1 and the range potentiometers VR_2, VR_3, etc. In a specific application of the system to the measurement of the co-efficient of linear expansion of materials, a differential transformer with a linear range of 0·1in was used in which the maximum sensitivity on each of the successive ranges was 5×10^{-4}in. Therefore, any change taking place as the 2in specimen was heated could be automatically recorded to 5μin; in practice a range sensitivity of 5×10^{-3}in was found adequate.

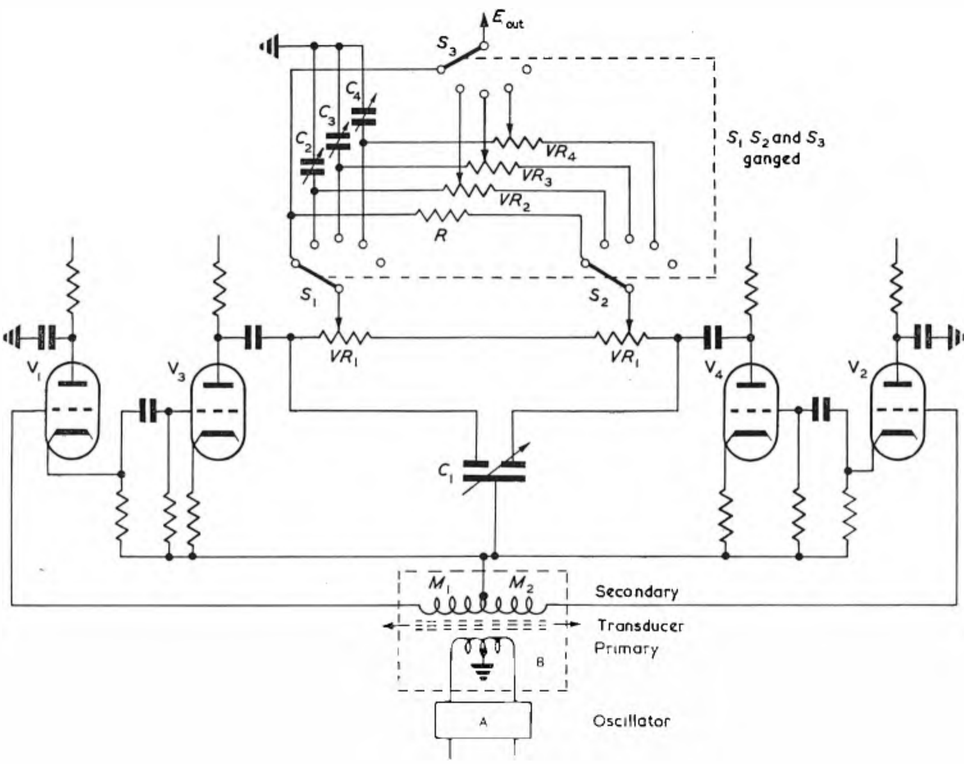

Fig. 5. A practical circuit developed from Fig. 2

Limitations

Although the analysis shows that a given set of ranges once set up will remain balanced for any part of the transducer linear range, in practice it has been found that if an attempt is made to rebalance the first range outside the limits of ± 50 per cent of the expansion covered by all the ranges, the zero for each successive range is no longer balanced, this has been shown to be due to:

(a) Eddy currents and hysteresis losses in the dust core and windings throwing a small variable impedance across the transducer thus creating a relative phase shift between the two secondary voltages (equation (15)).

(b) The reactive distribution along the balance potentiometer being hyperbolic.

Conclusions

This article is a condensed account of part of a programme of work undertaken in the application of electronic measurement technique to the study of behaviour of materials at high temperatures. The work shows that a very sensitive measuring technique has been designed in which phase changes associated with the transducer output voltages have little effect upon the measuring system.

The measuring system has been applied in this instance to an inductance strain gauge, but could equally be applied to many other forms of transducer.

Acknowledgment

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Control Apparatus for a Serial Drum Memory

By D. S. Kamat*, M.Sc.

This article describes an apparatus that has been developed and used successfully for obtaining design data for a faster track switching device for a serial drum memory. It gives circuit and performance details of the gates, triggers, switches and other components that have been developed. The apparatus is intended for further use for study of other computer functions.

IN an electronic digital computer[1,2] with a magnetic drum as its principal fast memory, and working in a serial mode the total memory storage capacity is shared between a number of tracks; where each track corresponds to one of the record-reproduce heads. The selection of a track or (record-reproduce head) is done in many cases by means of a group of electro-mechanical relays. It is felt that at least a few types of faster switching devices may be tried other than the usual relay tree track selection device. To conduct these experiments in particular and to study other design problems connected with serial memory devices in general a control apparatus was constructed.

This apparatus has three functions to perform, namely to generate coded information consisting of 32 bit word, to route this information to a given location on a memory track (or to extract information from a location and store it in a register) and to generate fast switching impulses that are to be used both in selection of a required track and in performing either of the record-reproduce functions.

A Synchronous Device and the Process of Transfer of Information

The apparatus is shown in a block schematic diagram in Fig. 1. The unit on the right-hand side of this diagram makes it possible to shift a word of 32 bits to and from a memory location whenever commanded. The unit has a shift register, a five-stage binary counter and an asynchronous to synchronous pulsing device that may start the shifting operation. Fig. 2 shows the circuit of this unit excluding the shift counter and the shift register. Suppose that there is a number already stored in the register and it is to be shifted by n places ($n < 32$). This n is set in the shift counter (with the help of manual switches) and whenever the switch S_1 is thrown over from 1 to 2 it unsets trigger T_1 (valve V_1) opens gates G_1 (valve V_3) and G_2 (valve V_5). Both these gates receive clock impulses (i.e. 32 selected clock pulses), the former directly and the latter delayed by a few microseconds via a delay trigger (valve V_7). The output of G_2 resets trigger T_1 to its normal state. As a result the gate G_1 allows only a single pulse to pass through which resets trigger T_2 (valve V_8). The gate G_0 (valve V_{11}) allows the 32 input pulses to pass which, after being amplified, simultaneously act as shift impulses for the shift register and input pulses to the shift counter. The shift counter, after completing a count of 32, sets trigger T_2 which in turn closes gate G_0, thus terminating both shifting and counting operations.

Pulse inputs ranging from a pulse repetition rate of 10kc/s to 100kc/s, as well as a memory drum giving clock pulses at 34·08kc/s were tried successfully in testing this operation.

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Selection of a Track and Associated Functions

A multi-position switch[3,4], using crystal rectifiers in association with a track address register, is used for selection of record-reproduce heads. Initially the multi-position switch is to select any one of the 16 positions, each corresponding to one of the 16 configurations possible in an address register with four binary stages. Fig. 3 is the circuit diagram of the multi-position switch. The lower row in the figure is the trigger assembly representing the address of the track. Both anodes of each of the triggers are connected separately to inputs of cathode-follower valves (V_{12} to V_{15}) the output levels of which control the switching action corresponding to the actual state of the triggers. Corresponding to the two levels (70V and 165V) on the trigger anodes, the cathode-follower cathodes rise to 85 and 185V respectively. These voltages after passing through a dividing network are fed to the grids of the switching valves (V_{16} to V_{30}). The switching valves either conduct fully or remain cut off.

It can be seen from the circuit diagram that a selected position (or terminal) has four 'and' gating diodes. The anodes of these diodes are raised to +h.t. (250V) via the 22kΩ resistors because the corresponding sections of the switching valves are not conducting, while the cathodes of these four gating diodes are each held to a voltage existing at the anodes of the rest of the (conducting) valves through the back resistances of 7 other diodes. The latter voltage level, what may be termed as the reference level, is adjustable by suitably fixing the screen voltage of the switching valves. In the case shown this reference level is at 70V, while the selected terminal is at 100V. Fig. 7(e) is an oscillogram of the transient impulse obtained at a switch position.

Later this pulse, on a selected switch position, is fed into a current switching device which gates either a waveform to be recorded or 32 clock pulses depending upon whether the record or reproduce function is to be performed.

Selection of Word Location on a Track[5,6]

On each track of the memory drum there may be more than one word location, the actual number depending upon the drum diameter and the digit packing density on the surface. In the present case the drum has 16 word locations per track. The identity of each of these 16 of words is established:

(1) By making the spacing or free interval between the first and the last word different and larger than that between the rest.

(2) By recourse to a synchronization device.

(3) By sensing the coincidence between the word address and the word counter when it occurs.

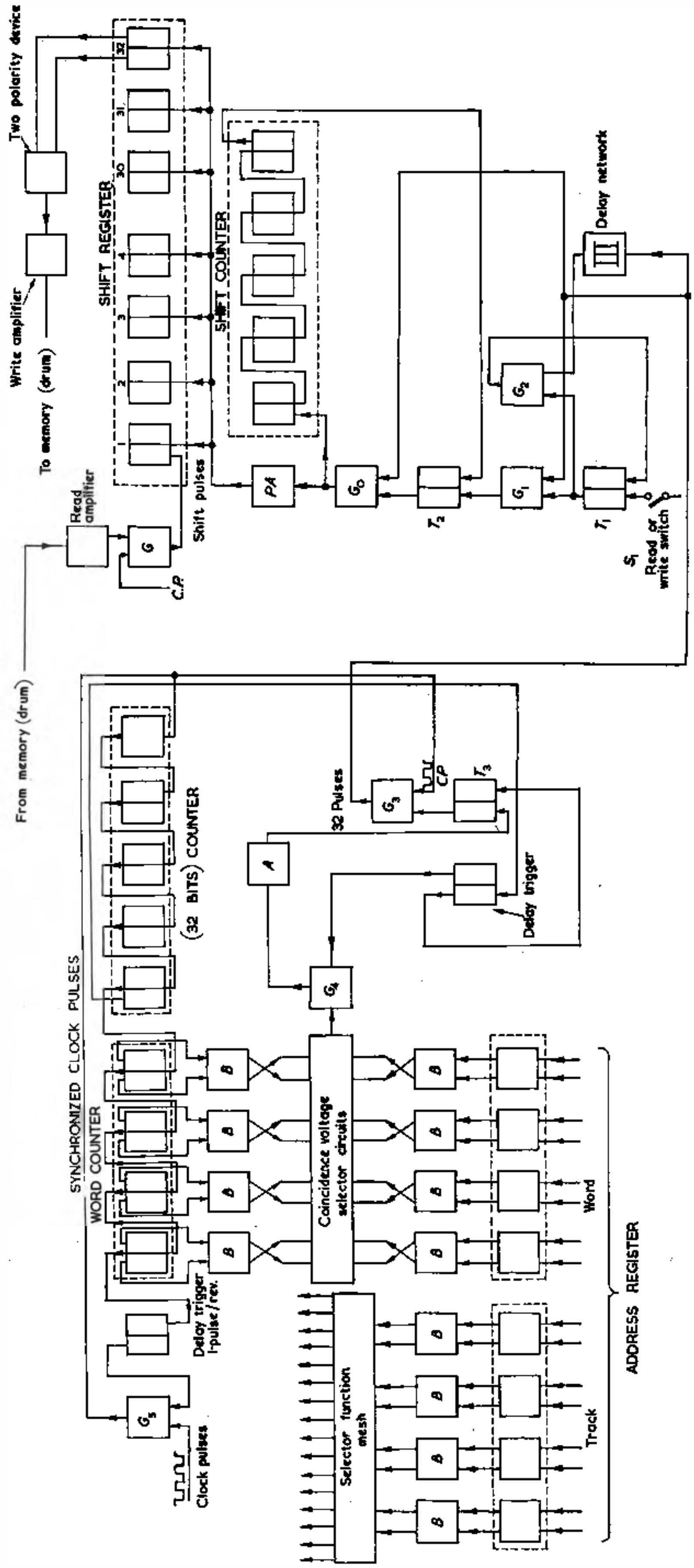


Fig. 1. Schematic of control for selection of location (address) and storing of operand in magnetic drum calculator

The pick-up head for clock pulses generates 16 groups of 32 pulses and these are spaced as stated above. These clock pulses are fed into a nine stage binary counter (see top left-hand portion in Fig. 1) and the counter generates a single pulse per drum revolution. A monostable multivibrator is turned 'on' by this pulse for a duration much longer than the spacing (or its equivalent in time) between all words except that between the last and the first. This multivibrator controls gate G_5 and as a consequence of the above it shuts off a portion of the pulses that normally get past. In a few revolutions the shutting off of output pulses from G_5 coincides with the end of each drum revolution, a state corresponding to complete synchronism.

The selection of 32 clock pulses corresponding to the required address location is done with the help of a scheme (as shown in Fig. 1) consisting of a coincidence voltage selector, trigger T_3 and gates G_4 and G_3. Each of the end of the word pulses coming from the 32 bits counter triggers a delay trigger which in turn generates a pulse that could be shaped to give a pulse coincident with input pulses and also another one delayed by about 100μsec. One of the latter 16 pulses is allowed through gate G_4 wherever the correct coincidence voltage level occurs. This pulse turns 'off' trigger T_3 which opens gate G_3 and thus allows input clock pulses to pass through. However, the next end of the word pulse turns the trigger T_3 'on' and thus closes gate G_3. So a maximum of 32 clock pulses pass through gate G_3 per drum revolution and are always coincident with the selected location as shown on an address register.

Triggers and Gates

Fig. 4(a) shows the trigger circuit[7,8] which stores binary bits '0' or '1'. A row of triggers, as shown in Fig. 3, when connected via gating diodes[1] (also shown in the same figure) constitutes a register for storing an address, a number or an instruction, all three generally being referred to as a 'word'. The d.c. voltage levels at the various electrodes are also shown. To turn the trigger in either of its stable states pulsed inputs at the grids of V_{1a} and V_{1b} are required. V_{1a} and V_{1b} are normally conducting valves. When anode current—anode voltage characteristics for 6SN7 valves for voltages in the range 30 to 40V are studied it is found that they have a constant slope. Thus for the analysis of the behaviour of the triode pair V_{2a}

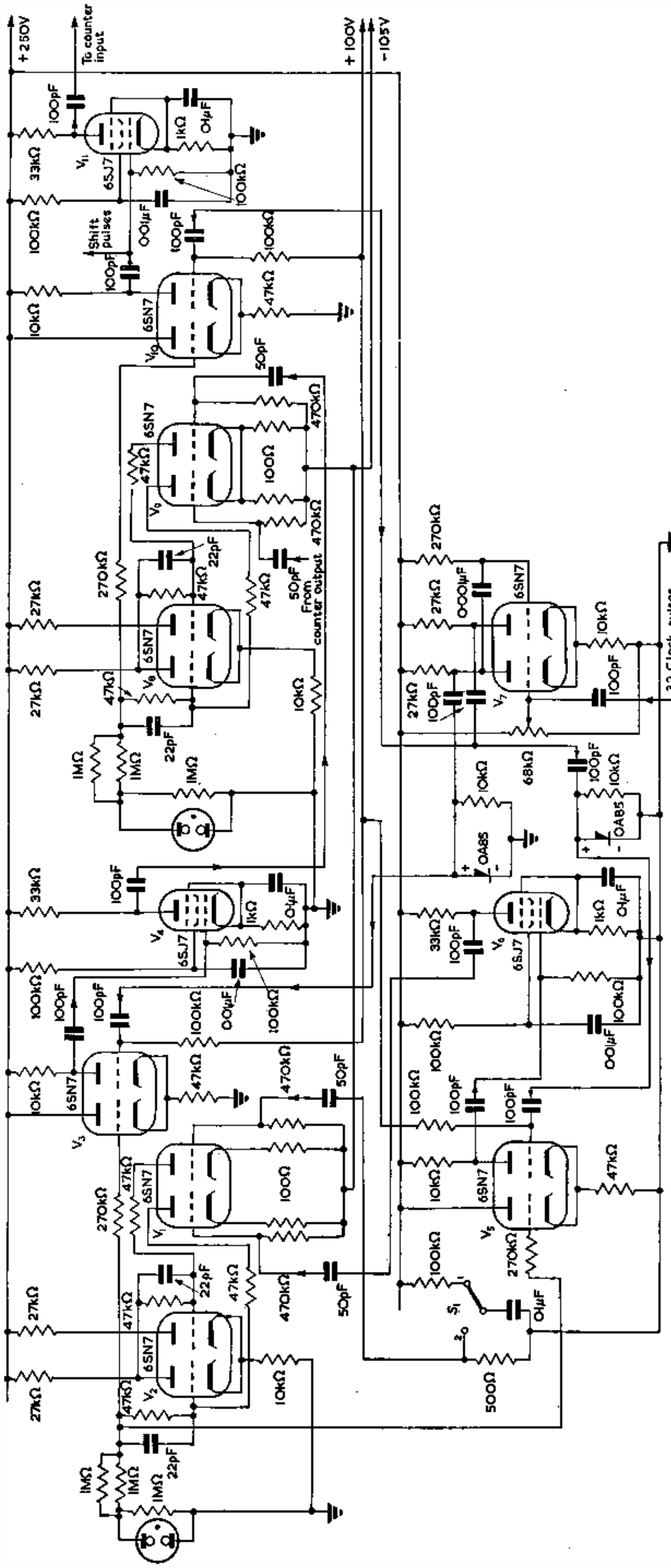


Fig. 2. Single pulse generating unit and control for shift operation

and V_{2b} both V_{1a} and V_{1b} could safely be regarded as constant resistances of equal value. The cathode degeneration introduced by a cathode resistance of 100Ω also helps in equalizing the transconductance of the valves for both the sections namely V_{1a} and V_{1b}. A simplified version of the circuit will then reduce to the one shown in Fig. 4(b). This approximates to the usual trigger circuits discussed[8] with the difference that the common cathode is grounded through a resistance that is not by-passed. This has an effect of greatly increasing the valve resistance and thus making each valve circuit effectively a constant current device. Fig. 5 shows a counter trigger circuit.

There is only one type of gate circuit that is used. In all places it is used in association with the d.c. voltage levels on a trigger anode. Whenever the relevant anode of a trigger (Fig. 6) or delay trigger is high (165V) the common cathode of the gate valve is raised to this voltage and hence the gate remains closed. When the grid of V_a is at 70V (corresponding to the altered state of the trigger) the common cathode can drop from 100V to this voltage whenever the grid of V_b is pulsed negatively, and then the output occurs.

The coincidence sensing gate for selection of word location on a chosen track is shown in Fig. 6(b). The two voltages that are summed at the grids of V_1 to V_4 are from opposite anodes of triggers representing bits of equal significance in the word counter and the word register. The common cathode of the summing valves will have a voltage of 105V whenever coincidence occurs, i.e. when the corresponding digital places in the counter and address register are of equal significance. At other times the cathode would be at 165V. Thus when the grid of V_5 is pulsed the cathode can drop from +135 to +105V only when coincidence occurs, at other times the gate will be held closed.

Performance Characteristics

The equipment was constructed with components and valve types that were readily available. It was desired that the switching speeds of various triggers and circuits should be such as to facilitate a wider degree of freedom for the study of computer functions such as auxiliary memory devices, input/output conversion, as well as a memory drum track switching device with pulse repetition rates up to 100kc/s.

Simulation of variation in computing speed is obtained by substituting in place of the magnetic drum a laboratory

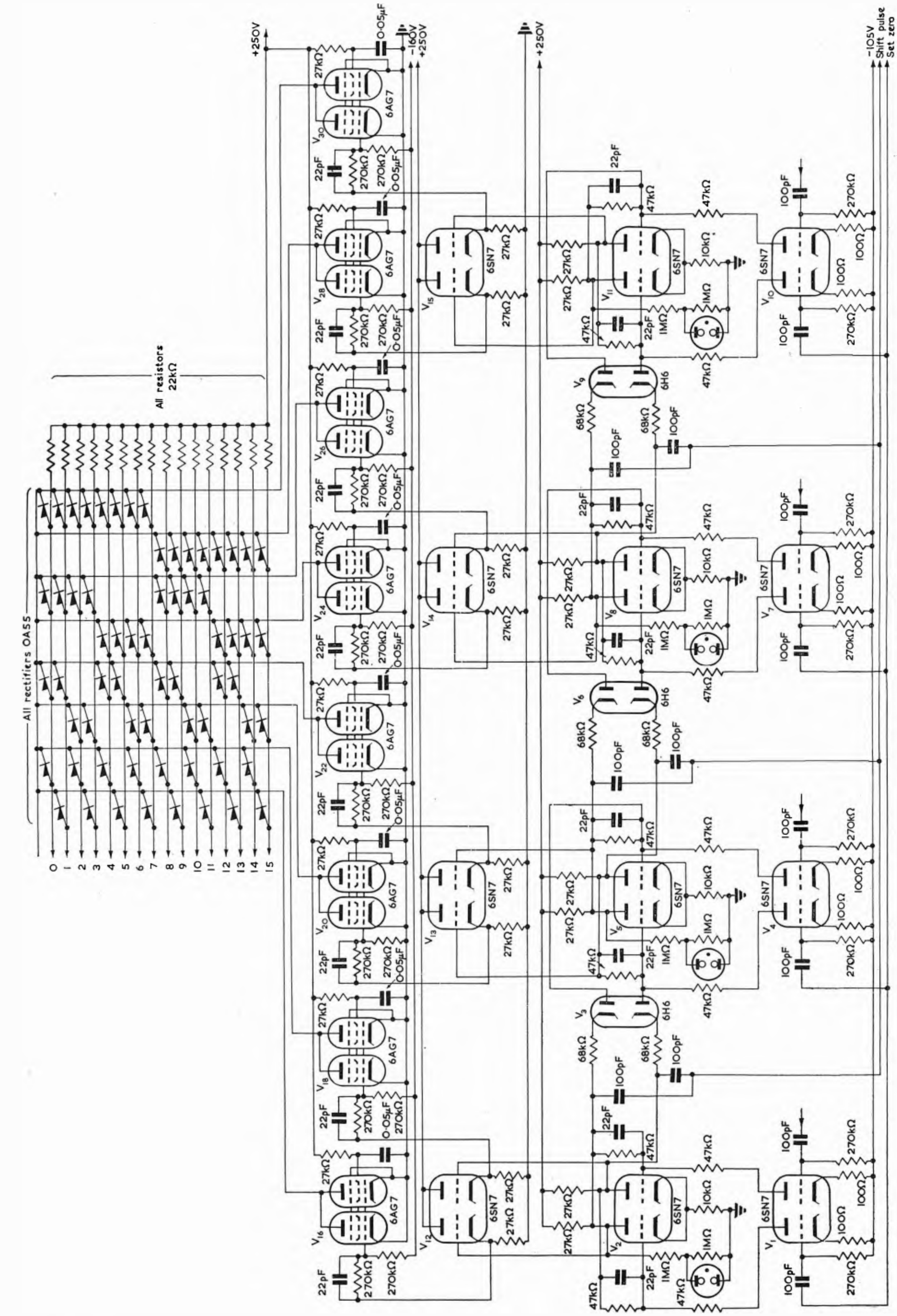


Fig. 3. Control circuit for track selection

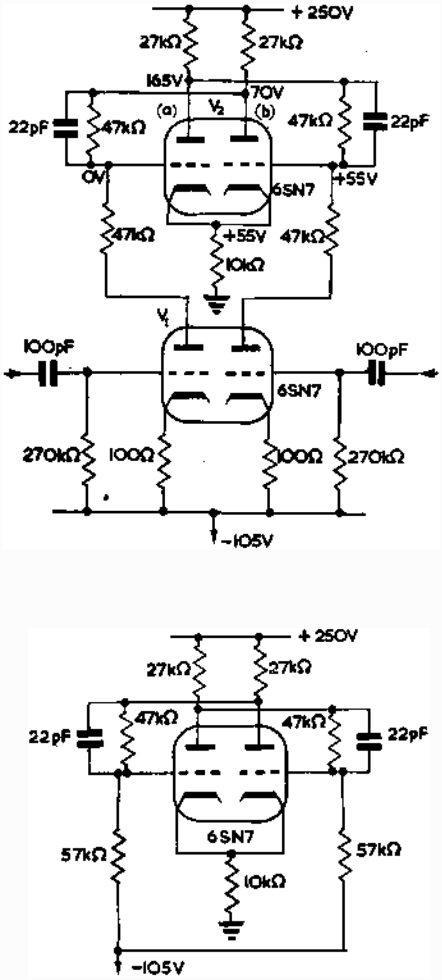


Fig. 4. (a) Trigger circuit; (b) Equivalent for the trigger circuit

pulse generator having a frequency range from 6kc/s to 150kc/s. When such a pulse source is used the synchronization circuit gives a burst of 512 pulses. The clock pulses corresponding to a drum location then obtained will not be 32 but less (as there is no gap between words). Fig. 7(a), (b), (c) shows such selected pulses at repetition rates of 100kc/s and 10kc/s and also the 32 selected pulses when a memory drum with clock pulse repetition rate of 34·08kc/s was used.

The response curves for the track selection device were

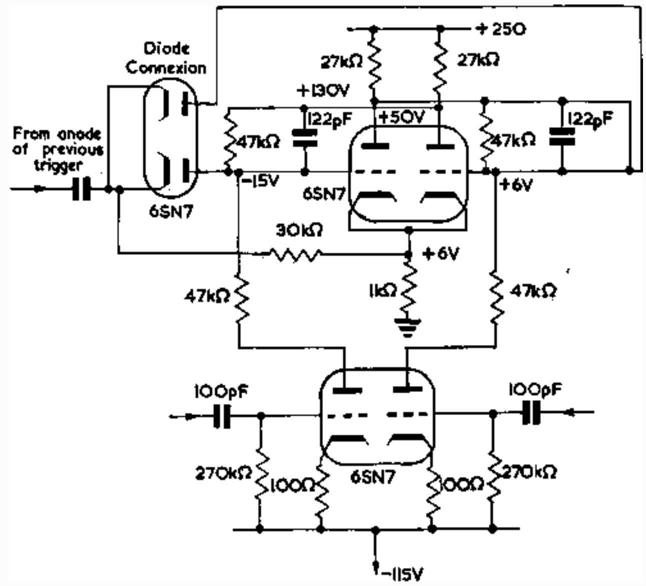
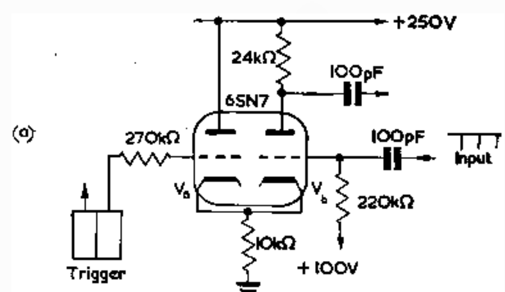


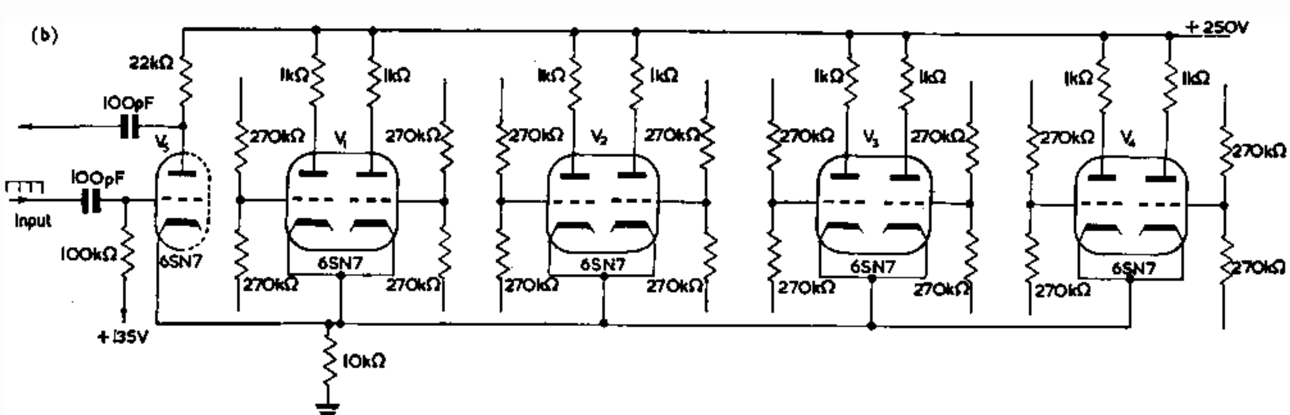
Fig. 5. Counter trigger circuit

obtained when the triggers of the track address register were set and reset within 2μsec intervals at 5kc/s.

The counter and register trigger and the gate circuits operate satisfactorily at pulse repetition rates and in association with other equipment in the manner stated above. Valves in the counter and register trigger cells each dissipate about 15 per cent of their rated maximum anode dissipation. Filament supply is maintained mostly at 6·3V a.c. although no special a.c. stabilizing methods are used. The 250V d.c. supply is made available through smaller series stabilized power units, each giving about 200mA. The total d.c. at 250V required for all the units as shown in Fig. 1 (excluding the record/reproduce functions) is

Fig. 6. (a) A gate circuit; (b) Coincidence sensing device





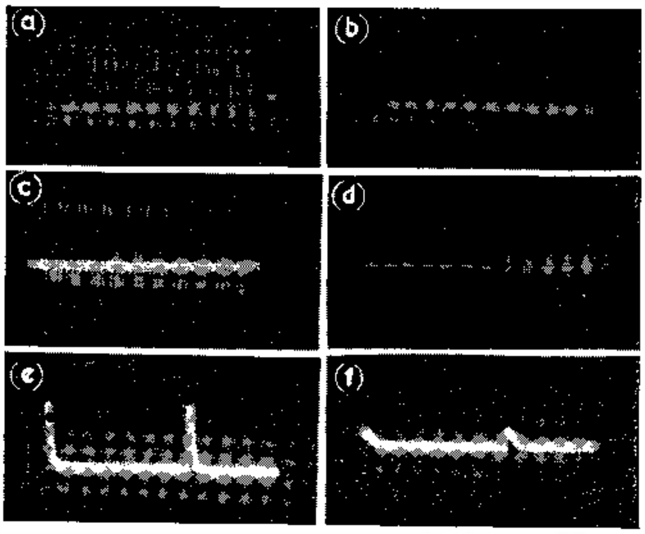


Fig. 7. Oscillograms obtained when using laboratory pulse generator

(a) 100kc/s pulses, synchronized and selected
(b) 10kc/s pulses, synchronized and selected
(c) 32 clock pulses form the memory drum (34·08kc/s) synchronized and selected
(d) 16 end of word pulses from the memory drum
(e) Transient response of the track selector switch. Output amplitude 30V, pulse width at bottom 2μsec
(f) Same switch position as in Fig. 7(e) but when not selected

nearly 1A. The valve types used are 6SN7, 6H6, 6AG7 and 6SJ7; out of these 6SN7's constitute about 80 per cent of the total. The failure among the germanium diodes used in the selector switch was relatively large (about 40 per cent), their back resistance in such cases varying between a few thousand ohms to about 100kΩ. These were replaced by newer types.

Acknowledgments

The author wishes to thank his many colleagues in the Electronics Laboratory and the Development Workshop of the Institute for their help in the work and in particular Mr. Bijan Mukherji and Mr. K. M. Patnaik for help in construction of this apparatus.

