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Commentary

IN recent years there has been growing concern at the ever-increasing flow of scientists from this country—notably to the U.S.A., where they are attracted by higher salaries and, superficially at least, high standards of living. Many advertisements appear, both in the technical and the daily press, which offer considerable inducements to men possessing the necessary qualifications to go and work abroad, particularly in the U.S.A. and Canada. Also, many American firms have the financial resources to sponsor research on a very large scale and to offer better laboratory facilities than can usually be found in this country.

Lord Chorley, the Secretary of the Association of University Teachers stated recently that a serious drain is being made on Britain's scientific manpower and that "if we are not careful a large part of the cream will be skimmed off and deposited in America". He went on to say "all my colleagues on the science side have been losing a large proportion of their best students to the United States over the past year. They go mostly on research scholarships after obtaining doctorates in Britain, and the number who come back is very small".

Of course, it is no doubt to the ultimate good of mankind that the peoples of all nations should mingle freely together and, perhaps, that there should be some measure of exchange from one country to another. There can also be no doubt that the experience gained by a young engineer in going overseas is of inestimable value to him, and also to his country—if he ultimately returns; while the spirit of adventure which prompts many young men to go abroad is indeed an admirable trait. It may also well be argued that it is a man's duty to provide the best living conditions of which he is capable for his dependents, whether it is in this country or another. Further, the prognostications of some of our senior politicians do not make the prospect of future life in this country, and consequently the inducement to stay here, very attractive. From a nationalistic viewpoint it is, of course, somewhat gratifying to know that other countries are so keen to obtain the services of British engineers and scientists.

The export of brains, and very often exceptionally good brains, is, however, a thing the country cannot afford. It has already been made abundantly clear that if this country is to survive economically at all, and certainly if our standards of living are to be raised, then we need an ever-increasing supply of first-class scientists and engineers. It is, indeed, of little avail to lay down ambitious schemes of education and training if the best products of such schemes are to be enticed away.

It would then appear to be essential to stop or, better still, reverse this flow and not only keep all the scientists we can train, but induce men from other countries to come and work here.

Men cannot be prohibited from leaving this country; the very thought of doing so is revolting to the English way of life and, indeed, to any democratic mind. To offer financial inducements to stay here is, to say the least, difficult in a country which needs so actively to combat inflation. It may, however, be said that although the salaries of the professional classes have increased considerably since, say, 1939, they have increased nowhere near in proportion with those of the lower skilled workers. If financial reward is "out", what then is left to prompt a sought after scientist to stay in a country which, as Sir Anthony Eden recently put it "is in mortal peril of poverty by stages"? There is one incentive, and one to which perhaps the well educated man should be particularly susceptible. That is conscience or, to put it more boldly, patriotism; the pride of being British and of being proud of our long tradition of fine craftsmanship and engineering and, further, of having a steadfast desire to continue that tradition and to keep Britain in the forefront as a great trading and manufacturing nation. These are sentiments which are comparatively easy to engender in time of war, but which appear to be far more difficult to arouse in times of peace, even though the need be just as great.

Transistor Frequency Meters

By L. R. Blake*, Ph.D., B.Sc., A.M.I.E.E., and A. R. Eames*, B.Sc.

The frequency meter described operates on the principle that if a capacitor C is switched at a frequency f so that during each cycle the capacitor is charged completely to a voltage V and then discharged completely, the charging current has the mean value CVf and is a measure of frequency. A transistor is used to perform the switching duty, and a moving-coil meter to indicate the charging current. The frequency meter described is capable of operating at up to 100kc/s, with six ranges from 0·3 to 100kc/s, each having a linear scale. The minimum signal power required is about a microwatt (12μA, 60mV) and the meter is quite insensitive to signal level or waveform. The size of the instrument is dictated mainly by the moving-coil meter employed, and the supply power requirement is 250mW, which can be derived from a battery or mains. An accuracy of 0·5 to 1 per cent can be realized simply, with stability of a high order.

THE transistor frequency meter employs the switched capacitor principle, illustrated in Fig. 1. This is a very useful circuit which has found application not only as a method of obtaining a frequency indication in terms of a direct current, but also for computation purposes, where it provides a simple method of obtaining the product of two or even three quantities and, in slightly modified form, a quotient or reciprocal. The technique is to switch the capacitor C at a frequency f so that during each cycle it is charged completely to the supply voltage V and then discharged completely. Under these conditions the mean value of the charging (or discharge) curent is $I = CVf$ and

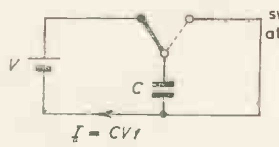

Fig. 1. Switched capacitor method of measuring frequency

becomes a measure of frequency provided CV is constant. It is sometimes more convenient to vary the capacitor voltage about the limits V and $V + \Delta V$ when the current is $C\Delta Vf$, and sometimes it is preferable to measure the capacitor current itself, its mean rectified value being $2C\Delta Vf$. An alternative method similar in principle is to switch a mutual inductor, changing its current by ΔI each cycle, which produces a mean rectified voltage of $2M\Delta If$.

Frequency meters have been constructed based on these principles in which the switching has been accomplished by mechanical relays and by electronic means, using gas-filled and vacuum tubes. Transistors are readily adapted as switching devices and a transistor frequency meter based on the switched capacitor principle has outstanding qualities with many advantages over the other methods. Two complete frequency meters have, in fact, been constructed, one using a point-contact transistor to switch the capacitor and the other using a junction transistor. Each are delightfully simple in construction and operation, a typical complete instrument being shown in Fig. 2. The general characteristics of the transistor frequency meter can be seen from Table 1 which gives the performance details of both meters. The figures given in the table are not necessarily limiting values, however.

It will be seen from Table 1 that the characteristic features of the transistor frequency meter are: a wide

Fig. 2. Complete frequency meter

frequency range—limited to about 500kc/s with available transistors, but this will undoubtedly be greater with future types; direct meter indication with any number of linear scales by connecting different values of measuring capacitor; reasonable accuracy—with the junction type transistor, accuracy to within 0·5 per cent is not difficult; very small input signal power requirements and complete independence of reading on signal voltage above the minimum value. This is achieved with three transistor stages in the point-contact transistor type and two stages in the junction transistor type. In each meter the first stage is a junction pre-amplifier to improve performance and versatility.

The main difference between the two instruments is that the junction transistor type is best suited for the measurement of oscillatory input signals, and the point-contact

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transistor type for pulse recurrence measurements; with this meter it is necessary to include the additional stage to enable it to deal also with oscillatory signals. A further difference is that the junction transistor type is simpler to calibrate, in this case the moving-coil meter is used to measure the capacitor charging voltage when setting it to the desired value, whereas in the point contact transistor type calibration is provided using the power mains frequency as a reference. Except at periods of load shedding, the frequency of the British grid makes a reasonably accurate reference, however. With the added refinement of automatic stabilization of the charging voltage, the junction transistor type shows possibilities as a high grade instrument needing no calibration facilities, and with a Kalium cell as source of reference potential, accuracy to within 0·5 per cent is possible with great freedom from drift.

Apart from the innumerable applications for the transistor frequency meter in the field of electronics generally, a useful and interesting application is that of tachometry in which the signal is derived from a vacuum photocell, photo-transistor or lead sulphide cell. A B.T.-H. lead sulphide cell or a S.T.C. type P50A photo-transistor, for example, will drive the frequency meter directly when placed in front of a stroboscopic disk under ordinary ambient (but not flickering) lighting conditions. In either instance a small d.c. supply of 20V or so is required, which is available from the frequency meter; in the case of the photo-transistor it is necessary to focus light on to the active element (which is 1mm^2) by a small lens placed in front of it. To measure the speed of a shaft, it is only necessary to splash part of the shaft with white (or black) paint and to hold the cell near it. A telephone type polarized earpiece is equally effective as a pick-up head if placed near a magnetic discontinuity, such as a key-way on the shaft.

Transistor Switching

To operate a frequency meter based on the switched capacitor principle, a circuit must be capable of converting an input signal to an output signal of the same frequency, or on exact multiple of that frequency, and having a voltage oscillating between accurately fixed limits. A few of the many circuits which are capable of doing this will now be examined.

Monostable Transistor Circuit

The basic circuit employed in the point contact transistor frequency meter is the emitter monostable transistor switching circuit, shown in Fig. 3(a)*. This circuit is very versatile and is in wide use[1,2,3]. It is ideal for pulse recurrence frequency measurement, since it is triggered from short pulses applied between emitter and base: measurements of an oscillatory signal requires an additional stage to produce the necessary triggering pulses.

Fig. 3(b) shows the input characteristic (that is, the input-voltage-current curve) of the monostable circuit, which is seen to consist of two positive resistance regions AB and CD separated by a negative resistance region BC.

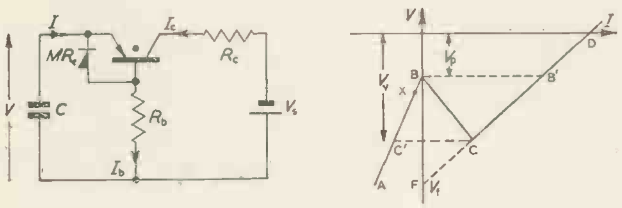
* *The circuits shown in this article employ both junction and point contact transistors and a means of distinguishing between them is desirable. The method adopted is to identify the point contact transistor by placing a dot between the collector and emitter connexions.*

TABLE 1

Summary of Performance of Transistor Frequency Meters

	POINT CONTACT TRANSISTOR FREQUENCY METER	JUNCTION TRANSISTOR FREQUENCY METER
Frequency range (kc/s)	0–30	0–100
Ranges provided (kc/s) (linear scales)	0·3, 1, 3, 10, 30	0·3, 1, 3, 10, 30, 100
Accuracy (per cent)	±1	±1
Minimum input signal (larger signals req. above 10kc/s) (mV r.m.s.)	600	60
(μA)	30	12
(μW)	18	0·7
Input impedance at minimum signal voltage (kΩ)	20	5
Maximum permissible series input impedance at 100V signal (MΩ)	3	8
Tolerable input signal voltage variation (V)	0·6–100 (higher voltages by adding external series resistor)	Range 1 0·06–30 Range 2 1·5–500 (higher voltages by adding external series resistor)
Operation with pulse input signal (f_m=Frequency at f.s.d.)	Satisfactory with pulses of duration greater than : 5μsec.	Satisfactory with pulses of duration greater than : 1/8 f_m. sec.
Meter employed	100μA f.s.d. with 5in mirror scale.	100μA f.s.d. with 5in mirror scale.
Supplies to meter—basic	40V. 36mA d.c. total. Derived from 250V. or 110V. 50 or 60c/s power mains.	20V. 36mA d.c. (with cell 20V 12mA d.c.) Derived from 250V. or 110V. 50 or 60c/s power mains.
Controls provided	Range switch Calibration potr. Input-cal. switch	Range switch Calibration potr. Input-cal. switch
Method of calibration	Set range to 0·3 kc/s and switch to CAL. Adjust pot. until meter reads 4× mains freq.	Set to CAL and adjust pot. until meter reads f.s.d.
Stability	Directly proportional to mains voltage variations (with Kalium cell less than 0·03 per cent/hr.)	Directly proportional to mains voltage variations (with Kalium cell less than 0·03 per cent/hr.)

Diode MR_e is included to keep the resistance low in region AB: without it the slope would have the high value corresponding to the resistance of the effective emitter diode in the open state. The circuit can be made to rest in a stable position, say X, by feeding a small negative current into the emitter. If, by means of a triggering impulse, the input voltage V is raised to the point B, the input current will suddenly rise to the value B′, and the capacitor will commence to charge negatively along B′C, aiming at the value V_f. At C the capacitor is unable to change its voltage suddenly so the current drops to the negative value C′, which starts to recharge the capacitor in a positive direction, back to the rest point X. In completing this cycle the capacitor voltage changes by ΔV between the limits V_p and V_v. If the cycle is repeated at the rate f, the mean rectified value of the capacitor current is $2C\Delta Vf$, provided the trigger pulse separation time is longer than the time required to complete the cycle, so that every trigger pulse is effective. Triggering need not necessarily be accomplished by raising the capacitor voltage from X to B, but can be achieved by temporarily depressing the peak point. In this case the rest point X must be stabilized and the capacitor given time to return to this point before re-triggering.



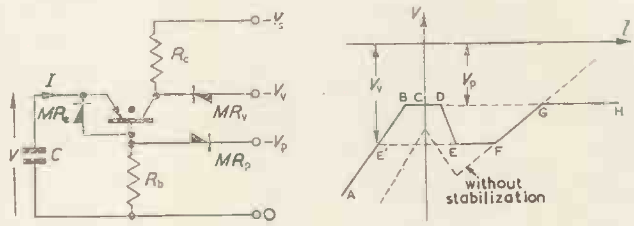
(a) (b)

Fig. 3. Basic monostable transistor switching circuit

The capacitor voltage cycling time depends on the charging resistances (slopes AB and CD), the value of capacitor C, and on the potentials V_f, V_v and V_p (or V_x). As the cycling time must be less than $1/f_m$, where f_m is the full-scale frequency for the appropriate range, and as the slopes R_{ab}, R_{cd} and the potentials V_f, V_v and V_p are fixed for a given transistor, the required conditions must be satisfied by adjustment of the value of capacitor C, which in turn defines the full-scale meter current. In practice this is about 200μA, and the meter resistance permissible is about 1kΩ, so a fairly rugged instrument can be employed.

A defect of the simple circuit of Fig. 3(a) is drift of the peak and valley potentials, due to change of transistor properties: V_p depends on transistor collector resistance r_c and V_v depends on r_c and on current gain α. Both α and r_c vary widely between transistors and with transistor temperature, which leads to variation of ΔV. Shea[3] shows a method of stabilizing the peak and valley points, and a similar though simpler circuit is shown in Fig. 4(a). This circuit more than stabilizes the peak and valley points: it increases the value of ΔV which can be employed without exceeding the rating of the transistor. Apart from the stabilization of the peak and valley points, controlled now mainly by the bias potentials $-V_p$ and $-V_v$, its behaviour on triggering is similar to the simple circuit of Fig. 3(a). Measurements on two dozen Mullard OC50 transistors showed a variation of peak voltage of 16 per cent, which was reduced to 0·8 per cent by peak stabilization and a variation of valley voltage of 6 per cent, which was reduced to 3 per cent by valley stabilization. The change of valley voltage is almost entirely responsible for the variation in ΔV of 10 per cent between different transistors. By operating the transistor at its limiting values, variation in ΔV can be reduced to about 3 per cent. The relatively ineffective stabilization of the valley voltage is due to resistance of the transistor and of diode MR_v. Drift of ΔV for a particular transistor is low, but calibration facilities are desirable due to possible variation.

Another form of monostable circuit suitable for pulse triggering is the collector monostable circuit, described by Anderson[2]. Its operation is much the same as the emitter



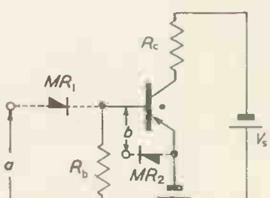
(a) (b)

Fig. 4. Stabilized monostable transistor switching circuit with input characteristic

monostable circuit and it appears no better as a means for eliminating drift of ΔV.

Point contact transistors can also be used in other types of switching circuits, a general form of which is shown in Fig. 5. By adjusting the base resistance R_b and the emitter bias voltage V_e, the transistor can be made stable in either the closed or the open state or in both states. The transistor can be triggered between base and emitter, and a diode in series with the triggering circuit may be used to ensure that a high impedance is presented when the transistor is on. In Fig. 5 a trigger voltage at a through diode MR_1 is obviously equivalent to triggering at b through diode MR_2. The voltage excursion of the collector can be limited and more accurately defined by diodes connected between collector and suitable bias potentials.

When the circuit shown in Fig. 5 is operated with V_e and R_b adjusted so that the transistor is stable in the "on" state, we have the "squaring circuit", two forms of which were described recently by Cooke-Yarborough[4] and by Chaplin[5]. "Squaring circuit" is used here to indicate a circuit which converts an input oscillatory wave of arbitrary shape into a flat topped wave. In these circuits the transistor can be opened by the trigger current, which must be maintained to keep the transistor in that state. Changing V_e and R_b enables the circuit to be stable in the "open" condition, when the circuit can be closed by a trigger current for as long as this current is applied. A further change of V_e and R_b produced a bistable circuit, switched on by one pulse and opened by another. All these circuits can be employed for use in a frequency meter.



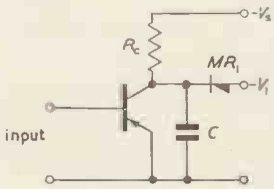


Fig. 5(left). General form of switching circuit

Fig. 6(right). Junction transistor operation as saturating amplifier with voltage swing of capacitor limited by diode

A characteristic of point contact transistors is their value of current gain $\alpha = \delta I_c/\delta I_e$, of greater than unity which gives rise to negative resistance, as in Fig. 3(a), in their characteristics viewed from any two terminals, if a positive feedback is provided in each case. In two-state circuits and monostable circuits using a single transistor this negative resistance is an essential requirement, but in squaring circuits it is not, and a saturating amplifier using a junction transistor with $\alpha < 1$ can be made to act in a similar manner, although even here the snap action switching of the point contact transistor, when the trigger current exceeds a well defined threshold value, limits the similarity.

Saturating Junction Transistor Amplifier

Although the snap action switching property of the point contact squaring circuit is absent in the junction transistor saturating amplifier, shown in Fig. 6, it has the advantage when used in a frequency meter of a low collector-emitter voltage in the "on" state, which in turn reduces the variation and drift of the voltage swing ΔV of the measuring capacitor. Measurements showed a collector-emitter voltage (with a base current of 200μA and a collector current of 4mA) of 0·11V ± 0·03V and 0·09 ± 0·03V for a score of Mullard OC70 and OC71 transistors respectively. As the maximum permissible collector-emitter voltage in the open state is 10V, the variation of ΔV between transistors due to this cause is about ± 0·3 per cent, which is a great improvement over the point contact transistor circuit; this variation can in any case be allowed for by the method of calibrating. A temperature increase of the transistor from 20°C to 45°C causes the "closed" collector-emitter voltage under the above conditions to decrease by 0·005V, which is negligible. Variation of the capacitor voltage with the transistor in the "open" state can arise due to inherent resistance of the catching diode MR_1 and changes in the open state transistor properties, mainly the collector current at zero base current, I_{co}', which has a value of about 150μA at 20°C and rises to 1mA at 45°C. This can cause a change of about 1mA in the catching diode and alter its voltage drop by 0·1V. At the same time a further change of the diode voltage drop of nearly 0·1V can arise due to the temperature rise of the diode itself, which adds to the previous effect and gives an overall change in the maximum capacitor voltage of 0·2V, or 2 per cent in ΔV.

These temperature effects can be compensated for readily by a thermistor in a resistor chain controlling the bias potential V_1, and if this is done, temperature stability of ΔV to ± 0·5 per cent for a 30°C rise is not difficult.

It is equally important to stabilize the potentials V_s and V_1, particularly the latter, and a "reference voltage diode"[6,7] allows this to be achieved very conveniently and effectively. When this is done, and temperature compensation added, if necessary, the frequency meter based on the circuit of Fig. 7 becomes an instrument which, without need for calibration, is capable of an accuracy to within ± 1 per cent or even better. Another method is to use a Kalium cell to provide the reference potential V_1, and though calibration facilities may then be necessary to be sure of ± 1 per cent or ± ½ per cent accuracy, this makes a very stable and satisfactory reference source. There is no doubt, however, that of the two methods, the use of the silicon reference voltage diode is preferable, since at this voltage level of 5 to 10V, it can be made with almost ideal characteristics and virtually zero voltage coefficient[7].