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The Effect of a Magnetic Field on Point-Contact Transistors

By K. K. Bose*

When a magnetic field is applied to a point-contact transistor at right-angles to the plane of the emitter and collector, interesting observations have been made of the alterations of all the characteristics of the transistor, both static and dynamic. The changes which take place in the operational characteristics of the transistor are positive or negative with respect to the normal values, according to the polarity of the applied field. Brown[1] first noticed that the frequency response of a point-contact transistor improved when it was magnetically biased with the proper polarity. In the present article not only the frequency response, but the changes in the output characteristic and amplification characteristic of the transistor due to a magnetic field, as determined experimentally, have been presented and an attempt has been made to give a suitable explanation for the peculiar behaviour.

THE action of the point-contact transistor is primarily determined by the nature of the current flowing from the emitter. When the emitter and the collector are placed in close proximity, a large part of the 'holes' injected by the emitter in the n-type transistor will flow to the collector and into the collector circuit, on account of the electric-field set up by the collector current. The carriers are drawn to the collector and add to the collector current. As mentioned later, this addition is not algebraic. The injected holes which concentrate around the collector point modify the barrier space-charge below the collector in such a way that the change in collector current is more than the change in the emitter current to which the former is due; or, in other words $\Delta I_c > \Delta I_e$ and α of the point contact transistor is greater than unity. The value of 'alpha' (α) depends not only on the spacing between the emitter and the collector, but also on the forming process etc., by which a p-layer is formed below the collector in the n-type germanium. The frequency response of the transistor is, however, primarily dependent on the geometrical spacing between the emitter and the collector, other things, e.g., temperature, mobility etc., remaining constant.

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The nature of the collector is normally such as to provide a high resistance barrier to the flow of electrons from the metal to the semiconductor (being reversely biased), but there is little impediment to the flow of holes. In addition, the hole-current causes an enhanced flow of electrons from the collector to the semiconductor.

If I_c be the collector current, the field (F) produced at a distance r from the collector:

$$F = \rho I_c / 2\pi r^2 \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

where ρ = resistivity of the germanium (ohm-cm).
The drift velocity of the holes flowing from the emitter to the collector is:

$$v = \mu_h F \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

where μ_h is the hole-mobility (cm^2/V.sec).

The time taken by a hole to reach the collector vicinity from the emitter source, i.e., the transit time (T),

$$T = \int dr/(\mu_h \, . \, F) = 2\pi/\mu_h \rho I_c \int_0^S r^2 \, dr$$

or,

$$T = \frac{2\pi S^3}{3\mu_h I_c \rho} \ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

where S is the separation between the emitter-contact and collector-contact.

Frequency Response

The transit time has an approximately inverse relationship to the frequency response. Thus the frequency response would be expected to vary inversely as the cube of the point spacing if other conditions were maintained constant. In practice the transit time is of the order of 10^{-8}sec. It is obvious that as the frequency is increased there will be an increased phase-shift as well as reduction of current-amplification (α), since the hole-current arriving at the collector from the emitter will be dispersed in phase.

Now it suggests itself that if a magnetic field is applied at right-angles to the plane of the emitter and the collector, the path of the hole current will be diverted from the original one, i.e., the effective value of the spacing S between the emitter and the collector will be changed. This will result in a change in the transit time T, thus altering the frequency response of the transistor.

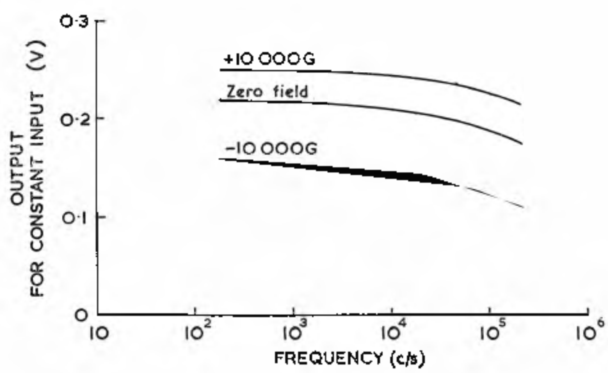


Fig. 1. Effect of magnetic field on frequency response

In Fig. 1 the effect of the magnetic bias of about 10kG is shown. The transistor used was the Western Electric, type WE-2A.

Depending on the polarity, the magnetic field tends, (1) either to compress the hole-current towards the surface reducing the average path length and the transit time, resulting in less phase dispersion, or (2) to force the hole-current to travel over widely varying paths. The hole-current is diverted from the surface downwards and forced to flow deep through the body of the germanium. This is possible, since Shockley and Shive[2] have shown independently by their famous slab type and wedge type germanium experiments that the flow of the hole-current is not confined to the contact surface only. The holes may flow through the surface layer as well as through the body of the semiconductor, as shown in Fig. 2.

The amount of deflexion of the path of the hole current will depend on the strength of the magnetic field, and the direction of the deflexion on the polarity.

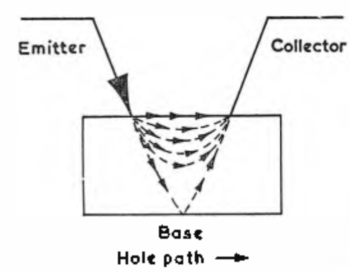
Another thing to be noticed in Fig. 1, is that the changes in the frequency response due to positive-field and due to negative-field are not equal. This is because the holes are deflected down from the surface by the negative field and the stronger the field the more will be the bending of their paths. The higher the value of the negative field, the more will be the downward deflexion of their flow-paths towards the collector. Moreover, some of the holes would be lost by their recombination with free electrons. This recombination will increase with transit time.

But the positive magnetic field tends to concentrate the holes towards the surface and cause them to flow direct to the collector. The path of the hole-current becomes shorter with the positive-field, as the latter is increased, but cannot be shorter than the geometrical value of S, or the distance between the emitter and the collector along the surface. With stronger positive fields little improvement takes place and saturation is reached. Moreover, the recombination of the holes with the electrons is much higher at the surface of the semiconductor than inside, and this accounts for the relatively smaller change by the positive field. The output seems to reach a saturation when the field is further increased.

Voltage Amplification and Output Characteristics

The theory of the hole-current explains how the change in emitter-current causes a change in the collector-current; but it is not fully understood how ΔI_c can be larger than ΔI_e. It is believed that there is some sort of alteration of the space charge in the barrier layer at the collector by the hole-current. It is better, according to Bardeen[3], to think of the hole-current from the emitter as modifying the current-voltage characteristics of the collector rather than as simply adding to the collector current. It has not been possible to detect, if the collector barrier-layer is directly

Fig. 2. The flow of 'holes'



affected by the applied magnetic field, but it has been observed that both voltage and current amplification do change by the transverse magnetic field. They increase or decrease according to the polarity of the magnetic field.

If F_b is the field in the barrier layer and μ_h the mobility, the drift velocity (v) is given by

$$v = F_b \cdot \mu_h \quad \text{(4)}$$

$$\text{The hole-current} = I_h = n_h \cdot e \cdot v \cdot A_c \quad \text{(5)}$$

where n_h = hole concentration in the barrier

A_c = area of collector contact.

From which

$$n_h = \frac{I_h}{e \cdot v \cdot A_c} \quad \text{(6)}$$

The numerical value of n_h in a point contact n-type germanium transistor is of the same order as the concentration of electrons, and it is quite possible that the hole-current alters the space charge in the barrier quite significantly, thus changing the collector current to such an extent that $\alpha > 1$. The net effect of the hole-current from the emitter of the n-type point contact transistor is to increase the conductivity of the germanium in the vicinity of the collector. The electrical conductivity is directly proportional to the product of carrier concentration (n_h) and the mobility (μ_h). In Fig. 3 the collector resistance (static) has been demonstrated in the $V_c - I_c$ curve under three conditions,

0 gauss, +10kG, and −10kG field strength, and, of course, for constant emitter current. In Fig. 2, one sees clearly how the collector current increases, or, decreases by the positive or negative magnetic field for different collector voltages. As shown in equation (6) the hole concentration, which causes the change in the collector resistance is directly proportional to the hole current. How the hole concentration

affects the barrier resistance at the collector is not very clearly understood. But it is obvious that the applied magnetic field controls the flow of holes from the emitter to the collector and indirectly changes the $V_c - I_c$ characteristic curve. Suhl and Shockley[4] had experimentally observed that transistor action is obtainable at the collector in the presence of a magnetic field when the collector is physically spaced further away from the emitter, than it is possible to space it when the magnetic field is turned off. In the presence of the magnetic field of proper polarity the diffusion of the holes is controlled and the hole-current is made to flow more nearly in a straight line. With the

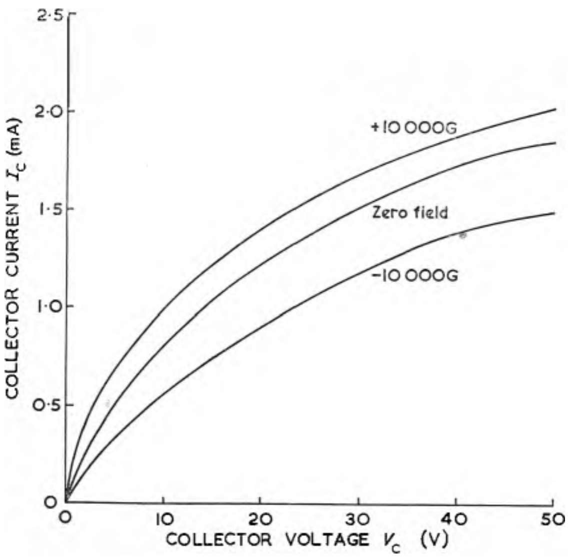


Fig. 3. Effect of magnetic field on conductivity

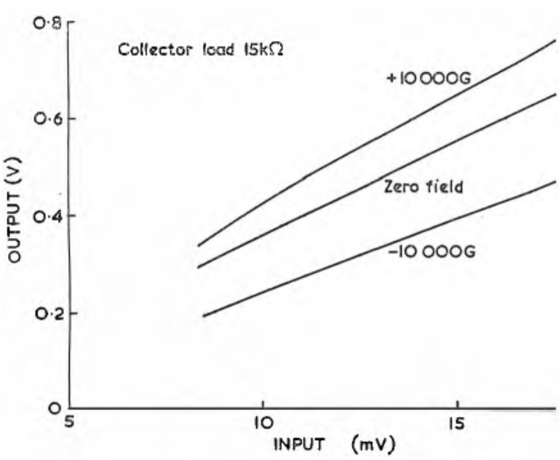


Fig. 4. Effect of magnetic field on voltage amplification

opposite polarity the path will be curved, as mentioned before.

But the distortion in the path of the hole-current alone is not responsible for the alteration of the collector-current. The deflected path increases the effective separation S between the emitter and the collector contact points, which results in phase dispersion, increase of the transit time of the carriers, and thus affects the frequency response of the transistor. The most important effect of the increase of transit time is the decay of the injected carriers by recombination with excess electrons. In other words, a strong magnetic field reduces the hole-current. Hole-current is attenuated exponentially by the recombination with electrons in going from the emitter to collector. If $N(o)$ is the original number of injected holes, the number existing after time t is

$$N(t) = N(o)\exp(-t/\tau_p) \quad \ldots\ldots\ldots\ldots (7)$$

where τ_p is the lift time of the positive carrier or holes. If T be the transit time of the holes, the hole-current will be decreased by a factor $\exp(-T/\tau_p)$.

τ_p is not a very constant quantity. It varies with temperature, and also from crystal to crystal due to the presence of chemical impurities and imperfections in the lattice. When the magnetic field is applied, the transit time (τ) increases, causing an attenuation of the hole-current, which results in the change of characteristics of the collector current.

In Fig. 4 the voltage amplification curves are shown for a constant frequency of 500c/s under positive and negative magnetic fields of 10kG. Here also the variation due to positive field is less than that due to negative field.

In all these experiments the field was reversed by rotating the transistor axis through 180°. The magnetic field had no effect at 90°, i.e., when the plane of emitter and collector coincided with the applied field.

Conclusion

This article has attempted to show how the basic mechanism of current flow in a point-contact transistor is affected by a magnetic field. It has been shown that all the changes in the transistor characteristics can be attributed to the disturbance of the flow of the injected carriers by the magnetic field. Briefly, the effects of the magnetic field on transistor operation are: (1) the effective separation S between the emitter and the collector is altered. (2) The concentration of the injected carriers in the vicinity of the collector is altered. (3) The recombination rate of the injected carriers with the excess electrons is also affected due to the changed transit-time of the carriers by the magnetic field.

The experiments have been limited to point-contact transistors only as junction-transistors with non-magnetic leads were not available. The lead wires of all the available junction transistors were highly magnetic and almost shielded the transistor against the field.

Because of the extremely small separation between the emitter and the collector (~0·005cm for 20dB gain) a comparatively high magnetic field is necessary to get an appreciable change at the output. Any deviation of the operational characteristic due to stray magnetic fields is out of the question. In the experiments carried out, both emitter current and collector current changed when the field was applied. The emitter current was then re-adjusted to its previous value in order to find out how much the collector current changes by the magnetic field alone. A high resistance was put in the emitter circuit to minimize the variation of I_e due to feedback action.

Acknowledgments

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Electronic Musical Instruments in Germany

By Alan Douglas, M.I.R.E., M.Amer.I.E.E.

Comparatively little has been disclosed about developments in electrical music abroad since the last war. This article presents a short summary of some of the more interesting German designs. These include the AWB organ, which is a full-scale model with two manuals and pedal and is an example of an additive sine wave circuit using valves. Several rather novel instruments are also described.

ALTHOUGH Duddell and de Forest produced musical sounds by electronic means prior to 1920, the first real musical instruments were designed by Oskar Vierling and Friederich Trautwein in 1928. At that time, gas tubes (which had reached a high state of development in Germany) were preferred as relaxation oscillators. A notable advance was made by Welte in 1936 when he produced a photo-electric instrument with rotating glass disks carrying fully preformed tone colours.

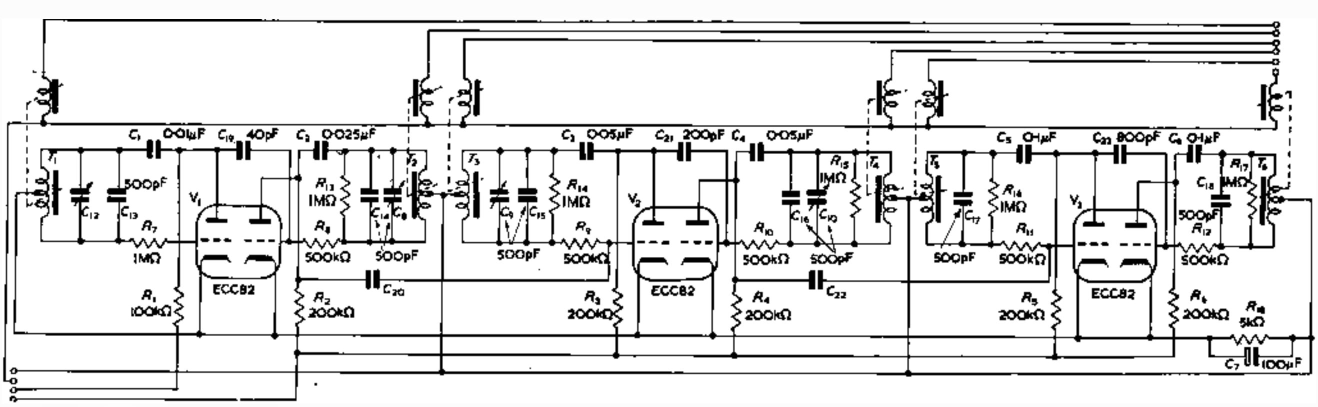
The production of non-essentials was slow in a country impoverished by war, but more recently some interesting instruments have been commercialized in Germany. Perhaps the best-known example is the AWB organ. This is a full-scale model with two manuals and pedal and is the only example of an additive sine wave circuit using valves. Fig. 1 shows the oscillator and dividers for one note sign; these latter are of the blocking type, each being tuned to the required frequency and synchronized with the previous one by the small capacitors of 40,100,200,400 and 800pF. A coupling coil, shown above the oscillatory circuits, conveys the signals to the busbar system.

Keying is carried out in a manner similar to that of the Hammond, six different pitches being available at the busbars. From the multiple key contacts the signals are taken to a tapped mixing transformer through 'antirobbing' resistors of 200Ω each as in Fig. 2, which shows part of the pre-amplifier circuit. The harmonic reduction and transient filters to ensure pure sine waves can be seen following the input transformers.

Extensive stabilization of the power supplies is required in an additive instrument, and the circuit for this (which incorporates the vibrato oscillator) is shown in Fig. 3. The use of mutations such as tierce and twelfth endows this organ with great brilliance of tone and without doubt the upper work is better developed in the AWB than in any other electric organ. Recently this instrument has been completely redesigned to incorporate transistor oscillators, although at the moment no exact details are available. A

The photograph shows the designer, Oskar Sala, at the keyboards of the concert model Trautonium

Fig. 1. AWB oscillator and dividers



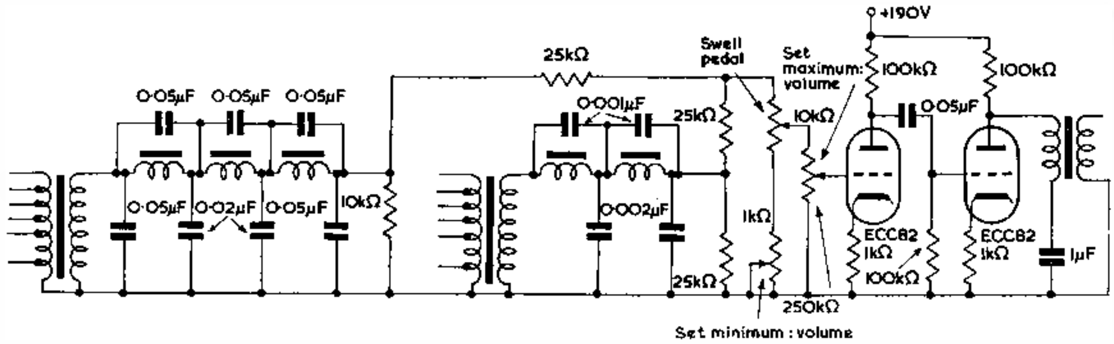


Fig. 2. Part of AWB pre-amplifier circuit

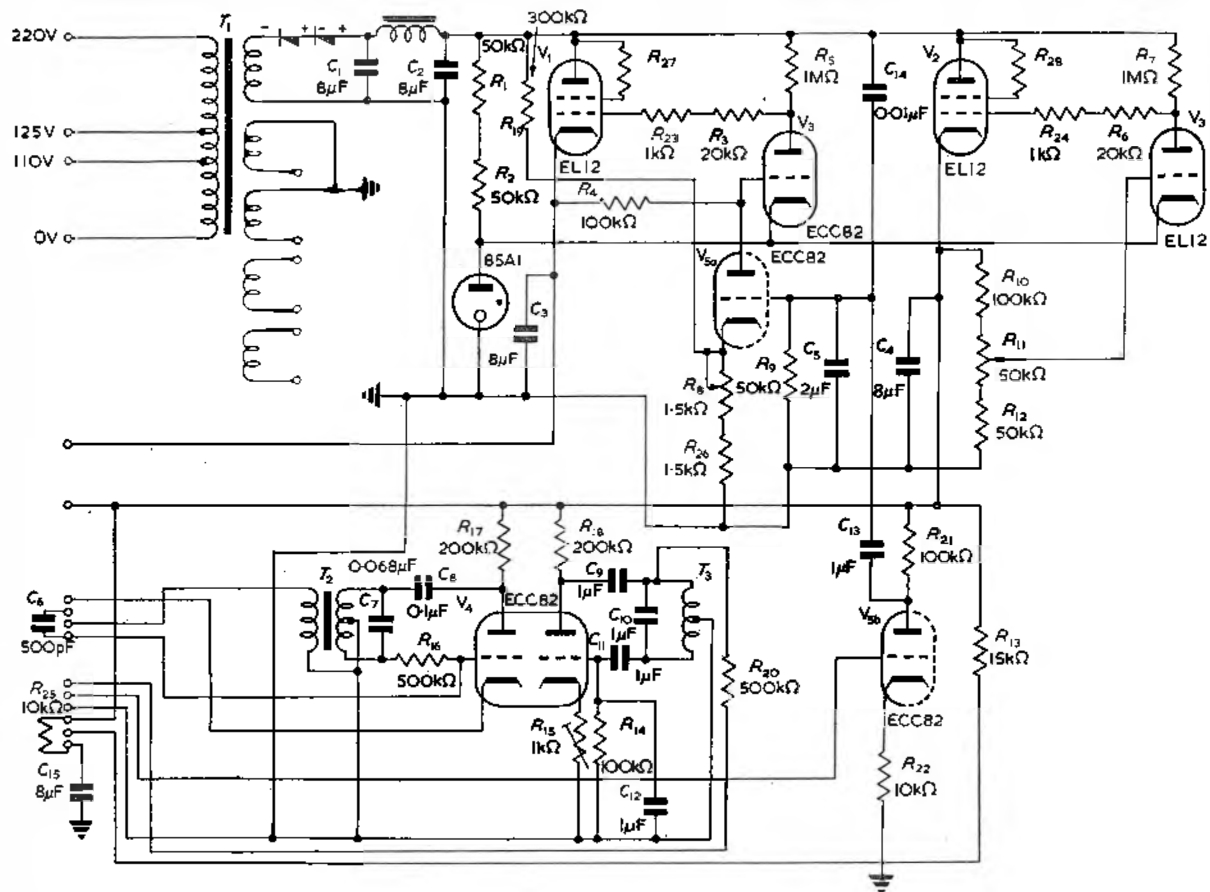


Fig. 3. Stabilized power supply and vibrato oscillator

similar, though not quite so elaborate, instrument designed by Harald Bode is marketed by Haeberlein of Munich.

The firm of Matth. Hohner AG, known for many years as makers of harmonicas, produces several electrical novelties. Apart from the Clavioline made under the Seybold patents and re-named the Electronium P1 when supplied in the usual form with a keyboard, another model is made in which the expression control is worthy of note. Fig. 4 is a photograph of the interior, from which it can be seen that the tone generator is played by accordion keys but the bellows fulfil a double function. If compressed at one end, a rheostat alters the volume, while if compressed at the other end, the harmonic content is changed by moving a coupling coil away from another coil fixed to this instrument. Otherwise the controls are arranged as for the Clavioline and this model may also be obtained with buttons in place of keys.

A rather quaint device is the Multimonika, shown in Fig. 5. Here there are two keyboards each of 41 notes compass. The lower one operates three sets of reeds which are contained within the case together with a small blowing fan. Different sets of reeds may be selected by pressing the buttons shown and the reed volume is regulated by a knee-operated shutter controlling louvred openings in the case. The upper manual is electronic and only one note at a time may be played thereon. The circuit, Fig. 6, is reminiscent of the earlier Solovox, the tone oscillator frequency being altered by removing capacitors from the main tuning circuit. The vibrato oscillator uses a gas tube and a tuning control appears on the front panel to set the pitch to that of the reeds; also on this panel are the tone switches.

Hohner also make a reed accordion with acoustic insulation and a microphone to pick up the sound. A pre-amplifier allows the reed tones to be altered in many ways by formant circuits. Another useful product is the Bassophon, designed to simulate either the string base or the tuba, but playable by keys. The controllable attack allows very rapid speech over its two and a half octaves, and additional amplifiers may be paralleled to increase the volume of sound.

Turning to more serious developments, the instruments used by the Cologne school of music research comprise the Melochord and Trautonium.

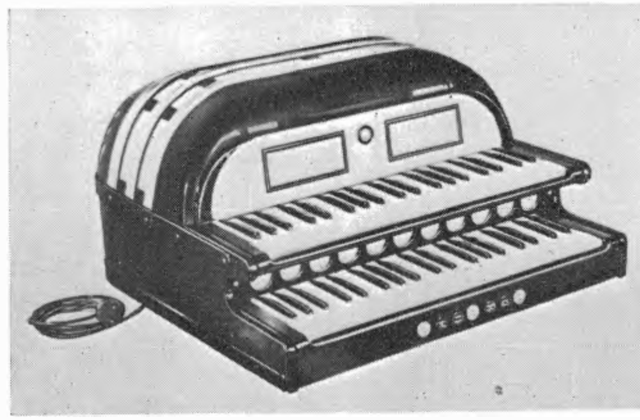


Fig. 5. Hohner Multimonika

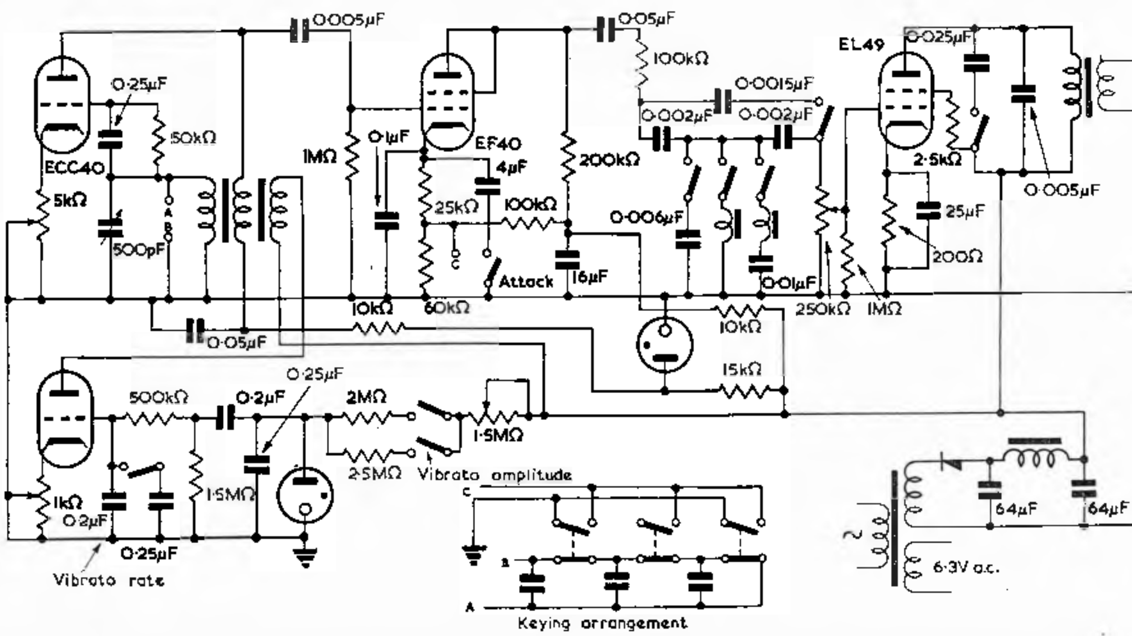


Fig. 6. Circuit of Multimonika

Fig. 4. Hohner Electronium showing expression control

Both of these devices have been described previously[1]. While their primary purpose is for research on musical tone qualities and the composition of 'new' music, a concert model of the Trautonium has been developed by Oskar Sala for use with a conventional orchestra, and has an amplifier of 150W output so that the power level can equal that of the orchestra if required[2]. The photograph on page 642 shows the designer at the keyboards of this Trautonium, the frequency dividers of which are unique in that the harmonic content per octave is controlled by the method of division, the object being to apply tone filters of varying characteristics for each octave and not, as is more usual, to feed the whole signal into common shaping networks. By this means, octaves below the fundamental can be heard with quite different timbres, forming some very attractive sound patterns.

REFERENCE

1. Douglas, A. Synthetic Music. *Electronic Engng. 28*, 208 (1956).
2. German Patent No. 917470.

Diode Hole Storage and 'Turn-on' and 'Turn-off' Time

By G. Grimsdell*

This article describes both the 'turn-on' and the 'turn-off' circuit effects in a typical germanium diode and shows why it is dangerous to compare diodes by their published hole storage times, without at the same time considering the conditions of measurement in each case.

VALVE diodes behave more or less according to their theoretical prediction when fed with a train of voltage pulses. The current through the diode starts and stops according to the instantaneous polarity across the diode, and only the small physical capacitance of the diode has any adverse effect, at any rate at ordinary frequencies when transit time has little significance.

With the semiconductor diode, however, the position is very different, and unwanted effects occur both at the beginning and end of the current pules.

'Turn-on' Time

When a diode which has been biased in the reverse (high impedance) direction suddenly has its polarity reversed (low impedance direction), the forward voltage drop across the diode for a given current is higher immediately

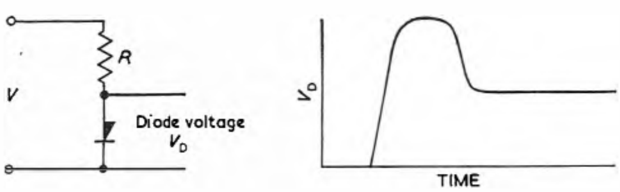


Fig. 1. The diode circuit and its 'turn-on' characteristic

after the reversal than it would be under steady state conditions (see Fig. 1).

This phenomenon is hardly noticeable in point contact diodes due to their limited maximum current and high forward voltage drop, but in gold-bonded and small-area junction diodes, where these factors have been improved, it can be the next important effect.

The phenomenon appears to be due to the fact that when a diode is conducting and its base is flooded with 'carriers' the base resistance is lower than when the first carriers pass the junction but have gone no further, i.e., the effect is due to the modulation of the base resistance by the injected carriers.

It is worse when the diode is used at high currents with low circuit impedances and sharp waveforms. Diodes specially produced for use in computers and other digital circuits must of course have this property controlled to a point where it does not affect the operation of the circuits using the diode.

'Turn-off' Time and 'Hole Storage'

When a diode has been carrying current in the forward direction and the voltage across it is suddenly reversed, the high current through the diode does not immediately cease but flows for a small time in the reverse direction.

The effect is due to the fact that when the diode is in forward conduction, minority carriers are injected into the base, and these flood a large volume of the germanium. When the polarity is then reversed, these minority carriers are thus available to provide a short burst of high reverse current, before the diode settles down to its normal low leakage current.

Assuming for the moment that the minority carriers do not recombine in the bulk of the base material, it will be seen that for any forward current there will be a fixed charge of carriers available to flow in the reverse direction on polarity reversal. The rate of reverse flow of this charge and therefore the time for the flow to decrease to a given value will depend among other things, on the external circuit resistance. Indeed, it would be possible to stop the forward current flow, wait a small fraction of time, and then to apply the reverse voltage and still get the hole storage current.

In fact, if the circuit impedance were low enough, very large currents could be drawn from the diode for very short times, the diode emptied of carriers and made ready for use as a normal diode much sooner.

There are, however, two other phenomena that must be taken into account. Firstly, spontaneous recombination (carrier lifetime) and secondly, minority carrier transportation by diffusion.

Recombination takes place all the time and minority carriers have themselves an average expectancy of life. Thus, if the base region of a diode is full of minority carriers, because the diode has been conducting in the forward direction, and the current is cut off, it will be found that on the later application of the reverse potential the available hole storage charge is less than was expected. In fact, whether the external circuit values affect the hole storage time of the diode or not, will depend upon whether the circuit drive pulse and rise-times are long or short compared with the average life-time of the minority carriers.

In the extreme case the carrier life-time is made so short that there can never be any measurable hole storage because they all recombine before they are recalled to the junction at the reversal of polarity.

In the same way, a reduction in the bulk of the germanium will reduce the hole storage simply because there is less space to store them in, quite apart from the closer proximity of the surface effects.

Specification of Hole Storage in Published Data

The actual hole storage properties of a diode can be specified in a number of ways: in terms of the total charge immediately available on potential reversal; as the time taken for the large reverse current to die away to a given value through a given resistance; the reverse current reached after a given time; or as the peak current in a given circuit. However, all these methods are dependent both on the measuring circuit parameters and on the ratio of contribution made to the hole storage performance of the diode under test by the volume and geometry of the base, by the life-time and other properties of the germanium, and by the forward current originally passed through the diode.

* *Mullard Ltd.*

Thus, two types of diode may be co-related between the various types of test at one or more measuring points, while a third diode, in which the contribution of the different diode phenomena occurs in different amounts, will fail to co-relate at all between various test circuits and will perhaps appear better than some other type in one test and worse in another test.

Much argument at present exists as to the best method of measuring hole storage and presenting it in published data. So far, although some standard tests exist, it is not really possible to infer diode performance in every circuit from one value of hole storage in one standard circuit.

In time, perhaps, suitable constants will be found to specify completely the diode's hole storage under all conditions.

In the meantime, all the manufacturer can do is to give the hole storage data in a specified circuit at specified currents which are as near as possible to the user's circuit requirements.

The user must, in his turn, realize that the hole storage time of a diode can only be meaningful if the circuit and current are also quoted.

The coming of small area junction diodes and the gold bonded diode makes this precaution even more necessary, since the range of their operating forward currents is much larger and hence hole storage becomes more important.

Acknowledgment

The author wishes to thank the Directors of Mullard Limited for permission to publish this article.

A Constant Voltage — Constant Current Stabilizer

By D. P. C. Thackeray*, B.Sc., A.Inst.P.

Uses are suggested for a stabilizing arrangement which will switch itself automatically between constant voltage and constant current operation, to suit the load or overload demands made upon it. A circuit is developed to suit the particular demands of an electronic photo-flash unit.

DESIGNERS and constructors of stabilized power supplies are well served by a considerable literature, describing in greater or lesser detail practically every permutation and combination of available components. It is perhaps not realized widely enough that completed stabilizers, by virtue of their stabilizing action, need to be designed against misuse.

For example, a constant voltage d.c. stabilized power supply unit, when connected to an uncharged capacitor, may readily switch itself into a condition of considerable current overload until the capacitor is charged. A similar voltage overload may appear when a constant current flow is established in an inductor. Neither application can be deemed unusual in general instrumentation practice, but clearly the damage done by such peak overloads will be more marked when high voltage or high current apparatus is concerned. Thus, while it may be sound practice to design medium voltage and medium current stabilizers with some leeway to meet peak demands, at higher voltages or currents it may be economically worthwhile to consider methods of automatic protection.

Automatic Overload Protection

Consider a simple case, for instance that of Fig. 1, where a battery may be used to charge a capacitor when a switch is closed. The peak current may be limited to a suitable safe value by choice of an appropriate value for the series resistor. Such a resistor inserted in series with a stabilized voltage supply would be a possible, though inelegant, protective device. In general, and depending upon whether the resistor followed or preceded the stabilizer, the source impedance of the combination would be high enough, either to spoil the regulation of the supply, or to present a large voltage drop at full load current. A better position for the resistor in such a supply is shown in Fig. 2, where it has been inserted in series with the cathode of the series control valve. Here the resistive effect of valve plus resistor R_k, as far as current changes are concerned, is now $[(r_a/R_k) + \mu]$ times that of a resistor R_k alone, where μ is the amplification factor of the series valve, and r_a its anode impedance. Hence the resistor may be smaller by this factor, for the same degree of protection, in which case it will drop only what may be a negligible fraction of the

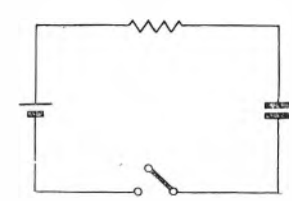


Fig. 1. Simple current limiting circuit using a resistor
The source impedance is increased and the charging current waveform constrained to an exponential shape

voltage drop of the simpler case. The insertion of this resistor results in some degradation of the stabilization obtainable, but this may readily be countered by a suitable increase in the gain of the shunt amplifier stages(s).

In the more elaborate form of this stabilizing circuit, it may be necessary, Fig. 3, to insert a diode to ensure that during overloads, the grid of the series valve is returned effectively to the correct point. In practice it is convenient in either circuit to make the inserted resistor variable, so that the peak overload current may be adjusted to a suitable value.

Constant Current—Constant Voltage Operation

It can be seen that, if the load on a modified stabilized supply, such as that of Figs. 2 and 3, is switched from an open-circuit to a short-circuit, the stabilizer will change from supplying a stable voltage, to supplying a current determined principally by the series valve and its cathode

* *Vickers Group Research Establishment, formerly Research Laboratory for the Physics and Chemistry of Solids, Department of Physics, Cambridge.*

resistor. Since such a combination provides a dynamic resistance, for current changes, many times the quotient voltage ÷ current, it can be considered as a current stabilizer, albeit of a simple type. Thus it may be used to charge a capacitor at a constant rate. When the capacitor has charged to a voltage equal to that at which the voltage stabilizer normally provides a controlled output, the charging current will fall to zero, and the stabilizer will revert to holding constant the voltage across the capacitor. The change from one condition to the other will be more or less abrupt, depending upon whether the shunt amplifier stage(s) has a high gain and is turned on rapidly from cut-off to its working point, or has a low gain and is turned on slowly. If it is desirable to set a limit to the power that may be drawn from such a device, the less abrupt characteristic may be tailored to reach this limit at a suitable point on the regulation characteristic. These properties may be exploited in, for instance, a power supply unit, Fig. 4, designed to charge the capacitor bank of an electronic flash unit. This unit has been used in the research programme of the laboratory for the photo-irradiation of explosive materials. The automatic recharging of the bank after each flash, to a constant voltage, will relieve the operator of an onerous duty, when it is desirable to hold this voltage to within a few per cent over long periods of operation. The constant rate of charge has two advantages, by comparison with resistive charging. First, the slower initial rate of charge allows the flash discharge tube better to deionize after the previous flash. Second, the faster ultimate rate of charge avoids the lengthy exponential approach to the final voltage.

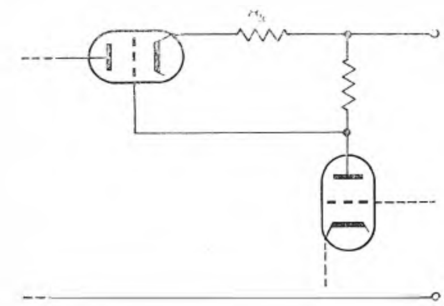


Fig. 2. Use of limiting resistor in cathode of series valve of a series-parallel stabilizer

This increases the dynamic impedance by μR_k, thereby reducing the overload currents that can be drawn from the stabilizer

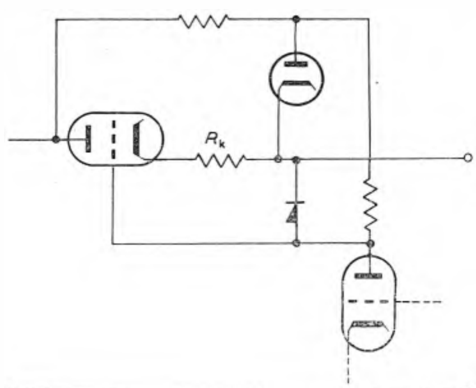
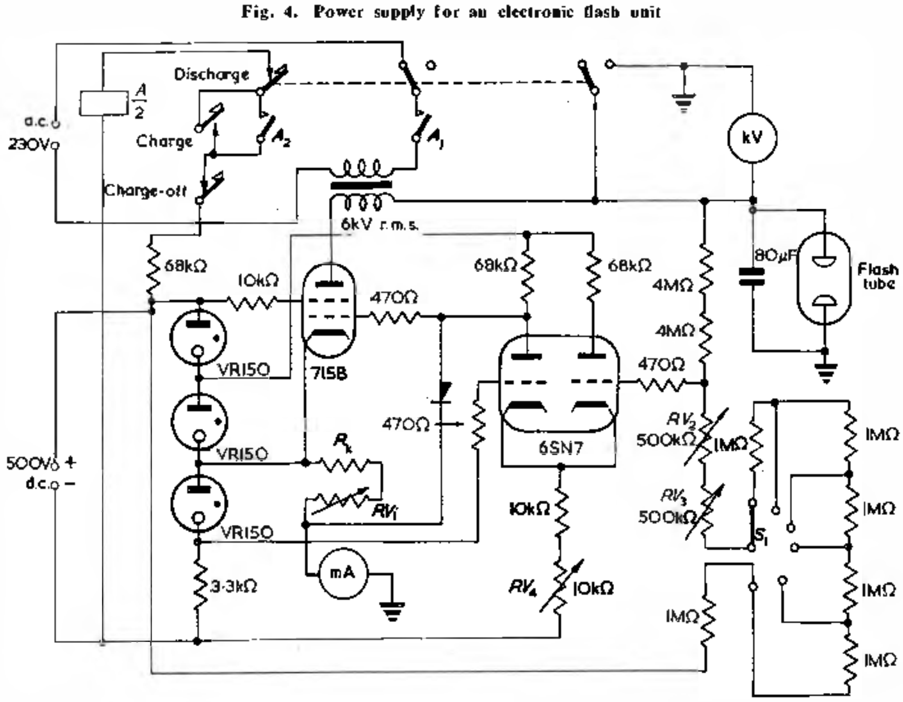


Fig. 3. Insertion of 'catching diode' to prevent grid voltage of series valve rising above output voltage of the unit

If the series valve is a tetrode or pentode its screen grid can be fed from the gas-filled voltage stabilizer to raise its dynamic impedance still further

Referring to Fig. 4, R_k and RV_1 together form the resistance controlling the charge rate, which is normally set to between 5 and 20mA. RV_2 and RV_3 and S_1 are used to set the controlled output voltage to any required value in the range −700V to −3·7kV. The series valve, a 715B, may dissipate up to 60W on its anode, a demand which this operation may frequently entail. It combines this function with that of a rectifier, since the addition of a separate rectifier serves no useful purpose in such a unit. Further to simplify operation, the unit is switched by means of push-buttons and a contactor, the latter interlocked with a discharge switch for the capacitor bank, and one of the main e.h.t. switches. Thus the unit when enclosed is proof against accidental mis-operation, and suffers no damage when the capacitor bank is discharged.

Fig. 4. Power supply for an electronic flash unit



Conclusion

It has been shown that a simple modification will protect stabilized supplies against accidental overloads. The device is readily applicable to stabilizers designed to work between conditions of zero current, stabilized voltage, and full load current at voltages below this. The transfer from one state to the other is automatic, and is a particularly useful feature for reliable equipment, and where the operator must devote his attention principally elsewhere.

Acknowledgments

Thanks are due to Dr. G. T. Rogers, for drawing the attention of the author to the potential value of an automatic power unit, and to S. Barton and J. Ward for constructional work on the prototype and final unit.

TIME-SYMMETRIC FILTERS

(Part 2)

Response to Gliding Tones

By L. R. O. Storey*, M.A., Ph.D., and J. K. Grierson*, M.Sc.

In Part 1 of this article, methods were described for simulating filters that have impulse responses that are symmetrical in time. Their application to the analysis of gliding tones is now discussed. The response of a narrow-band time-symmetric filter to a gliding tone is evaluated, and is shown to consist of an oscillation bounded by a slowly-varying envelope. The shape of the envelope is governed by a single dimensionless parameter, that involves the bandwidth of the filter and the rate of variation of the instantaneous frequency of the tone. There is an optimum value of this parameter that yields the sharpest response.

In general, the envelope of the response is symmetrical, and reaches a maximum at the time when the frequency of the tone is equal to the resonant frequency of the filter. This behaviour contrasts with that of a simple tuned circuit. Use of the time-symmetric filters, instead of simple tuned circuits, should lead to greater accuracy in the analysis of gliding tones.

THE aim of this study was to improve upon the techniques in current use for the analysis of gliding tones.

A gliding tone is a quasi-sinusoidal oscillation that can be represented as the resolved part of a vector of constant or slowly changing amplitude that rotates at a variable speed. Suppose, to be definite, that the tone is in the form of an electrical voltage. Then the variation of voltage with time is given by such an expression as

$$V(t) = V_0 \sin \phi(t) \tag{48}$$

The *instantaneous frequency* of the tone is defined as

$$f = \frac{1}{2\pi}\left(\frac{d\phi}{dt}\right) \tag{49}$$

and is assumed to vary only slowly with time: slowly, that is, compared with the oscillation itself. Variations in the amplitude V_0, if any, are assumed to be slow also.

The particular problem in hand is that of the analysis of certain experimental data that are recorded as waveforms on magnetic tape. Each record consists of a gliding tone, or a closely-spaced sequence of such tones, mixed with varying amounts of unwanted noise. The object of the analysis is to determine, for each tone, how its instantaneous frequency varies with time. This is done usually by means of narrow-band filters. The time when the instantaneous frequency takes a particular value is measured by passing the tone through a filter tuned to that frequency, and then noting when the response of the filter rises to a maximum. The complete relationship between frequency and time is determined by repeating this measurement at a number of different frequencies, either simultaneously using a number of filters, or, if a recording of the tone is available and can be played back many times, sequentially using a single filter tuned to the different frequencies in turn.

Questions arise as to what type of filter should be used, and what bandwidth it should have, in order to resolve adjacent tones most completely and to discriminate most effectively against interfering noise.

In the past much use has been made of the resonant LC circuit, because this is the simplest type of narrow-band filter. The response of this circuit to a gliding tone has been evaluated by Barber and Ursell[8], and by Barber[9]. They find that it consists of an oscillation bounded by a relatively slowly-varying envelope, and that the shape of the envelope, though complicated, depends on a single parameter k that involves the bandwidth of the filter and the rate of change of the instantaneous frequency of the tone; this parameter will be defined shortly. Fig. 6 shows how the envelope varies with time, for several values of k. The origin has been set at the moment of resonance; that is to say, the moment when the frequency of the tone is equal to the resonant frequency of the filter. When analysing a tone, each individual measurement consists of determining one such time by inspecting one such response.

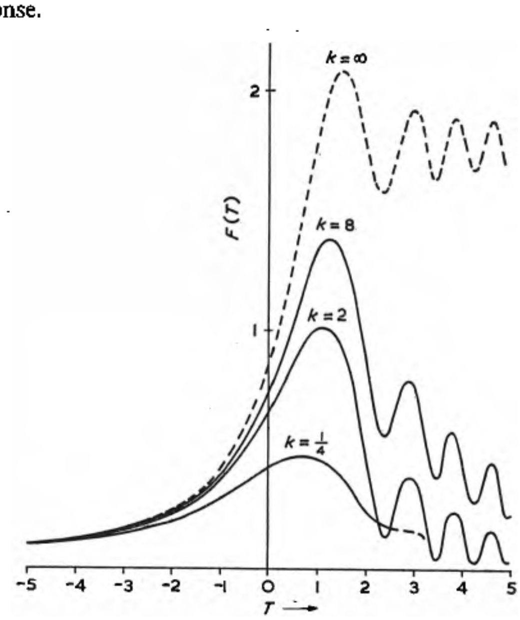


Fig. 6. Envelopes of the response of a simple tuned circuit to a gliding tone, for different values of the parameter k. The abscissa T is a normalized measure of time, and its origin is set at the moment of resonance (after Barber and Ursell)

In general, the response is asymmetrical in time, and reaches its maximum somewhat after resonance. Its leading edge rises smoothly, but its trailing edge contains pronounced ripples; these may be understood as beats between a forced oscillation, at the frequency of the tone, and a transient oscillation, at the natural frequency of the circuit, which is excited around resonance and subsequently decays.

Now responses of this type do not lend themselves to making very accurate analyses, for the following reasons:—

Firstly, because of the asymmetry, a correction must be applied to the observed time of maximum response to obtain the time of resonance. The amount of this correction depends on the value of k, being greatest when k is large.

* *Defence Research Telecommunications Establishment, Candaa.*

Secondly, the ripples on the trailing edge cause confusion when attempting to resolve a number of adjacent gliding tones. The envelope of the response then exhibits many closely-spaced peaks, and it is hard to tell which are due to multiple tones and which are just subsidiary peaks in the response to a single tone.

Clearly, it would be better if the main maximum of the response of the filter occurred at resonance for all values of k, and if the subsidiary maxima were smaller and less numerous.

Now the reason that the peak of the gliding tone response of the resonant LC circuit is delayed is that its frequency response contains phase-shifts, or, what amounts to the same thing, that its impulse response is asymmetrical in time. A narrow-band filter that had a time-symmetric impulse response would not produce this delay, but such filters are not physically realizable. Fortunately, however, simple methods exist for simulating their effects artificially. These methods, which are particularly applicable to tape-recorded data, have been described fully in Part 1 of this article, where results were presented to show how a narrow-band 'time-symmetric filter' could be derived from parallel-tuned LC circuit. Part 2 now presents a discussion of the applicability of this filter to the analysis of gliding tones.

The content of Part 2 is arranged as follows: First, Barber and Ursell's analysis for the straightforward case of the tuned LC circuit is reviewed and summarized. The properties of the time-symmetric filter are then quoted, and its response to an arbitrary waveform is shown to be related quite simply to that of the parallel-tuned circuit from which it is derived. On this basis, its gliding tone response is evaluated by making use of the results of the earlier analysis. Next, the question of the best value of k to use when analysing a gliding tone is discussed, both for the simple tuned circuit and for the time-symmetric filter; the optimum responses of the two filters are compared, and their predicted shapes are confirmed experimentally. The results are then summarized and discussed. Finally, in an appendix, the possibility of using a series-tuned circuit, instead of a parallel-tuned circuit, is examined.

The Response of a Parallel-tuned Circuit

Barber and Ursell, in their original paper[8], dealt with a mechanical resonant system, the electrical analogue of which is a series-tuned circuit; they calculated its response to a gliding tone of steadily falling pitch. In reproducing their analysis here, it was found convenient to translate it into electrical terms, and to modify it slightly so as to make it apply to the parallel-tuned instead of the series-tuned circuit, and to tones of rising as well as of falling pitch. Also it has been simplified by assuming throughout that the Q-factor of the resonant circuit is much greater than unity. The notation used is that of Part 1, and the argument proceeds as follows:—

The response of a filter to an arbitrary input $V_{in}(t)$ is given by the convolution integral

$$V_{out}(t) = \int_{-\infty}^{\infty} V_{in}(t-\tau)\, u(\tau)\, d\tau \qquad \text{(c.f. equation (1))}$$

where $u(t)$ is its impulse response. For a parallel-tuned circuit, the impulse response is

$$u(t) = \begin{cases} 0 & (t<0) \\ \tfrac{1}{2} u_0 & (t=0) \\ u_0 \exp(-\pi W t)\cos(2\pi f_0 t) & (t>0) \end{cases} \qquad \text{(c.f. equation (26))}$$

Accordingly, when the gliding tone of equation (48) is applied to the input, the output is

$$V_{out}(t) = V_0 u_0 \int_0^{\infty} \sin\phi(t-\tau)\exp(-\pi W\tau)\cos(2\pi f_0 \tau)\, d\tau \quad (50)$$

From here on, the time t is measured from the moment of resonance, when $f = f_0$; this choice of origin involves no loss of generality.

The convolution integral (50) may be evaluated approximately by expanding $\phi(t-\tau)$ as a power series in τ, and neglecting all higher terms than that involving τ^2. Thus:

$$\begin{aligned}\phi(t-\tau) &= \phi(t) - \tau\,\phi'(t) + \tfrac{1}{2}\tau^2\phi''(t) + \dots\dots\\ &\simeq \phi(t) - 2\pi\tau f(t) + \pi\tau^2 f'(t)\\ &\simeq \phi(t) - 2\pi f_0\tau + \pi f'[\tau^2 - 2t\tau] \quad \dots\dots(51)\end{aligned}$$

where

$$f' = df/dt \quad \dots\dots (52)$$

and use is made of the approximation

$$f(t) \simeq f_0 + f'\,t \quad \dots\dots (53)$$

Evidently the neglect of higher terms in the power series amounts to taking f' as constant, which is to assume that the instantaneous frequency varies linearly with time through the pass-band of the filter.

By substituting this approximation for $\phi(t)$ into the convolution integral, Barber and Ursell show that the output from the filter can be expressed as the sum of two oscillatory terms, one of which is in phase with the oscillations of the original gliding tone, while the other is in quadrature:—

$$V_{out}(t) = V_0 u_0\,[B(t)\sin\phi(t) + A(t)\cos\phi(t)] \quad \dots\dots(54)$$

The amplitude factors, which vary with time relatively slowly, are given by

$$\left.\begin{matrix}A(t)\\ B(t)\end{matrix}\right\} = \tfrac{1}{2}\int_0^{\infty}\exp(-\pi W\tau)\left\{\begin{matrix}\sin\\ \cos\end{matrix}\right\}[\pi f'(\tau^2 - 2t\tau)]\,d\tau \quad (55)$$

The two terms combine together to form an oscillation of intermediate phase that has an envelope given by

$$E(t) = V_0 u_0\left[\{A(t)\}^2 + \{B(t)\}^2\right]^{\frac{1}{2}} \quad \dots\dots (56)$$

To obtain the shape of this envelope, the integrals for the amplitudes $A(t)$ and $B(t)$ must be examined. Now the Admiralty Computing Service has prepared tables[10] of the functions

$$\left.\begin{matrix}S(T)\\ C(T)\end{matrix}\right\} = k^{\frac{1}{2}}\int_0^{\infty}\exp(-z)\left\{\begin{matrix}\sin\\ \cos\end{matrix}\right\}(2Tk^{\frac{1}{2}}z - kz^2)\,dz \quad \dots (57)$$

The unknown integrals can be reduced to this form by setting

$$z = \pi W\tau \quad \dots\dots(58)$$

$$k = |f'|/\pi W^2 \quad \dots\dots (59)$$

$$T = (\pi|f'|)^{\frac{1}{2}}t \quad \dots\dots (60)$$

The parameters z, k, and T are dimensionless, and have the following meanings:—

z is the ratio of the time displacement τ to the decay time of the impulse response of the filter (c.f. equation (26)).

k involves the bandwidth W, and also the modulus of the rate of change of frequency of the gliding tone. The latter quantity has dimensions of (frequency)^2, so that its square root $|f'|^{\frac{1}{2}}$ defines a bandwidth; this is in fact the range of frequencies around f_0 over which the different Fourier components of the tone are more or less in phase at $t = 0$. Thus k is proportional to the square of the ratio of this 'inherent bandwidth' of the gliding tone to the bandwidth of the filter[11].

T is proportional to the ratio of the time t to the duration of the period around $t = 0$ over which these same Fourier components more or less maintain their phase coherence. The meaning of T may be clarified, perhaps, by noting the identity

$$Tk^{\frac{1}{2}} = t|f'|/W$$
$$= \triangle f/W, \text{ say} \quad \ldots\ldots\ldots\ldots (61)$$

where

$$\triangle f = \pm (f - f_0) \ldots\ldots\ldots\ldots (62)$$

and where the positive sign is used for ascending tones, and the negative sign for descending tones. Thus the quantity $\triangle f$ is equal in magnitude to the difference between the instantaneous and resonant frequencies, but it has the same sign as t has, so that it is reckoned negative before and positive after resonance. Equation (61) shows that, for a given value of k, T is proportional to the ratio of $\triangle f$ to the bandwidth of the filter. In sum, T is a normalized measure of time.

On rewriting the integrals in terms of these parameters, the amplitudes are found to be given by

$$A(t, W, f') = \pm \frac{1}{2\pi W k^{\frac{1}{2}}} \quad S(T, k) \ldots\ldots (63)$$

$$B(t, W, f') = \frac{1}{2\pi W k^{\frac{1}{2}}} \; C(T, k) \quad \ldots\ldots \quad (64)$$

where, in equation (63), the positive sign applies to descending tones, and the negative sign to ascending tones, i.e., the sign is opposite to that of f'.

Thus the envelope of the gliding tone response is given by

$$E(t) = V_0 \frac{u_0}{2\pi W} k^{-\frac{1}{2}} \left[\left\{S(T)\right\}^2 + \left\{C(T)\right\}^2\right]^{\frac{1}{2}} \ldots\ldots (65)$$

Now, from Part 1

$$\frac{u_0}{2\pi W} = y_0 \; \ldots\ldots \text{(c.f. equation (28))}$$

where y_0 is the amplitude of the frequency response of the filter at resonance.

So, finally,

$$E(t) = V_0 \, y_0 \left\{k^{-\frac{1}{2}} F(T)\right\} \; \ldots\ldots\ldots\ldots (66)$$

where

$$F(T) = \left[\left\{S(T)\right\}^2 + \left\{C(T)\right\}^2\right]^{\frac{1}{2}} \ldots \; (67)$$

Values of the function $F(T)$ are given in the Admiralty Computing Service tables[10], and are reproduced in some tables prepared recently by Grierson and Parr[12]. Fig. 6 shows graphs of $F(T)$ for certain selected values of k; this figure is similar to Fig. 1 of Barber and Ursell's paper. The main features of these graphs have been discussed already, but it is worth noting again that the shape depends on the bandwidth W and the sweep rate f' only through their combination in the dimensionless parameter k. Also, as k involves only the magnitude of f' and not its sign, the envelope of the response has the same shape, at a given rate of frequency sweep, whether the frequency is rising or falling.

It is easy to show that, in the limit of very large k (high sweep rate or narrow bandwidth), the response near $T = 0$ reduces to the form indicated by the broken line in Fig. 6; this distribution is familiar in optics, where it is encountered in the problem of Fresnel diffraction at a straight edge. In the opposite limit of very small k (low sweep rate or wide bandwidth), the gliding tone response simply follows the frequency response curve of the filter (equation (21)).

The Time-Symmetric Filter

In Part 1 methods were presented for simulating a filter that has the following time-symmetric overall impulse response:

$$U(t) = U_0 \exp(-\pi W|t|) \cos(2\pi f_0 t) \quad \text{(c.f. equation (30))}$$

This is related to the impulse response of a parallel-tuned circuit by the expression

$$U(t) = \frac{U_0}{u_0} \left\{u(t) + u(-t)\right\} \qquad \text{(c.f. equation (29))}$$

where $u(t)$ is given by equation (26) and $u(-t)$ is its time-reverse.

Evidently, to find the overall response of the time-symmetric filter to any waveform, it is necessary only to calculate the corresponding responses for the simple parallel-tuned circuit, first with the impulse response the normal way round, and secondly with it reversed, then to add the two results, and finally, to multiply by the amplitude factor U_0/u_0. The gliding tone response of the parallel-tuned circuit has been evaluated in the preceding section, and in the section which now follows the result is applied to yield the overall gliding tone response for the time-symmetric filter.

The Overall Response

The first step in calculating the overall response of the time-symmetric filter is to find out how the gliding tone response of the simple parallel-tuned circuit would be altered if somehow its impulse response was reversed in time. The output from the circuit would be given by

$$V_{out}(t) = \int_{-\infty}^{\infty} V_{in}(t-\tau)\, u(-\tau)\, d\tau \qquad \text{(c.f. equation (1))}$$

$$= \int_{-\infty}^{\infty} V_{in}(t+\tau)\, u(\tau)\, d\tau \qquad (68)$$

where V_{in} is in the gliding tone of equation (48). On developing the analysis as before, the following result is obtained:

$$V_{out}(t) = V_0 u_0 [B(-t) \sin\phi(t) + A(-t) \cos\phi(t)] \qquad (69)$$

where $A(t)$ and $B(t)$ are given by equation (55) as before. This result may be compared with that of equation (54) which applies to the circuit when its impulse response is the normal way round. These two results may now be combined to give the overall response of the time-symmetric filter to the gliding tone:

$$V_{out}(t) = V_0 U_0 [\{B(t) + B(-t)\} \sin\phi(t) + \{A(t) + A(-t)\} \cos\phi(t)] \;..\; (70)$$

The envelope of the response is

$$E(t) = V_0\, U_0 [\{A(t) + A(-t)\}^2 + \{B(t) + B(-t)\}^2]^{\frac{1}{2}} \quad (71)$$

By manipulations similar to those by which equation (66) was obtained from equation (54) this expression may be reduced to the final form

$$E(t) = V_0\, Y_0 \left\{\tfrac{1}{2}\, k^{\frac{1}{2}}\, G(T)\right\} \quad \ldots\ldots\ldots (72)$$

where Y_0 is the amplitude of the overall frequency response at resonance, and

$$G(T) = [\{S(T) + S(-T)\}^2 + \{C(T) + C(-T)\}^2]^{\frac{1}{2}} \; (73)$$

This function represents the shape of the envelope. Its values have also been tabulated by Grierson and Parr[12]. In Fig. 7 $G(T)$ is plotted out for the same values of the parameter as apply to the graphs of Fig. 6. The envelope of the response is seen to have its maximum value at resonance ($T = 0$) for all values of k, and also to be symmetrical about this point, as is obvious from the form of equation (73). This behaviour contrasts with that of the simple tuned circuit. If the corresponding responses of the two filters are compared, the time-symmetric response is seen to be slightly broader in its main peak, but its subsidiary peaks are much reduced in relative amplitude.

In the limit of large k, $G(T)$ reduces to a constant value of $\pi^{\frac{1}{2}}$ (the broken line in Fig. 7); this result may be understood to mean that the filter simply picks out the Fourier component of the signal at the frequency f_0. In the limit of small k, $G(T)$ reproduces the overall frequency response curve of the filter (equation (23)).

Optimum Conditions for Analysis

The problem to be considered now is that of the best value of k to use when analysing a gliding tone with a filter of a given type. The simple tuned circuit and the time-symmetric filter are considered separately, and the optimum k is found for each; the gliding tone is assumed to be given, so that the sweep rate f' is fixed, but k may be varied by altering the bandwidth W of the filter (c.f. equation (59)). Later the optimum responses are compared, and in this way the relative merits of the two types of filter are examined.

Now the choice of k would present no problem if the record contained just a single gliding tone, without any noise. In such a case the time of maximum response could be measured exactly, whatever the value of k, so all values would be equally useful. But in the present application (see above) the analysis of each gliding tone is confused by interference from other, neighbouring tones and from spurious noise. The optimum k is that value which leads to the most accurate measurements in the presence of these sources of interference.

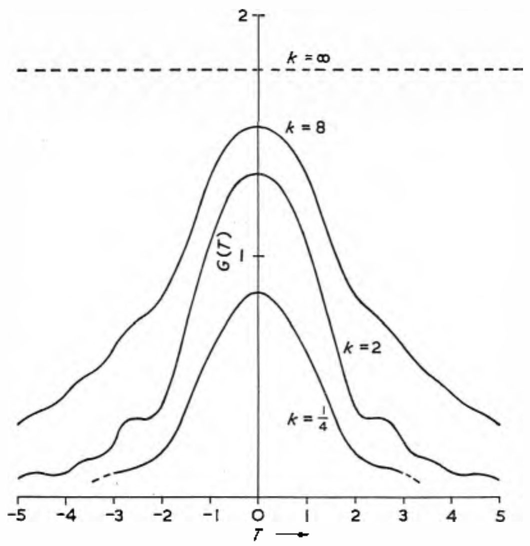


Fig. 7. Envelopes of the response of the time symmetric filter

Consider first the problem of discriminating against neighbouring tones. Clearly, for this purpose, the response should be as sharp as possible, i.e., its width should be a minimum. The measure of the width of the response that will be adopted here is the interval of normalized time T between the two '6dB points', that is to say, the points where the amplitude is half the maximum value. Fig. 8 shows how this quantity varies with the parameter k, both for the simple tuned circuit (curve A) and for the time-symmetric filter (curve B). From inspection of these curves it appears that the response of the tuned circuit has minimum width at about $k = 4$*, while the minimum for the time-symmetric filter is near $k = \frac{1}{2}$. Both minima are broad, so the choice of k is clearly not very critical.