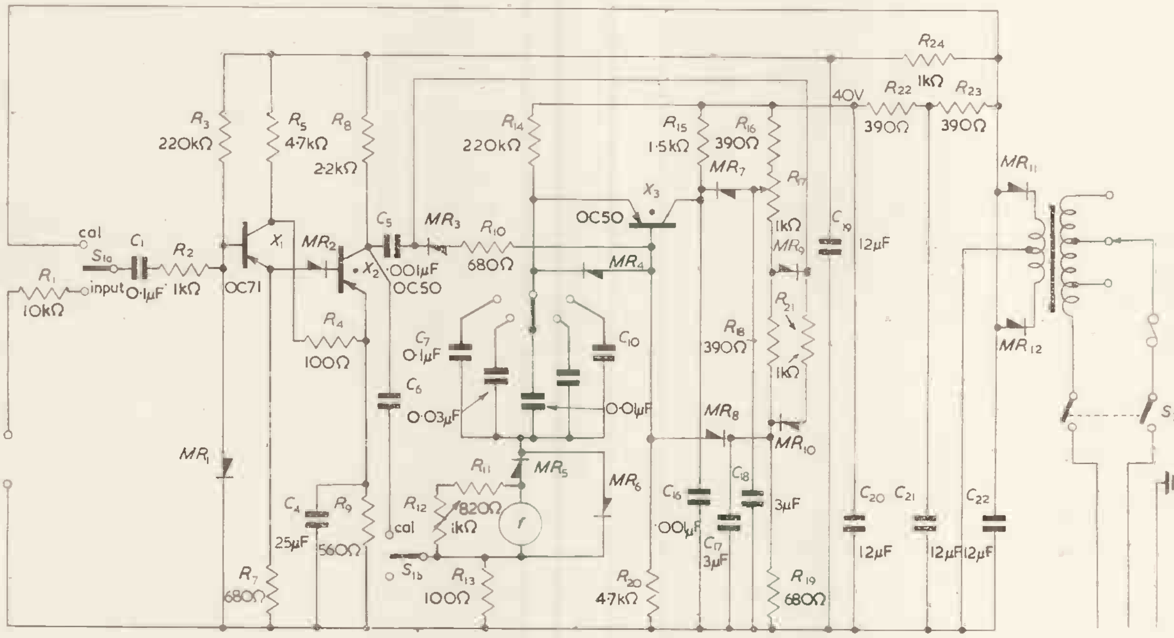


Fig. 7. Complete circuit of meter based on point contact transistor monostable switching circuit

Point Contact Transistor Frequency Meter

The complete circuit diagram of the point contact transistor frequency meter is shown in Fig. 7. It is seen to divide conveniently into four main parts: the junction transistor amplifier and impedance convertor, which is reasonably conventional; the point contact squaring circuit which is a variant of Fig. 5; the monostable point contact transistor circuit Fig. 4(a); and the d.c. supplies.

The instrument is calibrated by appropriately switching S_1 and adjusting the meter current using R_{12} until the meter indicates four times the mains frequency. Frequency quadrupling is desirable to give a larger deflexion on the 300c/s range to assist accurate calibrating, and is achieved in a very simple manner. In the calibrating position the base connexion of transistor X_1 is connected to the first smoothing capacitor of the d.c. supply which has, superimposed on the direct voltage, a 10V peak-to-peak 100c/s ripple which is more than adequate for triggering purposes. This provides the first frequency doubling stage. The second stage of frequency doubling is effected by triggering the monostable circuit twice per cycle instead of once, by triggering the emitter as well as the base. This is done by connecting, in the calibration position only, capacitor C_6 to the emitter circuit at the triggering resistor R_{13}; C_6 and R_{13} differentiate the square wave appearing at the collector of X_2.

The individual design details of the four stages will now be summarized. The grounded collector stage is provided to raise the input impedance from the low squaring circuit-value of a few hundred ohms, to 10kΩ at the base of X_1, or 20kΩ at the input terminals. There is no voltage gain at this stage, so the power gain is about 14dB. Resistor R_1 and diode MR_1 are included so that the input voltage can be raised to 100V r.m.s. without damaging the transistor. Resistor R_3 ensures that transistor X_1 remains in the closed state. A positive pulse applied at the positive input terminal then opens X_1, which causes a rise in the base current of point contact transistor X_2, which opens it. The resulting negative pulse at the collector of X_2 is differentiated and applied to the base of monostable circuit transistor X_3. Resistor R_4 sets the potential of the collector of X_1 about 0·5V negative with respect to the emitter of X_2, which is a convenient reference. Diode MR_2 is included to prevent relaxation oscillations caused by C_5, which tends to make the squaring circuit act as an astable collector circuit. Resistor R_9 is set so that X_2 closes, with the base biasing current not too large, so that triggering is sensitive yet reliable.

In the monostable circuit the meter is connected on the ground side of the measuring capacitors to minimize the effects of stray capacitance, which in this position tends to by-pass the meter. Resistor R_{12} is the calibration control. Half-wave rectification, with the diodes MR_5 and MR_6 connected as shown, is preferable to full wave rectification since it helps to keep down the resistance of the emitter circuit, particularly by removing the meter resistance from the circuit during the recharging period (Region C′B of Fig. 3(b)). This reduces the cycling time appreciably, since the time to charge from C′ to B virtually defines its value. Resistor R_{14} ensures that the transistor normally remains open (at position x in Fig. 3(a). Resistors R_{15} and R_{20} are adjusted to give maximum capacitor voltage swing ΔV without exceeding the transistor limiting current values. Diodes MR_7 and MR_8 are the valley and peak stabilizing diodes respectively and resistors $R_{16\text{-}21}$ form a reference voltage chain, also used with diodes MR_9 and MR_{10} to limit the triggering voltage to definite values, and prevent the triggering impulses from influencing the meter reading. Potentiometer R_{17} of this chain allows the valley stabilizing voltage to be pre-set according to the requirements of a particular transistor.

The upper frequency limit of 30kc/s for this meter is set by the finite switching time of transistors X_2 and X_3. In transistor X_3 the switching time becomes a noticeable part of the cycling time at 100kc/s, which causes a reduction of the voltage swing of the measuring capacitor and a drop in reading near full-scale deflexion. Equally important is the lack of sharpness of the triggering pulse from X_2, which also begins to influence the capacitor voltage at about 100kc/s, causing an increase in the reading. It is possible, however, to measure up to 100kc/s—and even up to 300kc/s with selected transistors—but the scale becomes a little non-linear and it is necessary to raise the minimum signal voltage.

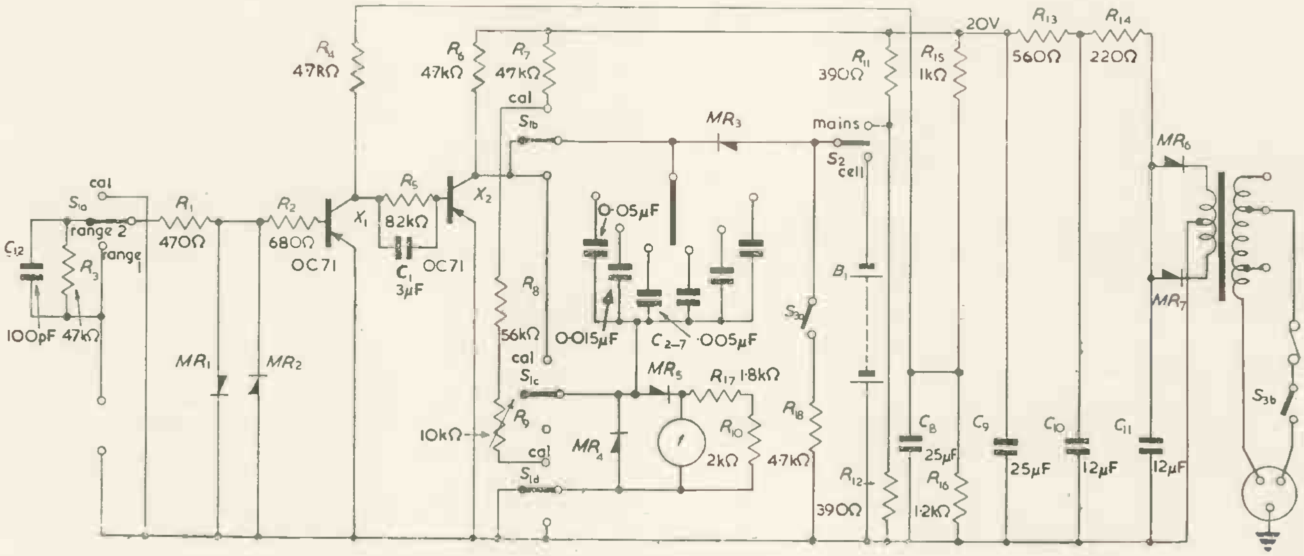


Fig. 8. Complete circuit of meter based on saturating junction transistor amplifier circuit

Junction Transistor Frequency Meter

The circuit of Fig. 8 shows an alternative form of frequency meter based on the saturating junction amplifier circuit of Fig. 6. It is not so suitable for short pulse recurrence frequency measurement as the previous meter, but it is an improvement for oscillatory signals. Its use enables a higher frequency to be measured, even though junction transistors are employed throughout, since the transistors are operated at the signal frequency and not effectively about ten times as high. Moreover, the minimum input signal requirements are lower (60mV and 12μA minimum) and calibration with respect to the mains is not necessary.

The frequency measuring saturating amplifier stage is basically as in Fig. 6, with the meter connected on the ground side of the measuring capacitor to limit stray capacitance effects. Half-wave rectification of the capacitor current is again employed, for two reasons in this case. When the transistor X_2 is open, the capacitor charges negatively through R_6, the capacitor aiming at the value -20V, but being cut-off by diode MR_3 at -10V. If the capacitor were charged through the meter resistance R_m, the voltage across R_m at the time of MR_3 closing would be

$-10(R_m/R_6)$ volts, so capacitor C would not be fully charged to -10V and would have to charge to its full value through R_m; this would materially increase the charging time to 99 per cent full value. It is better to charge the capacitor negatively through MR_4, when the capacitor is charged to within 1V when MR_3 closes. The second reason is to have some resistance in series with C when it discharges through X_2, to limit the discharge current to the specified safe values.

It was found that although each frequency range is linear to $\pm 0{\cdot}25$ per cent, limited by the meter itself, the measuring capacitors required were not in exact inverse proportion to the frequency range. This proved to be due to meter inductance. The effect of this is to make it necessary to trim each capacitor except one, or alternatively to shunt the meter with resistance and use larger capacitance values, which can then be exactly related to the frequency range. If a high impedance meter is desirable, one or other course must be taken. A low resistance and more sensitive meter overcomes this trouble, however.

The input impedance of the saturating amplifier is about 2kΩ so the value of an earthed collector input stage as impedance convertor is lost and it is better to use an earthed emitter amplifier to obtain a voltage as well as a current gain. Apart from the power amplification which the input stage provides, it also provides a simple method of triggering the second stage with an accurately defined signal over a wide range of input voltage; the input diodes MR_1 and MR_2 and resistors R_1 and R_3 further extend this voltage range to 30V and 500V respectively. The low signal requirements enable diodes MR_1 and MR_2 to be connected back-to-back without series resistance to act as a voltage limiter.

The upper frequency range of 100kc/s is limited by the transistors; at about 100kc/s in most transistors, but higher than this in some; the output voltage at the collector of X_2 is markedly trapezoidal, due presumably to hole storage and transit time effects; a further increase in frequency causes the waveshape to become triangular and the voltage swing to be limited.

A useful refinement of the circuit, which is included in Fig. 8, is to employ a dry cell to provide the reference voltage—inserted in place of R_{11} and R_{12}. Under these conditions diode MR_3 isolates the cell as long as X_2 is closed, and closes when X_2 opens to allow a charging current, equal to the current in R_6, to flow. Unfortunately, a Kalium cell is not very stable under conditions of charge, even at the low value of 2mA as in this case. A test showed that, when charged at 2mA for 30min a V0106 cell increases in potential by 2 per cent. It is necessary therefore to shunt the cell by resistor R_{18} choosing a value so that the current in it equals the maximum charging current. If this is done, the drain is continuous at 2mA with no signal, and with a symmetrical signal, the drain falls to a mean value of 1mA. With 2mA drain the life of a V0106 cell is about 1 000 hours. The addition of another pole to S_2 in series with the resistor across the cell prevents drain when the meter is not in use.

The stability of the instrument is demonstrated by the fact that when operating continuously for eight hours with an accurate 1kc/s signal, the reading fell by only 0·2 per cent, this being due to drop in cell potential. If rested, the cell recovers to its original voltage value. A Kalium cell also has the advantage that it makes the instrument insensitive to the supply voltage variations; for example, variation of the d.c. voltage supplied to R_6 in Fig. 8 by ± 20 per cent changes the meter reading by only ± 1 per cent.

Conclusions

The transistor frequency meter described has the advantages of a wide frequency range with any number of linear scales, small input signal power requirements, complete independence of reading on signal voltage, reasonable accuracy and stability, great simplicity, and small size depending mainly on the size of the moving-coil meter employed. It has the disadvantage of requiring an auxiliary power source, either a battery or the power mains, and the need for calibration, though this can be avoided. The instrument should be useful in innumerable applications in connexion with electronic and semiconductor circuits, for tachometry, and many other industrial purposes.

Addendum

After this account was written a voltage reference diode ("Zener diode"), was obtained and tested. As anticipated this proved to be by far the most satisfactory voltage reference and eliminates the need for calibration. The diode employed had an inverse cut-off of 7·7V which increased by 0·5 per cent/°C and an incremental resistance in the saturated region of 5Ω.

There are two convenient operating positions for this diode: connected directly across transistor X_2 in Fig. 8, when MR_3, R_{11}, R_{12}, R_{18}, B_1, S_2 and S_{3a} can be omitted together with, possibly, the calibration facility provided by R_7, R_8, R_9, S_{1b} and S_{1d}; or in place of resistor R_{12}, when R_{18}, B_1, S_{3a} and S_2 can be omitted and R_{11} raised in value. When across the transistor the effects of supply voltage variation are almost entirely eliminated, provided the diode operates in its saturated region, but unfortunately hole storage in the diode limits high frequency operation to the 30kc/s range. This restriction is not imposed when the diode replaces R_{12}; supply voltage stabilization is not then as good, but remains adequate with mains voltage variation effects reduced by 10:1, and the current drain is larger, due to the circuit R_{11} and the Zener diode. Thus, the simplest circuit is achieved with the diode connected directly across transistor X_2, when calibration facilities and voltage reference chain R_{11}, R_{12}, B_1, etc., become unnecessary. Moreover the current drain of the circuit then becomes so small (5 and 10mA with and without input signal respectively) that a dry battery becomes the most convenient method of supply, which brings further circuit simplification. Thus, in these circumstances the complications, introduced by supply difficulties and the need to remove errors by compensation and calibration, are eliminated and the circuit returns to virtually its ideal form.

Acknowledgments

The authors are indebted to colleagues in the Research Laboratory of the British Thomson-Houston Co. Ltd., for assistance in this work, in particular to Mr. R. G. Hibberd who provided the experimental voltage-reference "Zener" diodes.

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Waveguide Components with Non-Reciprocal Properties

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(Part 1)

Ferromagnetic Resonance and Faraday Rotation

When an electromagnetic wave travels through a magnetized magnetic material, the polarization rotates; the physical mechanism causing this Faraday rotation and a related effect, ferromagnetic resonance are explained in Part 1. Waveguide components, which depend on these effects for non-reciprocal properties are described in Part 2.

WAVEGUIDE components whose attenuation and phase characteristics depend on the direction of propagation and which therefore violate the reciprocity theorem have recently been developed. These components have many useful properties and will probably result in considerable changes in microwave techniques, particularly when common transmission and reception is desired from one aerial or when a waveguide switch is required. The two simplest elements are the so called "gyrator" which depends on the non-reciprocal phase shift and the "isolator" for which a wave travelling in one direction is only slightly attenuated while a wave in the opposite direction is greatly attenuated. Descriptions of these components and their uses will be given in Part 2.

Two methods of producing non-reciprocal properties have been suggested, the first requiring that a proportion of the energy in the waveguide should travel through a ferromagnetic material which is subjected to a strong steady magnetic field. In the second method, the ferromagnetic is replaced by a discharge tube and the behaviour of a wave travelling through the ionized gas in this tube is similar to that of a wave propagating in the ionosphere. This article is concerned with the first method and in Part 1, an account of the properties of suitable ferromagnetic materials and a qualitative discussion of the theory of propagation in such materials is given.

Ferromagnetic Materials Suitable for use at High Frequencies

An essential requirement for the production of non-reciprocal components using ferromagnetic materials is that propagation through the material should be possible without excessive attenuation. Most of the commonly used ferromagnetic materials are metallic alloys which are good conductors and in them any high frequency wave is rapidly attenuated because of the eddy currents which are excited. Such eddy currents lead to a conversion of the electromagnetic power to heat. A similar difficulty arises in the provision of magnetic cores for r.f. coils and has been solved to a certain extent by using dust-cores, formed by very small particles of a ferromagnetic alloy electrically insulated from each other. More recently, a class of magnetic materials, known as ferrites[1], which have very low conductivities have been developed for use as r.f. cores and it is the availability of these materials which has permitted the development of non-reciprocal waveguide components.

Chemically, the ferrites consist of compounds given by the formula $XO.Fe_2O_3$ in which X is a divalent metal, such as copper, zinc, lithium, cadmium, iron (in the ferrous state), magnesium, nickel, divalent manganese or cobalt. The earliest known magnetic material, lodestone, is ferrous ferrite. Ferrites cystallize in a form known as a "spinel" and detailed studies of the properties have been made by a number of workers[2,3]. Single crystals can be grown with considerable difficulty but practical forms of the material are polycrystalline ceramics. If the divalent metal is ferromagnetic so is the ferrite and by using mixtures of ferromagnetic and non-magnetic ferrites, the degree of ferromagnetism can be controlled within quite wide limits. A commonly used material is a mixture of nickel and zinc ferrites.

The ferrites can be quite easily magnetized, fields of the order of 1 600AT/m (20 oersteds) being sufficient to secure saturation. The flux density in a saturated material is of the order of 0·3Wb/m² (3 000G). The initial relative permeability at low frequencies is in the range 20 to 600 and falls to values close to unity at frequencies in the microwave region. The behaviour of a wave propagating in a ferrite depends also on the relative permittivity which lies in the range 10 to 15 for microwave frequencies. The attenuation arises from both dielectric and magnetic losses, the loss tangents being in the range 0·01 to 0·1 for the dielectric loss and extending up to at least unity values for the magnetic loss. The attenuation is therefore considerably greater than in dielectric materials such as polystyrene for which the loss tangent is 0·0003.

The reason for the importance of the ferrites lies in the variation of the magnetic properties at high frequencies which arises when a steady magnetic field is applied to the material. An indication of the effects which then occur is given in the next section.

The High Frequency Behaviour of a Ferromagnetic Material Subjected to a Steady Magnetic Field

An explanation of the high frequency behaviour of a magnetized material requires some knowledge of the relation of the magnetic properties of the material to its atomic structure. As is well known, any electric current generates a magnetic field and since the motions of the electrons of an atom constitute currents they must set up magnetic fields. There are two different types of motion to be considered, the movement of each electron in an orbit around the nucleus of the atom to which it is attached and a second motion which arises because the electron is to be regarded as a charged particle which is spinning about its axis. For most atoms the combined effect of the orbital motion of the electrons and their spins gives no magnetic field. The exceptions are the ferromagnetic elements, which have permanent magnetic properties because the magnetic

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fields caused by the electron spins do not balance out. Each spinning electron behaves as a elementary magnet whose axis is parallel to the axis of rotation, and when a magnetic field is applied, this axis tends to align itself parallel to the direction of the field. The material is saturated when all the electron axes, which contribute to the magnetism, are directed parallel to the exciting field.