Now consider the problem of discriminating against noise. The following general argument shows that, if white noise is present together with the gliding tone at the input to the filter, the best analysis is still obtained by using the value of k that yields the sharpest response. The measure that will be adopted for the detectability of the response in the presence of noise is the peak signal-to-noise ratio at the output, which may be defined thus:

$$\sigma = \begin{bmatrix}\text{Peak signal output power, in absence} \\ \text{of noise}\end{bmatrix} \div \begin{bmatrix}\text{Mean noise output power, in absence} \\ \text{of signal}\end{bmatrix}$$

$$= \tfrac{1}{2} E^2_{max} / V_n^2, \text{ say} \quad \ldots\ldots\ldots\ldots \quad (74)$$

Here the output voltages are supposed to be applied to some resistive load, and the power referred to is that dissipated in the load. The factor $\frac{1}{2}$ appears in equation (74), because, at the output from the filter, the signal and the noise are both quasi-sinusoidal oscillations, so that they must be compared on the same footing. Accordingly, when comparing it with the mean noise power, the peak signal power must be taken as the average of the signal power over a complete cycle of oscillation at the maximum of the response, rather than as

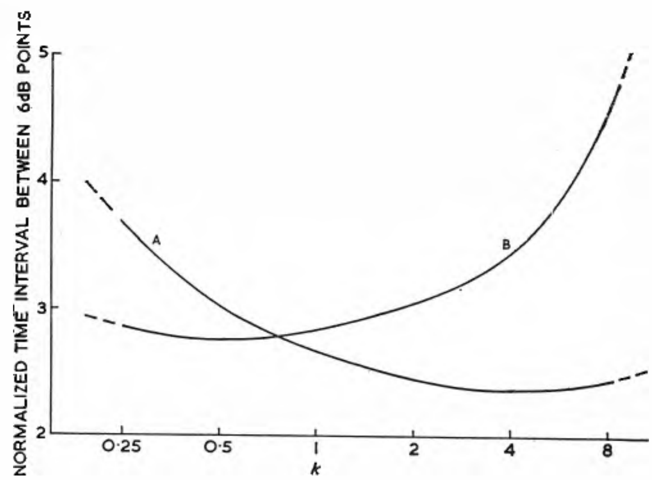


Fig. 8. Dependence of the widths of the response of the parameter k
Curve A. Simple tuned circuit
Curve B. Time-symmetric filter

the instantaneous power at the peak of a cycle; for a sinusoidal oscillation, the average power is half the instantaneous peak power.

The way in which σ varies with the filter bandwidth W is deduced as follows:—

If the frequency of the gliding tone varies smoothly over a wide range, then clearly its energy spectrum must be more or less level over any relatively narrow part of this range. Therefore the amount of energy which the filter accepts from the signal at its input is proportional to its bandwidth W. At its output, this same energy forms a response that has a height E_{max} and is spread over a certain time, δt say. Thus it follows from the law of conservation of energy that

$$E^2_{max}\, \delta t \propto W \quad \ldots\ldots\ldots\ldots \quad (75)$$

If the noise at the input to the filter has a level power spectrum, then the total noise power at the output, in the absence of signal, is also proportional to the bandwidth, i.e.:

$$V_n^2 \propto W \quad \ldots\ldots\ldots\ldots \quad (76)$$

Hence it appears that

$$\sigma \propto 1/\delta t \quad \ldots\ldots\ldots\ldots \quad (77)$$

so the adjustment for sharpest response also yields the highest signal-to-noise ratio. This conclusion is approximate only, since the form of the response as well as its width

* *From a slightly different criterion of sharpness, Barber and Ursell concluded that the value $k = 3{\cdot}1$ was optimum for the simple tuned circuit.*

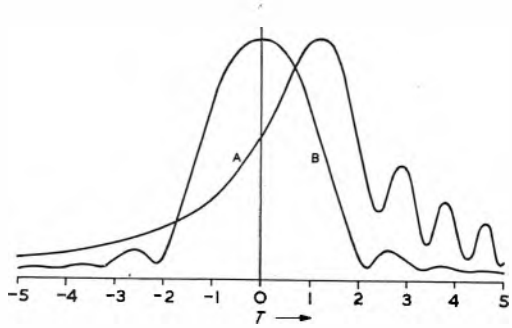


Fig. 9. Comparison of the theoretical optimum responses
Curve A. Simple tuned circuit ($k = 4$)
Curve B. Time-symmetric filter ($k = \frac{1}{2}$)

changes when k is altered, with the result that the proportionality (75) is inexact.

The optimum responses for the two types of filter are graphed in Fig. 9. The two responses have been plotted on the same scale of normalized time; also, they have been normalized to the same peak amplitude.

Experimental Results

To provide a qualitative test of the analysis, the optimum responses of the simple tuned circuit and of the time-symmetric filter were recorded by cathode-ray oscillograph on moving film.

An artificial gliding tone, in the form of a recording on magnetic tape, was used in the test. The original tone was obtained from a beat-frequency oscillator, in which part of the tuning capacitance for the variable oscillator was supplied by a motor-driven variable capacitor, so that the frequency of oscillation was swept back and forth continuously. The frequency sweep was substantially linear in the middle of its range, the sweep rate being $500\text{c/s}^2 \pm 5$ per cent. The tone was recorded on one track of a multi-track tape recorder, while time marker pulses were recorded simultaneously on another track. The responses of the two filters were both obtained from the same recording, so that their relative timing could be established by aligning the marker pulses.

The magnetic tape recorder and the filter system have been described in Part 1. The bandwidth of the filter was set at 6·3c/s to give $k = 4$, and at 17·8c/s for $k = \frac{1}{2}$, while its resonant frequency was 600c/s in both cases. Before entering the oscillograph, the output from the filter was passed through a half-wave rectifier, but not smoothed, and then the time marker pulses were added to it, using a combined adding and switching circuit which suppressed the signal during the marker pulses[13]. In this way the oscillograms were arranged to show the envelope of the output as displacements of the trace in one direction, and the timing marks as displacements in the opposite direction, from a fixed baseline .

The oscillograms of the two optimum responses are given in Fig. 10. The small timing marks are spaced at intervals of 10msec, and the large marks at intervals of 50msec. The larger interval corresponds roughly to an interval of 2 in normalized time T. Though the observed and calculated responses have not been compared quantitatively, their general agreement appears satisfactory.

Discussion and Conclusions

On comparing the optimum responses of the two filters, it becomes clear that the time-symmetric filter is more useful than the simple tuned circuit in the analysis of gliding tones, for the following reasons:—

(*a*) Its response is symmetrical and centred on the moment of resonance, so that no correction need be applied to the observed time of maximum response. This is its main advantage.

(*b*) The subsidiary maxima are considerably reduced in amplitude. The first two such maxima, which are symmetrically disposed on either side of resonance, have amplitudes of only 13 per cent of the value at resonance. Hence there is less likelihood of confusion between subsidiary maxima and the responses produced by adjacent tones.

(*c*) The general level of the skirts of the response is lower, so that it should be possible to resolve gliding tones at closer spacing.

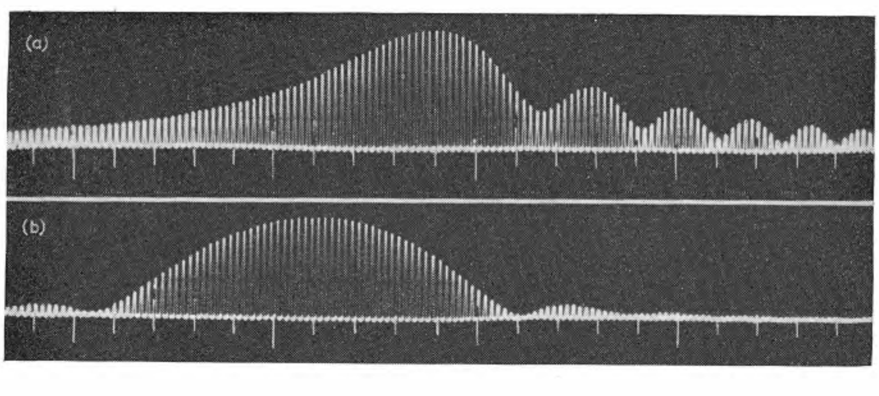
These remarks should be qualified by pointing out that the main maximum of the time-symmetric response is slightly the wider (c.f. Fig. 8), and that the effects at (*b*) and (*c*) are probably due more to the sharper cut-off of the time-symmetric filter, that is to say, its greater attenuation of frequencies well removed from resonance (c.f. Fig. 4 of Part 1), than to its freedom from phase-shifts, because similar improvements are obtained when the simple tuned circuit is replaced by two such circuits in cascade[14].

To conclude, the narrow-band time-symmetric filter is likely to prove itself useful in the analysis of gliding tones, whenever the highest accuracy of time measurement is called for, and the greater time required for this technique is not objectionable. The application envisaged originally was to the precise analysis of whistling atmospherics[15], but it might also be applicable to analysing the Doppler-shifted continuous-wave transmissions from the artificial earth satellites.

Acknowledgments

The authors wish to thank Mr. P. A. Field, Mr. W. L. Hatton and Dr. L. L. Campbell for their advice in the preparation of this article.

Fig. 10. Comparison of the observed optimum responses
(*a*) *Simple tuned circuit* ($k = 4$)
(*b*) *Time-symmetric filter* ($k = \frac{1}{2}$)



APPENDIX

THE SERIES-TUNED CIRCUIT

In the appendix to Part 1 it was shown how, starting from a series-tuned circuit, a filter can be simulated that has an overall impulse response somewhat resembling that of the time-symmetric filter described above; the envelope of the response is the same, but the oscillation within the envelope is a sine instead of a cosine wave, i.e.,

$$U(t) = U_0 \exp(-\pi W|t|) \sin(2\pi f_0 t)$$

(c.f. equation (43))

This overall impulse response is related to the impulse response $u(t)$ of the series-tuned circuit by the expression:

$$U(t) = \frac{U_0}{u_0}\left\{u(t) - u(-t)\right\} \quad \text{(c.f. equation (42))}$$

where $u(t)$ is given by

$$u(t) = \begin{cases} 0 & (t < 0) \\ u_0 \exp(-\pi W t)\sin(2\pi f_0 t) & (t \geqslant 0) \end{cases} \quad \text{(c.f. equation (39))}$$

Since this overall impulse response is so similar to the other, it is interesting to compare the gliding tone responses. Equation (42) shows that the gliding tone response for the present case can be obtained by evaluating the corresponding responses for the series-tuned circuit, first with the impulse response the normal way round and secondly with it reversed, then subtracting the two results, and finally multiplying by the amplitude factor U_0/u_0 in the same general way as before.

Barber and Ursell's analysis (which was, indeed, developed originally for the series-tuned circuit) gives the gliding tone response as

$$V_{\text{out}}(t) = V_0 u_0 \left[A(t)\sin\phi(t) - B(t)\cos\phi(t)\right] \quad \ldots (78)$$

where the amplitude factors $A(y)$, $B(t)$ retain their previous meaning (equation (55)), and the filter has its normal impulse response. The envelope of V_{out} is

$$E(t) = V_0 u_0 \left[\left\{A(t)\right\}^2 + \left\{B(t)\right\}^2\right]^{\frac{1}{2}} \quad \text{(c.f. equation (56))}$$

which is the same as for the parallel-tuned circuit.

If the impulse response of the filter were reversed somehow, the gliding tone response would be

$$V_{\text{out}}(t) = V_0 u_0\left[-A(-t)\sin\phi(t) + B(-t)\cos\phi(t)\right] \ldots (79)$$

On combining the results of equation (78) and (79), the overall gliding tone response of the simulated filter is found to be

$$V_{\text{out}}(t) = V_0 U_0 \left[\left\{A(t) + A(-t)\right\}\sin\phi(t) - \left\{B(t) + B(-t)\right\}\cos\phi(t)\right] \quad \ldots (80)$$

The difference between this response and that of the time-symmetric filter (equation (67)) lies only in the phase of the oscillatory terms. Their envelopes are the same, viz:—

$$E(t) = V_0 U_0 \left[\left\{A(t) + A(-t)\right\}^2 + \left\{B(t) + B(-t)\right\}^2\right]^{\frac{1}{2}} \quad \text{(c.f. equation (68))}$$

so the rest of the theory proceeds as before. Evidently this filter is equally applicable to the analysis of gliding tones.

Errata

Equation (39), Part 1, should read:

$$U(t) = \begin{cases} 0 & (t < 0) \\ U_0 \exp(-\pi W t)\sin(2\pi f_0 t) & (t \geqslant 0) \end{cases}$$

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International Colloquium on Nuclear Electronics

Electronic techniques have been applied to most, if not all, fields of scientific endeavour. In some of these, new techniques have had to be devised to meet the particular problems encountered and this, combined with the general growth of activity in electronics, has led to the establishment of separate specialized fields of electronics.

The emergence of Nuclear Electronics as a field on its own was marked by an International Colloquium on the subject held in Paris from 16 to 20 September. The Colloquium was organized, most admirably, by the Société des Radioelectriciens and was attended by some 500 delegates. Something in the region of 80 short papers were presented in nine sessions.

The first two sessions were concerned mainly with devices for the conversion of the nuclear radiation to electrical signals. The scintillation method is receiving a great deal of attention at present and a number of papers were presented giving details of work carried out with a view to understanding the basic processes in scintillators and in evaluating the performance of a variety of existing scintillators. Recent work on photomultipliers, particularly on the reduction of transit time spread was described. A paper was read on a gas Cerenkov counter and details were given of several gamma ray spectrometers.

One session was devoted to fast pulse techniques. At the present time, an electronic circuit is regarded as fast if the pulse rise-time is less than, say, 10 nanoseconds. A number of papers dealing with the generation, handling and measurement of such pulses were given. The application of slow pulse techniques to the nuclear field and particularly to the design of pulse height analysers was dealt with at another session entitled "Classical Electronics".

Several pulse height analysers were also described at the session on Data Processing. These reflected the growing interest in two and three-dimensional analysers. A relatively simple apparatus, in which two coincident pulses from separate counters are applied directly to the X and Y plates of a cathode-ray tube, has been found to be very useful as an instrument for chemical analysis. (1 part in 10^7 of manganese in rats liver has been detected). A 96 × 32 channel XY analyser with magnetic drum storage and a 256 × 256 × 256 channel XYZ analyser with a 25 channel magnetic tape store were also described.

Several papers on the processing and presentation of reactor data were read and two sessions were devoted to reactor instrumentation and simulation. Health monitoring and radioactivity survey instruments were the subject of another session.

One session was devoted to the use of transistors in nuclear electronic systems. The trend in this direction could be seen from the papers presented in other sessions and it was held to be highly desirable particularly from the aspect of reliability.

Crane-Approach Warning System

In situations where two or more overhead travelling cranes are operated on the same track there is constant danger of collision, because the crane driver must concentrate on his load and the signals of his slinger rather than on the movement of neighbouring cranes. To eliminate this danger a new electronic equipment has been developed (under the auspices of the British Iron and Steel Federation) by The British Thomson-Houston Co. Ltd, and an installation has been undergoing extensive trials at the Redbourn, Scunthorpe, iron and steel-works of Richard Thomas and Baldwins Ltd.

The BTH system is based fundamentally on the principle of the echo sounder. An oscillator circuit generates short electrical pulses at regular intervals which are emitted in the form of very high frequency sound by a loudspeaker fitted on the crane. The sound pulses are reflected by a special trihedral reflector fitted on the neighbouring crane and the echo is picked up by a microphone mounted alongside the sound source. The microphone signal is then amplified and fed to relays which may operate audible alarm or visual indicating circuits in the driver's cab, or may be used to initiate braking of the crane.

Automatic warning of approach can thus be given to one or more cranes operating on the same track, or to cranes operating at different levels.

The equipment can be adjusted to operate up to a distance of 60ft from the reflector. The oscillating frequency is 11kc/s and the time taken for a return echo to operate the protective system is 2msec/ft. Selective circuits are designed to make the alarm system unresponsive to any other signals than those of the genuine echo.

Design Curves for Simple Filters

By D. J. H. Maclean*

A series of charts is presented giving the normalized element values as a function of the allowable pass band variation for Butterworth or Chebyshev behaviour of the insertion loss. The values relate to LC filters of constant-k configuration working between certain resistive terminations, and can be read to an accuracy of 1 per cent. The charts are given for low-pass filters having from two to seven branches, and can be extended to high-pass, band-pass or band-stop filters by the well-known frequency transformations.

In the case of an odd number of branches, a simple procedure allows one to obtain the element values for a filter working between general finite and non-zero resistances from the element values given in the charts.

Graphical data is presented to show the number of branches required to meet a given pass band variation and stop band attenuation.

The minimum unloaded-Q required to realize the desired response shape is given for the three cases when the number of branches is odd.

Some examples illustrate the use of the charts.

THE article only considers a particularly simple structure of inductors and capacitors, namely single element series or shunt branches, connected alternately series-shunt-series and so on. This configuration, the ladder configuration, is, however, of great practical interest. The form of the low-pass ladder network is readily derived by considering the desired behaviour, namely no transmission loss at low frequencies and high loss beyond some arbitrary frequency. The basic low-pass structure is thus series inductors and shunt capacitors connected alternately (see Fig. 1). Such a network constructed from random elements of the appropriate kind would show some sort of low-pass behaviour. However, one is interested in a specific, predictable and realizable behaviour so that there must be definite relationships between the element values. To establish these, one must be given further information. Since only the amplitude-frequency characteristic is of interest here, the relevant information is the allowable variation of transmission in the pass band, the required minimum loss in the stop band, the frequency interval between the pass and stop band and finally the values of the terminating resistances. The primary design parameters are thus; allowable pass band variation, minimum stop band loss, frequency transition interval. Usually the pass band variation and the frequency interval are fixed, but the stop band specification represents a lower limit. If this is the case, two of the parameters can be combined to give one secondary parameter, the number of elements required. The approximation to constant transmission in the pass band can be done in several ways, two of which are normally used. The first historically is the Butterworth, or maximally flat, and the second the Chebyshev, or equal ripple approximation, both are illustrated in Fig. 2.

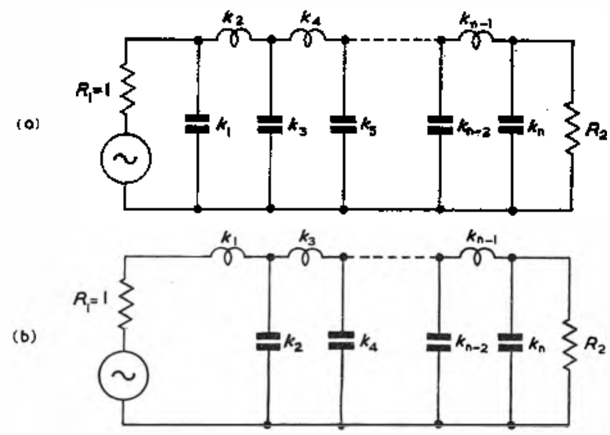

Fig. 1. The basic low-pass structure

In order to give design data in a usable form graphical presentation is desirable, but for this to be practicable some further restrictions are required. The restriction imposed here is one applying to the ratio of the terminations, and this will be considered later. Design data in graphical form has long been available for filters based on image parameter theories. Such information of itself does not enable an approximation to constant pass band transmission to be obtained without further extensive computation. Curves giving the element values as a function of the pass band insertion loss variation were given by Atiya[1], but not in convenient form. The present curves are offered as a direct method for the design of a restricted class of filter and terminations. The class of filters is the simple *LC* ladder with all its infinite loss points at infinite frequency. The terminations are restricted to be equal in the case of Butterworth response filters and are normalized to 1. For the Chebyshev approximation, the ratio of the terminations depends upon the number of branches, being unity if the number of branches is odd, and a definite, non-unity ratio for an even number of branches.

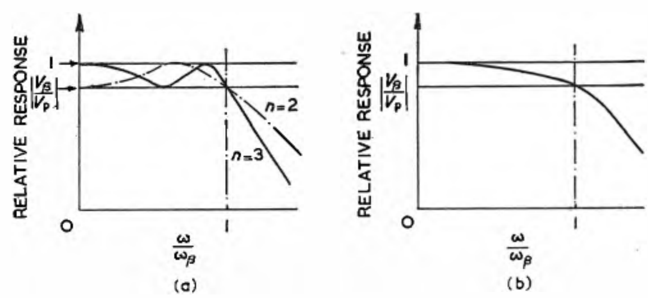

Fig. 2. The Butterworth and Chebyshev approximations

If there are an odd number of branches, the normalized element values given in the charts, which are for equal terminations, can readily be converted to the case of unequal terminations. The element values from the charts yield a symmetrical network which is then bisected along its plane of symmetry, giving two networks having mirror-image symmetry. This is an example of Bartlett's Bisection

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Theorem. The branch impedances of the right-hand network are then multiplied by the ratio R_2/R_1. The two networks are then rejoined yielding a filter having the same relative insertion loss, but working between resistances R_1, R_2 where $R_1 = R_2$. The insertion loss of the unsymmetrical filter is $\frac{1}{2}(1 + R_2/R_1)$ times that of the original types of response and the symbols used in the work.

The first two charts enable one to choose a suitable number of branches n from a given specification of the band-pass variation A_p and the minimum discrimination A_m. Chart 1 must be corrected for the 'pass band factor' shown in Chart 2. The smaller the permissible variation in

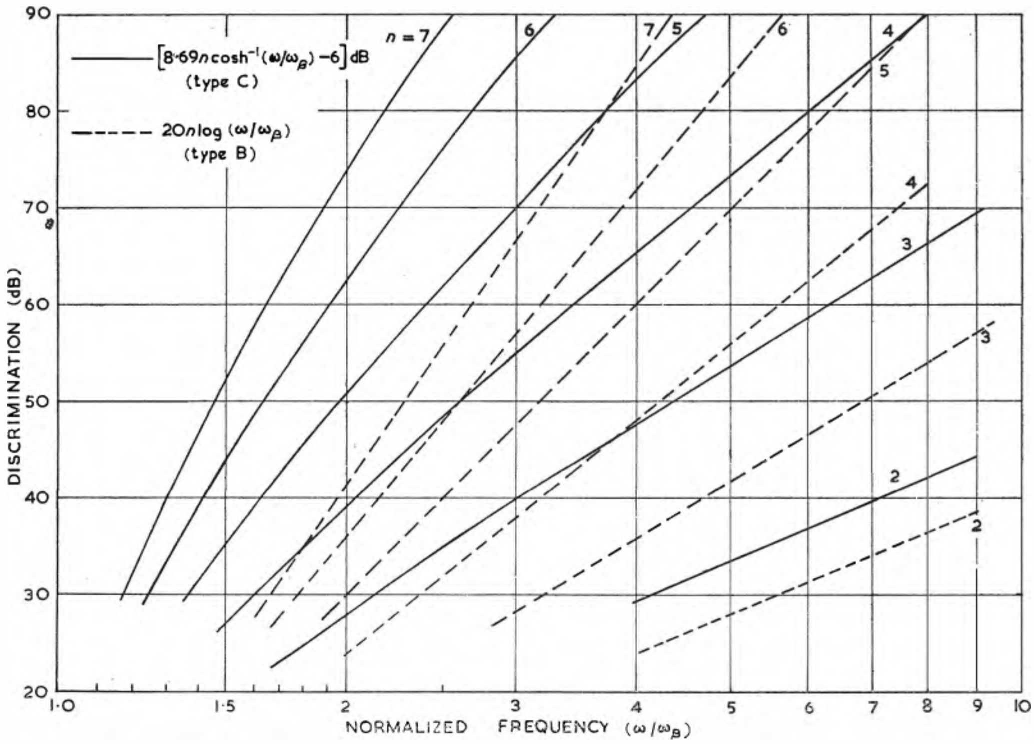


Chart 1. Discrimination/normalized frequency for different values of n

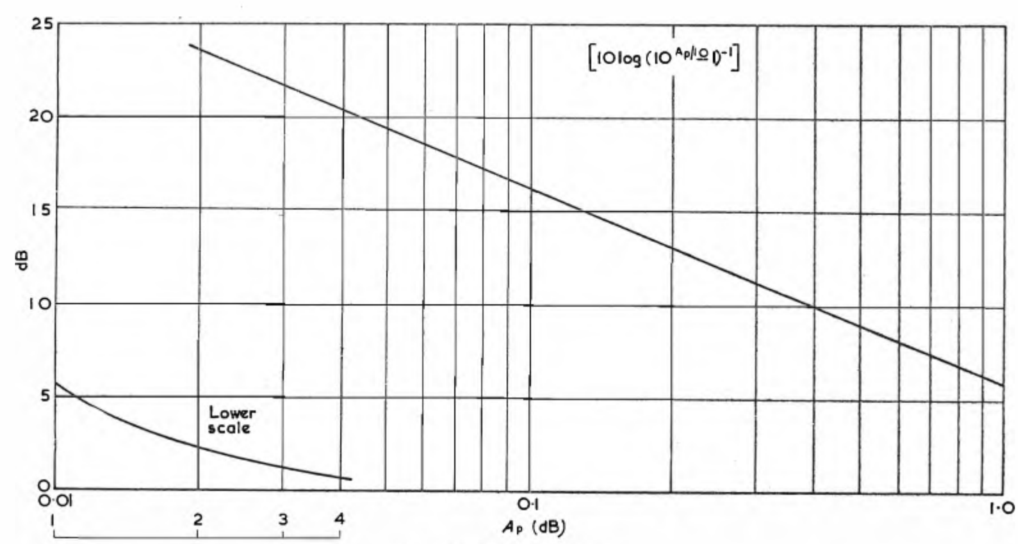


Chart 2. Pass band factor

symmetric network. Note that both R_1 and R_2 must be finite and non-zero. An illustrative example is given.

The Charts

Fig. 1 illustrates the types of filters under consideration, i.e., simple low-pass ladder filters. Fig. 2 shows the two

the pass band the lower is A_m for a given frequency and n.

The next set of Charts 3 to 8 give the element values as a function of A_p and of n for type B response. Charts 9 to 14 give the same information to provide type C response. The curves were obtained by extensive computation from the formulae shown in the appendix and the

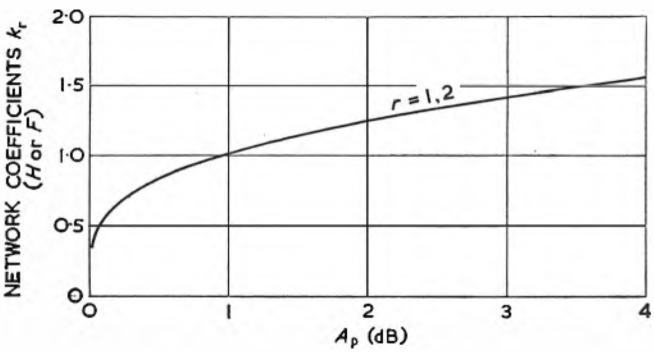


Chart 3. Element values for type *B* response. $n = 2$

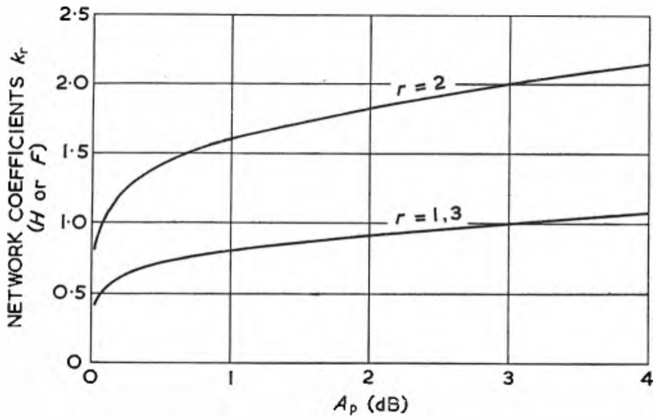


Chart 4. Element values for type *B* response. $n = 3$

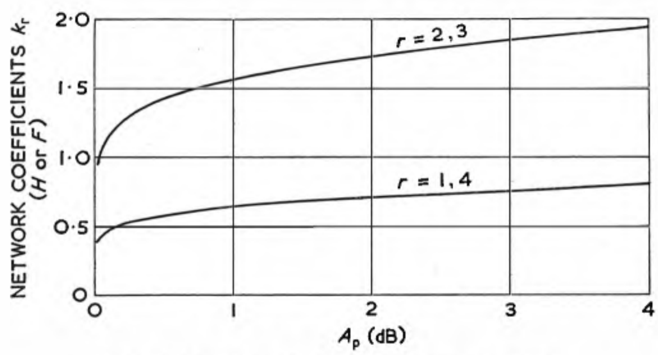


Chart 5. Element values for type *B* response. $n = 4$

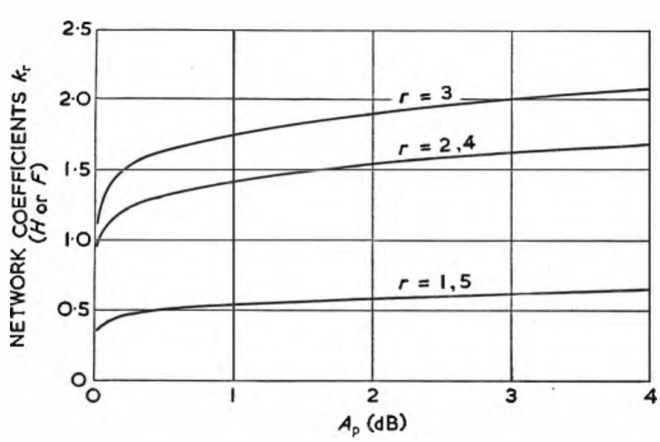


Chart 6. Element values for type *B* response. $n = 5$

working was done to an accuracy of about 0·3 per cent. Where possible the results have been checked against other sources, e.g. Mole[2] and Green[3]. It should be noted that when n is even and type C response is required, the shunt branch is on the side of larger R which in terms of Fig. 2 is R_2. It is, however, immaterial which side of the network is taken as the generator end. The value of R_2 for $R_1 = 1$ is given in Chart 15.

So far, ideal reactive elements have been assumed; in practice the Q of capacitors and inductors, particularly the latter, imposes a limitation on the performance which can be obtained. Following Dishal[4] one can express the limitation in terms of the 'minimum unloaded Q' that is, the lowest Q of the elements from which the filter will be constructed. If the available elements have an unloaded Q which is lower than the minimum value, the particular requirements cannot be met. The minimum Q for $n =$

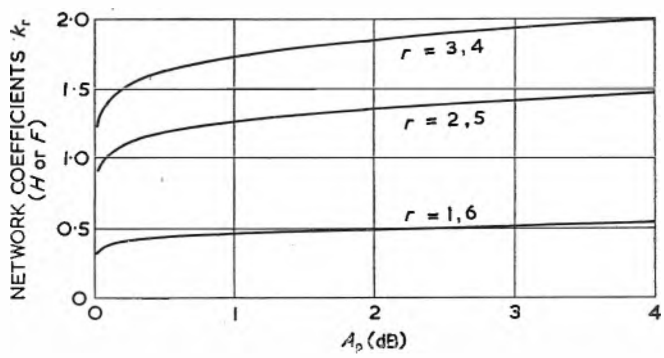


Chart 7. Element values for type *B* response. $n = 6$

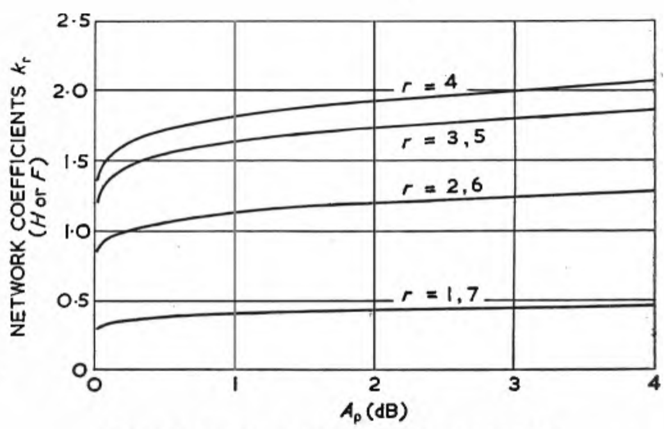


Chart 8. Element values for type *B* response. $n = 7$

3, 5, 7 and for both types of behaviour is shown in Charts 16, 17. It will be observed that type C behaviour imposes a much more stringent requirement on the minimum Q than does type B performance. These low-pass values can be readily converted to the band-pass or band-stop cases by multiplying by the factor $2/w$ where w is the fractional bandwidth.

To convert from the normalized element values found in the charts, one uses the formulae below:

$$L = kR/\omega_\beta,\ C = k/R\omega_\beta$$

Where k is the normalized inductance in henries or capacitance in farads. For conversion from low-pass values above one utilizes the relevant frequency transformation.

Example 1

A low-pass filter is to be designed to meet the following specification:

ATTENUATION	FREQUENCY INTERVAL	NORMALIZED INTERVAL
$< 0{\cdot}2$dB	$0 \leqslant f \leqslant 1\,000$ c/s	$0 \leqslant \omega/\omega_\beta \leqslant 1$
> 36dB	$2\,000 \leqslant f \leqslant \infty$	$2 \leqslant \omega/\omega_\beta \leqslant \infty$

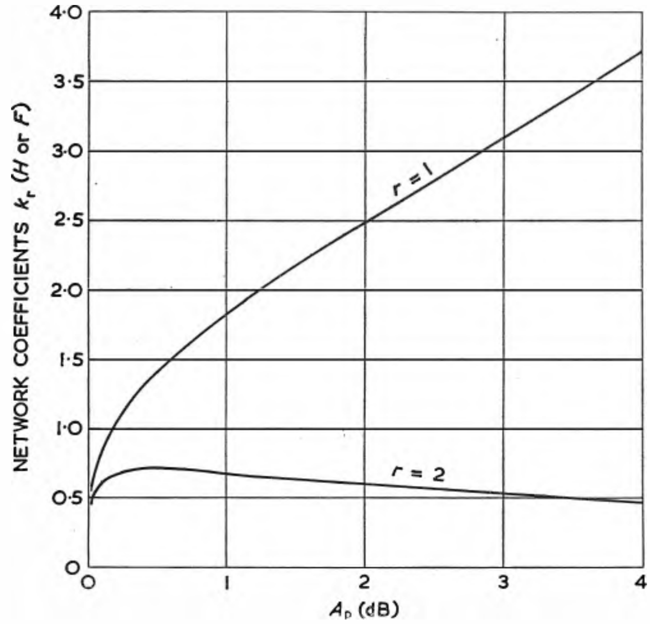


Chart 9. Element values for type C response. $n = 2$

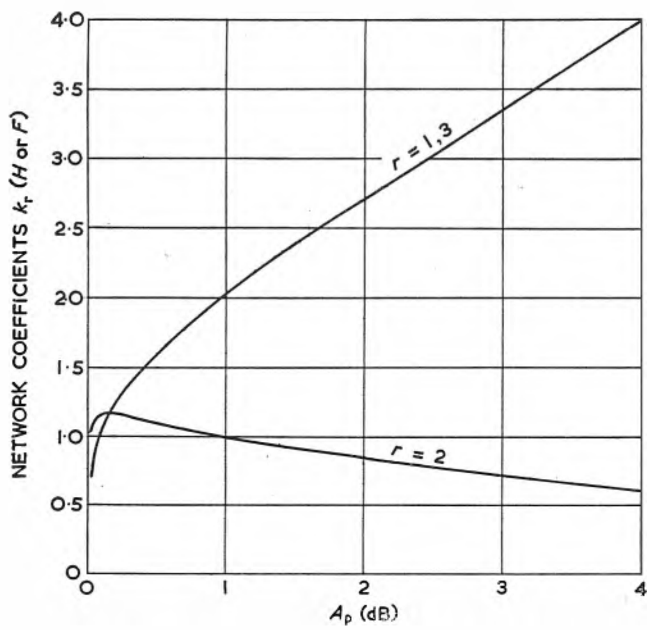


Chart 10. Element values for type C response. $n = 3$

The filter is to work between equal resistive terminations of value 1 000Ω. The available inductor Q is 100 and that of the capacitors comparatively infinite. A simple series L, shunt C ladder is required.

Design Procedure

(1) From Chart 2, the pass band factor for $A_p = 0{\cdot}2$dB is 13·2dB.

(2) The correction factor of 13·2dB is to be subtracted from all curves, alternatively the ordinate scale is suitably modified. At a normalized frequency of 2 one sees that the modified requirement of $36 + 13{\cdot}2 = 49{\cdot}2$dB could be met with $n = 5$ and type C behaviour. The corresponding number of elements for type B response is evidently $n = 9$.

(3) Since the specification includes a figure for Q, one must check that the above choice is suitable. From Chart 17 one sees that $Q_{min} = 7$ so that for $Q = 100$ the response will be little affected by finite Q. It might well be remarked that were a narrow band band-pass filter under consideration, the minimum fractional bandwidth w is given by:

$$(2/w)\,7 = 100$$

or $w = 0{\cdot}14$.

(4) The element values are obtained from Chart 12 as:

$$k_1 = k_5 = 1{\cdot}347$$
$$k_2 = k_4 = 1{\cdot}336$$
$$k_3 = 2{\cdot}219$$

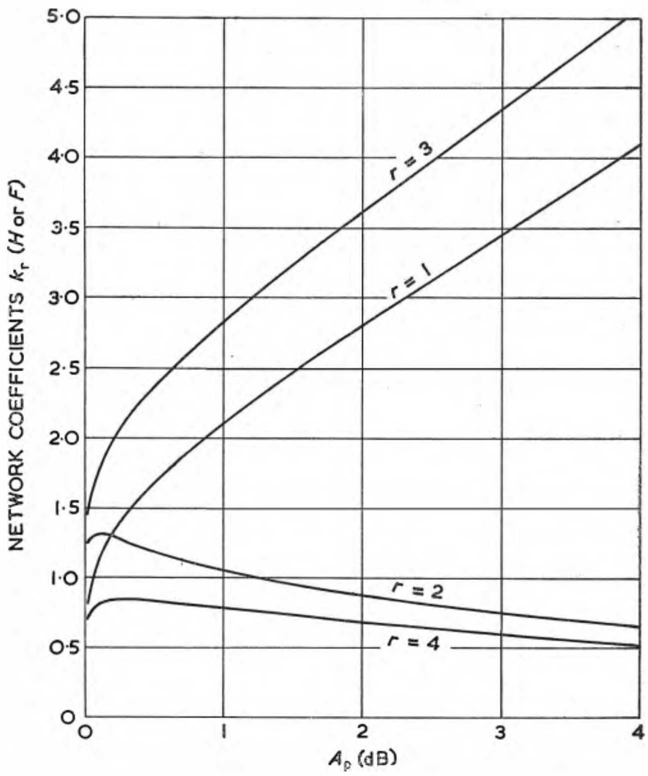


Chart 11. Element values for type C response. $n = 4$

(5) From the normalized values, the actual element values are obtained as follows:

$$C_1 = C_5 = 1{\cdot}347/2\pi 10^6 = 0{\cdot}224\mu\text{F}$$
$$L_2 = L_4 = 1{\cdot}336/2\pi = 0{\cdot}212\text{H}$$
$$C_3 = 2{\cdot}219/2\pi 10^6 = 0{\cdot}353\mu\text{F}$$

The dual network ($L_1 = L_5$, $C_2 = C_4$, L_3) is obtained similarly.

Example 2

A filter is required to meet the same specification as that of example (1) but is to operate between resistive terminations of $R_1 = 1\,000\Omega$ and $R_2 = 3\,000\Omega$.

One can commence immediately with the normalized element values (or with the final element values) given in step (4) above. The stages of the transformation from symmetric network to one terminated in 1Ω and 3Ω are illustrated in Fig. 3. The network (b) shows explicity the mirror image component structures. Network (c) is obtained by multiplying all the branch impedances of the right-hand structure by the ratio $R_2/R_1 = 3$, i.e. each inductance is multiplied by 3 and each capacitance by 1/3. The original left-hand section and the modified right-hand section are recombined to give a normalized network having the following element values.

$k_1 = 1{\cdot}347,\ k_2 = 1{\cdot}336,\ k_3 = 1{\cdot}48,\ k_4 = 4{\cdot}008,\ k_5 = 0{\cdot}449$

The actual element values are obtained from these in a precisely similar fashion to that of example (1).

The flat loss for this filter will be 0·5 (1 + 3) = 2 times that of the symmetric filter, which has a minimum insertion loss of 0dB, so that the asymmetric filter has a minimum insertion loss of 6dB.

EXAMPLE 3

A band-pass filter is required which meets the specification:

ATTENUATION	CENTRE FREQUENCY	FREQUENCY INTERVALS
< 0·5dB	500kc/s	$475 \leqslant f \leqslant 525$kc/s
⩾ 50dB		$450 \leqslant f \geqslant 550$kc/s

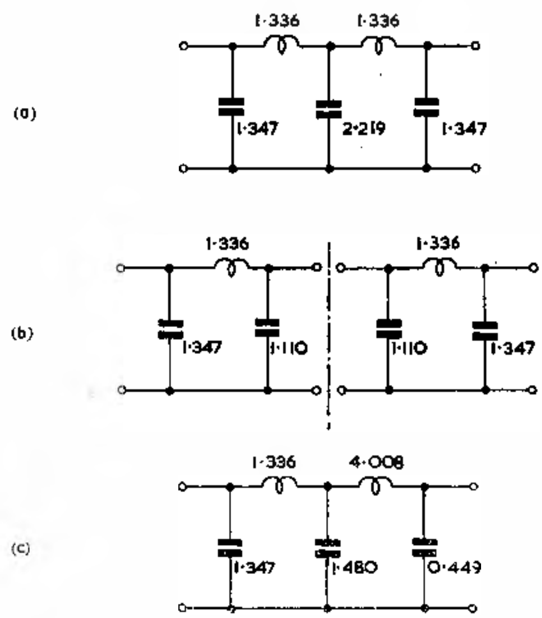


Fig. 3. Stages in the transformation from a symmetric network to one terminated in 1Ω and 3Ω

The available inductor Q at 500kc/s is 250.

A geometrically symmetrical response is desired so that the design can be based on a low-pass prototype network. This prototype network is required to meet the following specification:

ATTENUATION	FREQUENCY INTERVAL	NORMALIZED INTERVAL
< 0·5dB	$0 \leqslant f \leqslant 50$kc/s	$0 \leqslant \omega/\omega_\beta \leqslant 1$
⩾ 50dB	$100 \leqslant f \leqslant \infty$	$2 \leqslant \omega/\omega_\beta \leqslant \infty$

The equivalent Q is found as follows:

$$Q_e = (w/2)Q$$
$$= 50/1\,000 \times 250$$
$$= 12{\cdot}5$$

The computation can be carried out as shown below.

(1) *Pass band factor*

From Chart 2, the factor for $A_p = 0{\cdot}5$dB is 9dB.

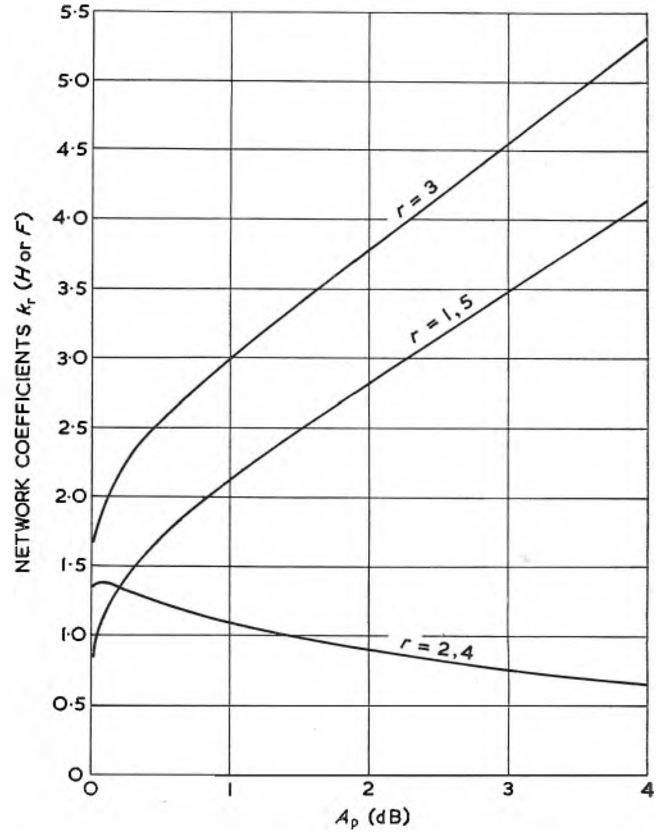
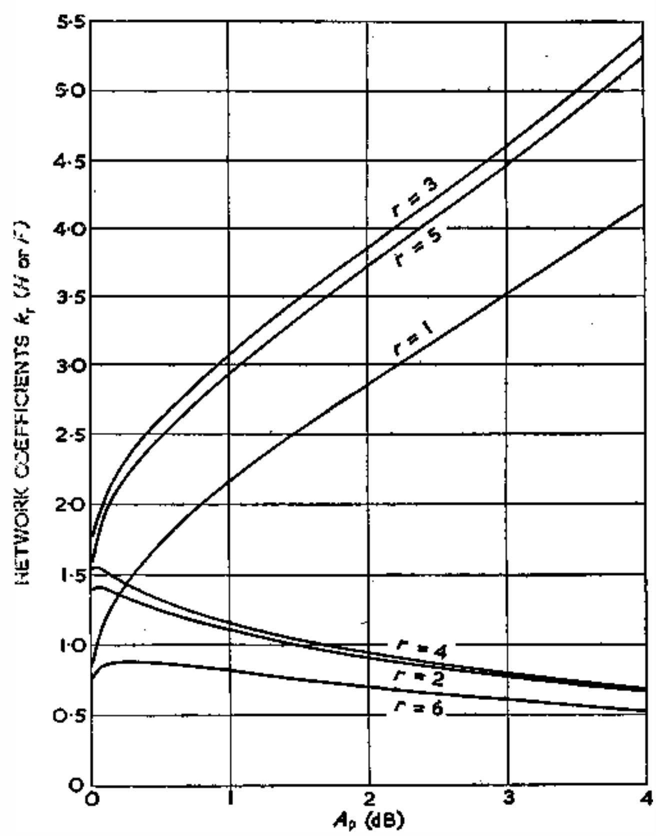


Chart 12. Element values for type C response. $n = 5$

Chart 13. Element values for type C response. $n = 6$



(2) *Estimate of number of elements required*

The figure for the discrimination to be used with Chart 1 is obviously 59dB. From the chart one sees that with $\omega/\omega_\beta = 2$, the required value for n is 6 if type C behaviour is chosen; it is evident that many more elements would be necessary to obtain type B behaviour. As a band-pass network is the ultimate object, the fact that n is even is not of importance.

(3) *Check on required Q minimum*

Chart 17 shows that for A_p = 0·5dB, the minimum unloaded Q is approximately 13, thus the desired behaviour cannot be realized as the design stands. It is evident, however, that if A_p is reduced the value of Q minimum is decreased. A suitable choice would be to alter A_p to 0·2dB. This alteration means that steps (1) and (2) must be

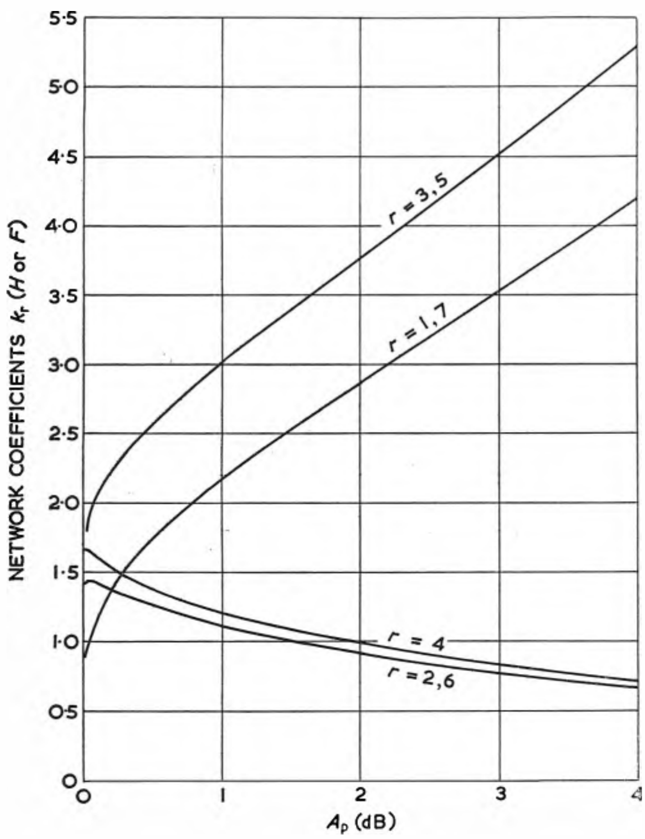


Chart 14. Element values for type C response. $n = 7$

repeated; the results are then:

(a) The pass band factor = 13·2dB (Chart 2).

(b) The discrimination of 63·2dB is just about met for n = 6, and is sufficiently close (Chart 1).

(c) The minimum Q is now 10 (Chart 17).

(4) *Element values*

From Chart 13, one finds:

$k_1 = 1{\cdot}350 \qquad k_2 = 1{\cdot}360$

$k_3 = 2{\cdot}230 \qquad k_4 = 1{\cdot}462$

$k_5 = 2{\cdot}091 \qquad k_6 = 0{\cdot}880$

If one chooses a network with series input (see Fig. 1(b)), the k's with odd subscripts correspond to inductance values and those with even subscripts to capacitance values.

(5) *The prototype low-pass filter*

The low-pass network from which the band-pass filter

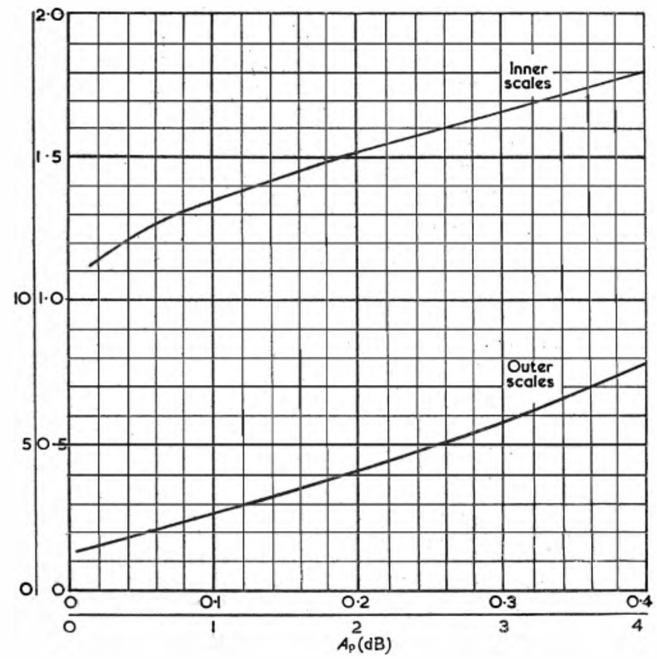


Chart 15. The value of R_2 for $R_1 = 1$

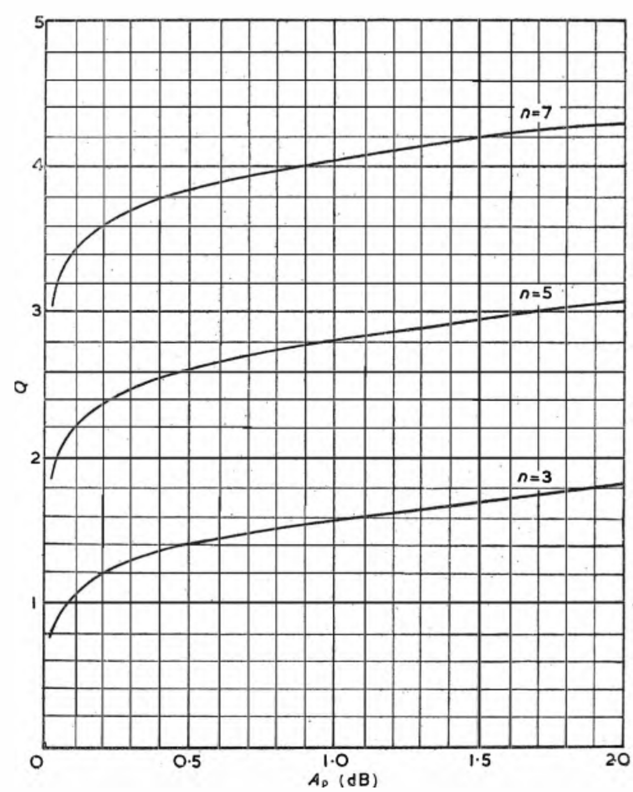


Chart 16. Minimum Q for type B response

will be derived is readily obtained from the normalized values found in step (4). The values are found from the formulae:

$L_i = (k_i R/\omega_\beta)$ H, i = 1, 3, 5; $C_j = (k_j/R\omega_\beta)$ F, j = 2, 4, 6, which yield:

L_1 = 4·30mH, $\qquad C_2$ = 4 330pF

$$L_3 = 7{\cdot}10\text{mH}, \qquad C_4 = 4\,660\text{pF}$$
$$L_5 = 6{\cdot}66\text{mH}, \qquad C_6 = 2\,803\text{pF}$$

for $R_1 = 1\text{k}\Omega$, $R_2 = 1{\cdot}52\text{k}\Omega$ (from Chart 15), and $\omega_\beta = 2\pi \,.\, 50 \,.\, 10^3$.

Summarizing the progress so far, a low-pass filter having a bandwidth of 50kc/s to 0·2dB points and 100kc/s to 50dB points has been designed.

(6) *The band-pass filter*

The method for obtaining this from the prototype low-pass filter is very simple. The mid-frequency has to be translated from zero to the geometric mean of 475kc/s and 525kc/s. This is accomplished by resonating each inductor with a series capacitor, and each capacitor with a shunt inductor, all circuits being tuned to 500kc/s (very closely the geometric mean frequency). The required element values are quickly obtained from the *LC*-product at 500kc/s, that is from the value of $1/\omega_r^2$.

$$LC\text{-product} = 101{\cdot}32\text{mH} \times \text{pF}.$$

$L_1 = 4{\cdot}30\text{mH}\,(C_1 = 23{\cdot}6\text{pF})$; $C_2 = 4\,330\text{pF}\,(L_2 = 23{\cdot}4\mu\text{H})$

$L_3 = 7{\cdot}10\text{mH}\,(C_3 = 14{\cdot}28\text{pF})$; $C_4 = 4\,660\text{pF}\,(L_4 = 21{\cdot}8\mu\text{H})$

$L_5 = 6{\cdot}66\text{mH}\,(C_5 = 15{\cdot}24\text{pF})$; $C_6 = 2\,803\text{pF}\,(L_6 = 36{\cdot}2\mu\text{H})$

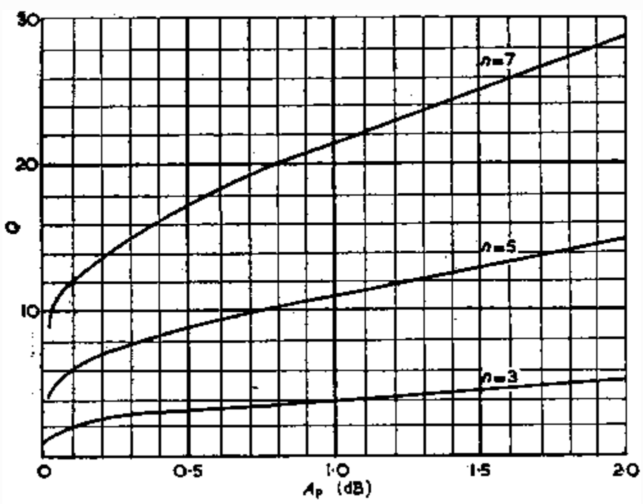


Chart 17. Minimum *Q* for type *C* response

(7) *Some remarks of a practical nature on the above design may be of interest*

In the first place, since the actual *Q* is by no means comparatively infinite, the measured response will not show a flat top of 0·2dB ripple but will droop near the edges of the pass-band. To a certain extent, this could be compensated by modifying the terminations. There will also be a loss at mid-band due to finite element *Q*.

Secondly, band-pass filters offer a great deal of freedom in their realization, and this can be used to modify the element values in some convenient manner for example. In addition, transformer action can be incorporated, and thus one could use equal terminations if so desired.

The coils used in the above design would require to have a fairly low self-capacitance, but ferrite pot cores would probably be satisfactory.

As a rough guide to component tolerances, a figure of ± 1 per cent on both inductors and capacitors is reasonable, and perhaps $\pm 0{\cdot}2$ per cent on the resonant frequency.

Acknowledgment

The author wishes to thank the directors of Barr & Stroud Ltd for permission to publish this article.

Appendix

Stop Band Attenuation, Type B Response

The response of this type of filter is given by the equation:

$$|V_p/V|^2 = 1 + \left\{ |V_p/V_\beta|^2 - 1 \right\} (\omega/\omega_\beta)^{2n}$$

When the second member on the right-hand side greatly exceeds unity, which will be the case in the stop band, one can write:

$$|V_p/V|^2 \simeq \left\{ |V_p/V_\beta|^2 - 1 \right\}^{1/2} (\omega/\omega_\beta)^n$$

whence:

$$20 \log |V_p/V| = 10 \log \left\{ |V_p/V_\beta|^2 - 1 \right\} + 20\, n \log (\omega/\omega_\beta)$$

The first member on the right-hand side is the 'pass-band factor' given in Chart 2. The remaining equation when plotted on log-linear paper reduces to straight lines of slope *n*. The results are plotted on Chart 2.

Stop Band Attenuation, Type C Response

In this case, the response is given by the equation:

$$|V_p/V|^2 = 1 + \left\{ |V_p/V_\beta|^2 - 1 \right\} [T_n(\omega/\omega_\beta)]^2$$

In the stop band, the second member greatly exceeds unity, and the appropriate form of the Chebyshev function $T_n(\omega/\omega_\beta)$ is:

$$\cosh (n \cosh^{-1} \omega/\omega_\beta)$$

Thus, taking the exponential approximation to the cosh function and expressing the results in dB, one finds:

$$20 \log |V_p/V| = \left\{ 8{\cdot}68\, n\, [\cosh^{-1}(\omega/\omega_\beta)] - 6 \right\} + 10 \log \left\{ |V_p/V_\beta|^2 - 1 \right\}, \text{dB}$$

As before, the second term is the 'pass-band factor' given in Chart 2. The remaining term is plotted in Chart 1.

The formulae used in the preparation of the curves giving the element values for type *C* performance are due to Orchard[5]. His notation has been slightly modified. The formulae used in computing the values for type *B* behaviour are based on Atiya's article[1].

Type B Response

$$R_1 = R_2 = 1 \text{ for all } n$$
$$k_r = 2(10^{A_p/10} - 1)^{1/(2n)} \sin (2r - 1)\, \pi/2n$$

Type C Response

$$R_1 = 1 \text{ for all } n$$
$$k_1 = (2a_1/\gamma)$$
$$k_r k_{r-1} = \frac{4a_r a_{r-1}}{b_{r-1}}$$
$$R_2 = 1 \text{ for } n \text{ odd}; \; R_2 = R_1 \coth^2 (nb/2) \text{ for } n \text{ even}$$
$$\text{check: } k_1 = R_2 k_n$$

where:

$$20 \log \coth nb = A_p \quad \text{dB}$$
$$= \sinh b$$
$$a_r = \sin (2r - 1)\, \pi/2n$$
$$b_r = \gamma^2 + \sin^2 (2r\pi/2n)$$

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An Electrical Multiplier Utilizing the Hall Effect in Indium Arsenide

By R. P. Chasmar* and E. Cohen*

The advantages pertaining to the use of semi-conducting indium arsenide as the basis of an electrical multiplier are discussed. A particular merit is the low value of the temperature coefficient of the Hall constant (0·06 per cent/°C) between −40°C and +100°C. The design of Hall plates is considered and certain figures of merit are proposed. Details are given of the construction and performance of a multiplier in which an indium arsenide Hall plate is mounted in the gap in a ferrite core. The harmonics produced in this multiplier at peak output were less than 1 per cent of the fundamental.

A DEVICE capable of the accurate multiplication of two electrical quantities could be of considerable importance to the electronic industry. Frequency changing utilizing the multiplication inherent in a non-linear valve characteristic dates back to very early days, but the more recent advent of the analogue computer has emphasized the need for a cheap and accurate device that can multiply without the production of unwanted frequency components. Various ingenious methods have been devised to meet this problem, but in order to obtain the necessary linearity and frequency response several valves and an associated number of diodes are required[1,2].

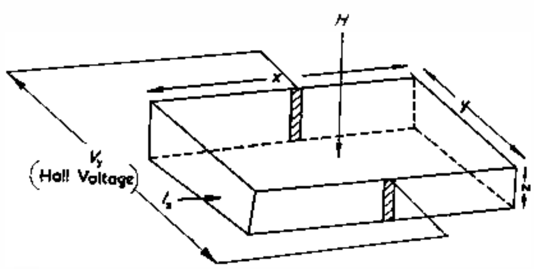


Fig. 1. Illustration of Hall effect

The possibility of using the Hall effect in a semiconductor as the basis of a multiplier has already been considered[3,4,5,6,7,8] since the Hall voltage measures the product of a current through the semiconductor specimen and a magnetic field which may itself be derived from a current through a magnet winding. This method offers several attractive features:—

(1) The multiplier is small, robust and relatively simple to manufacture as it consists primarily of a coil wound on some suitable ferromagnetic core with a semiconductor plate mounted in the gap.

(2) Linearity better than 1 per cent is obtainable and this is comparable with most electronic multipliers.

(3) The frequency response of the Hall effect is virtually unlimited. The multiplier will therefore work from d.c. to such frequencies as are determined by the electric and magnetic circuit arrangements. For very high frequencies an air-cored coil may be used.