Suppose now that a ferromagnetic material is saturated by a steady magnetic field H_o and that an additional field is momentarily applied in a direction at right-angles to that of H_o. All the spinning electrons which contribute to the magnetization originally have their axes parallel to the direction of H_o and the effect of the extra field is the same as that of the force F applied at right-angles to the axis of the spinning top of Fig. 1(a). Under these conditions, the axis of the top moves in a direction perpendicular to both the axis and the direction of the force F. Further, when the force F is removed, the top does not return to its equilibrium position in which the axis is vertical but precesses, i.e. the axis sweeps out a cone and the frequency of the precession is determined

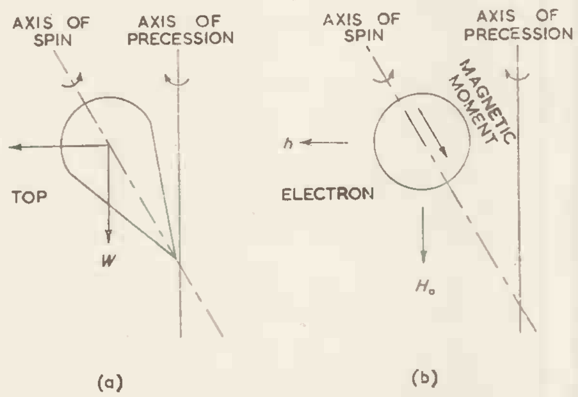

Fig. 1(a). Spinning top precession
(b). Spinning electron precession

by the moment of inertia and the gravitational field (W in Fig. 1(a)). Similarly, when the axis of the spinning electron (Fig. 1(b)) is displaced from its equilibrium position parallel to the steady magnetic field H_o, precession occurs: the restoring couple exerted by H_o on the displaced magnetic moment plays the same part as the couple provided by the gravitational force and the reaction at the point of rotation of the top. From the equivalence between the two, the direction in which the electron axis precesses can be found and is most conveniently stated as being the direction in which current must flow to establish the steady field H_o.

The frequency, f_o, with which the axis of the electron precesses, can be calculated from mechanical considerations and is given by the equation:

$$2\pi f_o = \gamma H_o = \omega_o \quad \text{(1)}$$

where γ is a constant called the gyromagnetic ratio and is the ratio of the magnetic moment caused by the electron spin to the angular momentum of the electron*. The value of γ may be calculated from the known properties of electrons and $\gamma/2\pi$ is approximately 0·035 if f_o is expressed in megacycles per second and H_o in amperes per metre. In Gaussian units, $\gamma/2\pi$ equals 2·8Mc/s/oersted.

Equation (1) applies for an isolated spinning electron under the action of the magnetic field, H_o, but an identical equation applies to ferromagnetic materials provided that

(a) The applied field is sufficiently large to saturate the material, i.e. to align the axes of the spinning electrons which contribute to the magnetization.

(b) The field H_o is interpreted as the field acting on each electron. This "internal field" will differ from the field applied to the material as a whole because of the additional fields arising from the magnetic moments of all the other electrons.

When a magnetic material is saturated by a steady field, H_o, and an alternating magnetic field is applied at right-angles to the steady field, the axes of the electrons precess at the frequency of the alternating field. This means that the direction of the magnetic moment of each electron and hence of the material as a whole may lie on the surface of a cone whose axis is parallel to the steady field, H_o. The flux density, B, within the material is always given by the equation:

$$B = \mu_o H + M \quad \text{(2)}$$

where M is the magnetization, i.e. the magnetic moment per unit volume in the material. The quantities B, H and M are all vectors since each has a direction as well as a

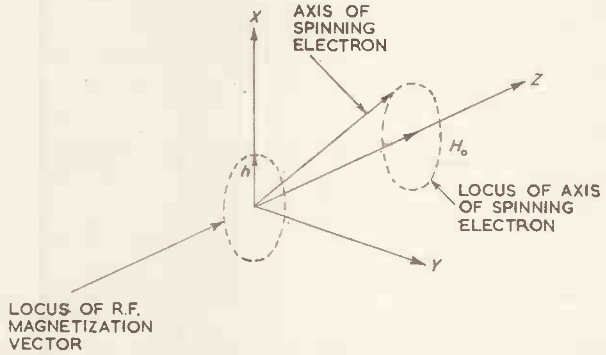

Fig. 2. Effect of the precession of the electron axis on the magnetization

magnitude. In the case being considered, it is sufficient to examine the relation between the time varying quantities. Suppose the directions of the alternating magnetic field, h, and of the steady magnetic field H_o are taken as the x- and z-axes respectively. Then, the magnetization has alternating components in both the x- and the y-directions (Fig. 2) because of the precession about the z-axis. It is clear from the figure that these two components must have a phase difference of 90° and a detailed analysis carried out by Polder[4] shows that the x-component of the magnetization is either in phase or in antiphase with the alternating magnetic field, h_x. In general therefore, the flux density will have a component b_x in phase or in anti-phase with h_x and a component b_y in phase quadrature with h_x. This gives the pair of equations:

$$b_x = \mu h_x : b_y = jkh_x \quad \text{(3)}$$

μ and k both being real quantities.

Similarly, if an alternating magnetic field is applied in the y-direction, components of flux density exist in both the x- and y-directions and if the steady magnetic field is again H_o in the direction of the z-axis, these components are given by:

$$b_x = -jkh_y : b_y = \mu h_y \quad \text{(4)}$$

An alternating magnetic field in the direction of the steady magnetic field produces an alternating flux in this direction only. Since the material is assumed to be saturated, there can be no change in the magnetization so that equation (2)

* *The magnetic moment and angular momentum of the spinning electron are oppositely directed, so that γ is sometimes taken as negative. In the present article, γ is defined in terms of magnitude only, so that it is positive.*

gives:

$$b_z = \mu_0 h_z \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (5)$$

The material may be acted on by a magnetic field with components in the directions of each of the co-ordinate axes and then the flux density has to be calculated using the three equations (3) to (5) which may be combined to give:

$$\begin{aligned} b_x &= \mu h_x - \mathrm{j}k h_y \\ b_y &= \mathrm{j}k h_x + \mu h_y \\ b_z &= \mu_0 h_z \end{aligned} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (6)$$

These equations may be written in matrix form as:

$$\begin{bmatrix} b_x \\ b_y \\ b_z \end{bmatrix} = \begin{bmatrix} \mu & -\mathrm{j}k & 0 \\ \mathrm{j}k & \mu & 0 \\ 0 & 0 & \mu_0 \end{bmatrix} \begin{bmatrix} h_x \\ h_y \\ h_z \end{bmatrix} \quad \ldots\ldots \quad (7)$$

The square matrix on the right is usually referred to as the permeability tensor: it should be remembered that this tensor only describes the properties of the material when it is magnetized by a steady field in the direction of the z-axis. The magnetic properties of a non-isotropic material must also be represented by a tensor, even in the absence of a steady magnetizing field, but the tensor is then symmetrical, i.e. corresponding off-diagonal elements are equal. In equation (7), the elements for row 1, column 2 and row 2, column 1 are not equal because of the change in sign, and it is this change which leads to the non-reciprocal properties.

Polder[4], in the analysis already referred to, has calculated the values of the constants μ, k and has obtained the equations:

$$\mu = \mu_0 + \frac{\gamma^2 H_0 M_0}{\gamma^2 H_0^2 - \omega^2} = \mu_0 + \frac{\gamma \omega_0 M_0}{\omega_0^2 - \omega^2} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (8)$$

$$k = -\frac{M_0 \gamma \omega}{\gamma^2 H_0^2 - \omega^2} = -\frac{M_0 \gamma \omega}{\omega_0^2 - \omega^2} \quad \ldots\ldots \quad (9)$$

in which M_0 is the steady magnetization and is therefore the value for saturation.

ω, and ω_0 are respectively the angular frequency of the alternating fields and the angular precessional frequency as defined by equation (1). Equations (8) and (9) have been derived for materials without losses, but in practice damping forces act on the precessional motion of the electrons and when allowance is made for these, the constants μ and k become complex. A modified form of equations (8) and (9), which is valid when the material is not saturated, has been developed by Rado[5].

Propagation in a Magnetized Ferromagnetic Material

The simplest type of wave which can occur in a homogeneous loss-free material is a linearly polarized plane wave, whose properties will be recalled as an introduction to propagation in materials such as ferrites. The principal properties are:

(1) The electric and magnetic fields are constant in magnitude and direction at every point in a plane perpendicular to the direction of propagation;

(2) The directions of the electric field, the magnetic field, and propagation are mutually perpendicular;

(3) The magnitudes of the electric and magnetic fields at any point in the material satisfy the equation:

$$|e| = Z_m\, |h| \quad \ldots\ldots\ldots\ldots\ldots \quad (10)$$

where Z_m, the wave impedance of the material, has the value $\sqrt{(\mu_m/\epsilon_m)}$, μ_m and ϵ_m being respectively the permeability and permittivity in M.K.S. units;

(4) If propagation occurs in the direction of the positive z-axis, the electric and magnetic fields are both proportional to $\exp(-\mathrm{j}\beta z)$, where the phase factor β satisfies the equation:

$$\beta = \omega(\epsilon_m \mu_m)^{\frac{1}{2}} \quad \ldots\ldots\ldots\ldots \quad (11)$$

The mechanism which supports the propagation of a plane wave is a combination of the generation of an electric field by a changing magnetic flux according to Faraday's Law and of the generation of a magnetic field by a displacement current according to Ampere's Law. Consider an alternating magnetic field acting in the y-direction and an electric field in the x-direction: the alternating magnetic field gives a flux density, given by:

$$b_y = \mu_m h_y \quad \ldots\ldots\ldots\ldots\ldots \quad (12)$$

If both fields depend only on z, then Faraday's Law shows that the e.m.f. induced by the changing flux gives a change of electric field with z according to the equation:

$$(\partial e_x/\partial z) = (\partial b_y/\partial t) \quad \ldots\ldots\ldots\ldots \quad (13)$$

The electric field is also varying with time and gives a displacement current:

$$i_x = \epsilon_m(\partial e_x/\partial t) \quad \ldots\ldots\ldots\ldots\ldots \quad (14)$$

which by Ampere's Law sets up an m.m.f. leading to a change in the magnetic field h_y with z, viz.:

$$(\partial h_y/\partial z) = \epsilon_m(\partial e_x/\partial t) \quad \ldots\ldots\ldots\ldots \quad (15)$$

The plane wave described above is obtained as one solution of the set of equations (12) to (15): the only other solution is a similar wave travelling in the direction of the negative z-axis.

An attempt to apply the same analysis to a magnetized ferromagnetic material fails because equation (12) must be replaced by the pair of equations in (4), i.e. h_y gives an x- and a y-component of flux density. The x-component of flux density generates a y-component of electric field, so that the resulting wave must have a changing direction of polarization. The simplest procedure for determining the form of this wave rests on replacing equation (12) by a similar relation between flux and magnetic field, which can be done for periodic fields. It is easily seen from equation (6) that:

$$\mathrm{j}b_x + b_y = (\mu - k)\,(\mathrm{j}h_x + h_y) \quad \ldots\ldots \quad (16)$$

which is similar in form to equation (12), except that h_y is replaced by $\mathrm{j}h_x + h_y$ and b_y by $\mathrm{j}b_x + b_y$.

In place of equation (13), there now appears the equation giving the two electric field components arising from the two changing flux components: this is:

$$(\partial/\partial z)\,(e_x - \mathrm{j}e_y) = -\mathrm{j}\omega\,(\mathrm{j}b_x + b_y) \quad \ldots\ldots \quad (17)$$

The y-component of electric field also gives a displacement current and finally equation (15) is replaced by:

$$(\partial/\partial z)\,(\mathrm{j}h_x + h_y) = \mathrm{j}\omega\epsilon_m\,(e_x - \mathrm{j}e_y) \quad \ldots\ldots \quad (18)$$

Equations (16) to (18) are identical in form to equations (12), (14) and (15), the changes being:

$$\begin{aligned} e_x &\to e_x - \mathrm{j}e_y \\ h_y &\to \mathrm{j}h_x + h_y \\ \mu_m &\to \mu - k \end{aligned}$$

The solution of equations (16) to (18) can therefore be found directly by comparison with that for the ordinary medium and is still a plane wave propagating in the direction of the z-axis or of the negative z-axis: the differences are that the effective permeability to be used in calculating the phase constant and the wave impedance is $(\mu - k)$ and that since both the electric and magnetic fields have two components in phase quadrature, the wave is no longer linearly polarized. At any point in space, the directions of the fields follow the sequence shown in Fig. 3. The ends of the vectors depicting the fields lie on a circle, i.e. the wave is circularly polarized, and the direction of rotation of the vector is clockwise when looking in the direction of the positive z-axis and is independent of the direction of propagation.

A second wave may be found by repeating the above

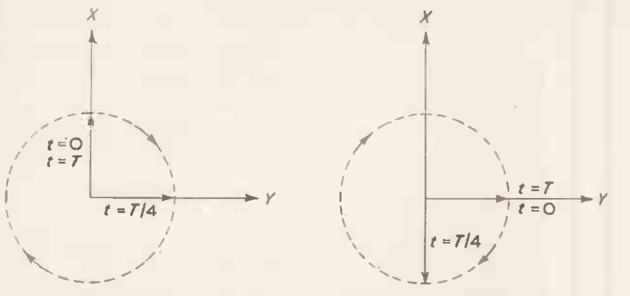

Fig. 3. Sequence of field directions for positive circularly polarized wave

analysis and using the result obtained from equation (6) that:

$$-jb_x + b_y = (\mu + k)(-jh_x + h_y) \quad \ldots\ldots \quad (19)$$

This wave is circularly polarized with the field vectors rotating counter-clockwise, when looking in the direction of the positive z-axis, and its phase velocity and wave impedance are calculated using $(\mu + k)$ as the effective permeability.

The simplest solutions for wave propagation in a magnetized ferrite therefore involve circularly polarized plane waves: the phase velocity and wave impedance for each wave depends on the direction of rotation of the field vectors with respect to the direction of the applied steady magnetic field. The wave for which the rotation is clockwise with respect to the direction of the steady field is called the positive wave and the other the negative wave. The direction of rotation for the positive wave is the same

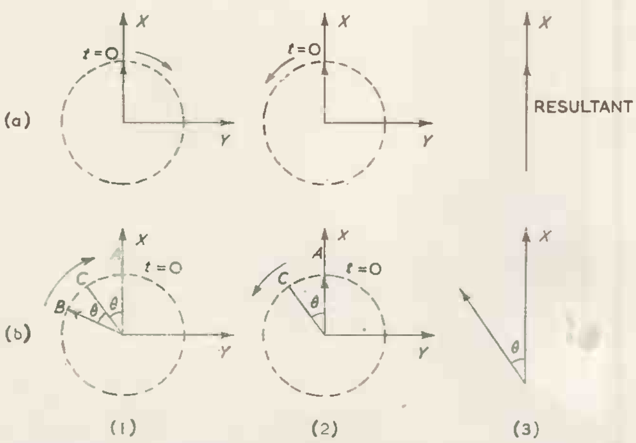

Fig. 4. Electric field components at (a) the incident face and (b) the rear face of the slab of ferrite

(1) and (2) respectively the positive and negative circularly polarized components, (3) the linearly polarized resultant

as that in which a current must flow to establish the steady magnetic field. It follows from the arguments above that the phase constants, β_+ and β_-, for the positive and negative waves respectively, are

$$\beta_+ = \omega \sqrt{[\epsilon(\mu - k)]} \quad \ldots\ldots\ldots \quad (20)$$

$$\beta_- = \omega \sqrt{[\epsilon(\mu + k)]} \quad \ldots\ldots\ldots \quad (21)$$

Faraday Rotation

Suppose a linearly polarized plane wave is incident on the surface of a slab of magnetized ferrite. This linearly polarized wave may be resolved into positive and negative circularly polarized waves, each of which then propagates through the ferrite with the phase velocity given by equation (20) or (21) as appropriate. At the rear face of the slab, the two circularly polarized waves combine to give a linearly polarized wave whose direction of polarization is that for which the fields of the two circularly polarized waves are in phase. Since the two circularly polarized waves do not travel with the same phase velocities, this direction will differ from that of the polarization of the incident wave. The direction of the polarization of the wave is therefore rotated during transmission through the ferrite.

The angle through which the direction of polarization is shifted depends on the difference between the phase constants for the positive and negative circularly polarized waves. Suppose the incident wave is vertically polarized so that it may be resolved into positive and negative circularly polarized waves so phased that the electric fields are in phase when they are vertical, Fig. 4(a). Now, the phase constant for the positive wave is usually less than that for the negative wave and the positive wave lags at the rear face by $(\beta_- - \beta_+)\, l$ radians on the negative wave, where l is the length of the ferrite slab. Accordingly, when the electric field of the negative wave at the rear face is directed vertically, *OA* in Fig. 4(b), the electric field for the positive wave is directed along *OB*, where the angle *AOB* $= (\beta_- - \beta_+)\, l$ radians. The two waves will have their electric fields in the same direction when these fields have both reached *OC* mid-way between *OA* and *OB*, and this is therefore the direction of polarization of the emerging linearity polarized wave. The angle of rotation of the polarization is therefore:

$$\theta = \tfrac{1}{2}(\beta_- - \beta_+)\, l \quad \ldots\ldots\ldots\ldots \quad (22)$$

and is in the counter-clockwise direction, when looking along the direction of the applied magnetic field, provided β_+ is less than β_-. This phenomenon is known as Faraday rotation, and since the direction of the steady magnetic field determines which is the positive and which the negative circularly polarized wave, it also determines the direction of the rotation of the polarization. The Faraday rotation therefore occurs in the same sense irrespective of the direction of propagation so that if a perfect reflector is placed at the rear face of the slab, the reflected wave which appears at the front face will have its polarization twisted by an angle 2θ with respect to the incident wave. This violates the reciprocity condition so that Faraday rotation is one non-reciprocal effect which may occur in magnetized ferromagnetics. If the results of equations (8), (9), (20) to (22), are combined, the Faraday rotation per unit length can be shown to be:

$$\theta' = \tfrac{1}{2}\gamma M_o(\epsilon/\mu_o)^{1/2} \quad \ldots\ldots\ldots\ldots \quad (23)$$

provided that $\omega_o \ll \omega$ and that $\gamma M_o \ll \mu_o\omega$. Equation (23) is therefore applicable only when the frequency of operation is much larger than the frequency of the electron precession.

Ferromagnetic Resonance

In the above discussion of Faraday rotation, μ and k have been regarded as constants of the medium, but reference to equations (8) and (9) shows that μ and k depend on both the frequency of the alternating fields and the magnetization of the material. Further it appears that both μ and k become infinite if the operating frequency equals the electron precessional frequency. In practice, the damp-

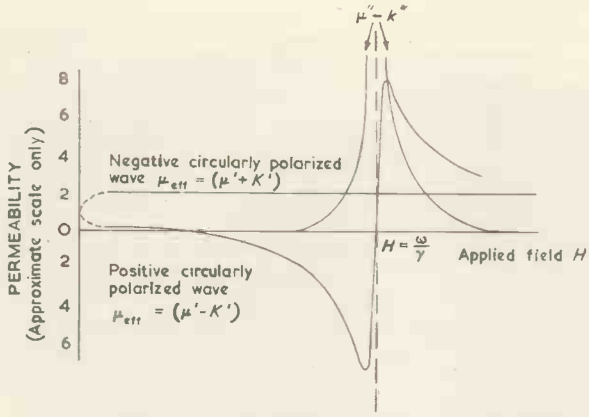

Fig. 5. The behaviour as a function of applied field of the effective permeabilities $\mu-k$ and $\mu+k$

ing forces on the electron spins keep the values finite and the variation of μ and k with frequency has a similar form to the response curve of a tuned circuit. The dependence of these quantities on the applied steady magnetic field follows much the same form because the natural precessional frequency of the electrons is proportional to this field. Of more interest than μ and k themselves are the effective permeabilities for the positive and negative circularly polarized waves and curves showing their dependence on the applied steady magnetic field are shown in Fig. 5. The solid curves are based on the Polder equations, corrected for the damping forces, while the dotted portion takes into account the fact that the material is not saturated at low values of applied magnetic field. The imaginary part of $\mu'' - k''$, contributes to the attenuation of the positive circularly polarized wave.