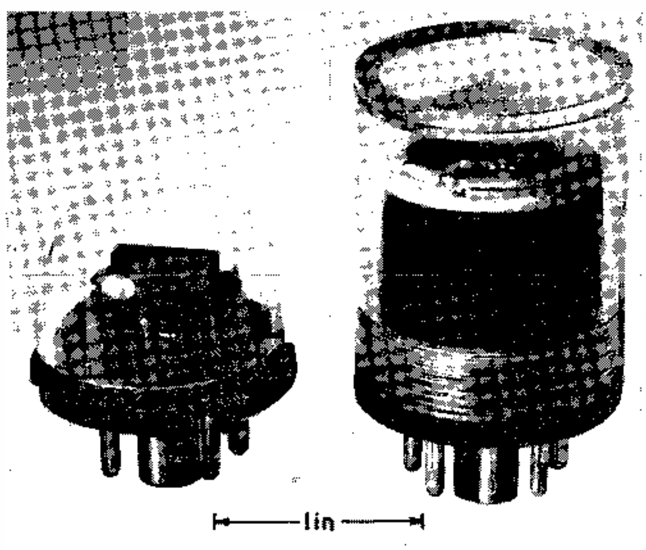
The Hall Effect

Consider a slab of semiconductor (Fig. 1) placed at right-angles to a magnetic field H and carrying a current I_x. Due to the presence of the magnetic field the charge carriers are deflected to the edge of the slab and build up a potential V_y which may be detected between leads connected to the xz faces of the slab.

* *Metropolitan-Vickers Electrical Co. Ltd.*

Equilibrium is reached when this potential rises to such a value that it just annuls the effect of the magnetic field and subsequent carriers are able to pass through the slab undeflected. If the electric fields in the x and y directions are E_x and E_y (in e.m.u.) then the balance of forces on the charge carriers can be expressed by

$$E_y \,.\, e = H \,.\, e \,.\, v.$$



Hall effect multiplier

e = electron charge

v = drift velocity of the electrons in the electric field.

If the electron mobility is μ then

$$v = E_x \,.\, \mu$$

so that

$$E_y = H \,.\, \mu \,.\, E_x$$

and

$$V_y = yE_y = y \,.\, H \,.\, \mu \,.\, (r \,.\, I_x/x)$$

where r is the resistance of the slab in the x direction and can be expressed as

$$r = (1/\sigma) \,.\, x/(y \,.\, z) = 1/ne\mu \,.\, x/yz$$

where σ = specific conductivity of the semi-conductor

n = number of charge carriers per cubic centimetre in the semiconductor

Thus

$$V_y = (H \,.\, I_x)/z \,.\, 1/ne \text{ e.m.u.} \qquad \ldots\ldots\ldots\ (1)$$

V_y is known as the Hall potential, after its discoverer E. H. Hall, (1879) and the factor $(1/ne)$ is known as the Hall coefficient. The Hall coefficient is a constant of the material since it depends on the charge carrier concentration of the semiconductor.

For normal metals, where n is of the order of 10^{23} electrons/cm^3 the Hall coefficient has a very small value and the Hall effect is of little practical significance. Due to recent advances in semiconductor technology, advantage can be taken of the much lower electron concentrations in these materials and an increase of the order of 10^7 in the magnitude of the Hall coefficient obtained.

Hall Effect Multipliers

Open-Cirauit Operation

It is seen from equation (1) that the Hall potential is proportional to the product of the exciting current, the magnetic field and the Hall coefficient. Thus if the Hall plate and the coil producing the magnetic field are driven from constant current sources the Hall potential, measured across a load resistance high compared with the internal resistance of the Hall plate, will accurately represent the product. This accuracy will be maintained over a range of ambient temperatures only if the Hall coefficient of the material is independent of temperature.

The recently developed compound semiconductor, indium arsenide, appears to be particularly favourable since the temperature coefficient of the Hall effect between $-40°C$ and $+100°C$ is only 0·06 per cent/°C for material of Hall coefficient = 100cm^3/C. The exceptionally high electron mobility (25 000cm^2/V sec) and consequent high conductivity of the indium arsenide provide a low internal resistance suitable for operation into a grounded emitter transistor without appreciable departure from the open-circuit condition. Also it has been found that relatively crude polycrystalline material is adequate to make satisfactory multipliers.

Matched Operation

Circumstances may arise in which it is necessary to draw power from the Hall plate. The available power will be a maximum under the usual condition that the load resistance is equal to the internal resistance of the plate between the Hall probes. With this condition and with the plate driven from a constant current source it has been shown[9] that the conversion efficiency is proportional to the square of the carrier mobility in the material. Indium arsenide and indium antimonide have the distinction of possessing the highest electron mobility among known semiconductors. Typical values are given in Table 1.

The efficiency using indium arsenide is thus fifty times greater than with germanium and several hundred times greater than silicon. It should be noted, however, that the above-mentioned relationship between efficiency and mobility only holds for weak magnetic fields where the magneto-resistive effect is small. It is in fact never possible to obtain efficiencies greater than 17 per cent even with the highest mobilities and fields[10].

A disadvantage of the matched type of operation as compared with the open-circuit method is a certain loss of linearity. This is, of course, of prime importance in an electrical multiplier and open-circuit working is therefore to be recommended whenever possible. The non-linearity arises due to the dependence of the resistance of the semiconductor on the magnetic field. The resistance increases with increasing magnetic field and within certain limitations the increase is proportional to the square of the product of the magnetic field and carrier mobility. Thus as the field increases the Hall output departs more and more from matched conditions giving rise to a non-linear characteristic. This problem has been discussed by Kuhrt[4] who claims that by special construction of the Hall plate some compensation can be obtained and a linearity of 1 per cent achieved at an output of 80 per cent of the maximum available power.

TABLE 1

Electron mobility

MATERIAL	ELECTRON MOBILITY AT 20°C (cm^2/V.sec)
Germanium	3 600
Silicon	1 300
Indium antimonide ..	65 000
Indium arsenide	25 000

Hall Plate Design Considerations

Consider the voltage output into an open-circuit of a Hall plate. The response of the plate to changes in exciting current or magnetic field may be represented by a quantity R such that from equation (1) expressed in practical units

$$R = V_H/H \,.\, I = R_H/t \,.\, 10^{-5}$$

where V_H = Hall voltage (mV)

I = plate current (mA)

H = magnetic field (kG)

R_H = Hall coefficient (cm^3/C)

t = plate thickness (cm)

so that

$$\Delta V_H/\Delta I = R \,.\, H$$

and

$$\Delta V_H/\Delta H = R \,.\, I$$

R will be called the 'responsivity' of the plate given in units of mV/(mA . kG).

The maximum output obtainable from a Hall plate is limited by the maximum permissible heating effect produced in the plate by the exciting current. If the power dissipated in the plate by this means is P mW, the resistivity of the semiconductor ρ Ω-cm, the plate width w cm and plate area A mm^2, then using equation (1) it can be shown that

$$(V_H/H)\sqrt{(A/P)} = 3{\cdot}16 \,.\, 10^{-3} \,.\, R_H \,.\, w\sqrt{(2/t\rho)} \quad \ldots. \quad (2)$$

The quantity $V_H/H\sqrt{(A/P)}$ will be called the 'dissipation constant' of the plate and will be represented by the symbol M in units of mV/kG . $\sqrt{(\text{mm}^2/\text{mW})}$. The value of M for a Hall plate corresponds to the Hall voltage output in millivolts obtained with a magnetic field of one kilogauss and a plate dissipation of 1mW/mm^2. This represents approximately the maximum allowable dissipation for a Hall plate mounted in free air.

It will be seen that both R and M will be increased by reducing the thickness of the plate. For devices such as multipliers with ferromagnetic cores a further advantage is obtained by using a thin plate in that the core gap is reduced permitting a greater magnetic field for a given coil current. Experience in this laboratory has shown that a practical limit to the minimum thickness of a Hall plate mounted on a rigid backing is about ·002in.

A further increase in M can be obtained by increasing the width of the plate (equation (2)). At the same time the length must also be increased in proportion to eliminate the undesirable effects due to the partial short-circuiting of the Hall voltage by the current electrodes[4]. The increased area of the Hall plate permits a greater power to be dissipated and a consequent increase in maximum output.

Construction of Multiplier

Hall effect multipliers can be constructed using air-cored coils where operation at high frequencies is desired. The output with such coils is small owing to the low value of the flux density obtained with normal values of coil current. For most analogue computer applications a frequency

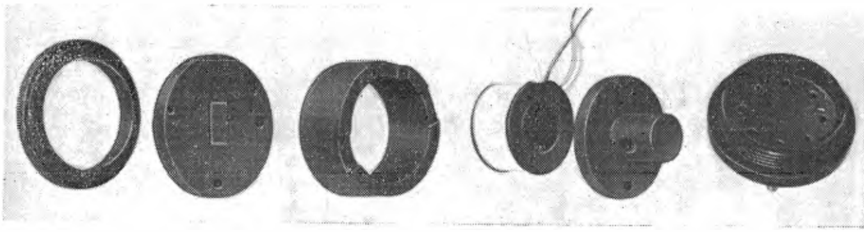


Fig. 2. Exploded view of multiplier

response ranging from d.c. to 100c/s is quite adequate and the use of a gapped ferromagnetic core to increase the flux is permissible.

The magnetic ferrite commonly known as Ferroxcube A.1 has some attractive features. The pot core assembly permits of a simple form of construction while the high resistivity (100Ω-cm) enables the Hall plate to be mounted directly on the magnetic material without the need for insulation which would increase the gap and reduce the flux. Also the initial permeability is greater and hysteresis losses less than silicon steel. On the other hand a disadvantage is the low saturation flux density ($B_{max} \simeq 3{\cdot}4$kG at

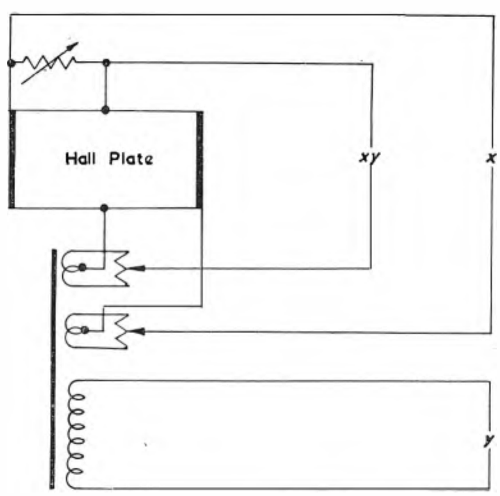


Fig. 3. Circuit of Hall effect multiplier

20°C). A multiplier utilizing a Ferroxcube pot core assembly is shown in exploded form in Fig. 2 and assembled multipliers are shown on page 661. The characteristics of a typical multiplier are given in Table 2.

The circuit diagram of the multiplier is given in Fig. 3.

With a.c. operation inductive coupling from the coil to the plate circuit and Hall circuit is inevitable owing to the finite size of the plate. This can be balanced out by the method suggested by Keister[6] in which additional small coils, shunted by very low resistance potentiometers are connected in the appropriate circuits. These are built into the multiplier and can be preset on test.

Exact positioning of the Hall probes on the specimen so that they lie on an equipotential in the absence of a magnetic field is extremely difficult. In order to eliminate any 'standing voltage' due to this cause a variable resistance is connected between one current connexion and one Hall connexion. By suitable adjustment of the resistance, the 'standing voltage' can be reduced to zero.

TABLE 2

Characteristics of typical indium arsenide multiplier

Electrical conductivity	$250\Omega\text{-cm}^{-1}$
Hall coefficient	$100\text{cm}^3/\text{C}$
Plate dimensions	0·8cm × 0·4cm × 0·005cm
Resistance—current connexions	5·4Ω
Resistance—Hall connexions ..	3·9Ω
Magnetic core gap	0·1mm
Coil winding data	3 000 turns 44 s.w.g.
Coil inductance	4·7H
Coil resistance	450Ω
Coil resonant frequency ..	6kc/s

D.C. characteristics of this multiplier are given in Figs. 4 and 5. Fig. 4 shows the Hall output plotted against coil current for a plate current of 50mA; this relation is linear up to a coil current of 13mA. The departure from linearity at high coil currents is due to saturation of the Ferroxcube. Fig. 5 gives the dependence of Hall output on plate current at various values of field current; again the relationship is linear.

The field in the gap can be calculated as follows:—

$$\text{Flux density} = B = 4\pi/10 \ . \ \frac{N\mu}{(L + (\mu - 1)g)}$$

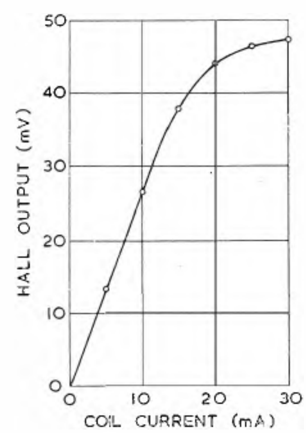


Fig. 4. Variation of Hall output with field current for indium arsenide multiplier

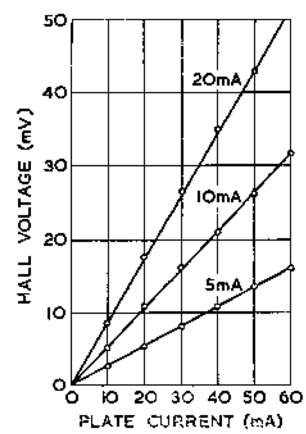


Fig. 5. Variation of multiplier output with plate current for three values of coil current

N = ampere-turns of field coil

L = length of iron circuit (cm)

g = gap (cm)

μ = permeability.

Taking $L \simeq 5$cm and $\mu \simeq 750$ then for a coil current of 13mA giving 33·5mV.

$B = 2{\cdot}9$kG

$$\text{Responsivity} = R = V_H/BI = \frac{33{\cdot}5}{2{\cdot}9 \times 50} = 0{\cdot}23$$

This is in reasonable agreement with the value of R calculated from the Hall constant and estimated plate thickness.

From the Hall output and measured resistances of the plate the dissipation constant for the multiplier is

$$M = 20(\text{mV/kG}) \sqrt{(\text{mm}^2/\text{mW})}.$$

Such a Hall plate mounted in contact with the pole piece could operate at a dissipation of 4(mW/mm²) which would correspond to a maximum Hall output of 150mV for a plate current of 220mA and coil current of 13mA.

The chief source of harmonic components in the Hall voltage with a.c. circulating in the core is the non-linear relationship between magnetizing field and magnetic induction. For a closed ring of Ferroxcube the shape of the *BH* loop is approximately rectangular, but if a gap is included in the ring the *BH* loop is sheared and ultimately degenerates into an approximately straight line for sufficiently large gaps. That this is so under d.c. excitation is seen in Fig. 4.

Measurements have been made of the harmonic content in the output of a Hall multiplier with d.c. supplied to the plate and a.c. to the coil at fundamental frequencies 80c/s and 300c/s. The harmonic content of the Hall output and of the source oscillator were measured using a Muirhead-Pamatrada wave analyser and the results compared. Unfortunately with the best oscillator available the harmonics present in the source signal were comparable with those introduced by the multiplier and so an accurate estimate of the distortion produced by the multiplier was not possible. The results at 80c/s indicated that where the peak value of the coil current did not exceed the limits of the straight portion of the d.c. characteristic (Fig. 4), the ratio of fundamental to second harmonic was about 40dB. At low values of coil current this rose to 60dB. An improvement of several decibels over these figures was obtained with a 300c/s fundamental.

Acknowledgments

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Multiplication by Semiconductors

By C. Hilsum*, B.Sc.,A.Inst.P.

Analogue multipliers can be made using either the Hall effect or magneto-resistance in semiconductors. The possible modes of operation are considered, and the choice of materials discussed. Details are given of several experimental multipliers and the results obtained with them.

THE operation of multiplication in an analogue computer has always been one of the more difficult stages, and there has not so far been available a device which could do this in a simple and reliable fashion. Nearly always complicated circuits are involved, and an accuracy of 1 per cent is seldom achieved. A promising relation to use as the basis for a multiplier is the force acting on a charged particle moving across a magnetic field, this force being proportional to the product of magnetic field strength and the particle velocity. Such a multiplier has in fact been constructed in a special cathode-ray tube[1]. A simpler solution is to use the corresponding effects occurring in a solid viz. the Hall effect, and magneto-resistance.

Hall discovered in 1879 that if a current was passed through a metal at right-angles to a magnetic field a voltage appeared in the mutually perpendicular direction. A simple arrangement for observing the effect is shown in Fig. 1. The Hall voltage is given by

$$V_H = R_H \frac{BI}{t.10^8} \text{ volts}$$

where I is in amperes, B in oersteds and t in centimetres; it may be shown that R_H, the Hall constant, is the reciprocal of the product of the charge on the carriers and their concentration. Since the carrier concentration is much less for semiconductors than for metals, the voltages developed are larger. If the charge carriers have a high

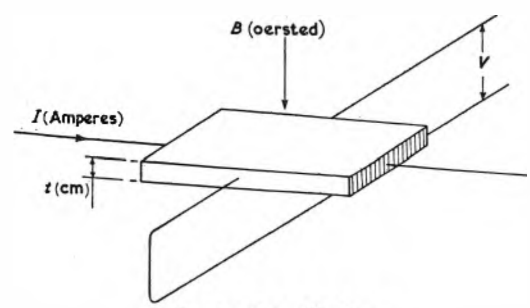

Fig. 1. The Hall effect

mobility a good proportion of the input power to the specimen may be diverted to the Hall terminals.

When the Hall effect is to be used in a multiplier the magnetic field is obtained by feeding one of the input voltages into a magnetizing coil; three questions then arise.

(1) Which semiconductor is the best?

(2) Is the magnetic field proportional to the voltage developing it?

(3) Should the device operate as a voltage or a current multiplier?

The third question provides in some respects an answer to the first. If the multiplier is required to feed into a matched load the choice of materials largely depends on considerations of available power, and this restricts one to the high mobility semiconductors, since the efficiency

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of power transfer is proportional to the square of the mobility, and independent of the Hall constant[2]. If a voltage output into a high resistance is required the mobility is less important, and the numerical value of the Hall constant comes into the relations. It is necessary also to decide whether the multiplier is to use voltage or current input into the semiconductor, remembering that the resistivity is dependent on both the magnetic field and the ambient temperature. The magneto-resistance, the increase of resistance in a magnetic field, is of much smaller magnitude with the lower mobility semiconductors.

Four semiconductors offer themselves for this application, though further development of materials may provide others suitable for the purpose. The four considered may be divided into two classes, silicon and germanium being described as of medium mobility and high Hall constant, and indium antimonide and indium arsenide of high mobility and medium Hall constant. If the temperature stability required is specified then Ge and Si will give the best voltage sensitivity, and InAs the best power transfer.

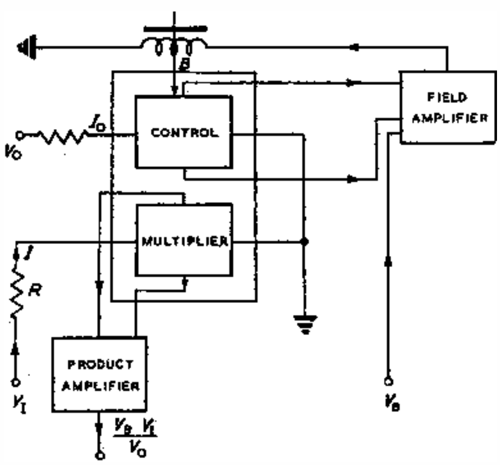

Fig. 2. Block diagram of a multiplier (after Lofgren)

InSb gives the best power transfer of all, but shows pronounced temperature effects.

The second question posed is answered more directly. With a ferromagnetic core with a fairly large air-gap there is a working range over which the field in the gap is proportional to the magnetizing current, but at fields above about 5 000 oersted non-linear effects will be observed. Since the Hall voltage may be made proportional to the magnetic field up to at least 10 000 oersted some workers have thought it worth using special compensation techniques to ensure field proportionality over this range.

The current research on Hall effect multipliers may be illustrated by three examples. The most advanced multiplier has been made by Lofgren[3]. He intended the device to work into a valve amplifier and so chose the regime, current input-voltage output. His initial work was on germanium but recently he has published a description of a silicon multiplier which he considers as superior. It is shown in Fig. 2. One feature worth noting is the method for ensuring magnetic field proportionality. A control specimen is identical to the multiplier specimen. It is driven from a constant current source, and the Hall voltage output applied in opposition to the voltage V_B at the input of a high gain amplifier which provides the magnetizing current. Lofgren shows that the magnetic field produced is almost exactly proportional to V_B.

The regime is made voltage input by putting a large resistance in series with the multiplier specimen, or by driving it from an amplifier designed to give a current output proportional to the voltage input.

Typical operating conditions are:—

Crystal current = 5mA.
Field = 5 000 oersted
Hall voltage = 0·5V
Amplifier inputs = ± 100V maximum
Product amplifier output = ± 20V maximum
Linearity = 0·1 per cent.

The frequency response is limited by the field part of the circuit since the crystal current may be modulated at several megacycles per second. The response time of the field is proportional to the voltage input and with a Ferroxcube core the device responds to maximum input voltage in 1msec. The maximum field is then restricted to 3 000 oersted by saturation in the ferrite. With an iron core the maximum field can be 10 000 oersted but the response is 10 to 20 times slower.

A simpler multiplier using InAs has been developed at the Metropolitan-Vickers Research Laboratory by Chasmar and Cohen and a complete account of their work is given

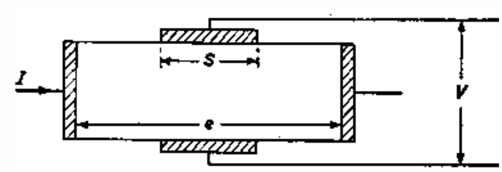

Fig. 3. Hall specimen with widened electrodes

elsewhere in this issue[4]. The unit uses current input and voltage output and the magnetic field is provided via a Ferroxcube core. The maximum field is therefore 3 000 oersted which is obtained with 15mA through the magnetizing coils. No proportionality correction is used but the linearity is better than 1 per cent. Typical operating conditions are

Crystal current = 200mA
Field = 2 500 oersted
Hall voltage = 0·12V.

The indium arsenide specimen has a resistance of 5·6Ω and the coil a resistance of 500Ω and an inductance of 5H. The resistance between the Hall probes is 4·1Ω.

Another indium arsenide multiplier has been described by Kuhrt[5], of the Siemens Laboratories. He was interested in obtaining power transfer and gives considerable detail of the techniques necessary for obtaining linearity in current input-current output regime. It was found that the magneto-resistive effects were large enough to cause errors when feeding into a matched load, the specimen resistance increasing with the magnetic field. A square shaped specimen of InAs more than doubles its resistance in a field of 10 000 oersted and this displays itself as a departure from linearity in the multiplier output at the higher fields. The effect can be compensated by using Hall electrodes of a finite width as shown in Fig. 3. The experimental results are illustrated in Fig. 4, where it is apparent that by using electrodes of one tenth the specimen length and working into a load resistance three times the zero-field specimen resistance a linearity error of <1 per cent can be obtained. An alternative method is shown in Fig. 5. An InAs control specimen is placed in the field air-gap and connected in parallel with the Hall specimen. The control is shaped to have a greater magneto-resistive effect than the Hall unit, and as a result a greater proportion of the driving current passes into the Hall unit at the higher fields. The resistances shown are adjusted so that this extra current is of correct magnitude to counterbalance the non-linearity.

The magneto-resistive effects have been mentioned earlier as 'nuisance' effects, tending to make the Hall effect multipliers inaccurate. Simpson[6] has shown that an accurate multiplier may be made by use of these same effects. His device is shown in Fig. 6. Two identical specimens of indium antimonide or indium arsenide are arranged in air-gaps in a magnetic core so that a biasing field B and a mutiplier field W can be applied to them. For one specimen the multiplier field supplements the bias, and for the other it is subtracted from it. The change of resistance ΔR is proportional to the square of the magnetic field. The two specimens are connected in a Wheatstone bridge in one of two arrangements (Fig. 7(a) and (b)). The resistances of the other two arms and of the load are chosen to give the most linear operation. The first arrangement gives excellent linearity for voltage or current output but the input impedance is slightly dependent on the magnetic field. The alternative system is not quite as linear, but the input impedance is more constant. Little experimental work has been done on this device but it provides an interesting alternative to the Hall effect multipliers.

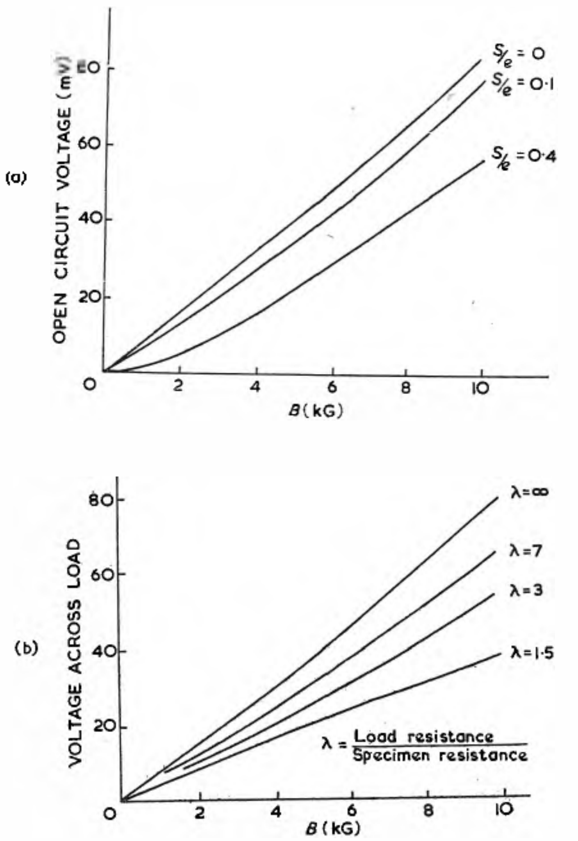

Fig. 4. Obtaining linearity by use of wide Hall electrodes (after Kuhrt)
(a) Open-circuit characteristics
(b) Effect of varying load resistance on linearity S/e = 0·1

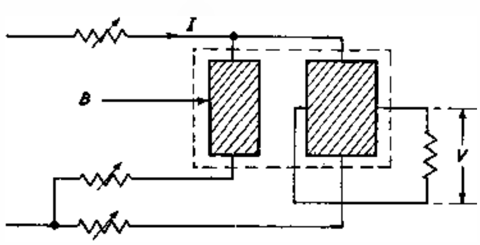

Fig. 5. Correcting for non-linearity with a magnetoresistive control (after Kuhrt)

In this short account no mention has been made of the difficulties involved in the construction of the multiplier

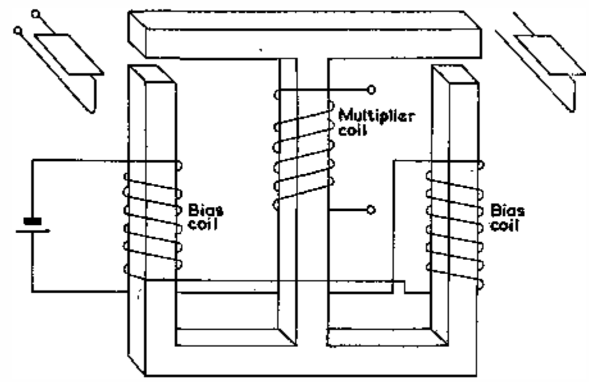

Fig. 6. The magneto-resistive multiplier

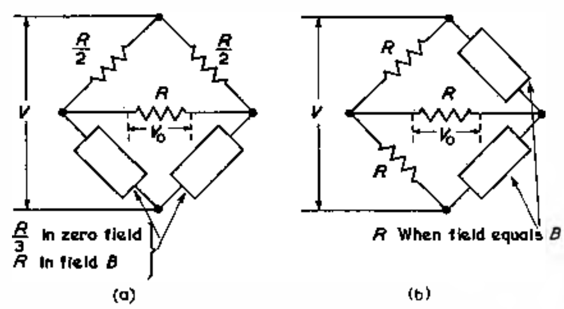

Fig. 7. Connexion of magneto-resistive multiplier

(a) For B = 3kG, an input field C causes an output

$$V_0 = \frac{V(C/B)}{2/1+C^4/4B^4/}$$

An input of 1kG gives a non-linearity of 0·3 per cent and one of 500G, a non-linearity of 002 per cent. (b) A voltage $V_{..}$ applied to the multiplier coil gives a field C, less than the field B

$$V_0 = \frac{V(C/B)}{3+7/6\ C^2/B^2}$$

B = 3kG, C = ±500G *for 1 per cent non-linearity*

units. The contacts must be made non-rectifying, and for Hall specimens there must be no residual output voltage in zero magnetic field. For devices using two elements there are the obvious problems of matching. All of these points are being investigated and there is little doubt that the problems will be solved. The designer of analogue computers will then have available a multiplier which will ease his task considerably.

Acknowledgment

The author is grateful to the Admiralty for permission to publish this article.

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Short News Items

The College of Aeronautics, Cranfield, held an Open Day recently. The aircraft industry, Ministries and Services, airlines, universities, colleges, press and others were represented. The formation of the College was first officially proposed by Sir Stafford Cripps, then Minister of Aircraft Production. The College was established in January 1946 in the former Royal Air Force station at Cranfield. The purpose of the College is to provide engineering and scientific training in aeronautics at post-graduate level, to enable students to develop a capacity to fill in due time senior and responsible positions in the aircraft industry, civil aviation, the Services, education and research.

The Board of Directors of Electric & Musical Industries Ltd has decided to close the E.M.I. College of Electronics with effect from July 1959. Over the past few years E.M.I. Electronics Ltd has found an increasing necessity to meet the training requirements of its technical staff, and steps have been taken to increase the training facilities within its own organization at Hayes and its various branches. It is also taking greater advantage of the increasing variety of courses now being offered by universities and technical colleges. Intending students on one-year courses in the Principles and Practice of Radio and Television may, however, still apply for enrolment with the E.M.I. College for courses commencing on 6 January and 14 April 1959. The E.M.I. College will be responsible for the instruction of students until 29 July 1959. After that date, they may, if they so wish, complete their courses at the Pembridge College of Electronics, 34a Hereford Road, London, W.2, which Mr. J. B. McMillan, M.A., B.Sc., the present Director of Studies of the E.M.I. College of Electronics, is founding and in which he will be assisted by members of his present staff. The Pembridge College which will open in September 1959, after Mr. McMillan and his staff have relinquished their present positions with the E.M.I. College will conduct, as one of its activities, courses based on the one-year course syllabus at present used by the E.M.I. College of Electronics.

The BBC has placed an order with E.M.I. Electronics Ltd for six f.m. transmitters of a new type for the v.h.f. sound station now being built at Llanddona, North Wales. The station, which will provide a full service of Home, Light and Third programmes, will serve some 180 000 people living in Anglesey and North Wales. The BBC has also ordered six similar transmitters for the new station to be built at NetherButton, Orkney. The new method of f.m. used provides an improved signal-to-noise ratio with reduced harmonic distortion.

The Institution of Telecommunication Engineers (India) is to hold a Second Technical Convention on 27-28 December. Further particulars may be obtained from the Honorary Secretary, The Institution of Telecommunication Engineers, Post Box No. 481, New Delhi.

The Council of the Institution of Electrical Engineers, with the encouragement of members of the medical profession, have instituted a Medical Electronics Discussion Group, to provide a forum in which electrical engineers specializing in electronics can meet members of the medical profession to discuss problems in which they have a common interest. The Group will hold regular meetings having the nature of colloquia at which subjects coming within the field of medical electronics will be discussed. The affairs of the Group will be guided by a committee broadly representative of members of both professions who are prominent in this field. The Council have noted with interest the outcome of the International Conference on Medical Electronics which, on the initiative of Dr. V. K. Zworykin, was held in Paris in June of this year, and feel that the second International Conference which it is planned to hold in 1959, is worthy of support. They have therefore agreed that the Group Committee should receive on behalf of Dr. C. N. Smyth, a Vice-President of the Interim Committee of the International Conference and Chairman of its Papers Committee, notice of any papers coming from British sources. Any person interested in the activities of the Group should send his name and address to the Secretary of the Institution of Electrical Engineers, Savoy Place, London, W.C.2, for inclusion in a mailing list for details of the Group's activities, which is now being compiled.

A Short General Purpose Analogue Computer is to be installed in the Department of Electrical Engineering at McGill University, Montreal. It will be used for research on electrical circuit design and control system analysis. The instrument is being shipped to Canada as a joint venture by Short Brothers and Harland Ltd and The Bristol Aeroplane Co of Canada Ltd. It is hoped that as well as being used for research at McGill, it will be available for use by outside organizations and for demonstration to some of the smaller industrial groups in the provinces of Quebec and Ontario. The computer will be equipped with the latest continuously drift-corrected units, designed for high-speed repetitive computation. Though basically similar to Short computers already in service in the U.K. and overseas, it has been specially adapted for operation on the higher frequency supply system used in North America.

The European Division of Electronic Associates, Inc, at 43 rue de la Science, Brussels, has announced the sale of a large general purpose Type 231R PACE analogue computer to the United Kingdom Atomic Energy Research Centre at Harwell. This system contains high accuracy linear and non-linear computing components, modular patch panel and ADIOS system which allows the computer to be programmed by paper tape or electric typewriter. The company maintains sales, service and training facilities at Brussels and, in addition, provides rental service and consulting engineering advice on a PACE computer installed there.

The BEAMA Publicity Conference 1958 will be held on 11 and 12 November. Attendance is open to BEAMA member firms only.

The Institution of Electrical Engineers mention that the Faraday Lecture 1958/59 will be presented by Dr. H. A. Thomas at the Royal Festival Hall on Monday, 26 January, 1959, at 6 p.m., on the subject of "Automation". Dr. Thomas is Manager of the Instrumentation and Control Section of Unilever's Engineering Department. Admission will be free by ticket available from the Institution, Savoy Place, London, W.C.2. Applicants are requested to enclose a stamped addressed envelope.

Hayward & Martin Ltd of Bromley, Kent, offer a complete service for engineering and electronic concerns, covering the preparation of instruction material, with particular reference to handbooks, sectioned and perspective illustrations, in line, half-tone and full colour. They also deal with the production of sales promotional material and technical publications from the commercial standpoint. They are officially accredited contractors to H.M. Government, including the U.K. Atomic Energy Authority and the Ministry of Supply. Further details may be obtained from 34, High Street, Bromley, Kent.

Lewis Berger (Great Britain) Ltd mention that they frequently receive inquiries regarding the supply and use of strain sensitive lacquers which probably arise from the fact that they were cited as suppliers some years ago in the monograph *Resistance Strain Gauges* by J. Yarnell, published by us. (Morgan Bros (Publishers) Ltd). The position is that, while they may, some while ago, have supplied an experimental sample for such use, they do not regularly manufacture, and have no technical background of experience in the use of these materials which would enable them to advise.

LETTERS TO THE EDITOR

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

Error-Integration in Servomechanisms

DEAR SIR,—The benefits of error-integration in servomechanisms are well known. However, its use can lead to trouble in handling large disturbances when torque-saturation occurs. I have recently been working on a servo including error integration, which went

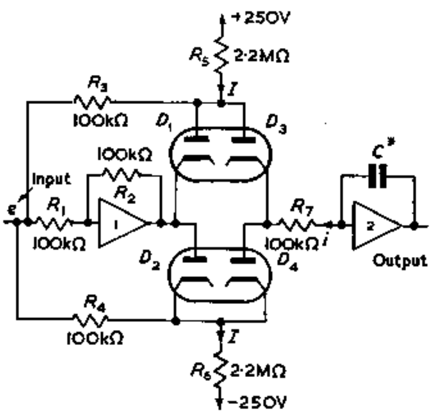


Fig. 1. Circuit diagram (*replaced by 100kΩ resistor for obtaining Fig. 2)

viciously unstable after application of a large step demand. To overcome this the integrator was disabled for error signals exceeding a fairly small value. This idea is not new, but the circuit employed (Fig. 1) may be of interest, and has the advantage that very good clamping of the integrator may be obtained without the necessity for carefully matched components. The experimental characteristic shown in Fig. 2 was obtained using a 100kΩ resistor instead of *C*.

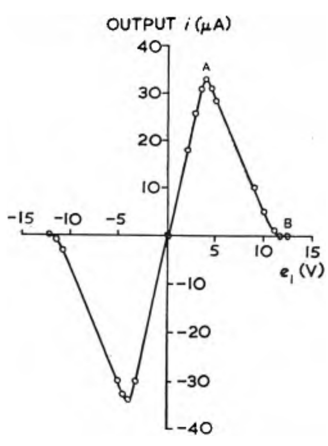


Fig. 2. Input-output characteristic

The d.c. amplifiers used in the experimental circuit had a voltage gain of about 500 times, and cathode-follower outputs, although these could doubtless be simplified if required.

Explanation may be simplified by assuming the resistors R_5, R_6 with the h.t. lines to be sources of constant current *I*. For small signals the current *I* divides between all the diodes, which therefore couple the inverting amplifier 1 to the integrator. The shunting effect of R_3 and R_4 is minimized by the low output impedance of amplifier 1.

As the input signal *e* increases in the positive direction, the currents in D_2 and D_3 diminish, becoming zero at the point marked A on the characteristic of Fig. 2. The current *i* in R_7 has meanwhile been increasing in the sense indicated.

A further increase in *e* decreases *i* (the invertor amplifier being now disconnected from R_7), and eventually when *i* reaches the value *I*, D_4 ceases to conduct. For any further increase in *C*, R_7 is effectively disconnected at the diode end. It may be noted that the integrator is thus effectively clamped even if there is a zero error in amplifier 2. A further test with a 1μF capacitor for *C* indicated a current in it of about 10^{-9}A (more or less independent of the zero settings) for inputs outside ±12V.

To date two specimens of this circuit have been constructed with no attempt to select or match components, and have given the above performance without any difficulty.

Yours faithfully,

R. S. TAUNTON

The Staveley Research Department,
Bedford.

Some Methods of Phase Measurement used in Transfer Function Analysis

DEAR SIR,—The article by D. J. Collins and J. E. Smith, which appeared in your April 1958 issue and in which the authors surveyed among others some of the techniques of phase-difference angle estimation using a single-beam cathode-ray oscilloscope, particularly interested me.

Since measurements of this sort are limited in accuracy by such considerations as deflexion linearity, spot size, and harmonics, the appeal of such methods must lie in the ease with which they can be employed. In this light, I suggest that two additions to the authors' section (1(b)) are needed.

First, in the right-hand drawing of Fig. 2 of Collins and Smith, not only is

$$\text{Phase shift } \phi = 2 \tan^{-1} B/A$$

but the angle between rays joining the ends of the minor diameter and one end of the major diameter is indeed the phase shift angle ϕ, which may be measured directly with a protractor.

Second, since most oscilloscopes are equipped with a graticule, a method I have described[1,2] is most attractive. Briefly, if the Lissajous ellipse is inscribed in a square 24 divisions on a side, direct approximation, within one degree due to method, of angles between zero and 70° is possible. Further, a correction which may be applied by mental computation extends the usefulness of the method to include all angles, the error due to method lying between zero and plus one degree or within plus or minus a half degree.

Very truly yours,

CARL R. WISCHMEYER

The Rice Institute,
Houston, Texas.

REFERENCES

1. WISCHMEYER, C. R., PFEIFFER, P. E. Direct Reading of Phase Angles on an Oscilloscope. *Rev. Sci. Instrum.* 25, 41 (1954).
2. WISCHMEYER, C. R. *Rev. Sci. Instrum.* 29, 535 (1958).

The Authors' reply:

DEAR SIR,—We are indebted to Mr. Wischmeyer for his constructive note. He appears to have evolved a measurement system which is relatively simple but requires an oscilloscope with special features.

In general, the oscilloscope should have exceptionally good deflexion linearity, small spot size and have identical *X* and *Y* systems with continuously variable gain controls.

However, even with this special oscilloscope it is very debatable as to whether a 1 degree accuracy can be achieved. If, for example, we have a 5in cathode-ray tube with a value for *R* of 2in, the scale factor is 0·022in per degree, which necessitates careful measurements.

The use of the oscilloscope for phase measurement is best suited to the higher frequencies where other techniques fail and we would emphasize that large errors can occur if the measured voltages have harmonic content[1].

Yours faithfully,

D. J. COLLINS
J. E. SMITH

Solartron Research & Development Ltd.

REFERENCE

1. ROTHSCHILD, RAYMOND. Precision Phase Measurements. Publication of Industrial Test Equipment Co., 55, East 11th Street, New York.

Gating of Threshold Signals

DEAR SIR,—The output when taken either from the cathode or the anode of a conventional gating circuit, is superimposed on a pedestal, due to the gating

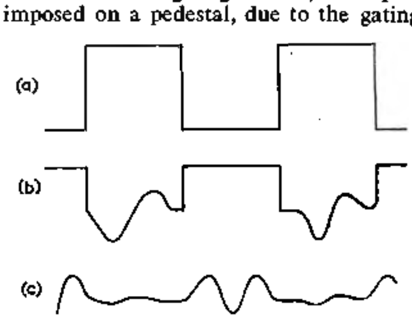


Fig. 1. Comparison between the output waveforms of the threshold and conventional gates
(a) Gating waveform, (b) Cathode output from conventional gate, (c) Cathode output from circuit of Fig. 2

waveform as shown in Fig. 1. When the signal to be gated is very small, this pedestal may be intolerable.

A simple re-arrangement of the circuit enables a gated output to be obtained, substantially free from the pedestal, and is of considerable use in applications involving threshold signals.

In the circuit described below gating is obtained by gain discrimination between the open and closed condition. In a typical circuit, discrimination of 10 to 20 to 1 can be obtained and this is usually sufficient for most applications.

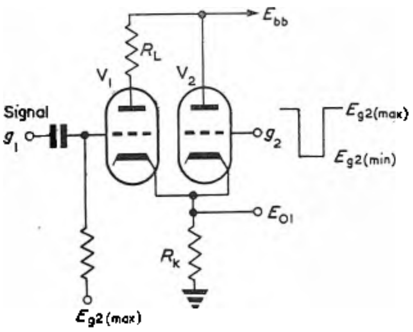


Fig. 2. Threshold signal gate

The circuit and voltage levels are shown in Fig. 2 and in Fig. 1 its output waveform is compared with that of a conventional gate.

In the conventional double triode gate the signal grid bias is arranged to lie somewhere between the variations in grid potential of the switching waveform. Thus, when the gate is open V_1 is conducting and V_2 is cut off and when the gate is closed the opposite occurs.

However, from Fig. 2 it will be observed that the bias on the signal grid is made equal to the maximum switching potential on the second grid. The output is taken from the cathode.

When the switching potential is high both valves are conducting and the cathode potential is slightly higher than either grid potential. The circuit operates

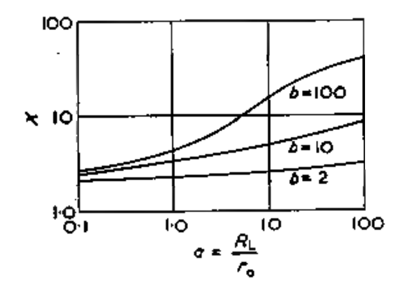


Fig. 3. Variation of gain discrimination ratio as a function of a and b

as a conventional difference amplifier, a very small signal appearing from cathode to ground.

On the other hand, when the switching potential is low, the cathode potential remains just above the potential of grid 1. There is a considerable increase in the anode current of the first valve but little change in the current in the cathode resistor. The second valve is cut off and the first acts like a cathode-follower.

Let E_{01} = the output voltage when the gating voltage is high (gate closed)

and E_{02} = the output voltage when the gating voltage is low (gate open)

The voltage discrimination ratio

$$X = (E_{02}/E_{01})$$

and is given approximately

$$\text{by } X = 1 + \frac{b(1+a)}{1+a+b}$$

where $a = (R_L/r_a)$ and $b = g_m R_K$

The above relationship is graphed in Fig. 3 and it can be seen that for R_L less than $10r_a$ there is little point in making R_K exceed $10/g_m$.

Practical values of X are about 10 times and in order to exceed this it is necessary to have very high values of R_L and R_K.

Similarly, an output impedance discrimination ratio may be defined as $Y = (Z_{02}/Z_{01})$ and for this gate

$$Y = \frac{1}{1 + \dfrac{b(1+a)}{1+a+b}}$$

When the outputs of n gates are to be paralleled it can be proved that $XY/(n-1)$ should be $\gg 1$.

Unfortunately as $XY = 1$ for this threshold gate, before such gates can be joined in parallel a buffer valve output is necessary for each one.

Yours faithfully,

G. W. Taylor
South Woodford,
London, E.18.

On Measuring the Admittances of a Four-Terminal Network

Dear Sir,—It is often desirable to consider an electrical or electronic device as a four-terminal network, and to measure some set of four-terminal properties. A convenient set is the admittance parameters, which are matrix elements given by

$$\begin{bmatrix} i_1 \\ i_2 \end{bmatrix} = \begin{bmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{bmatrix} \begin{bmatrix} v_1 \\ v_2 \end{bmatrix} \qquad (1)$$

Here i_1 and i_2 are the currents into the four-terminal at each end, V_1 and V_2 the voltages applied at each end, and the y's the admittances.

One way of measuring all these values is to measure the input admittance at each end of the network, the other end being first open-circuit and then short-circuit. If the network is known to be reciprocal, so that $y_{21} = y_{12}$, one of these four measurements is redundant; while if the network is reciprocal and symmetrical, two non-equivalent measurements will suffice.

Active devices, such as thermionic valves and transistors, will, in general, be neither reciprocal nor symmetrical. Thus all four measurements would be required. However, there may be difficulty in achieving some of the open- and short-circuit terminations. It is difficult, for example, to get an open-circuit (impedance much larger than the internal impedance) for the grid of a thermionic valve; the collector of a transistor may present the same problem.

Some of these difficulties may be overcome by a measurement which seems to have been little noticed, namely that of the admittance, as a two-terminal, of the four terminal with input and output connected in parallel. It is apparent that the admittance measured would be just the sum of the four admittances from the matrix in equation (1). Then three other measurements would determine matters completely; and, in some cases, it might be possible to dispense with some of them.

A valve at moderate frequencies, has $y_{21} = g_m$ and $y_{22} = 1/r_a$, g_m being the transconductance and r_a the anode resistance. The other y's are much less than these; and, moreover, they will be predominantly imaginary, while g_m and $1/r_a$ are real. Accordingly, a measurement of input admittance to the anode with the grid shorted to the cathode, and to the anode and grid tied together, will determine these quantities.

For a transistor in the common-emitter configuration, all four y's will often be wanted. A measurement of the input admittance to the base with the collector shorted, to the collector with the base first open and then shorted, and to the base and collector tied together, will give the desired information. The input admittance to the base is ordinarily of such a magnitude that it is not difficult to achieve either an open- or short-circuit termination, whereas it might be difficult or impossible to achieve a true open-circuit termination at the collector.

The valve and transistor are really three terminal networks. For a true four-terminal network there is another possible measurement, in which the input and output are connected in parallel, but one is reversed. This measurement would give $y_{11} - y_{12} - y_{21} + y_{22}$. It might be possible, in some cases, to get the same result with a three-terminal network by using transformers.

It should not be overlooked, of course, that this process might be reversed, measurements of transfer admittances and impedances being used to deduce input and output admittances or impedances. This might be worthwhile, since equipment for measuring transfer admittances and impedances is available.

In conclusion, it is believed that the measurement of the input admittance to the two ends of a four-terminal network connected together can be a handy method of avoiding a more difficult measurement. This technique has surely been noticed before, but the writer does not recall ever seeing any mention of it, and a reminder may thus be helpful.

This discussion was prepared while the writer was with Pacific Semiconductors, Inc., Culver City, California, U.S.A.

Yours faithfully,

H. L. Armstrong
Department of Physics,
Queen's University,
Kingston, Ontario.

BOOK REVIEWS

Switching Circuits and Logical Design

By S. H. Caldwell. 686 pp. 300 figs. Demy 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd, London. 1958. Price 112s.

THIS book is a major contribution to the rather limited range of basic text-books on this subject, and some useful methods of analysis and synthesis presented in it have only previously been published in rather inaccessible papers.

The presentation of the subject is very clear, and tables and diagrams are carefully explained. The engineer dealing with many day to day problems may feel that the methods developed in the earlier chapters are rather like taking a sledge-hammer to crack a nut, but it should be pointed out that the book is based on lectures given to students at M.I.T.

There are two introductory chapters, the first indicating the applications and objectives of switching circuits, and the second describing some of the commonly used switching components. Switching algebra is then introduced and references made to the work of Shannon and to the algebra of logic developed by Boole. After dealing with the postulates and basic theorems as an abstract structure, they are repeated using relay contacts to give a physical significance to the algebra. After a chapter which extends the theory and includes analysis and synthesis of contact networks, we come to the problem of minimization. This is a difficult problem if a reasonable number of variables is involved, and one of the methods described seems to represent a considerable advance. It is a modification by McCluskey of the Quine method, and results in a considerable saving of labour and simplification in handling the comparison stages of the process. The various methods yield an expression in the form of a minimum sum (or product). This may or may not be a unique minimum and in any case does not necessarily represent an optimum engineering solution. The designer must still rely on art and experience at this stage and a completely analytical solution seems unlikely, particularly as an optimum design is often difficult to define. There are five further chapters on combinational networks, one of them extending the theory to cover most of the high speed switching elements. Rectifiers, valves, transistors and magnetic cores are discussed and some of the engineering limitations pointed out.

The remainder of the book deals with sequential circuits and here again the basic methods seem to be an advance on those employed by many engineers. Electronic and solid-state devices are again discussed both in sequential and pulsed sequential circuits.

The use of the terms 'add gate' and 'multiply gate' in place of 'or gate' and 'and gate' is of little consequence; but the introduction of special symbols for gates using grids one and three of a pentode, and double triodes with commoned anodes, is in the reviewer's opinion confusing and quite unjustified in a logical diagram.

The student with no previous knowledge of the subject will have no great difficulty with this book and a useful range of problems is included at the end of each chapter; it is however a very long book and a thorough study of it is a major undertaking. It will certainly lead most engineers to think more clearly about the design methods they employ, but in many cases (certainly in high speed switching) the detailed formulation of the problem often goes hand in hand with trial solutions, and text-books can never replace clear thinking and experience for this type of design work.

F. BECKETT

Progress in Semi-conductors—Volume 3

Edited by A. F. Gibson, P. Aigrain and R. E. Burgess. 210 pp. 44 figs. Demy 8vo. Heywood & Co. Ltd. 1958. Price 55s.

AS in the two previous volumes in this series a wide variety of topics is presented as may be seen from the chapter headings: The Magnetoresistivity of Germanium and Silicon; the Chemical Purification of Germanium and Silicon; Electronic Conductivity of Silver Halide Crystals; Silicon Junction Diodes; Lifetime of Excess Carriers in Semiconductors; Scattering and Drift Mobility of Carriers in Germanium and finally Electronic Processes in Cadmium Sulphide. In calling on specialist authors for contributions, the editors have weighted the book in favour of germanium and silicon in view of their prime technological importance, but at the same time have included material on compound semiconductors.

The chapters on magnetoresistivity and bulk scattering processes are critical review articles which are admirable surveys of existing knowledge: they show that there is still much work to be done before a full understanding of the fundamental physics of the elemental semiconductors is achieved. That on lifetime is largely concerned with carrier recombination via intermediate centres which arise either from bulk imperfections or surface states in wide-gap semiconductors; the case of band-to-band transitions which is relevant for narrow-gap semiconductors is, however, also considered. The one paper dealing with devices—silicon junction diodes—deals adequately with their preparation; on the theoretical side, however, it is somewhat limited because of the lack of detailed agreement shown between experimental characteristics and those based on theory.

While the present standard is maintained it is clear that the usefulness of this series of annual volumes to the pure research worker, device designer and teacher is very great. Although not all the topics will be of direct interest to any one reader the specialist cannot afford to be unaware of the reviews presented here, which cover fields adjacent to his own. For those who are mainly interested in the performance of devices considerable benefit is to be gained from a study of those reviews dealing with the fundamental electronic processes.

F. J. HYDE

Transient Response from Frequency Response

By V. V. Solodovnikov, Yu. I. Topcheev, and G. V. Krutikova. 193 pp. 69 figs. Super Royal 8vo. Infosearch Ltd. distributed by Cleaver-Hulme Press Ltd., London. 1958. Price 42s.

WITH scientific and official co-operation from the U.S.S.R., Infosearch Ltd. of London has recently undertaken the production of a series of English translations of books by leading Russian scientists in the fields of physics, electronics, chemistry, metallurgy, and engineering. Some of these are now ready and are being distributed by Cleaver-Hulme Press Ltd. Among the first to appear is *Transient Response from Frequency Response* which has been translated and edited by A. Gelbtuch; the original was published in Moscow in 1955. Professor Solodovnikov, the first of the three authors, is at present in charge of the Central Research Institute for Complex Automation of the U.S.S.R. Academy of Sciences.

The book presents a development of G. F. Floyd's procedure for determining the transient response of a linear automatic control system from its frequency response. The system may be of minimum or non-minimum phase type with discrete or distributed parameters and may have a time delay. The frequency response of the system is first determined analytically or experimentally. By a graphical process, a set of trapezoids is then derived from the frequency response curve and a graphical representation of the transient response is obtained with the aid of tables of the function

$$h\kappa(t) = 2/\pi \left\{ S_i(\kappa t) + \frac{1}{1-\kappa}\left[S_i(t) - S_i(\kappa t) + \frac{\cos t - \cos\kappa t}{t} \right] \right\}$$

The primary purpose of the book is to provide fuller tables of this function than have previously been available. The tables are given for a hundred different values of κ from 0 to 1·00 at increments of 0·01 and the values of t range from 0 to 50·0sec at increments of 0·2sec.

The book contains three chapters. The first introduces the theory of the method and the second provides a wide selection of examples illustrating how the method is applied. The third chapter contains over a hundred pages of tables and nomographs to facilitate computation; the nomographs are printed on separate sheets and stored in a pocket at the back of the book.

S. R. DEARDS

Dilogarithms and Associated Functions

By L. Lewin. 353 pp. Demy 8vo. Macdonald and Co. (Publishers) Ltd. 1958. Price 65s.

THE title of this book is certain to fascinate anyone who sets eyes on it and he will not be disappointed. The author was stimulated by mathematical curiosity to investigate these functions so that, although later on writing up his results he found that many had been obtained before, his book has all the zest and enthusiasm of the pioneer.

Thus we can add a new name to the list of 'recognized' functions such as those of Legendre and Bessel. If any reader of this review is wondering—just as the reviewer did—what a dilogarithm is, he ought to read this book, for surely it is not for the reviewer to tell him.

Most of us have probably come across the function known as a dilogarithm in the course of our calculations but, until the reviewer read this book, he was quite unaware in how many different places it has occurred, including electrical network theory. In fact, the field of applications has been so wide that perhaps only Dr. Lewin, whose interests led him to explore them all, could have appreciated that the time was ripe for a book.

The dilogarithm, the trilogarithm and indeed the logarithm are all special cases of a class of functions studied by many mathematicians and the first monograph on them was written as long ago as 1809 by Spencer. Until this book appeared there has not been another, but Dr. Lewin has studied the literature exhaustively and not only collected and ordered the properties of the functions and many curious related facts, but included many new results and connexions of his own. The result is a valuable monograph including numerical tables of the main functions, which is not only extremely useful but will certainly stimulate those who read it.

G. J. KYNCH

Basics of Digital Computers

By John S. Murphy. 196 pp. (3 vols.). 80 figs. Demy 8vo. John F. Rider Publisher, Inc., New York. Price $6.95.

VOLUME 1 of this three-volume course reviews the development of computers and then explains the basic theory of computer arithmetic, data representation, circuits and control. Volume 2 discusses the logical elements and volume 3 deals with the system aspects of computers with discussions of types of memory, control systems, and input-output equipment.

Radio Valve Data

Compiled by 'Wireless World.' 136 pp. Medium 4to. 6th Edition. Iliffe & Sons Ltd. 1958. Price 5s.

THE latest edition of this widely used reference book has been enlarged and made easier to use. It now contains operating data on over 3 000 British and American radio valves, transistors, rectifiers and cathode-ray tubes. Twenty British manufacturers are represented, all of whom have co-operated with Wireless World to ensure that the information given is accurate, comprehensive, and up to date. A new feature of this edition is that the valve base connexion codes have been included in the index, as well as being retained in the main tables.

Elementary Engineering Science

By A. Morley and E. Hughes. 381 pp. 182 figs. Demy 8vo. 4th Edition. Longmans, Green & Co. Ltd., New York, Toronto, London. 1958. Price 11s. 9d.

THIS volume covers the Mechanics, Heat and Electricity sections of the syllabus in Science (Building and Engineering), Ordinary Level, of the Associated Examining Board for the General Certificate of Education and may therefore be found suitable for use in preparing pupils for this examination. It also covers the first year syllabus in Electrical Engineering Principles of the City and Guilds of London Institute's Courses for Electrical Technicians. The symbols and nomenclature are in accordance with the recommendations of the British Standards Institution. Twenty years have elapsed since the first edition of this book was published.

Chamber's Technical Dictionary

1028 pp. Demy 8vo. 3rd revised edition. W. & R. Chambers Ltd. 1958. Price 35s.

THIS new edition contains 55 000 entries drawn from more than 100 branches of scientific and industrial activity. It includes a 74 page supplement with nearly 5 000 definitions of important new terms covering the latest advances in science and technology.

Among the additions are aeronautical terms dealing with supersonic flight, guided missiles, aero engines; new chemical substances; definitions of electronic terms used in connexion with computers, automation, and colour television.

The Principles of Technical Electricity

By M. Nelkon. 250 pp. Demy 8vo. 3rd Edition. Blackie & Son Ltd. 1958. Price 15s.

THIS exposition of the fundamentals of electricity deals with the subject to the standard of the City and Guilds Telecommunications examination, Principles I and II. The usual introductory topics are covered and lead up to a.c. circuit theory, valves, typical radio circuits, and the cathode-ray oscillograph. For this third edition the text has been rewritten using the m.k.s. system of units which is now widely adopted. A comparison with c.g.s. units is given in the appendix.

ELECTRONIC EQUIPMENT

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

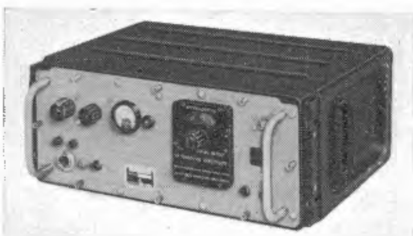
TACAN PERFORMANCE TESTER

(Illustrated below)

G. & E. Bradley Ltd., Electrical House, Neasden Lane, London, N.W.10.

THE Tacan performance tester is a portable 'go no-go test set' designed to check the efficiency of the airborne equipment used with both the British and American Tacan air navigation systems. The test set operates on channel 39, and is capable of checking the complete aircraft transmitter-receiver installation. It also provides an indication of the power output on all 126 channels.

Measurements of receiver sensitivity or transmitter power on channel 39 are referred back to the transmitter-receiver and presented as an 'accept/reject' indication of its performance. This is achieved by simulating the complex video, audio and r.f. waveforms generated by the beacons normally used



with the Tacan system. Receiver sensitivity at a frequency of 1 000Mc/s is checked by the test set at a simulated range of 5 or 105 miles, and at bearings of 140° or 320°. Both range and bearing facilities are selected by switches on the front panel. The point at which the aircraft indicators 'lock on' to the signals generated by the test set is interpreted as a pass or reject mark, dependent upon the sensitivity of the receiver. Normal beacon identity tone at 1 350c/s is maintained at the pilot's headset during all receiver measurements.

The test set will also provide an accurate measurement of the aircraft transmitter power at a frequency of 1 063Mc/s, again using the same accept/reject principle, and based on a minimum peak pulse power of 1kW. An assessment of the power generated on any of the 126 channels may also be made, but without the pass mark facility. Connexion between the test set and the Tacan equipment is made via an 18ft length of coaxial cable and a calibrated fixed attenuator, which is used for tests carried out either at the aircraft or in the workshop.

Portability and simplicity of operation are features of the test set which may be operated wherever a suitable power supply is available. It will operate from a normal 400c/s 115V aircraft power supply, or at 180 and 200V, 320 to 1 760c/s. Its power consumption at 400c/s is 100W.

The equipment is contained in a carrying case which is designed to withstand complete immersion in water, and proofed against the corrosive effect of salt water.

The test set is designed and built to meet the requirements of the Services, and has full type approval.

EE 3751 *for further details*

SPECTRUM ANALYSER

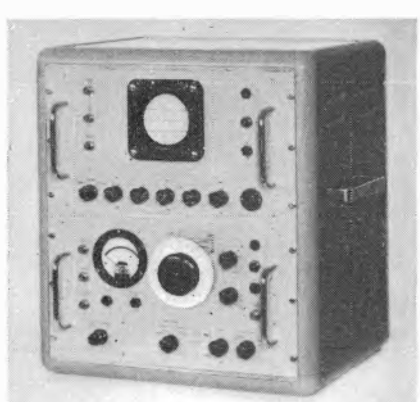
(Illustrated below)

James Scott & Co. (Electrical Engineers) Ltd., 68 Brockville Street, Carntyne Industrial Estate, Glasgow, E.2.

THIS instrument can be used at audio and supersonic frequencies to analyse the spectrum of a complex or simple signal. The frequency range of the standard instrument is from 1kc/s to 90kc/s, but custom-built equipment for other frequency ranges is manufactured to order.

The analyser provides two alternative presentations of the spectrum. For rapid qualitative analysis, the spectrum is displayed on a cathode-ray oscilloscope. The 'x' co-ordinate of the trace represents the frequency scale and the 'y' co-ordinate the power level at any frequency. Accurate calibration of the frequency scale is made possible by imposing a marker 'blip' on the trace derived from a calibrated manually controlled oscillator.

In order to make an accurate analysis of the power distribution at any part of the spectrum, the marker blip is lined up on the point of interest and the analyser switched to 'manual' operation. The manual oscillator setting now determines the centre frequency at which the measurment is made by means of a thermocouple milliameter. The signal power level is determined by using an accurate, calibrating oscillator, the meter



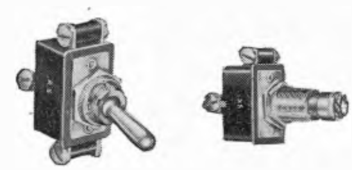
indication always being set to a constant value by attenuator adjustment.

The automatically swept oscillator can be switched to scan either 1 to 7kc/s or 1 to 100kc/s across the tube at a rate which may be controlled by the user.

The power in a part of the spectrum can only be expressed with reference to the bandwidth in which this is measured, in this equipment 1kc/s or 70c/s. The response of the narrow-band filter represents a close approach to the ideal 'plateau' as it incorporates no less than six staggered crystal filter elements.