The type of behaviour shown is known as ferromagnetic resonance and it is clear that the behaviour of waves propagating in a ferrite can be considerably influenced by changing the magnitude of the steady magnetic field. Furthermore, if the magnetic field is adjusted to the peak of the attenuation curve, the positive circularly polarized wave will be strongly absorbed, but the negative circularly polarized wave will be relatively unaffected. This forms the basis of one type of isolator which will be described in Part 2.

The observed variation with applied magnetic field of the properties of a ferrite is seldom so simple as predicted by the theoretical curves in Fig. 5 and additional attenuation peaks occur at low values of applied field and also for values corresponding to the negative wave. These additional peaks complicate the design of waveguide components, but do not lead to any change in the general principles of operation.

General Discussion

The possibility of non-reciprocal propagation in a ferrite arises when a steady magnetic field is applied to the material, this field being applied in the direction of propagation. A linearly polarized plane wave is subject to a rotation of the direction of polarization (Faraday rotation) proportional to the magnetization of the material. The sense of this rotation changes if the direction of the steady field is reversed but is not altered by a reversal of the direction of propagation. If the frequency of the propagated wave equals the precession frequency of the electrons in the material, the positive circularly polarized wave is strongly attenuated, but the negative wave is not affected.

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(To be continued)

Temperature Recording in Creep Testing

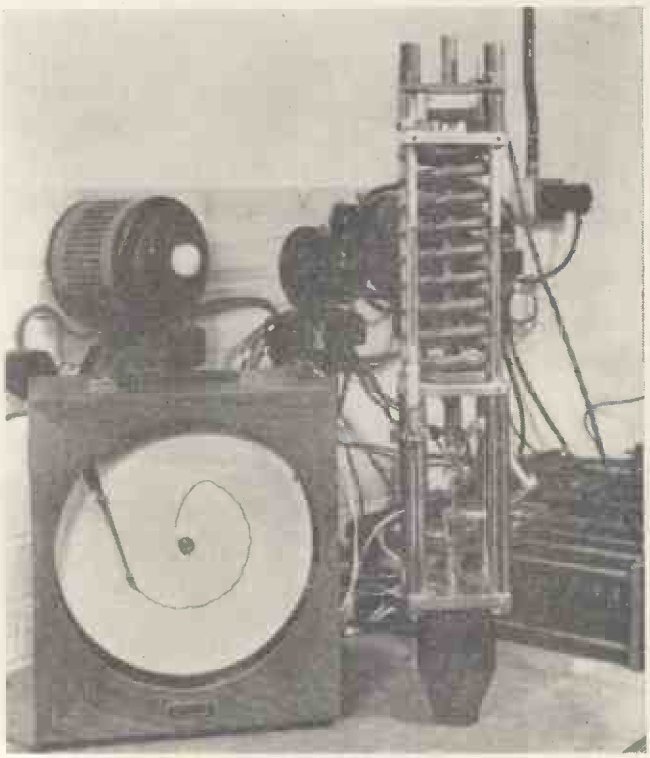
One of the uses of the 'Servograph' Mk. II is as a temperature recorder in conjunction with a creep test loading device in the testing of adhesively bonded lap joints. The combined equipment is shown in the illustration.

Two 1in wide strips of aluminium are bonded together with 'Redux' adhesive, with an overlap of ½in, giving a bonded joint area of ½in². One end of the jointed strips is anchored at the bottom of the machine and the other end to the central tie rod, which is pushed upwards by the spring, compressed to exert, in this case, a pull of 500lb on the joined strip. A shear load of 1 000lb/in² is thus applied to the ½in² bonded area of the joint.

The temperature at the joint is raised by heating the bonded strips above and below the overlap by means of electrical resistance heating elements clamped either side of the strip, the voltage applied to them being regulated by a 'Variac' transformer to maintain, in this case, a joint temperature of 100°C.

An iron-constantan thermocouple generating 5mV per 100°C clipped on at the overlap is used to measure the temperature of the joint. The voltage generated deflects the meter in the 'Servograph' recorder, and this deflexion is sensed by the Fielden capacitance proximity method which, by means of a capacitance bridge and amplifier, controls the servo power which moves the recording pen and plots a trace on the circular recording chart revolving once in 24 hours, any variation in the temperature of the loaded joint being shown on the temperature trace on the recorder chart.

(Fielden Electronics Ltd gratefully acknowledge the co-operation of Aero Research Ltd, Cambridge, by whose kind permission this note is published.)



Current Derived Resistance-Capacitance Oscillators using Junction Transistors

By D. E. Hooper*, B.E.E., and A. E. Jackets*

Oscillator circuits using resistance-capacitance feedback networks are often used, particularly at low frequencies. Most of the circuits discussed in the available literature are based on thermionic valves. The input and output impedances of junction transistor amplifiers differ substantially from those of thermionic valve amplifiers and it is necessary to design new networks in order to obtain stable and predictable operation.

The basic criteria of resistance-capacitance oscillators are to be discussed. Two particular circuits, one employing a 180° phase shift and the other a 0° phase shift, are considered. Most of the circuits use G.E.C. type EW59 transistors.

The oscillators using 180° phase shift networks have the advantage of simplicity and the use of only one transistor, but cannot be used at frequencies in excess of a few kilocycles per second. These oscillators may, however, be used to produce stable outputs having frequencies of a fraction of a cycle per second.

The oscillator using the 0° phase shift network allows a maximum frequency of about 30kc/s to be reached at the expense of an extra transistor and more complicated circuits. Due to the resistance-capacitance coupling within the two stage amplifier, this type of circuit will not function at frequencies of the order of a fraction of a cycle per second.

Subject to certain frequency limitations, which are discussed, and also to some "impedance discontinuity" conditions either of these two circuit types can be designed to give good performance.

THE use of resistance-capacitance circuits has become standard practice in low frequency thermionic valve feedback oscillators; the resistance-capacitance phase shift oscillator, Wein bridge oscillator, and twin-T oscillator are typical examples[1]. In valve circuits, the feedback network is often fed from a low impedance (voltage) source and feeds into a high impedance load. This is usually achieved by feeding the network from the anode or cathode of a triode and returning the output of the network to the high impedance grid circuit. At a given frequency, the loop phase shift becomes zero and if, at this frequency, the voltage gain of the amplifier compensates for the voltage attenuation of the network, oscillation will occur.

Since it is difficult[2] to produce simple transistor amplifiers with high input impedance, low output impedance, and substantial voltage gain, the principles used in thermionic valve amplifiers cannot be used directly in transistor circuits.

Any feedback oscillator consists basically of two components, a frequency sensitive passive network which has inherent losses and an amplifier to compensate for these losses. In general it is difficult to separate the circuit into these two components. However, if "impedance discontinuities" exist between amplifier and network, as discussed later, the two can be considered separately and the design procedure is simplified. The circuits discussed in this article as well as most thermionic valve circuits have been developed on this basis.

For small signal conditions, which determine the threshold of oscillation, the amplifier's performance can be conveniently specified in terms of the four quantities:

Input impedance $(Z_i)_A$
Output impedance $(Z_o)_A$
Current gain G_I
Voltage gain G_V

These quantities are defined in Fig. 1(a) where reference directions of voltage and current are shown. Typical values of the parameters for the three possible single transistor amplifiers are given in Table 1.

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If the passive network has input and output impedances $(Z_i)_N$ and $(Z_o)_N$ respectively, the design of circuits becomes simpler because of impedance discontinuities if one of the

$$G_I = \frac{I_2}{I_1} \text{ With output short-circuited}$$

$$G_V = \frac{V_2}{V_1} \text{ With output open-circuited}$$

$$(Z_i)_A = \frac{V_1}{I_1} \text{ With output open-circuited}$$

$$(Z_o)_A = \frac{-V_2}{I_2} \text{ With input short-circuited}$$

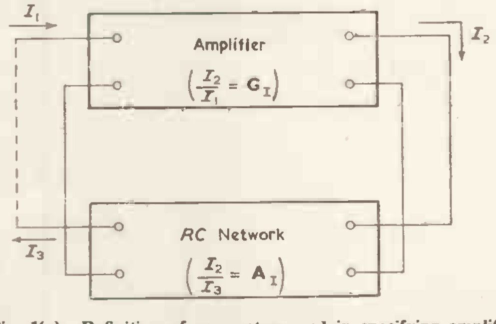


Fig. 1(a). Definition of parameters used in specifying amplifier performance (b). Block diagram of current-operated oscillator

TABLE 1

	$(Z_i)_A$	$(Z_o)_A$	G_I	G_V
Common Emitter	500 Ω	100 kΩ	−30	−500
Common Base	25 Ω	2 MΩ	0·97	500
Common Collector	700 kΩ	50 Ω	25	0·98

following conditions holds:

(a) $(Z_o)_A \gg (Z_i)_N$, $(Z_o)_N \gg (Z_i)_A$ — i.e. the network is fed from a current source and terminated in a short-circuit. The amplifier requires a current gain greater than unity.

(b) $(Z_o)_A \gg (Z_i)_N$, $(Z_o)_N \ll (Z_i)_A$ — i.e. the network is fed from a current source and terminated in an open-circuit. The amplifier requires a high transfer admittance and hence a current gain greater than unity.

(c) $(Z_o)_A \ll (Z_i)_N$, $(Z_o)_N \ll (Z_i)_A$ — i.e. the network is fed from a voltage source and terminated in an open-circuit. The amplifier requires a voltage gain greater than unity.

(d) $(Z_o)_A \ll (Z_i)_N$, $(Z_o)_N \gg (Z_i)_A$ — i.e. the network is fed from a voltage source and terminated in a short-circuit. The amplifier requires a high transfer impedance and hence, a voltage gain greater than unity.

For simple resistance-capacitance networks of the type considered here $(Z_i)_N$ and $(Z_o)_N$ are of the same order. Therefore, from a consideration of the properties of the various possible amplifier circuits (Table 1), and the desirable simplifying conditions [(a), (b), (c) and (d)] it appears that a combination of the common emitter amplifier and a network satisfying condition (a) will produce a useful oscillator circuit. In the discussion that follows this is the arrangement used, except where otherwise stated. A system satisfying condition (c) is usually used in thermionic valve resistance-capacitance oscillators.

The simplest transistor oscillator may then be considered as a 'current-operated' circuit of the type shown in Fig. 1(b), whereas the corresponding thermionic valve oscillators are best considered as 'voltage-operated'. The transistor oscillator may be considered as the dual of the thermionic valve oscillator[3].

The criterion for oscillation of the circuit shown in Fig. 1(b) is that the locus of the current loop gain vector, $\mathbf{G_L}$, given by

$$\mathbf{G_L} = \frac{\mathbf{G_I}}{\mathbf{A_I}} \quad \text{(1)}$$

encloses the point $1 + j0$ as the frequency varies from 0 to ∞ where $\mathbf{A_I}$ is the current attenuation in the network[4].

Now the amplifier current gain, $\mathbf{G_I}$, and the network current attenuation, $\mathbf{A_I}$, can be expressed in terms of their moduli and phase-angles as follows:

$$\mathbf{G_I} = G_I \angle\phi_A \quad \text{(2a)}$$

$$\mathbf{A_I} = A_I \angle\phi_N \quad \text{(2b)}$$

and the limiting criterion for oscillation becomes:

$$G_I / A_I = 1 \quad \text{(3a)}$$

$$\phi_A = \phi_N \quad \text{(3b)}$$

Since ϕ_A is either 180° or 360° at low frequencies (depending on the number of common emitter stages in the amplifier) it is convenient to consider separately networks giving 180° phase shift and networks giving 0° phase shift.

Resistance-Capacitance Oscillators Using 180° Phase Shift Networks

(a) Six Element Ladder Network

The network used is similar to the conventional resistance-capacitance ladder network used in thermionic valve oscillators[5], and, as with valve oscillators, some advantage is obtained by varying the impedance level along the ladder. It is possible to determine the optimum network attenuation[6] but, for convenience, this analysis assumes that there is a linear "tapering" of the impedance level along the ladder[7].

It is useful to consider the effect of various values of n, the increase in admittance per section as defined in Fig. 2(a), on the design of the oscillator circuit.

It may be shown that:

$$\mathbf{A_I} = (I_2/I_3) = [1 - (3 + (2/n))\,\rho\,(\omega_o/\omega)^2] + j\rho^{3/2}[(\omega_o/\omega)^3 - (\omega_o/\omega)] \quad \text{(4a)}$$

where:

$$\rho = [3 + (2/n) + (1/n^2)] \quad \text{(4b)}$$

$$\omega_o = (1/\sqrt{\rho}\,.\,RC) \quad \text{(4c)}$$

$$\text{and } \omega = 2\pi f \quad \text{(4d)}$$

where f is the frequency.

Now, when $\omega = \omega_o$:

$$(\mathbf{A_I})\omega_o = [1 - (3 + (2/n))\,\rho] \angle 0°$$
$$= [(3 + (2/n))\,\rho - 1] \angle 180°$$

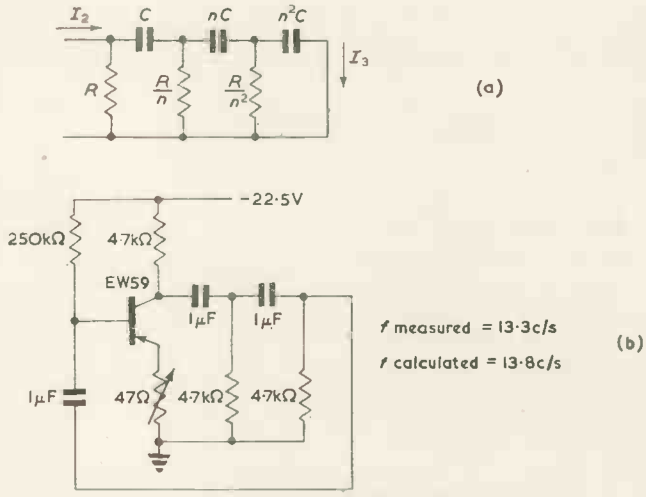


Fig. 2(a). Six element ladder network
(b). Low frequency oscillator using six element ladder

Fig. 3(a) shows the variation of the magnitude of $(\mathbf{A_I})\omega_o$ with n. As n varies from 1 to ∞, the magnitude of $(\mathbf{A_I})\omega_o$ is reduced from 29 to 8.

From the point of view of gain requirements, n should be as high as possible. In practice other factors set an upper limit to the value of n and it is necessary to consider its effect on these factors.

The frequency, f_o at which $\phi_N = 180°$ is given by:

$$f_o = (1/2\pi\sqrt{\rho})\,(1/RC) = K\,(1/RC) \quad \text{(5)}$$

where the factor K, plotted as a function of n in Fig. 3(b), shows that the frequency is not highly dependent on n.

The upper value of n is limited by the minimum reactance of the capacitor of value n^2C. In order that condition (a) holds $1/\omega_o n^2 C$ should be about ten times greater in magnitude than the amplifier input impedance:

i.e.:

$$(1/\omega_o n^2 C) \geqslant 10\,|(Z_i)_A| \quad \text{(6)}$$

In practice this value will be about 5 to 10kΩ.

From equations 4(b) and 4(c), the factor $1/\omega_o n^2 C$ is related to R, as follows:

$$(1/\omega_o n^2 C) = \sqrt{(\rho R)}/n^2 = \sqrt{[(5/n^4) + (2/n^5) + (1/n^6)]}R$$
$$= S.R. \quad \text{(7)}$$

This factor S is plotted as a function of n in Fig. 3(c). The value of R can be determined from the practical requirement that:

$$|(Z_i)_N| \leqslant (1/10)\ |(Z_o)_A| \quad \ldots\ldots\ldots\ldots \quad (8)$$

and the relationship between the magnitude of $(Z_i)_N$ and R is given by the expression:

$$|(Z_i)_N| = \left[\frac{(2 + (1/n))^2 + 4\rho(1 + (1/n))^2}{\rho(3 - (1/\rho) + (2/n))^2}\right]^{\frac{1}{2}} R$$

$$= U.R. \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (9)$$

The factor U is plotted as a function of n in Fig. 3(d) and has a value almost independent of n.

A study of the effect of n on the values of S and U and the requirements given by inequalities (6) and (8) show that it is not generally advisable to use a value of n greater than about 1·5.

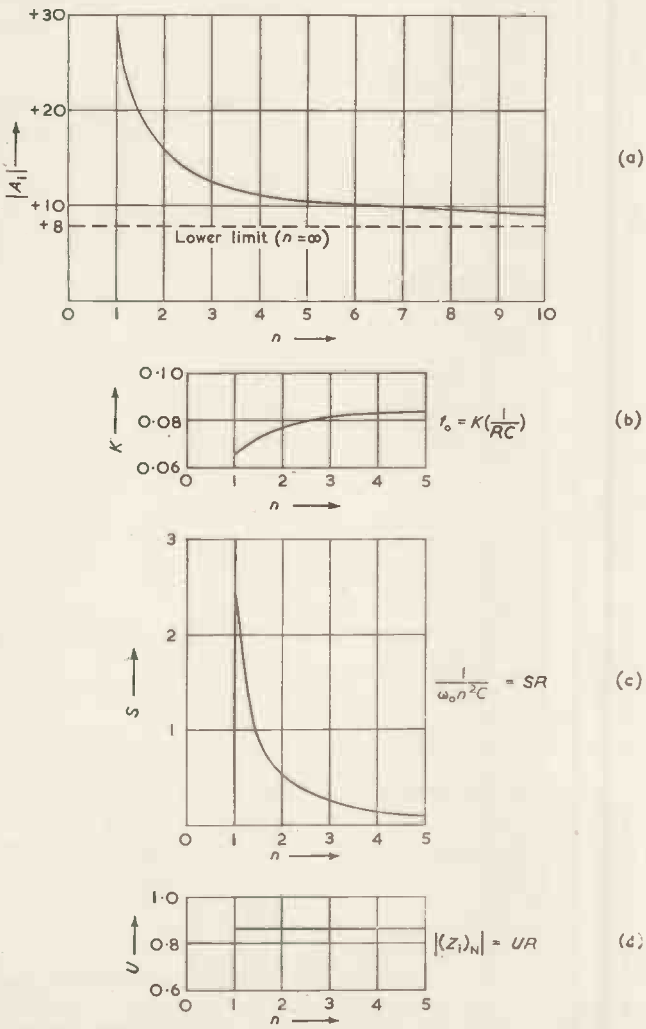


Fig. 3(a). Variation of magnitude of current attenuation in six element ladder network with factor n

(b). Variation of K with n for six element ladder network

(c). Variation of S with n for six element ladder network

(d). Variation of U with n for six element ladder network

A practical circuit of a very low frequency oscillator using this type of network ($n = 1$) is shown in Fig. 2(b). This circuit shows how the first resistor of the ladder can be used as the transistor collector load resistor. The variable emitter resistance can be used as an elementary form of amplitude control. For precision performance a low power thermistor or some form of automatic gain control would be desirable, but the simple circuit shown exhibits reasonable amplitude stability and produces little distortion. Since the impedance level at the collector is high a common collector buffer stage will be necessary when the oscillator is to work into a low impedance.