In addition to its function as a spectrum analyser at frequencies up to 90kc/s, the instrument can also be employed for work at higher frequencies by the use of a suitable superheterodyne receiver in which the i.f. must naturally have a minimum bandwidth of 100kc/s.

EE 3752 *for further details*



HEAVY DUTY TOGGLE SWITCHES

(Illustrated above)

Radiospares Ltd., 4-8 Maple Street, London, W.1.

INCLUDED in the large number of components listed in their new catalogue Radiospares Ltd. have a new range of heavy duty toggle switches.

These switches are rated at 250V, 3A a.c.; they are for single hole, 7/16in diameter, fixing and are fitted with screw terminals. They are available in s.p.s.t., d.p.s.t., and d.p.d.t. types.

A push toggle version is also available. This has a d.p.s.t., push-to-make, push-to-break action and is also rated at 250V, 3A a.c.

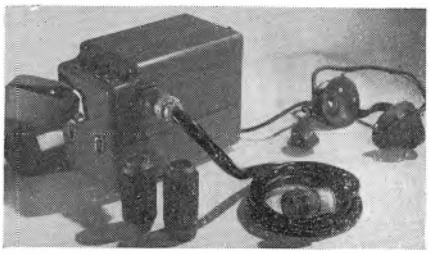
EE 3753 *for further details*

PORTABLE CABLE TESTER

(Illustrated above right)

De Havilland Propellers Ltd., Hatfield, Hertfordshire.

THE de Havilland cable test set provides, in easily portable form, the facilities for checking the continuity and insulation of multi-core cable looms hitherto available only in fixed or cumbersome equipment, Although initially designed for checking electrical systems in aircraft, the unit is of particular value in any type of installation where cabling is already in place, or where the cable runs are in confined spaces: its applications thus extend to many types of industry, including shipping, railways, mining, communications systems and factory maintenance.



Either a continuity or an insulation check can be carried out from one end of a cable run, the far end being terminated by a small unit weighing less than ½lb (450g). A test is initiated by a press button and the instrument automatically selects each wire in turn. The identification of the core undergoing test is indicated in a window which is illuminated by a red light if failure occurs, in which case the selector stops at the faulty core. Successful completion of a test sequence is indicated by the extinguishing of a separate indicator lamp. A reset button prepares the test set for a further test sequence.

A single outlet from the test set can be adapted for use with any cable by means of a suitable coupler supplied against the customer's requirements. In the case of a connector having a multi-position keyway (e.g. Plessey Mk IV) the coding is checked by a clearly marked movable keyway on both the coupler and termination units.

To enable long cable runs to be tested in conditions where the termination unit is more conveniently attached by a second operator, an intercommunication amplifier is incorporated in the instrument, connexion being made via the cable under test. Standard service style headsets may be used.

When this system is in use the operators receive an additional audible signal as each test is made and can distinguish when a cable failure is detected.

The speed of operation is 5 conductors/sec on continuity and up to 1 conductor/6sec on insulation, the test voltage being 500V d.c. The connector capacity is up to 25 ways per test.

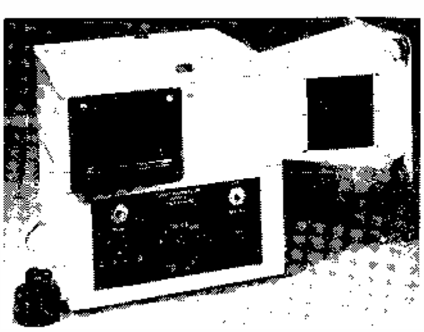
EE 3754 *for further details*

OSCILLOSCOPE CAMERA

(Illustrated below)

Cardiac Recorders Ltd., 63 Weymouth Street, London, W.1.

THE type 23A oscilloscope camera has been designed to bridge the gap between galvanometer multi-channel recorders and normal 35mm oscilloscope cameras. The magazine holds 50ft of unperforated 60mm recording paper, the whole width of which is available for recording.

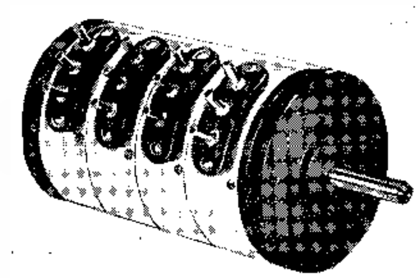


Two type of take-up cassette are available, a push feed for lengths up to 15ft and a power driven cassette for up to 50ft. Paper speeds, 0·02, 0·06, 0·2, 0·6, 2, 6in/sec may be changed during a recording, either by the speed change switch, or by electrical remote control.

A special feature of the camera is an automatic single frame advance, facilitating the taking of single sweep records. A built-in time marker exposes lines across part or the whole width of the paper at intervals of 1/25, 1/5, 1, 10sec, appropriate intervals are automatically selected by the speed change mechanism.

An $f3{\cdot}5$ bloomed lens with adjustable diaphragm in focusing mount is fitted as standard, other lenses to special order.

EE 3755 *for further details*



PRECISION POTENTIOMETERS

(Illustrated above)

P. X. Fox Ltd., Hawksworth Road, Horsforth, Yorkshire.

THE range of potentiometers manufactured by P. X. Fox Ltd has been extended to include precision type potentiometers with servo mountings.

Types with the servo mountings are of a completely new design using high precision windings in a special alloy aluminium housing to ensure a high rate of heat dissipation while giving the previous low torque characteristics, and they are available in the following types:—

SF. of 1·1"in o.d. in single or any required number of ganged form, and resistance per winding of up to 50kΩ.

SG. of 1·586in o.d. again in single or any required ganged units of resistance per winding of up to 50kΩ.

SB. of 2·528in o.d. available as above in any required single or ganged form of resistance per winding up to 100kΩ.

Torque varies with the number of ganged units but ranges from 0·5g./cm.

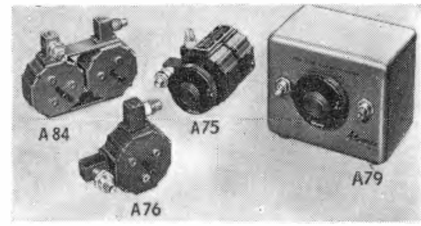
EE 3756 *for further details*

WIDEBAND ATTENUATORS

(Illustrated above right)

Advance Components Ltd., Roebuck Road, Hainault, Ilford, Essex.

THESE attenuators are a development the model A38, and employ resistive ladder networks at a constant



impedance of 75Ω. The construction and switch mechanism is similar to the model A38 and the fixing centres have been maintained in the case of the models A75, and A76, so that replacement with the new models is simplified. Special attention has been given in the design to obtain good v.s.w.r. performance. The standard sockets accept Plessey Minor Plugs.

Type A75, consists of two ten position networks providing an attenuation range of 99dB in 1dB steps. The two sections are mounted with concentric spindles and the attenuation setting appears in a window on the control knobs.

Type A76, consists of one ten position network providing an attenuation range of 90dB in 10dB steps.

Type A79, consists of the type A75, mounted in a metal case and is suitable for bench use in the laboratory or for any l.f. video, r.f. or v.h.f. application where a separate screened coaxial attenuator is required.

Type A84, consists of two ten position networks providing an attenuation range of 99dB in 1dB steps. The two sections are mounted side by side as required for example in a signal generator.

EE 3757 *for further details*

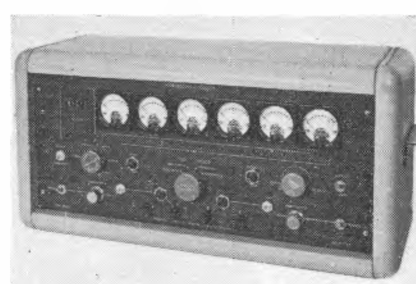
MICROSECOND STOPCLOCK

(Illustrated below)

Venner Electronics Ltd., Kingston-By-Pass, New Malden, Surrey.

THIS equipment, which uses transistors throughout, employs a new range of high speed counting stages to form a comprehensive time measuring equipment. The time range of the basic equipment is 3μsec to 1sec but this can be extended to 27·8h if required. The input impedance is 250kΩ and it can be arranged for single or two line working.

The equipment is capable of measuring pulse widths, mark to space ratios, period of pulse or other waveforms having amplitudes between 0·75V to 500V peak. Facilities are also embodied so that virtually any type of time measurement can be carried out on



contacts without the use of ancillary power supplies. The equipment is inherently self-checking on the applied waveform and has, in addition, six internally generated waveforms which are brought via a switch and emitter follower to a panel mounted socket.

The frequencies concerned are: 1Mc/s, 100kc/s, 10kc/s, 1kc/s, 100c/s, 10c/s.

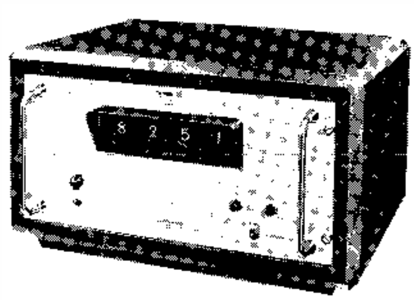
EE 3758 *for further details*

DIGITAL COUNTER

(Illustrated below)

J. Langham Thompson Ltd., Springland Laboratories, Bushey Heath, Hertfordshire.

J. LANGHAM THOMPSON LTD, have introduced a new digital counter incorporating several design features not usually associated with this type of instrument.



It is available in several forms to cater for different applications; it can also be supplied in a single form combining all these variations. Another interesting detail is the provision for up to two remote indicators.

The circuit employs the latest electronic techniques, and incorporates a crystal controlled cyclic timer with an accuracy of $\pm 0{\cdot}005$ per cent. The counter is therefore very stable and has an overall accuracy of ± 1 count.

Considerable attention has been paid to presentation and the brilliant neon display provides an in-line readout readable at 30ft.

The two versions of this instrument at present in production are a tachometer (see illustration), and a counter for random pulses up to 10kc/s.

EE 3759 *for further details*

PULSE HEIGHT ANALYSER

Sunvic Controls Ltd., 10 Essex Street, London, W.C.2.

THE prototype pulse height analyser shown by Sunvic Controls Limited at the Geneva 'Atoms for Peace' exhibition represents a considerable advance in flexibility of both arrangement and display. The instrument can be in the form of a desk type console measuring 62in × 29in × 40in approximately, or as one 6ft. or two 2ft 6in standard racks with display units at the top. Totals can be displayed in analogue form on a c.r.t., read in decimal form on Dekatron counters or automatically printed out.

A ferrite core storage matrix is used and provides 100 channels each with a maximum capacity of 16 binary digits (a decimal count of 65 535).

The measurement range is 10V with a channel width of 100mV. Minimum pulse amplitude which can be measured is approximately 0·75V.

When totals are displayed on the c.r.t. the six most significant digits are given in analogue form, horizontal deflexion being proportional to channel number and vertical deflexion to the counts in each channel. When the Dekatron indicators are used, any given channel can be selected or alternatively the instrument will read each channel in turn.

Unit construction, largely transistorized, is used, interconnexions being by multicore cables with plugs and sockets.

EE 3760 *for further details*

SERVO UNITS

(Illustrated below)

Servo Units Ltd., Farnborough Road, Farnborough, Hampshire.

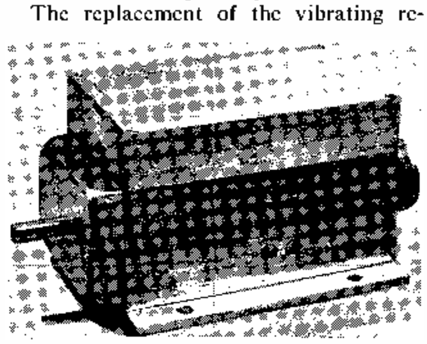
SERVO Units Ltd, the new subsidiary of Harvey Electronics Ltd. have produced an improved version of their servo units incorporating all the features of the earlier Harvey servos namely light weight, accuracy and high torque with the minimum of moving parts.

The previous version relied upon a vibrating relay to provide high torque without overshoot. In the new series the motor is fed via power transistors controlled either from a d.c. transistor amplifier or a thermionic amplifier supplied from a transistor h.t. supply all fed from the original d.c. supply. Mains a.c. supply units are also available.

The size of the smallest, type O.TR-6v, of these servos is 3·0in × 2·2in × 2·25 in, has a stalled torque of 1·0lb-ft. with a rate of response of 35°/sec, a repeatability of $\pm 0{\cdot}3$ per cent full scale on light load and on 90 per cent full torque 2·5 per cent. Type I.TR-12v is 4·10in × 2·55in × 2·50in, stalled torque 5·0lb-ft rate of response 50°/sec, repeatability $\pm 0{\cdot}3$ per cent and under 90 per cent full load 2·0 per cent 360° rotation can also be provided.

The servos are available with various gear ratios giving up to twenty times the above speed of response with proportionately less torque. Gears are also available separately for experimental work in conjunction with a standard unit. The gear box unit is also available separately.

The replacement of the vibrating relay by power transistors makes these units considerably more attractive for industrial applications and for fields where minimum servicing is essential, e.g. computers, handling of radioactive materials, spinning and weaving, paper manufacturing, rubber processing or any automatic or semi-automatic process.



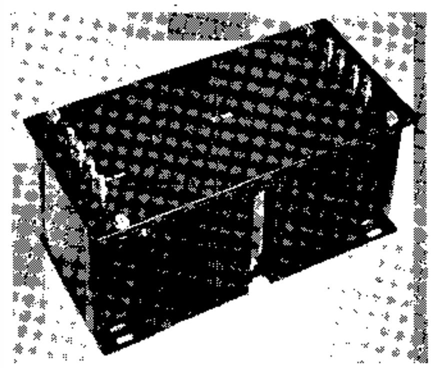
EE 3761 *for further details*

LINEAR SATURABLE REACTORS

(Illustrated below)

Haddon Transformers Ltd, Measons Avenue, Wealdstone, Middlesex

HADDON Transformers are now producing a range of linear saturable reactors up to 100kVA at unity power factor with wide range VA output and very low d.c. control power. The reactors are available in open or closed construction and may be processed to meet any climatic condition.



The 10kVA size, which is typical of the range, has an output VA at unity power factor of 3 to 91 per cent and requires a control power of 20W d.c.

EE 3762 *for further details*

MULTI-CHANNEL DIGITAL MILLIVOLT RECORDER

Bristol Aircraft Ltd, Filton, Bristol

THIS equipment will sample automatically 240 channels of thermocouples, recording on printed and punched paper tape, but can be used for voltage recording at all levels. The scanning system is entirely relay operated and digitizing is direct by a stepping potentiometer working with a 5-2-1-1 binary-coded-decimal step relationship. The detecting amplifier employs transistors and will operate at signal levels of 15μV. (This implies that the least significant of the three decimal digit presentations can correspond to 1/3°C for copper/constantan couples.) Alternative terminations in the form of typewriter or card punch may be used.

Many applications can be envisaged, from the collating of plant instrument data to the measurement of strains with resistive gauges.

The rates of operation are:

Punched card (3 channels/card) 60 channels/minute

Printed and punched tape 90 channels/minute

Typewriter punched tape 144 channels/minute

Punched tape 240 channels/minute

EE 3763 *for further details*

Notes from ————

NORTH AMERICA

Digital Simulation for Speech and Television Research

According to several papers read at the recent Western Electronics Conference by members of the Bell Telephone Laboratories, the use of general purpose digital computers in the simulation of new coding and transmission devices shows promise of broadening and accelerating speech and television research.

Complex experimental apparatus must often be built to test theories or design concepts in speech and video research. This construction is costly and time consuming. In addition, in the experimental stage, it is frequently difficult to distinguish equipment errors from theoretical deficiencies.

Simulation of new devices by large scale digital computers can greatly reduce expense and time lags, and thus make it easy to investigate a large number of approaches to coding and transmission problems. It even becomes feasible to investigate highly speculative methods which might otherwise have to be ignored. One of the greatest virtues of simulation, however, is versatility, according to the authors, who emphasized that computers can simulate any explicit operation on a speech or video signal.

Drs E. E. David, M. V. Mathews and H. S. McDonald, all of Bell Telephone Laboratories, described the use of computers in speech research. Speech is sampled, each sample is quantized into 10 bits or 1 024 amplitude levels and delivered to a magnetic tape recorder. These coded samples are recorded in seven parallel tracks, with 200 characters to the inch of tape. These tapes are then fed into the computer, where they are processed according to pre-assigned programs based on the coding or transmission scheme being investigated. The processed signals are then re-recorded, decoded, and played back for analysis and listener evaluation.

Computer memory requirements for speech processing are severe due to the large amount of data generated by even a short section of speech. Rapid access memory units must have a capacity of several seconds of speech to be useful. For example, an 8 000 32-bit word memory can conveniently hold only about 2·4sec of speech, at a sampling rate of 10 000 ten-bit samples per second.

A speech transmission scheme known as the 'Extremal' method has been studied at Bell Laboratories extensively by simulation, and illustrates graphically the advantages of the new technique. In its simplest form, only the extremes, or peaks and valleys, of a speech wave are sampled. The amplitudes and time of occurrence of these points are then transmitted, instead of a detailed representation of the entire wave. At the receiver, an approximation of the speech wave is generated by interpolating a suitable mathematical function between these points.

Listener evaluation of the simulated speech produced in initial tests showed that intelligibility is high—above 90 per cent sentence intelligibility—but that the quality is somewhat below that of commercial telephones.

Digital simulation has been applied to a number of other speech problems at Bell Laboratories. These include the design and evaluation of delta-modulation coders, pulse - code - modulation systems, detection of vocal pitch, and automatic recognition of speech.

Picture coding research has also been carried on by computer simulation. Mr. R. E. Graham and Dr. J. L. Kelly of the laboratories described a technique which has been used to evaluate a number of proposed television coding schemes, and to improve others. In order to hold machine time and memory requirements to a reasonable level, their system has used an input picture of 100 × 100 elements, corresponding to an area about 1/25 that of a conventional television frame. This 10 000 element 'window' has proved to be of sufficient size to allow critical evaluation of the processed images. For typical coding schemes, the total computer time required is 5 to 10sec per picture.

A magnetic tape recording of the video signal is prepared by scanning a square picture with 100 scanning lines in 2·4sec. Each picture dot is quantized to 10-bit accuracy, providing 1 024 amplitude levels, and recorded in the same form as the speech samples described above. The resulting signal is rooted, mixed with a synchronizing waveform in the conventional manner, and band limited to 2·5kc/s.

In playback, the computed picture signal is converted back to analogue form. It is then passed through a low-pass filter, and displayed on a monitor with two kinescope tubes. One of these tubes has a slow phosphor for direct viewing, while the other has a fast phosphor for photography.

A coding scheme which may have widespread significance in television transmission, known as 'Predictive Quantizing', has been studied by both simulation and conventional methods by Mr. Graham. This scheme takes advantage of the relatively low level of viewer perception during periods of scene change or motion, and in areas of picture confusion, these being the only regions in which predictive coding systems make significant errors. The method involves quantizing the difference between the original continuous signal and a predicted version of that signal. It employs fine quantum steps for small errors, and coarse steps for large errors, where the predictor and the viewer are surprised. This tapering of steps in the quantizing staircase allows the use of a smaller number of total levels, and thus reduces the channel capacity requirements.

A limiting case of predictive quantizing, in which the predictor is simply a one-sample delay, has been tested both with the simulation equipment described and with standard 525-line television, and found to afford a picture not significantly degraded from the original, and requiring only three bits per picture dot.

Analogue Computer

A new analogue computer of advanced design, and known as the model 200 has been introduced by the Colorado Research Corporation.

The basic model 200 is enclosed in a console unit equivalent in size to three standard racks. Included in the console are the patchbay assembly, a 0·01 per cent reference voltage divider, a precision panel mounted valve-voltmeter, and all the indicating lights, internal wiring, power and logic controls essential to the operation of an analogue computer system.

A writing shelf, at convenient desk height, extends across the console unit. Occupying the centre of the shelf is a raised turret supporting the control panel which places all the computer control and monitoring switches at the operator's fingertips, and all indicators are centralized and readily visible.

The control panel of the model 200 computer offers many new features and is very flexible. Component selection for adjustment or monitoring is achieved by sequential address entry on a keyboard of ten keys. Projection type indicators display the selected address, and any selection of erroneous or fictitious addresses is immediately revealed through a change in indicator field colour.

Another keyboard on the control panel is used for sequential entry of potentiometer values in the servo set potentiometer system which is utilized in the 200 Computer series. This CRC developed potentiometer setting system involves no clutches and can set ten turn potentiometers to the nearest turn of wire in an average time of less than two seconds. Panel mounted projection indicators record a four place potentiometer setting selection, with field colour showing when the desired value has been set.

Automatic scanning of computer components is a built in capability of the 200 console, and the equipment is easily adapted to tape programming and readout. A choice of AMP patching systems is available, ranging from a 1600 contact unshielded board to a 3600 contact shielded system with a patchboard assembled from individual nylon blocks. Passive computing elements are mounted directly behind the patchbay, and may be enclosed in a temperature regulated oven at the customer's option.

Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 26 November. Time: 6 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
The Annual General Meeting of the Institution and the Annual General Meeting of Subscribers to Brit.I.R.E. Benevolent Fund, followed by Presidential Address of Professor E. E. Zepler.

South Wales Section

Date: 12 November. Time: 6.30 p.m.
Held at: Glamorgan College of Technology, Treforest.
Lecture: Problems of Television Servicing.
By: E. A. W. Spreadbury.

South Western Section

Date: 25 November. Time: 7 p.m.
Held at: The School of Management Studies, University Street, Bristol 1.
Lecture: Electronic Transducers.
By: G. F. N. Knewstub.

South Midlands Section

Date: 28 November. Time: 7 p.m.
Held at: North Gloucestershire Technical College, Cheltenham.
Lecture: Television Recording.
By: P. J. Guy.

West Midlands Section

Date: 12 November. Time: 7.15 p.m.
Held at: Wolverhampton and Staffordshire College of Technology, Wulfruna Street, Wolverhampton.
Lecture: Transistors in Television Receivers.
By: B. R. Overton.

North Western Section

Date: 6 November. Time: 6.30 p.m.
Held at: The Reynolds Hall, College of Technology, Sackville Street, Manchester 1.
Lecture: The Determination of Aspect by a Shipborne Radar.
By: E. L. Crouchman.

Merseyside Section

Date: 28 November. Time: 7 p.m.
Held at: The University Club, Liverpool.
Lecture: The First Transatlantic Telephone Cable.
By: F. Scowen.

BRITISH SOUND RECORDING ASSOCIATION

Date: 21 November. Time: 7.15 p.m.
Held at: The Royal Society of Arts, John Adam Street, Adelphi, London, W.C.2.
Lecture: The Design of Stereophonic Pick-up Cartridges.
By: Stanley Kelly.

THE INSTITUTE OF NAVIGATION

Date: 21 November. Time: 5.15 p.m.
Held at: The Royal Geographical Society, 1 Kensington Gore, London, S.W.7.
Lecture: Navigating Across the Antarctic.
By: George Lowe.

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings, unless otherwise stated, will be held at the Institution, commencing at 5.30 p.m.

Ordinary Meeting

Date: 6 November.
Lecture: The Recognition of Moving Vehicles by Electronic Means.
By: T. S. Pick and A. Readman.

Measurement and Control Section

Date: 4 November.
Lecture: Operating Experience with a Transistor Digital Computer.
By: R. C. M. Barnes and J. H. Stephen.
and: A New High-Speed Digital Technique for Computer Use.
By: D. Eldridge.
Date: 18 November.
Discussion: Ferro-Electrics.
Opened by: L. A. Thomas.

Radio and Telecommunication Section

Date: 19 November.
Lecture: Television Recording: A Survey of the Problems and the Methods Currently in Use.
By: J. Redmond.
Date: 24 November.
Lectures: The Use of Dispersive Dielectrics in a Beam-Scanning Prism.
By: J. S. Seeley and J. Brown.
The Quarter-Wave Matching of Dispersive Materials.
By: J. S. Seeley.
Theory of Reflections from the Rodded-Type Artificial Dielectric.
By: A. Carne and J. Brown.

East Anglian Sub-Centre

Date: 17 November. Time: 6.30 p.m.
Held at: The Technical College, Cambridge.
Lecture: The ZETA Reactor. (Joint meeting with the Eastern Branch of The Institution of Mechanical Engineers.)
By: R. Carruthers.
Date: 24 November. Time: 6.30 p.m.
Held at: The Crown and Anchor Hotel, Ipswich.
Lecture: British Columbia-Vancouver Island 138kV Submarine Power Cable.
By: T. Ingledow, R. M. Fairfield, E. L. Davey, K. S. Brazier and J. N. Gibson.

Mersey and North Wales Centre

Date: 17 November. Time: 6.30 p.m.
Informal talk: The Importance of Research in Hearing and Seeing to the Future of Telecommunication Engineering.
By: E. C. Cherry.
Date: 24 November. Time: 6.30 p.m.
Held at: Chester Town Hall.
Lecture: The Recognition of Moving Vehicles by Electronic Means.
By: T. S. Pick and A. Readman.

North-Eastern Radio and Measurement Group

Date: 3 November. Time: 6.15 p.m.
Held at: King's College, Newcastle upon Tyne.
Lecture: Some Case Histories of Business Computers in the U.S.A.
By: A. T. Starr.
Date: 17 November. (Time and place as above.)
Lecture: A New Cathode-Ray Tube for Monochrome and Colour Television.
By: D. Gabor, P. R. Stuart and P. G. Kalman.

North-Western Radio and Telecommunication Group

Date: 12 November. Time: 6.45 p.m.
Lecture: The Relation between Picture Size, Viewing Distance and Picture Quality, with special reference to Colour Television and to Spot-Wobble Techniques.
By: L. C. Jesty.

South-East Scotland Sub-Centre

Date: 4 November. Time: 7 p.m.
Held at: The Carlton Hotel, North Bridge, Edinburgh.
Lecture: Speed Control of Large Wind Tunnels, with particular reference to the R.A.E. 8ft × 8ft High-Speed Wind Tunnel.
By: L. S. Drake, J. A. Fox and G. H. A. Gunnell.
Date: 5 November. (Time and place as above.)
Lectures: Operating Experience with a Transistor Digital Computer.
By: R. C. M. Barnes and J. H. Stephen.
and: A Basic Circuit for the Construction of Digital-Computing Systems.
By: P. L. Cloot.

South-West Scotland Sub-Centre

Date: 4 November. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow, C.2.
Lectures: Operating Experience with a Transistor Digital Computer.
By: R. C. M. Barnes and J. H. Stephen.
and: A Basic Transistor Circuit for the Construction of Digital-Computing Systems.
By: P. L. Cloot.
Date: 11 November. (Time and place as above).
Lecture: The Recognition of Moving Vehicles by Electronic Means.
By: T. S. Pick and A. Readman.

South Midland Radio and Measurement Group

Date: 24 November. Time: 6 p.m.
Held at: The James Watt Memorial Institute, Birmingham.
Lecture: The development of a Guided Missile.
By: W. R. Thomas.

Rugby Sub-Centre

Date: 11 November. Time: 6.30 p.m.
Held at: Rugby College of Technoloy and Arts.
Lecture: A New Cathode-Ray Tube for Monochrome and Colour Television.
By: D. Gabor, P. R Stuart and P. G. Kalman.
Date: 26 November. (Time and place as above).
Lecture: Thermo-Nuclear Reactions.
By: S. Kaufman.

Southern Centre

Date: 5 November. Time: 7 p.m.
Held at: Southampton University.
Informal Evening: The Use of Transistors in Radio and Television.
Talk by: E. Wolfendale.
Date: 12 November. Time: 6.30 p.m.
Held at: The C.E.G.B. Offices, 111 High Street, Portsmouth.
Lecture: A Survey of Harbour Approach Aids.
By: A. L. P. Milwright.
Date: 19 November. Time: 6.30 p.m.
Held at: S.E.B. Showrooms, 17 New Canal, Salisbury.
Lecture: Colour Television.
By: C. J. Stubbington.
Date: 26 November. Time: 6.30 p.m.
Held at: R.A.E. Technical College, Farnborough.
Lecture: Operational Experience at Calder Hall.
By: K. L. Stretch.
Date: 28 November. Time: 6.30 p.m.
Held at: South Dorset Technical College, Weymouth.
Lecture: The Design of Transistors.
By: J. R. Tillman.

Western Centre

Date: 27 November. Time: 6.45 p.m.
Held at: Colston Hall, Bristol.
Faraday Lecture: Automation.
By: H. A. Thomas.

West Wales (Swansea) Sub-Centre

Date: 25 November Time: 6 p.m.
Held at: The Brangwyn Hall, Swansea.
Faraday Lecture: Automation.
By: H. A. Thomas.

THE SOCIETY OF INSTRUMENT TECHNOLOGY

Date: 12 November. Time: 6 p.m.
Held at: Manson House, Portland Place, London, W.1.
Lecture: Some Applications of Control Technique in Astronomy.
By: P. Fellgett.
Date: 19 November. (Place as above) Time: 2.30 p.m.
Symposium on Analytical Instrumentation Developed at Harwell.
Date: 25 November. (Time and place as above).
Lecture: Digital Data-Recording for Nuclear Experiments.
By: R. C. M. Barnes.

THE TELEVISION SOCIETY

Date: 6 November. Time: 7 p.m.
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.
Lecture. Advertising in Relation to Television.
By: D. Ingman.
Date: 21 November. (Time and place as above).
Lecture: The European Television Network—Some Operational Problems.
By: J. Treeby Dickinson.

PUBLICATIONS RECEIVED

RUSSIAN-ENGLISH GLOSSARY OF ELECTRONICS AND PHYSICS is a compilation of Russian terms used in the various specialized fields of electronics. It includes terms from the most recent issues of Soviet journals, including Automation and Remote Control, Journal of Technical Physics, Radio Engineering and Electronics, etc. To this has been added an appendix containing circuit notation. American equivalents for vacuum tubes, etc. Consultants Bureau, Inc, 227 West 17th Street, New York. 11. Price $10.

A GUIDE TO NUCLEAR ENERGY by R. F. K. Belchem attempts to explain to persons with limited specialist knowledge how nuclear reactors function, and includes a general description of the more important types now being constructed or studied. The survey includes a description of the constructional materials of importance to the nuclear energy industry. Sir Isaac Pitman & Sons Ltd, 39 Parker Street, Kingsway, London, W.C.2. Price 10s.

THE RADIO AMATEUR'S HANDBOOK, 35th edition, has recently been published by the American Radio Relay League, West Hartford 7. Connecticut, U.S.A. It is revised and restyled in the light of current needs as a radio construction manual, reference work, and training text for class or home study. Price $4.50.

WORKED RADIO CALCULATIONS by Alfred T. Witts, 2nd edition. In this revised edition the text has been adapted to the m.k.s. system of units with the object of bringing it into line with the recommended procedure for teaching in Technical Schools. Substantial additions have been made to the text to include calculations on a variety of radio receiver problems that the modern service engineer and radio student is called upon to solve. Sir Isaac Pitman & Sons Ltd, 39 Parker Street, Kingsway, London, W.C.2. Price 12s. 6d.

MESURE ET DETECTION DES RAYONNEMENTS NUCLEAIRES by J. Sharpe and D. Taylor is the French translation and adaptation by J. Chatelet of the books Nuclear Radiation Detectors and The Measurement of Radioisotopes published by Methuen and Co. in 1957. M. Chatelet is chief physicist at the Laboratoire National d'Essais. Dunod Editeur, 92 Rue Bonapart, Paris 6. Price 3 400F.