If the capacitors in the circuit of Fig. 2(b) are reduced in value to 1 330pF the circuit oscillates at 1 040c/s, but on further reducing the value of these components, the particular circuit used was found not to oscillate. This upper frequency limitation is a general feature of this type of circuit and is discussed in more detail later.

(b) Eight Element Ladder Network

The above discussion shows that it is generally impracticable to use the six-element network with n having a value greater than about 1·5. The eight-element network shown in Fig. 4(a) has approximately the same attenuation for 180° phase shift as does the six-element network with $n = 1·5$. There is little point in "tapering" this network due to the necessity of satisfying condition (a) and results are given for the untapered case only.

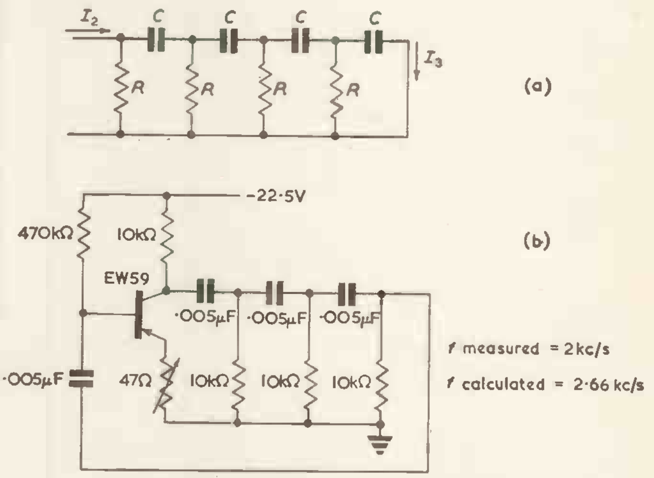


Fig. 4(a). Eight element ladder network
(b). Oscillator using eight element ladder network

It may be shown that:

$$A_I = (I_2/I_3) = [1 + \rho^2(\omega_o/\omega)^4 - 15\rho(\omega_o/\omega)^2] +$$
$$j\rho^{1/2}[7\rho(\omega_o/\omega)^3 - 10(\omega_o/\omega)] \quad \ldots\ldots \quad (10a)$$

when $\rho = (10/7)$ $\ldots\ldots\ldots\ldots$ (10b)

$$\omega_o = \sqrt{0·7}\,.\,(1/RC) \quad \ldots\ldots \quad (10c)$$

and $\omega = 2\pi f$ $\ldots\ldots\ldots\ldots$ (10d)

where f is the frequency.

Now:

$$f_o = (\sqrt{0·7}/2\pi)\,.\,(1/RC) = 0·133\,(1/RC) \quad \ldots \quad (11)$$

Hence for the same resistance-capacitance product the frequency corresponding to 180° phase shift is slightly greater for this network than for the six element network. However, the input impedance of the two networks is of the same order and to satisfy condition (a) it is necessary that:

$$R \leqslant \frac{|(Z_o)_A|}{10} \quad \ldots\ldots\ldots\ldots \quad (12)$$

Again when $f = f_o$:

$$A_I = 18·4\ \angle 180° \quad \ldots\ldots\ldots \quad (13)$$

Thus, at the expense of two extra components, this circuit is comparable with the "tapered" six element ladder with $n = 1·5$. A typical oscillator using this circuit is shown in Fig. 4(b). It is has a slight advantage in the maximum frequency obtainable for the reasons given later.

(c) Limitations of Maximum Oscillation Frequency

The frequency cut-off that occurs in phase shift oscillators of the type described above is due to the amplifier phase shift, ϕ_A, becoming less than 180° at the higher frequencies. In the limiting case, when oscillations will be just sustained, equation 3(b) shows that ϕ_N must adjust

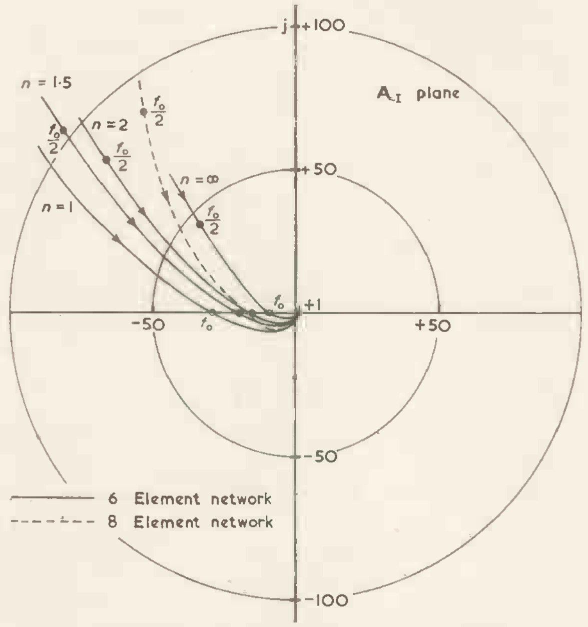
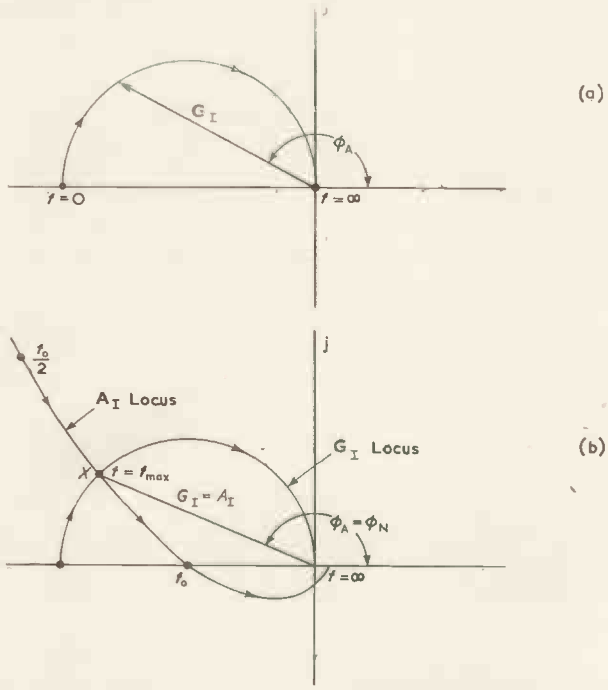


Fig. 5. Loci of A_I for 180° phase shift networks

Fig. 6. Frequency cut-off in phase shift oscillators



itself to the value ϕ_A. When ϕ_N falls below 180° the network attenuation increases above its value at $\phi_N = 180°$ and, if f_o for the network is increased by a slight amount, G_I becomes less than A_I and oscillations will no longer occur.

The loci of A_I, the attenuation of the network as a function of frequency as given by equation (4a) (for the six-element network) and equation (10a) (for the eight-element network) are plotted in Fig. 5 and show the increase in A_I as ϕ_N falls below 180°.

The mechanism of frequency cut-off can be explained qualitatively by reference to Fig. 6. Fig. 6(a) serves to describe approximately the behaviour of a common emitter junction transistor amplifier over the frequency range from 0 to ∞. It is emphasized that this response locus is very much idealized, particularly at very high frequencies, but it does serve to explain qualitatively the frequency cut-off[8].

In Fig. 6(b) a typical A_I locus is superimposed on the G_I locus assuming the limiting condition for oscillations. In this limiting condition the system adjusts itself to oscillate at a frequency given by $f = f_{max}$, where f_{max} is the frequency giving an amplifier phase shift, ϕ_A, equal to the network phase shift, ϕ_N. The frequency f_{max} is lower than f_o. Increasing f_o by redesigning the phase shift network makes the criterion for oscillation, given by equations (3a) and (3b), impossible to obtain. With EW59 type transistors, having α_{OB} cut-off frequencies of about 35kc/s, the practical value of f_{max} is about 1kc/s for the six-element network ($n=1$) and about 2kc/s for the eight-element network.

If these networks are modified by interchanging the resistors and capacitors, the value of A_I is reduced as ϕ_N falls below 180° with increasing frequency. In this case too the oscillation frequency is lower than f_o but, for a given transistor, the value of f_{max} is much higher for the

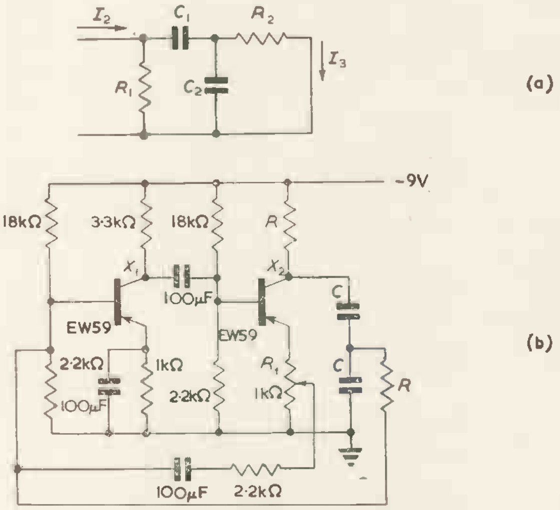


Fig. 7(a). 0° phase shift network
(b). RC oscillator using 0° phase shifting network

modified networks. In practice it is found that a value of f_{max} of about 30kc/s can be achieved with EW59 type transistors.

However the modified networks are of limited practical use because of the difficulty in providing d.c. supplies to the transistor. Furthermore the oscillation frequency becomes very dependent on the transistor cut-off frequency at frequencies in excess of a few kilocycles and thus the circuit is not "designable" at these frequencies. In fact it is possible to arrange a circuit which will not oscillate at low frequencies but at the higher frequencies where A_I is reduced, oscillation will occur. For these reasons the two-transistor circuit described later is to be preferred for use at frequencies between a few hundred cycles and 30kc/s and the simple circuits of Figs. 2(b) and 4(b) are preferred for use at the lower frequencies where the actual oscillation frequency agrees very well with the theoretically predicted value (as shown in Table 2 for a typical circuit using standard 10 per cent components).

Resistance-Capacitance Oscillators Using 0° Phase Shift Networks

A current controlled network, which is related to net-

TABLE 2

R (kΩ)	C (μF)	f actual (c/s)	f calculated (c/s)
10	16	0·42	0·406
4·7	8	1·64	1·72
4·7	0·1	121	138
4·7	0·02	680	690
4·7	0·0167	800	825
4·7	0·0133	1050	1050

TABLE 3

R (kΩ)	C (μF)	f actual (c/s)	f calculated (c/s)
3	2	28·3	26·5
3	0·5	104	106
3	0·01	5280	5310
3	0·005	10600	10600
1	0·005	27000	31900
1	0·000047	90000	339000

works that have been described for thermionic valve oscillators[9,10], is shown in Fig. 7(a).

The current attenuation is given by:

$$\mathbf{A}_I = (R_2/C_1)\left\{\left(C_1/R_2 + \frac{C_1 + C_2}{R_1}\right) + j\left(\frac{1}{\omega R_1 R_2} - \omega C_1 C_2\right)\right\} \quad \ldots\ldots\ldots\ldots\ldots (14)$$

It is often convenient to have:

$$R_1 = R_2 = R \text{ and } C_1 = C_2 = C$$

In this case equation (14) becomes:

$$\mathbf{A}_I = 3 + j\,[(\omega_o/\omega) - (\omega/\omega_o)] = 3 + j\,[(f_o/f) - (f/f_o)] \quad \ldots\ldots\ldots\ldots\ldots (15a)$$

where:

$$f_o = (\omega_o/2\pi) = (1/2\pi RC) \quad \ldots\ldots\ldots (15a)$$

when $f = f_o$:

$$(\mathbf{A}_I)_{f_o} = 3\angle 0^\circ \quad \ldots\ldots\ldots\ldots\ldots (16)$$

The value of G_I required for oscillation is thus only 3 but, since ϕ_A must be 360°, two common emitter stages are required in the amplifier. Feedback can be applied in the amplifier circuit to reduce the overall current gain, and, at the same time, decrease the input impedance and increase the output impedance to values of about 5Ω and 2MΩ respectively. The two-stage low current gain amplifier then becomes an almost ideal current amplifier.

When $f = f_o$:

$$(Z_i)_N = 0{\cdot}748R \quad \ldots\ldots\ldots\ldots\ldots (17)$$

and for the required "impedance discontinuity" it is necessary that:

$$0{\cdot}748R \leqslant (1/10)\ |(Z_o)_A| \quad \ldots\ldots\ldots (18a)$$

and:

$$R \geqslant 10\ |(Z_i)_A| \quad \ldots\ldots\ldots\ldots (18b)$$

where the factor 10 is again introduced as a useful practical value.

A typical circuit is shown in Fig. 7(b), while the effect of using various values of resistance and capacitance in this circuit is shown by the results given in Table 3.

In Table 3 the last result represents the order of the maximum frequency obtainable with this circuit and it can be noted that, in this case, the agreement between actual and predicted frequency is very poor. However, for frequencies below about 30kc/s the circuit will oscillate at a frequency close to the predicted value.

The initial divergence from the theoretical values of frequency, at about 30kc/s, and subsequent frequency cut-off, at about 100kc/s, is due to the effect of phase shift and loss of current gain in the amplifier at high frequencies. Due to a combination of overall amplifier feedback and internal feedback within the transistors the analysis of the high frequency performance of this circuit is difficult.

A low impedance output may be obtained from the emitter of X_2 and the feedback potentiometer R_f, which controls the effective current gain of the two-stage amplifier, may be adjusted to ensure that the circuit is just on the point of oscillation.

It is possible to use the network of Fig. 7(a) in conjunction with a single common base point-contact transistor amplifier having a current gain in excess of 3. This circuit is not of much practical use as the output impedance of the amplifier is much lower than that of the comparable junction transistor unit and it is difficult to satisfy the "impedance discontinuity" conditions.

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Cosmic Radiation Measurements

Electronic calculating and accounting equipment is playing an important part in the study of cosmic radiation, the subject of an important physical experiment now being conducted at the United Kingdom Atomic Energy Research Establishment, Harwell.

The constant stream of cosmic rays travelling through space represent a problem which must be solved before man and machine venture into outer space. Biologists, too are interested in cosmic radiation as it is felt that the rays have a bearing on the evolution of life itself on earth.

Hollerith punched card accounting machines, developed by The British Tabulating Machine Company Limited, are recording the velocity and distribution of these days.

Due to nuclear explosions in outer space, the earth is subjected to continuous bombardment by cosmic rays which are, as yet, the only material link between our planet and the universe beyond. Scientists are anxious to determine the origin of the rays and the source of their energy, believing that research in this project will result in increased knowledge of the workings of the universe.

The Harwell experiment is being conducted, with the aid of Hollerith office machinery, to collate data on cosmic radiation. It represents the first example of complete automation in this field.

Special detector equipment, in the form of 91 Geiger-Muller counters disposed at 100 yard intervals in an equilateral triangle, measure the strength and intensity of these rays as they reach the earth, setting up electrical impulses which are relayed through electronic circuits to a Hollerith Automatic Punch. The intensity of each cosmic ray shower is recorded by the number of holes punched in the Hollerith card, each hole representing an impulse from one particular counter unit.

Card recordings are so arranged that it is possible to determine the exact points within the triangle at which the cosmic rays registered. Other information tabulated includes the time, barometric pressure and air temperature of each recording.

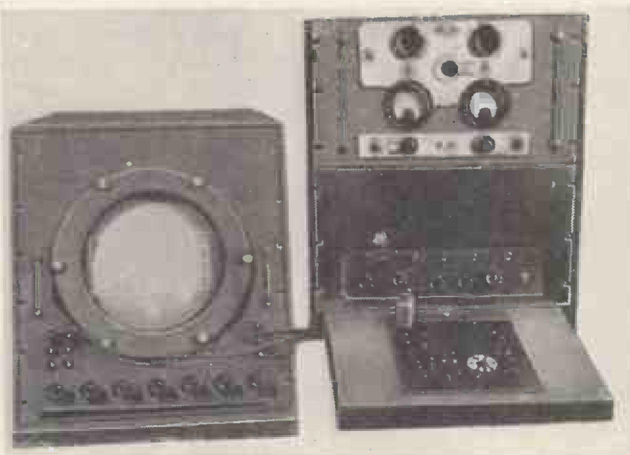
Once created, the punched cards not only give a visual picture of the pattern of a cosmic ray shower but can be sorted and analysed automatically at any time, with the aid of other Hollerith machinery, to indicate the conditions associated with such phenomena.

A Valve Curve Tracer

By R. H. James*, A.M.I.E.E.

The instrument which is described has been designed and constructed for use as a visual training aid at the Civil Aviation Signals Training Establishment of the Ministry of Transport and Civil Aviation. It displays, on a long persistence cathode-ray tube, families of mutual and anode characteristic curves together with calibration graph lines and a variable load line. In the case of pentode valves, screen grid and suppressor grid characteristics may be displayed.

THE basic circuits for tracing valve characteristics on a cathode-ray tube are well known, and other instruments have been constructed for this purpose. The main new feature of the one described is the display of calibration lines (graph paper), which improves the overall accuracy. All signal and calibration voltages applied to the indicator are developed across close tolerance resistors and all d.c. power supplies are stabilized. Valve curve voltages and calibration voltages pass through the same amplifying channels so that any distortion in these channels and in the cathode-ray tube deflecting system affects both valve curves and calibration lines equally. Accuracy is therefore not affected by drift or variations of amplifier gain, and the display may be adjusted to any convenient size and shape without affecting the calibration.



The complete valve curve tracer

Each display consists of 25 lines and curves traced sequentially on a long persistence cathode-ray tube screen as the result of the switching action of a four-bank uniselector. Fourteen positions of the uniselector are used to connect appropriate voltages to X and Y c.r.t. amplifiers to trace the graph lines, the 15th position is used to produce the variable position load line and the remaining 10 positions are used to obtain the valve curves.

Switching transients which would otherwise spoil the display are prevented from doing so by a blanking voltage applied to the c.r.t. grid. Additionally, a suitably shaped waveform is applied to the c.r.t. grid to make the trace, which would otherwise be bright at the ends and faint in the middle due to the sinusoidally derived sweep signals, more uniformly brilliant.

Types of Characteristics Which May be Displayed

The vertical axis of the c.r.t. screen may represent either anode current, or screen grid current. Four switched ranges permit the graph lines to represent 0 to 10mA up to 0 to 100mA.

The horizontal axis to the right of the centre vertical line may represent either anode voltage or screen grid voltage, the scale being fixed at 0 to 300V.

The horizontal axis to the left of the centre vertical line may represent either control grid voltage or suppressor grid voltage. There are four switched ranges from 0 to −4V to to 0 to −40V.

Valve curves at the left-hand side of the display are therefore "mutual" characteristic curves and at the right-hand side anode characteristic curves.

Voltage increments on the anode or screen grid, for the left-hand curves, are 50V, from 50V to 250V.

Voltage increments on the control grid or suppressor grid for the right-hand curves are switched by a range switch that also alters the scale of the horizontal axis to the left of the centre line. The voltage increments are the same as the increments of the scale to the left of the centre vertical line. Thus if the grid voltage range switch is placed in the 4V position, the vertical graph lines to the left of the centre line represent −1, −2, −3, and −4V, and the grid voltages corresponding to the five anode characteristic curves are 0, −1, −2, −3, and −4V.

When screen or anode characteristics of pentode and tetrode valves are being plotted, the electrode whose characteristics are not being plotted is connected to a stabilized supply that may be set to any value between 0 and 250V, and the effect of varying this voltage may be observed.

Similarly, when control grid or suppressor grid characteristics are being plotted, the electrode whose characteristics are not being plotted is connected to a calibrated source that may be varied between 0 and −40V and the effect of varying this voltage may be observed.

Provision is also made for the manual, instead of automatic, variation of anode voltage applied to the valve under examination. Valve curves can then be 'plotted' directly on to the cathode-ray tube, and the position of the spot correlated with the readings of calibration controls and built-in meters, for demonstration purposes.

Examples of the types of characteristics that may be displayed are shown in Fig. 1.

Circuit Details

A complete circuit is shown in Fig. 2 and simplified circuits in Figs. 3 to 14.

Calibration Lines

During the first 15 positions of the uniselector, when the graph paper and load line are being traced, the valve under

* *Ministry of Transport and Civil Aviation.*

test is biased to -40V on its grid and has a low voltage (50V) on its anode.

The first 5 positions of the uniselector are used to produce the horizontal lines of the graph paper. A simplified circuit showing how this is done is given in Fig. 3. The X amplifier for the c.r.t. is supplied with an alternating voltage of 8V peak-to-peak obtained from a potentiometer connected across a 6·3V a.c. supply. The Y amplifier is supplied sequentially with direct voltages of -4, -3, -2, -1 and 0V. Both amplifiers are d.c. amplifiers and the Y amplifier produces vertical deflexion upwards for negative input voltages.

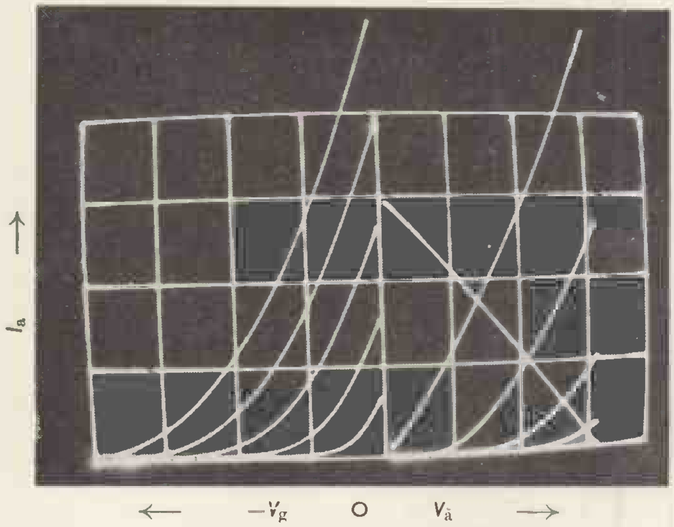


Fig. 1(a). Characteristics of 6C4 (CV133)

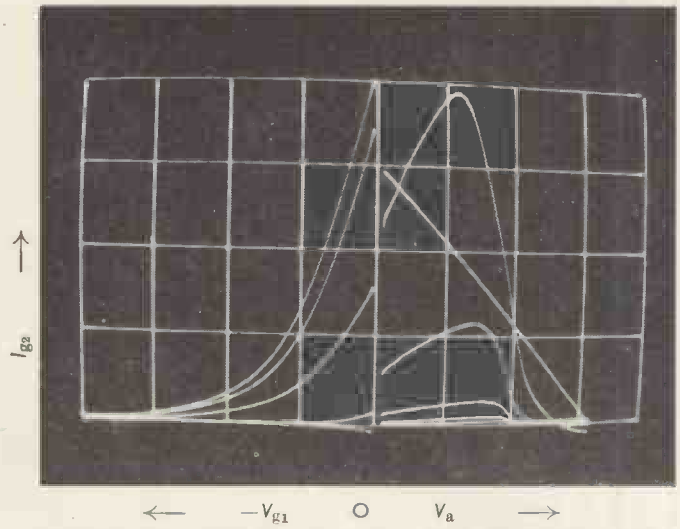


Fig. 1(b). "Screen-grid" valve characteristics [EF91 (CV138) with g_2 and g_3 strapped]

A similar method is used, employing positions 6 to 14 of the uniselector, to obtain vertical graph lines. Fig. 4 is a simplified circuit. In this case the Y amplifier is supplied with -2V d.c. having 4V peak-to-peak a.c. superimposed. This produces a vertical line which lies between the horizontal lines previously traced. Successive direct voltages applied to the X amplifier move this line from the extreme left-hand end of the horizontal lines to the extreme right of the horizontal lines in nine steps.

Load Line

Position 15 of the uniselector is used for the load line. In this position the h.t. supply is swept sinusoidally between 0 and $+300$V, and from a potentiometer at the bottom end of a divider chain a variable amplitude portion of this voltage is fed to the X amplifier. Simultaneously a variable voltage obtained from the source used for vertical lines is applied to the Y amplifier, and an adjustable load line is thus traced on the c.r.t. screen. This arrangement is illustrated in Fig. 5.

Valve Curves

In positions 16 to 20 of the uniselector, anode characteristic curves are traced (see Fig. 6). The anode of the valve under test is supplied with a sinusoidally varying voltage of 0 to 300V approximately and the grid voltage is stepped from 0 negatively according to the grid voltage range selected. A voltage proportional to the anode voltage is obtained from the potential divider across the h.t. supply and applied to the X amplifier, and a voltage proportional to the anode current is obtained from a resistor in the cathode return circuit and applied to the Y amplifier. As the grid voltage changes in steps a family of five I_a/V_a curves is produced. The anode current range switch alters the value of cathode circuit return resistor and this is equivalent to altering the scale of the "graph paper" Y axis.

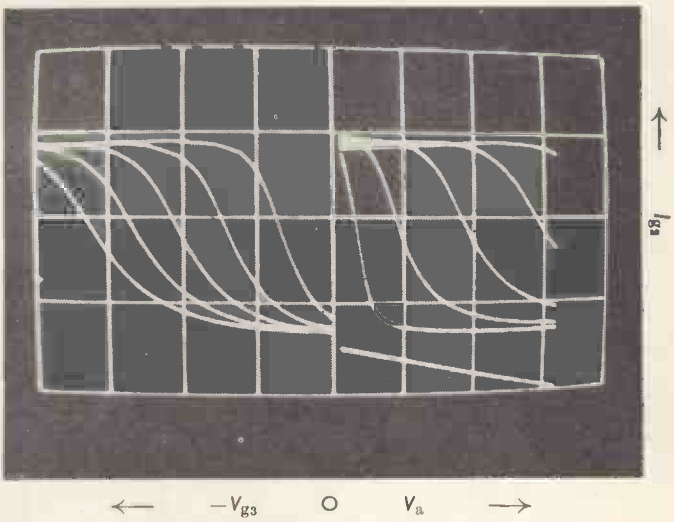


Fig. 1(c). Suppressor grid characteristics of EF91 (CV138)

Mutual Characteristic Curves

These are produced when the uniselector is in positions 21-25 (see Fig. 7). The anode voltage of the valve under test is increased in 5 steps of 50V from 50V to 250V. The grid is supplied with a sinusoidal voltage varying between 0 and a negative value depending on the position of the grid voltage range switch.

The Y amplifier is supplied with a voltage proportional to anode current as before, and the X amplifier with a voltage directly proportional to the instantaneous grid voltage. A family of five I_a/V_g curves is thus produced on the c.r.t. screen as the anode voltage changes in steps.

Additional Facilities

An "anode current-screen current" switch is provided on the power unit (see Fig. 8). This switch connects the current monitoring resistor in either the anode supply or screen supply return circuit so that the vertical axis of the displayed curves may be either anode current or screen current. The supply which is not being monitored is returned directly from earth.

A 'function' switch is fitted on the control unit chassis. This switch enables pentode valves to be connected in

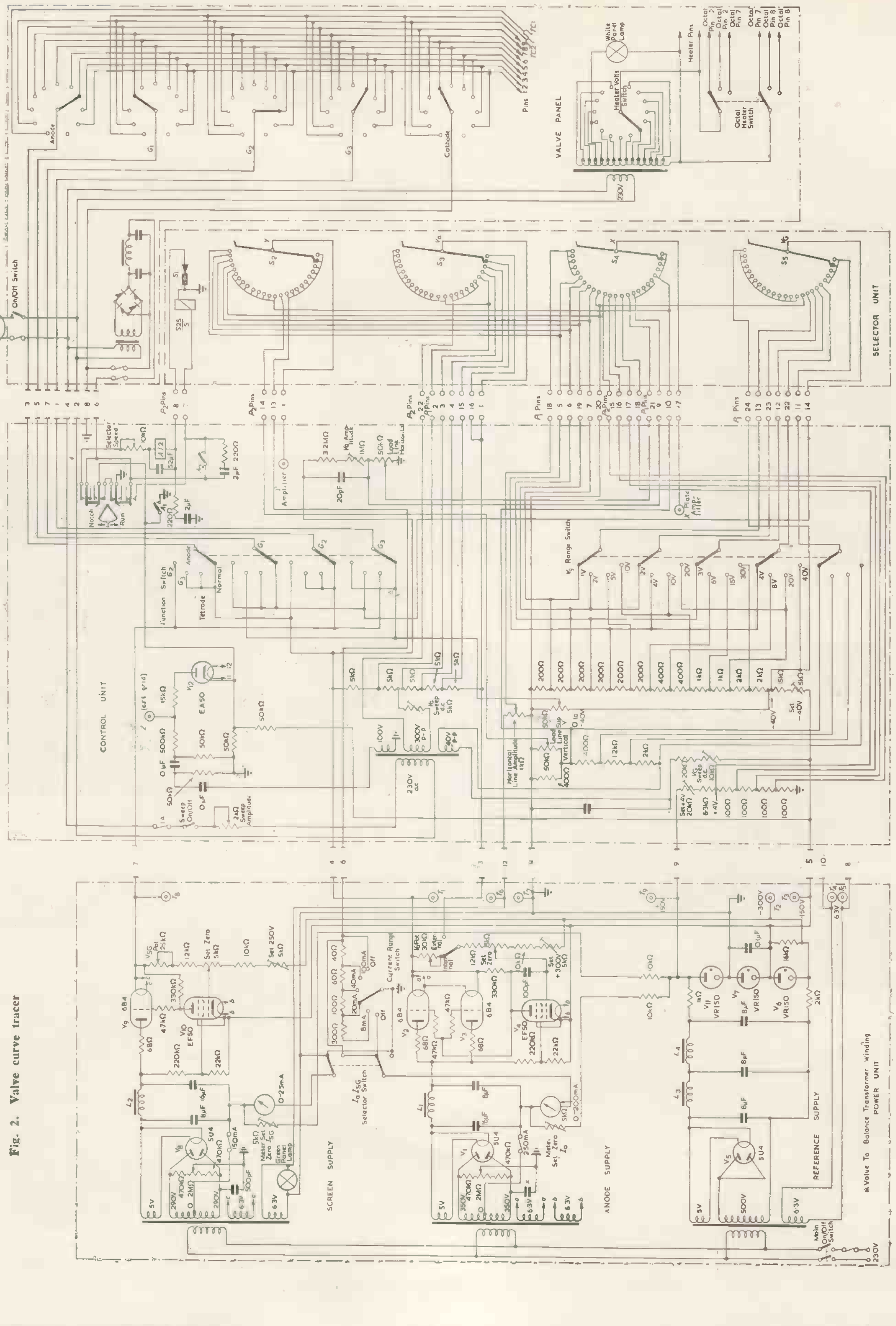


Fig. 2. Valve curve tracer

various ways for displaying characteristics other than I_a/V_a and I_a/V_g (see Fig. 9). The first position of this switch is called 'normal' and I_a/V_a, I_a/V_g or I_{g2}/V_a, I_{g2}/V_g curves may be displayed by operating the anode current-screen current switch.

The second position connects screen and suppressor grids together for demonstrating screen grid valve characteristics as in the preceding paragraph.

In the third position the control grid and suppressor grid connexions are interchanged so that curves of I_a/V_a, I_a/V_{g3} or I_{g2}/V_a, I_{g2}/V_{g3} are plotted.

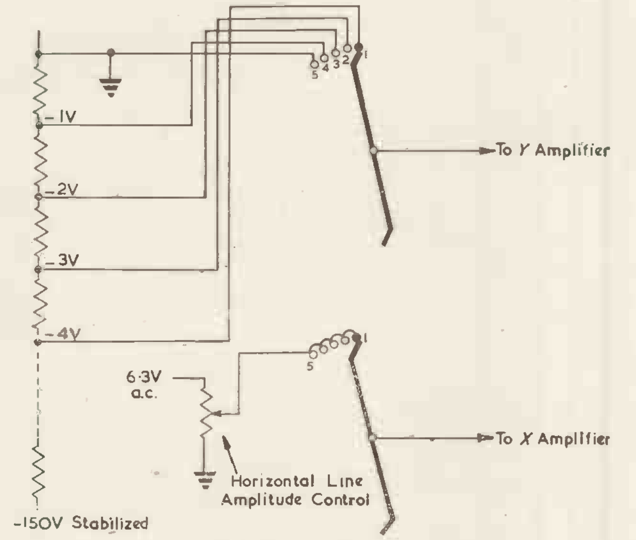


Fig. 3. Production of horizontal graph lines

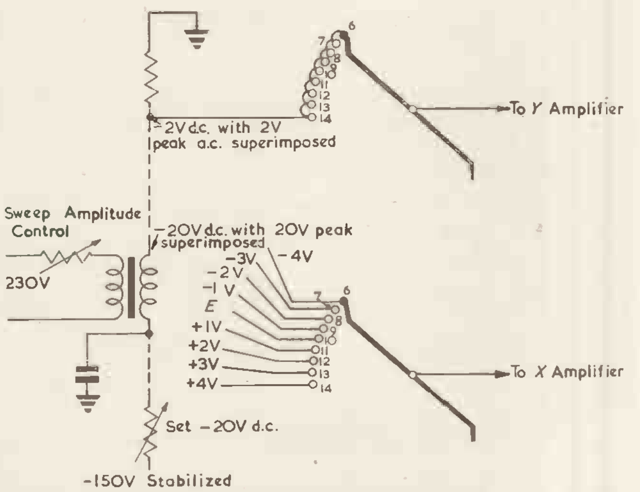


Fig. 4. Production of vertical graph lines

The fourth position connects the anode supply to the screen grid, and vice versa, enabling curves of I_a/V_{g2}, and I_a/V_g or I_{g2}/V_{g2} and I_{g2}/V_g to be displayed.

Valve Panel

A standard type of valve panel with built-in heater transformer is used. This may be set up for all normal types of receiver valves.

Switching and Blanking

The uniselector which performs the switching is operated in a simple control circuit illustrated in Fig. 10. The uniselector may be stationary in any one of its 25 positions, 'notched' one position at a time by pushing the control key up or run continuously at a variable speed when the key is pushed downwards.

Blanking of the c.r.t. trace is arranged via relay contact A_1 which causes the application of a cut-off bias to the c.r.t. grid during stepping of the uniselector. Brilliance of the c.r.t. trace is controlled by varying the cathode bias.

During the time that the display is visible the velocity of the spot on the c.r.t. in most cases varies approximately sinusoidally since on at least one axis a sinusoidal deflexion

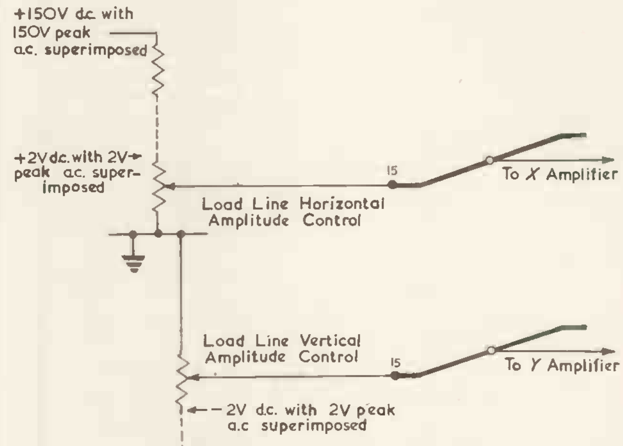


Fig. 5. Production of variable position load line

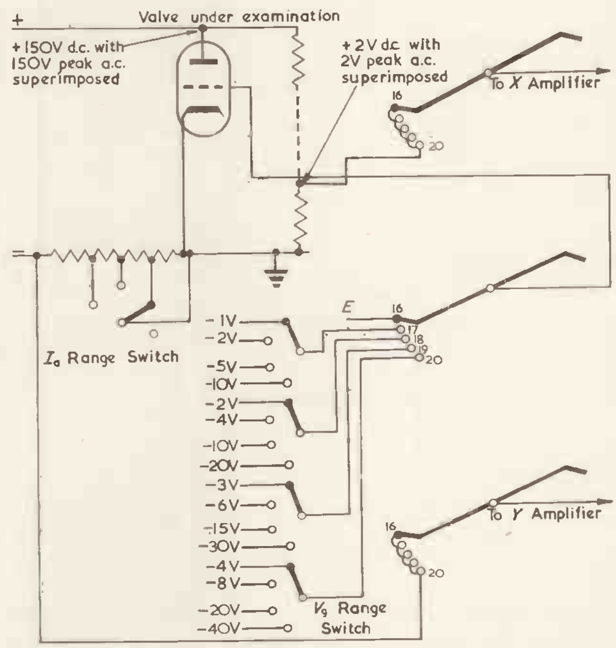


Fig. 6. Production of anode characteristic curves

signal is being applied. This means that the spot is travelling slowly at each end of the trace and relatively quickly at the centre. Consequently, the ends tend to appear bright and the centre fainter. To combat this, the circuit of V_{12} has been included. A 50c/s voltage is applied through a phase shift network to produce a voltage 90° out of phase with the sweep voltage. This is then rectified in a circuit which by virtue of R_1 leaves a bow on the top of the waveform. This waveform is then applied to the c.r.t. grid via an amplitude control and makes the trace brilliance acceptably even, suppressing return sweeps. The waveforms are illustrated in Fig. 11.

Power Supplies

Three separate stabilized supplies are used, one for the anode, one for the screen and one for the grid and reference potentials.

The first two are versions of the well-known series-shunt regulator illustrated in Fig. 12.

It can be shown that with suitable design the output voltage is directly proportional to R_1 and this is the method by which the anode voltage is controlled. When variation of anode potential is required the uniselector switches resistors of different value into the position occupied by R_1 in Fig. 12.

A variation of this system which is believed to be novel is used to obtain the 0 to 300V sinusoidally varying anode voltage for the I_a/V_a curves. This is shown in Fig. 13. In this case the resistor R_1 of Fig. 12 is replaced by a resistor and the secondary winding of a transformer whose total resistance makes the output voltage 150V when no a.c. is applied to the transformer primary winding. However, when the transformer is energized its secondary winding virtually makes R_1 vary sinusoidally between the values required for 0 and 300V and the output voltage varies accordingly.

An 'Internal-External' switch on the power unit makes it possible for the anode voltage to be varied manually by substituting a calibrated potentiometer for the uniselector switched resistors when the switch is put in the Internal position. This permits demonstration in slow motion of the way the curves are traced and facilitates the checking of calibration.

A resistor is included in the negative return to the rectifier circuit in order to obtain a voltage proportional to the instantaneous anode current in the valve under test. Currents flowing in the regulator circuit also flow through this resistor, but these are balanced out by a "set zero" circuit illustrated in Fig. 14. Current flowing in the shunt regulator valve varies with output voltage, but the variation is reduced to negligible proportions by connecting the anode load of the shunt regulator to the output terminal of the regulator, and by making this anode load resistor high. Additionally it is necessary to balance the two halves of the secondary winding of the h.t. transformer with respect to earth so as to prevent unbalanced alternating current flowing in the current monitoring resistor.

The basic circuit of the screen supply is as shown in Fig. 12. In this case a calibrated potentiometer in position

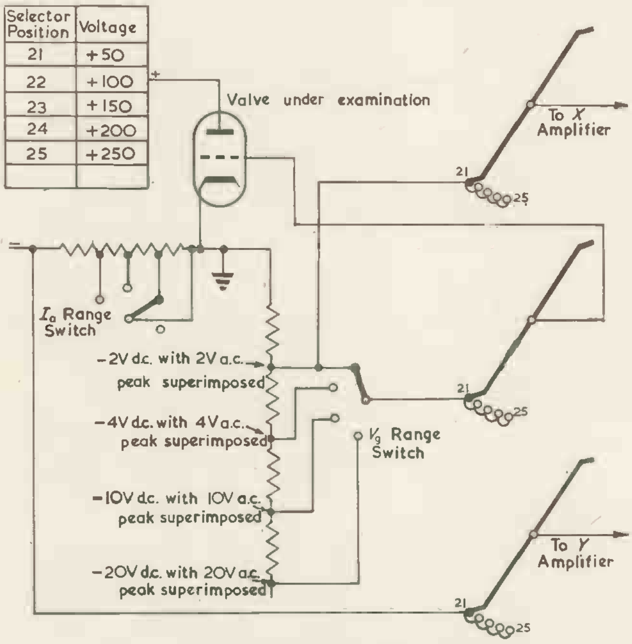


Fig. 7. Production of mutual characteristic curves

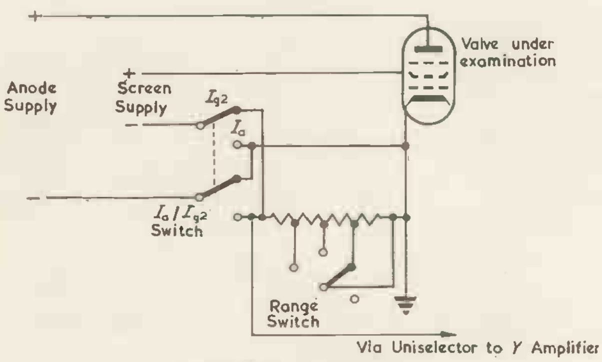


Fig. 8. I_a/I_{g_2} switching circuit

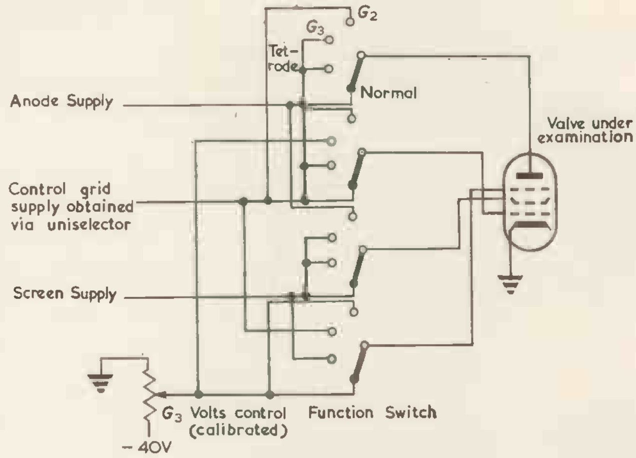


Fig. 9. Circuit of function switch

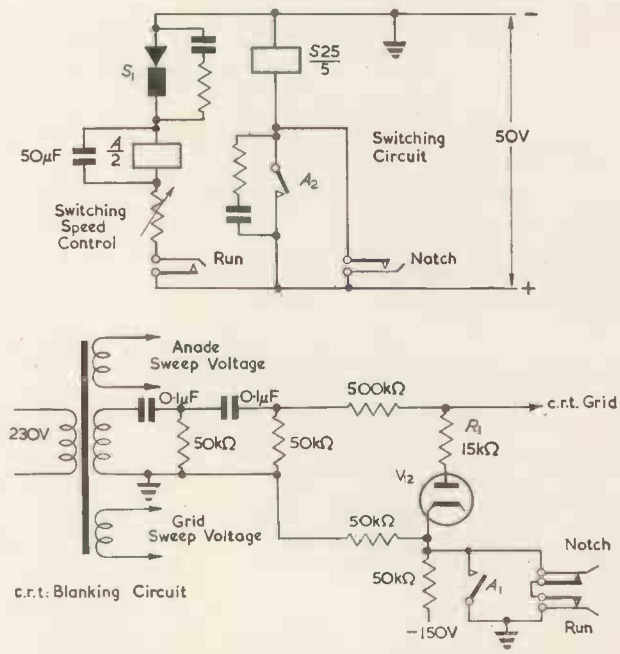


Fig. 10. Uniselector switching and c.r.t. blanking circuits

R_1 allows the screen voltage to be set to any value between 0 and 250V.

For the grid and reference voltages, three neon stabilizers connected in series across the output of a half-wave rectifier provide voltages of +150, −150, and −300V.

The +150V supply is used for balancing out the regulator circuit currents in the current monitoring resistor as already described and for obtaining +1, +2, +3 and +4V stabilized for the vertical anode voltage graph lines.

The −150V supply is used as the reference voltage for the anode and screen supply shunt regulator valve cathodes and for the control and suppressor grid voltages to the

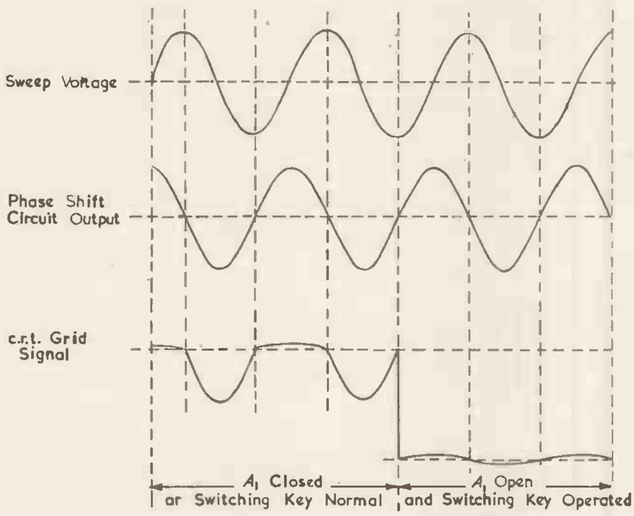


Fig. 11. Derivation of c.r.t. grid signal

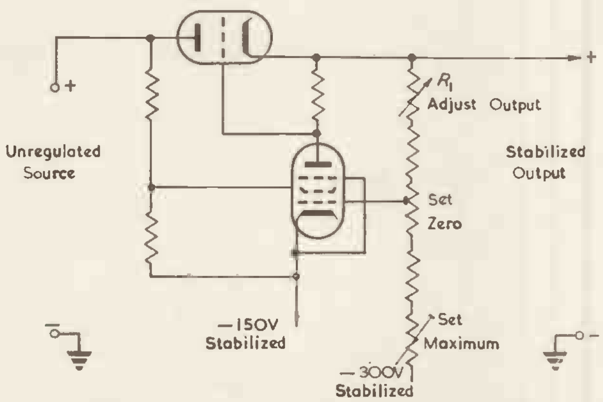


Fig. 12. Basic circuit of anode and screen supply regulators

valve under examination. The −300V supply is used as the reference voltage for the grids of the anode and screen supply shunt regulator valves.

Indicator

The indicator consists of a 12in, long persistence cathode-ray tube with associated d.c. amplifiers for deflexion. Power supplies are obtained from a separate power unit.

Accuracy

In the application for which the equipment was designed, i.e. as a visual training aid, extreme accuracy is not essential. Nevertheless, the design is based on circuit principles which are inherently capable of giving good accuracy if reasonable attention is given to detail.

The overall accuracy depends mainly upon (1) the closeness of tolerance of the resistors used in the voltage divider chains and in the voltage regulator circuits, (2) the accuracy of setting up the preset controls and (3) the characteristics of the deflexion amplifiers. In the latter case it is essential that the amplifiers should be capable of handling d.c. and low audio frequencies without appreciable phase/frequency or amplitude/frequency variation if the valve curves are to be reproduced faithfully.

Stability

The employment of stabilized d.c. power supplies and wire-wound resistors ensures good stability, and it is only

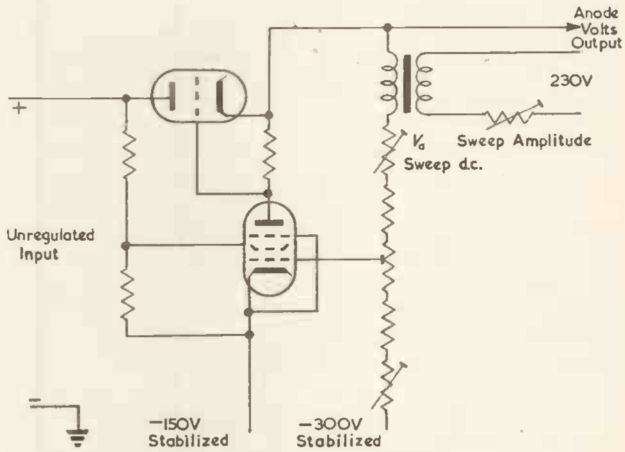


Fig. 13. Generation of sinusoidally varying anode voltage

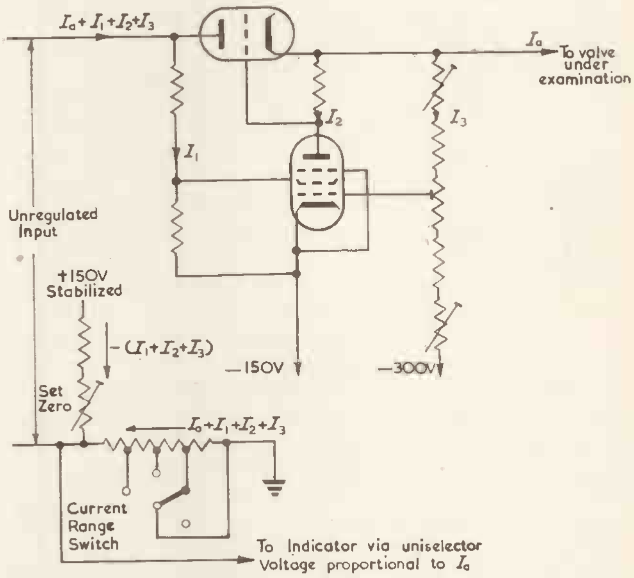


Fig. 14. Circuit for balancing out regulator currents in valve current monitoring resistor

occasionally that checks reveal the need for re-adjustments to compensate for the ageing of valves.

Conclusion

The equipment serves adequately the purpose for which it was constructed. It might be of value as a measuring instrument.

Acknowledgments

The author is indebted to Mr. D. W. McQue who contributed to the designs and to Messrs. S. C. Langford and D. G. Marsh who assisted with the construction and diagrams. All illustrations in this article are Crown copyright.

A Nine's Complement Decade Counter with Recorder

By J. A. Phillips*, B.Sc.

A brief outline of decimal counting using weighted binary digits is given with special reference to systems giving complements of nine. A binary decade electronic counter arranged so as to allow the reading of nine's complements, which may be used to represent negative numbers is then described. In the circuit arrangement used in this counter the maximum counting rate remains the same as that of a simple binary counter. Recording of the number counted is made on "Teledeltos" paper, the record and counter reset being carried out simultaneously.

THE binary decade counter and recorder were produced as part of the development programme for a multi-channel digital strain recorder. Since both positive and negative strains were to be measured it was considered advantageous to be able to present the results as a positive or a negative number respectively. The actual number representing strain was wholly positive, being a series of pulses passed by an electronic gate. A survey of available literature did not disclose the existence of a system similar to that outlined below which is based upon a commonly used binary counter circuit which had already been built in the laboratories.

General Considerations

Electronic counting circuits are in common use wherever rapidly recurring pulses of an electrical nature are to be counted or where data in some form is to be represented digitally to enable it to be analysed and processed more readily. The accuracy of the data in the digital form is, neglecting non-linearities, dependent on the number of digits used to represent a given change in value of the input data. Hence it is possible to increase the accuracy to the point where a single digit represents a change in the input of the same order as the noise in the data or its analogue.

By far the most common and reliable form of electronic counting circuit is the Eccles-Jordan, bistable or scale-of-two circuit[1,2] of Fig. 1. This circuit possesses two stable states of operation each characterized by one valve being at cut-off while the other is saturated. The injection of a pulse of sufficient magnitude into the input terminal causes the circuit to transfer from one state to the other. Hence two input pulses are required to make the scale-of-two complete a cycle of operation and produce an output pulse of the same polarity as the input pulse, i.e. the input pulses can be scaled down by a factor of two. In the circuit shown, negative input pulses are required to operate the scale-of-two. The output pulses of one binary stage may be fed into a second stage to be further scaled down. If lamps are connected across the anode loads of the valves in the scale-of-two, as shown in Fig. 1, they will indicate the state of the stage and if some means is available to place all the stages in a counter in a particular state prior to the arrival of a group of pulses, then the lamps may be used to indicate the number of pulses in the group. Resetting may be obtained by opening the switch S so as to insert an additional resistor which causes the scale-of-two to transfer to or remain in one state.

As has been shown in the preceding paragraph a counting device may be made by using a series of scales-of-two connected together; as shown in Fig. 2. Where there are n stages 2^{n-1} input pulses are required to bring about a transition of the n^{th} stage. Thus the counter records the number of pulses in binary form, i.e. in powers of two. If

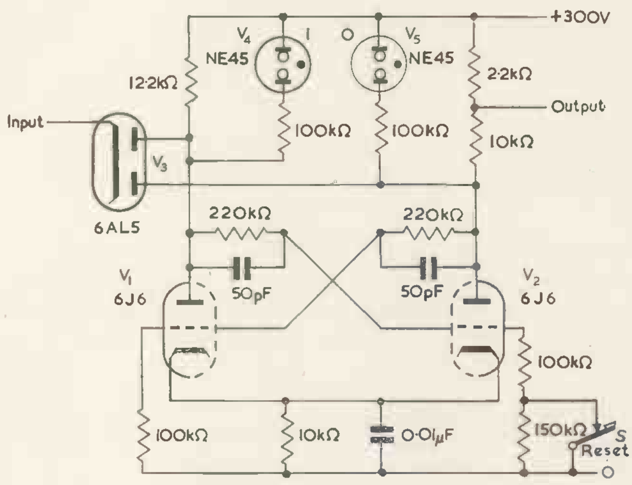


Fig. 1. A scale-of-two circuit

Fig. 2. Cascaded binary stages

the stages of the counter are set to a given state before making the count this state is called the 0 state and the other state occurring on the 1st, 3rd etc. input pulse to each stage, the 1 state. Thus if we have, for example, five binary stages in a counter they will represent the decimal numbers 16, 8, 4, 2 and 1 when each is in the 1 state, e.g. in the binary notation 10110 will represent the decimal number 22.

While the use of binary counting systems gives great operational simplicity to many counting and computing systems, it involves a considerable amount of mental labour where direct reading out is to be done, due to the fact that an individual trained to carry out arithmetic operations in the decimal system will usually go through the process of converting any binary number presented to him into decimals. Thus where further processing of the data is to be done by computers, with or without the aid of decimal desk machines, it is preferable to modify the binary counting system so that it will give a decimal form of presentation.

Since there are ten digits to be represented in the decimal system a minimum number of four binary stages will be

* *Australian Defence Scientific Service.*

required, these having 16 combinations as shown in Table 1. Thus six of the binary numbers must be absorbed within a binary decade in the process of producing the decimal number. Stating the case more generally there are 16!/6!, i.e. $2{\cdot}9 \times 10^{10}$ ways in which the ten decimal digits

TABLE 1

BINARY SEQUENCE	STATE OF BINARY XI	XI	XI	XI
0	0	0	0	0
1	0	0	0	1
2	0	0	1	0
3	0	0	1	1
4	0	1	0	0
5	0	1	0	1
6	0	1	1	0
7	0	1	1	1
8	1	0	0	0
9	1	0	0	1
10	1	0	1	0
11	1	0	1	1
12	1	1	0	0
13	1	1	0	1
14	1	1	1	0
15	1	1	1	1

can be coded using four binary digits. However, only $16!/(384 \times 6!)$ or $7{\cdot}6 \times 10^{7}$ basic coding systems are required in obtaining the total number of codes since there is a total of 384 input transformations, all of which produce codes which are similar. For the sake of simplicity it is usual to assign to each binary digit a constant weighting number and where visual read-out of the count is to be made the four weights are made positive. It can be shown[3] that the sum of the four constant weights must equal nine if the coding system is to produce nine's complements by the direct inversion, i.e. interchange of 0's and 1's of the binary digits. This with the restriction that the weights must be positive, limits the choice of weights to four groups. The arrangement of each group may be altered so that there is a total of 14 systems of four binary digit decimal codes having positive weights and giving nine's complements by direct inversion. These groups and systems are given in Table 2.

Nine's complements of decimal numbers are formed by subtracting each digit from nine, e.g. the nine's complement of 751 is 248. It will be seen that 248 is, with an error of one, the complement of 10 of 751, i.e. 751 + 249 is a power of ten. Hence it can be seen that in a computing device subtraction can be achieved by taking the nine's complement of the subtrahend, performing an addition and adding one. The last step can be produced by "end around carry".

In the counter being considered here it was required that a mechanical strain, measured and recorded as a count of pulses, i.e. a number wholly positive, should be represented by a positive or negative number analogous to the strain itself. Of course, binary digits can be readily complemented so as to represent negative numbers. If the number 10000 is taken as base then 10110 represents a positive binary number 110, while 01010 represents, with the error of one, a negative binary number 110. However, where visual read out is required a decimal system, giving positive and negative numbers, is the obvious choice. It can be seen that if a power of ten is taken to represent zero strain then positive strains will be represented by a number greater than the power of ten used as a base while negative strains will, with an error of one, be represented by the nine's complement of a number less than the base. For example, if 1000 is taken as the base, i.e. zero strain, then a count of 1 751 represents a positive strain proportional

TABLE 2

DECIMAL NUMBER	POSITIVE WEIGHTS ASSIGNED TO BINARY DIGITS 1 1 2 5	1 1 3 4	1 2 2*4	1 2 3 3
0	0 0 0 0	0 0 0 0	**0 0 0 0**	0 0 0 0
1	1 0 0 0	1 0 0 0	**1 0 0 0**	1 0 0 0
2	1 1 0 0	1 1 0 0	**0 1 0 0**	0 1 0 0
3	1 0 1 0	0 0 1 0	**1 0 1 0**	0 0 1 0
4	1 1 1 0	1 0 1 0	**0 1 1 0**	1 0 1 0
5	0 0 0 1	0 1 0 1	**1 0 0 1**	0 1 0 1
6	0 1 0 1	1 1 0 1	**0 1 0 1**	1 1 0 1
7	0 0 1 1	0 0 1 1	**1 0 1 1**	1 0 1 1
8	0 1 1 1	0 1 1 1	**0 1 1 1**	0 1 1 1
9	1 1 1 1	1 1 1 1	**1 1 1 1**	1 1 1 1
0	0 0 0 0	0 0 0 0	0 0 0 0	0 0 0 0
1	1 0 0 0	1 0 0 0	1 0 0 0	1 0 0 0
2	1 1 0 0	1 1 0 0	0 1 0 0	0 1 0 0
3	0 1 1 0	0 0 1 0	1 1 0 0	0 0 1 0
4	1 1 1 0	0 0 0 1	0 1 1 0	1 0 0 1
5	0 0 0 1	1 1 1 0	1 0 0 1	0 1 1 0
6	1 0 0 1	1 1 0 1	0 0 1 1	1 1 0 1
7	0 0 1 1	0 0 1 1	1 0 1 1	1 0 1 1
8	0 1 1 1	0 1 1 1	0 1 1 1	0 1 1 1
9	1 1 1 1	1 1 1 1	1 1 1 1	1 1 1 1
0	0 0 0 0	0 0 0 0	0 0 0 0	0 0 0 0
1	1 0 0 0	1 0 0 0	1 0 0 0	1 0 0 0
2	0 0 1 0	1 1 0 0	0 1 0 0	0 1 0 0
3	1 0 1 0	0 0 1 0	1 0 1 0	1 1 0 0
4	1 1 1 0	0 1 1 0	0 0 0 1	1 0 1 0
5	0 0 0 1	1 0 0 1	1 1 1 0	0 1 0 1
6	0 1 0 1	1 1 0 1	0 1 0 1	0 0 1 1
7	1 1 0 1	0 0 1 1	1 0 1 1	1 0 1 1
8	0 1 1 1	0 1 1 1	0 1 1 1	0 1 1 1
9	1 1 1 1	1 1 1 1	1 1 1 1	1 1 1 1
0	0 0 0 0		0 0 0 0	
1	1 0 0 0		1 0 0 0	
2	0 0 1 0		0 1 0 0	
3	0 1 1 0		1 1 0 0	
4	1 1 1 0		0 0 0 1	
5	0 0 0 1		1 1 1 0	
6	1 0 0 1		0 0 1 1	
7	1 1 0 1		1 0 1 1	
8	0 1 1 1		0 1 1 1	
9	1 1 1 1		1 1 1 1	

to 751 while a count of 280 is analogous to a negative strain of 719 + 1 = 720; for 1000 − 280 = 720.

There are a number of binary decade counting circuits in common use[2,4,5]. Decimalization is obtained by feeding back additional pulses within the counter so that for ten input pulses the circuit generates the equivalent of six extra pulses to make up the total of 16. However, none of these circuits provide nine's complements by complementing the states of the binary stages. Therefore a study was made of the 14 systems of four binary digit decimal codes giving nine's complements in an attempt to find a circuit to which normal scale-of-two circuits could readily be adapted. The counter described in the next section resulted from this study.

Nine's Complement Binary Decade Counter

Of the 14 systems mentioned in the previous section those belonging to the 1, 1, 2, 5 group seemed most attractive since the arithmetic required in obtaining the decimal number is so simple. However, no simple circuit arrangement seemed possible. The 1, 2, 2*, 4 group was then studied and a circuit, devised on the system which is printed in bold type in Table 2.

With reference to the relevant part of Table 2 it will be seen that the states of stages 1 and 2 alternate, except when the count is zero or nine, and the binary stages weighted 2, 2* and 4 operate as a normal binary counter. Normal binary counter action also occurs for all stages for the 7th

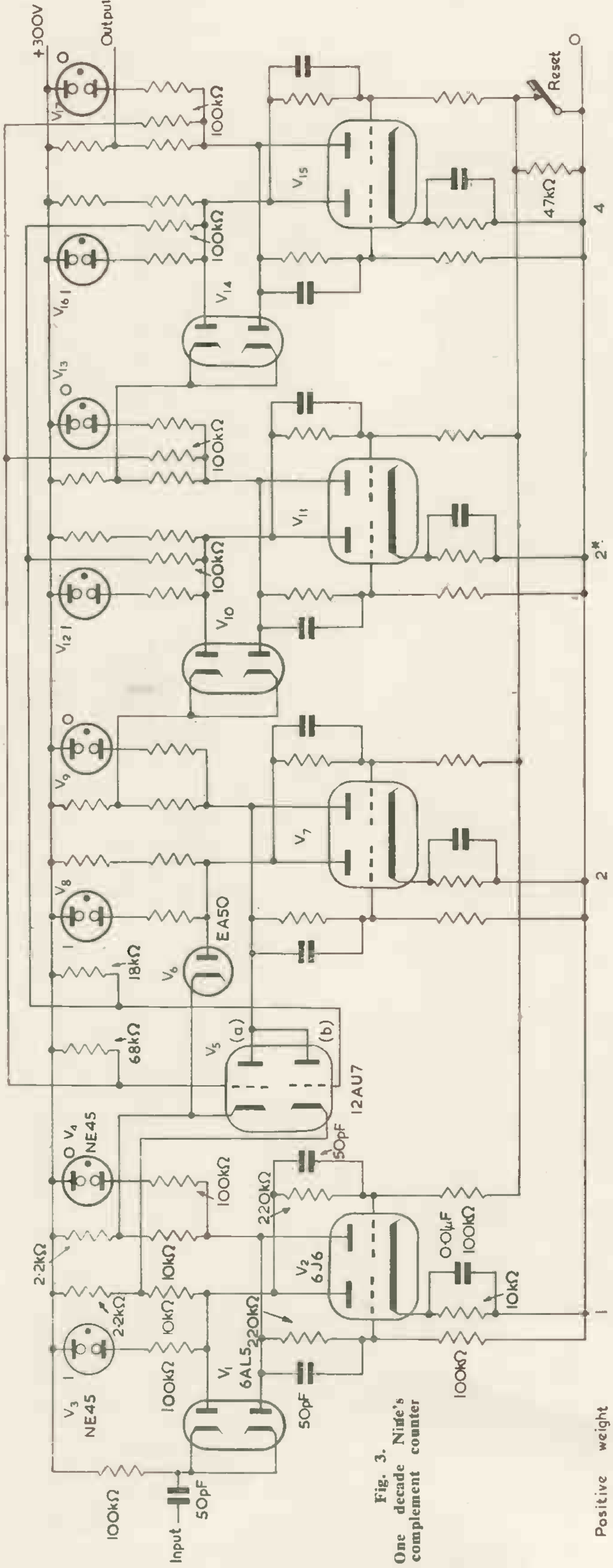


Fig. 3. One decade Nine's complement counter

through to 10th (i.e. reset to zero) input pulses. Therefore if the states of stages 1 and 2 are made to alternate up to a count of 7 and then the decade is caused to operate as a normal 4-stage binary counter the system is achieved.

The four binary stages of the circuit Fig. 3 are labelled 1, 2, 2*, 4; 2* being used to differentiate between the two stages having the same assigned weight. Interstage coupling is normal except for the coupling between the stages 1 and 2. Here the double diode is replaced by a twin triode, which acts as a gate, and a single diode. The conditions at the control grids of the triodes are set by the states of stages 2* and 4. When either or both of stages 2* and 4 are in the 0 state stage 2 is driven by stage 1 via V_{5b} and V_6, which ensures that when one stage is in the 0 state the other is in the 1 state and vice versa. The control grid of V_{5b} is held near zero grid-cathode potential by the voltage divider between the relevant anodes of stages 2* and 4 and the + 300V line. Similarly V_{5a} is held at cut-off by its voltage divider. As soon as both stages 2* and 4 attain the 1 state the conditions on V_{5b} and V_{5a} are reversed. Stage 2 is now driven in normal binary fashion via V_{5a} and V_6, until the counter decade resets to zero on the tenth input pulse. As the decade returns to zero on the tenth pulse a negative pulse appears at the output terminal to trigger the following decade; stages 2* and 4 returning to the 0 state apply saturation and cut-off voltages to V_{5b} and V_{5a} respectively in preparation for the following group of ten input pulses.

As has been mentioned previously the binary decade counter was designed about an existing binary counter. This counter was required to count a maximum of 2 000 input pulses, which required 11 binary stages, i.e.:

$$2 \times 10^3 = 10^{3 \cdot 301} = 2^n$$

$$n = 3{\cdot}301 \log 2$$

$$= 11$$

The base count was taken as $2^{10} = 1\,024$ and positive and negative numbers were taken about this base.

In the binary decade counter three complete decades and one stage to indicate 1 000 are required if a total of 2 000 input pulses is to be counted. Since four binary stages are required per decade the total number of binary stages is increased to $(3 \times 4) + 1 = 13$. A count of 1 000 was taken as the base and the positive and negative numbers were taken directly and by nine's complements from this.

Lamps to indicate the state of each stage are placed in anode circuits of the valves as shown in Fig. 3, appearing on the front panel of the counter with the designations given in Fig. 4. Zero strain is indicated by the base count of 1 000, i.e. the left-hand lamp indicates the 10^3 stage as being in the 1 state while all the other stages are in the 0 state. Positive strains are indicated by counts greater than 1 000, that is the 10^3 stage and some of the other stages in the three decades are in the 1 state. The decimal number indicated by the 1 state lamps in the three decades is then proportional to the positive strain. The count is less than 1 000 for negative strain so that the 10^3 stage is in the 0 state with the other 1 state lamps indicating the actual number of pulses counted. The decimal number indicated by the 0 state lamps in the three decades is the nine's complement of the number of pulses and is thus proportional, with an error of one, to the negative strain. Thus it is possible to mark the 10^3 stage 1 state and 0 state lamps + and − respectively to show the strain sign and which row of lamps is to be used to indicate its magnitude. There will be an error of one in the negative strain number which can be eliminated by giving the 10^3 stage 0 lamp a weight of − 1 which has to be added to the nine's complement, taken also as negative, or alternatively zero strain may be taken as a base count of 999 and "end around carry" used by injecting a single pulse into the counter input when the 10^3 stage goes to the 1 state, e.g.

a count of $999 \equiv 0$ strain

$$998 \equiv -1 \text{ strain}$$

$$1000 + 1 = 1001 = +1 \text{ strain}$$

where the 1 is the "end around carry" digit.

However, in the case of the Multi-channel Digital Strain Recorder there is an inherent error of one in counting as the counter pulses are derived from an oscillator whose output is gated for a time interval proportional to strain. Therefore if the recorder is adjusted so that at zero strain the count is 999 or 1000 there will be an error of $+1/-0$ in negative count by neglecting the error of one in taking nine's complements and an error of $+0/-1$ in positive count because of the pulse gating circuit. This total error of ± 1 in count could be tolerated.

The original binary counter had a maximum counting rate of 120kp/s. When converted to a decimal counting system just described it was found that the counting rate had not been altered. It may be concluded from this that the time taken for a pulse to be transmitted through a decade and the grid potentials of V_{5b} and V_{5a} to be reversed is less than the time interval of two input pulses, i.e. $2/(120 \times 10^3)$ sec.

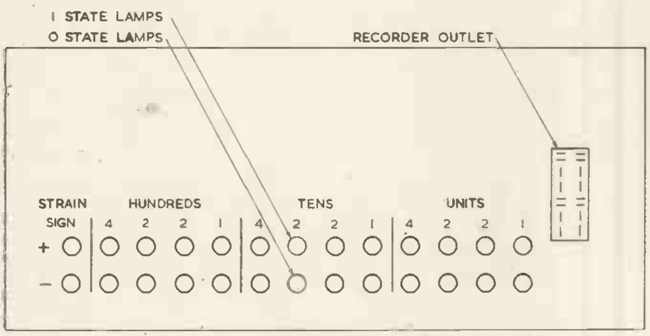


Fig. 4. Counter front panel

Recorder and Reader

It has been shown in the previous section how a number of pulses may be counted in such a way as to enable it to be read out as either a positive or negative number referred to some base. In order that these numbers may be retained for future processing some form of recording device is required which may be attached to the counter. At the time when construction was begun on the strain recorder there was no suitable form of printing recorder available which would operate at a speed of 12½ records per second. On this account a recorder was built in which a row of dots, corresponding to the counter stages in the 1 state, is made on "Teledeltos" paper.

A small shaded-pole induction motor drives, through reduction gearing, the paper drive roller and take-up spool. The "Teledeltos" paper is carried over the drive roller and beneath the 13 pens which are grouped in three sets of four pens corresponding to the decades, and a single pen for strain sign indication. The paper is driven at approximately 12½ft per minute giving a spacing between each record of about 0·2in. Fig. 5 is a photograph showing the layout of the recorder. The recorder pens may be lifted from the drive roller to allow loading of paper. A cable terminated in a multi-point plug to fit the counter connects the recorder pens and recorder roller to the counter stages and recorder supply while a 3-core cable with 3-pin plug connects 230V a.c. and an earth to the motor and recorder case respectively.

As will be seen from the above, positive strains will be recorded as the presence of a dot in the sign column with dots for the stages in the 1 state in each decade. Assigning the correct weight to each dot gives the positive strain number. With negative strains there will be a space in the sign column and it is then necessary to give the spaces in the decade columns the weights to obtain the negative strain number.

The Reader, yet to be built, will consist of a take-off and manually driven take-up spool with the paper travelling between them and beneath a semi-cylindrical lens. The plane underside of the lens is to have lines engraved on it to break the dots into the decades and sign columns while the case is to be engraved above the decade dot positions with the positive weights assigned to each. The strain signs and counts will thus be read off and tabulated or punched on to cards for computer processing.

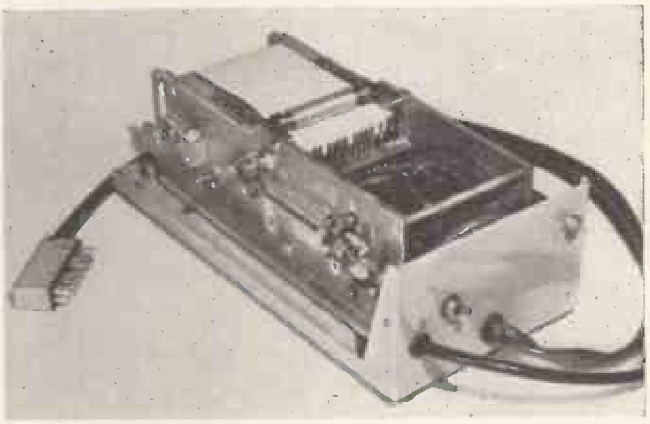


Fig. 5. Recorder, showing drive motor pens and pen lifting mechanism

Recording and Counter Resetting

As will be seen in Fig. 6, there is a type 2D21 thyratron associated with each binary stage. The control grid potential of the thyratron is set by the state of its stage via a voltage divider so that the thyratron can conduct only when the scale-of-two is in the 1 state. The cathode of the thyratron is at earth potential and the anode is connected to the appropriate pen in the recorder via a current limiting resistor. The circuit for all the pens is completed via the recorder roller, a 500V r.m.s. transformer winding and a control 2D21. The control thyratron is held at cut-off by the application of an anti-phase voltage to its control grid.

On the completion of a count, a relay, whose contact unit is designated X_1, is energized at a zero intercept of the 50c/s mains voltage cycle. The relay contacts close approximately 1msec after this, energizing the counter reset relay Y and applying a positive pulse via a 10pF capacitor to the control grid of the recorder control valve. Since the instantaneous voltage applied to the recorder is then approximately $500\ \sqrt{2} \sin \pi/10 = 220$V the control and pen thyratrons ignite and the record is made. The reset relay Y operates about 1msec after X_1 closes, energy being supplied by the 2μF capacitor which has been charged via the 10kΩ resistor from the -50V line. The discharge keeps relay Y operated for approximately 10msec, the current through the resistor being insufficient to hold the relay. The reset relay contact Y_1 inserts a resistor into the common lead to one grid of each binary stage upsetting

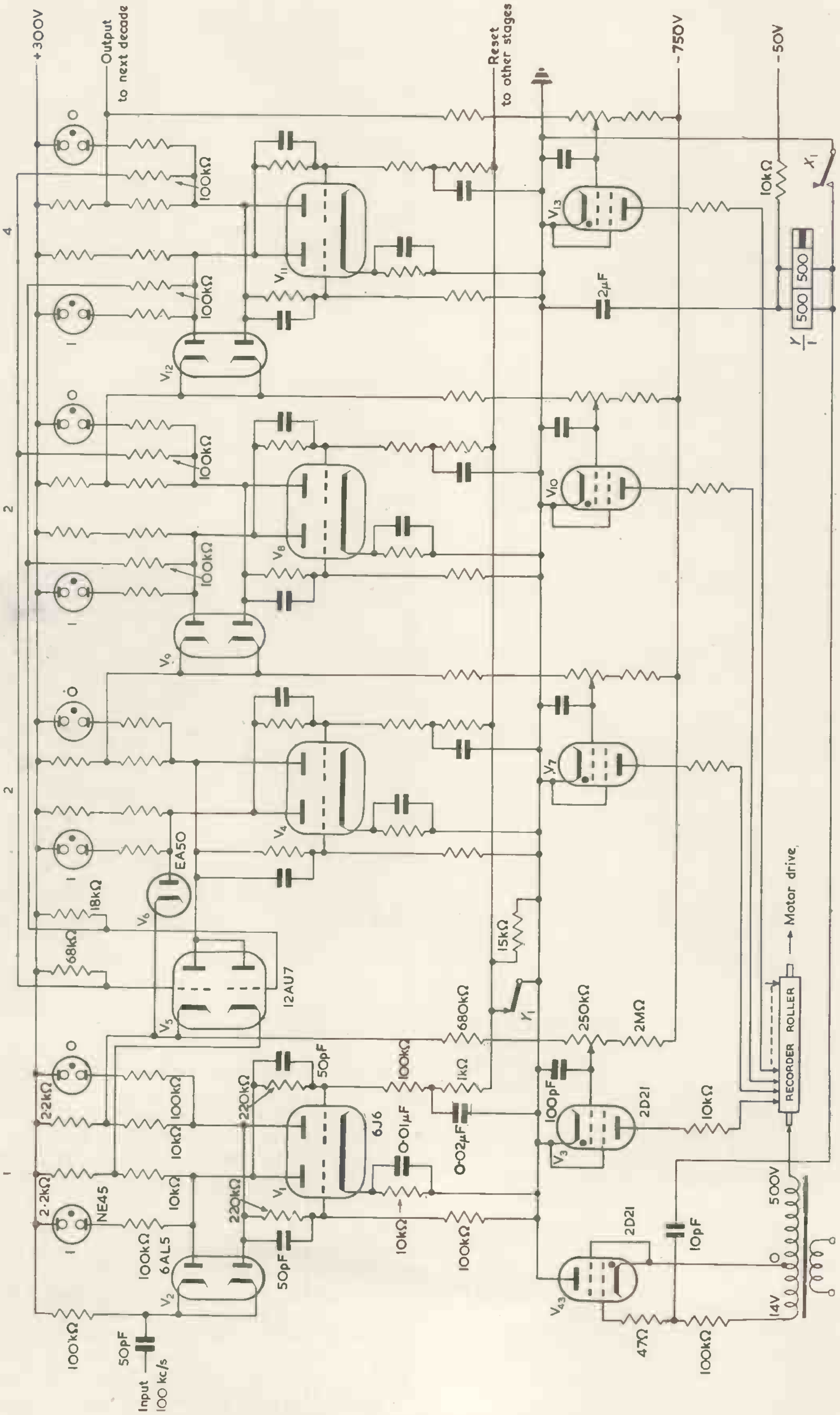


Fig. 6. Counter (one decade) and recorder

it so that it changes to the 0 state. Because the recorder thyratrons have already ignited it is quite satisfactory to reset the counter during the recording interval. This reduces the record and counter reset time to about ½ cycle of mains frequency, i.e. 10msec. If the counter input can be timed to occur in synchronism with the mains frequency the total time taken in counting and recording up to 2 000 pulses of 10^5 pulses per second may be reduced to 30msec or say 35msec if one introduces a space between count and record to allow for variations of the instant at which counting commences.

Conclusion

A binary decade counter and recorder have been described which will count pulses and present the number in decimal form in such a way that complements of nine of the number counted are also available. The circuit does not decrease the maximum counting rate from that of one of the binary stages as do some feedback arrangements, yet it is comparatively simple and does not appear to be critical as regards components and voltages. The method of recording and counter resetting is such that these operations can be concluded in a minimum of time.

Acknowledgment

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Industrial Television at Steel Rolling Mill

The Steel Company of Wales has long been aware of the potentialities of closed circuit television. Some while ago, experimental demonstrations by various manufacturers of such equipments were carried out at Abbey Works, Port Talbot and as a result, an order was placed with Marconi's Wireless Telegraph Co. Ltd.

But before the equipment could be delivered the operation of the 45in Slabbing Mill was very seriously affected by the failure of a slab manipulator, which would take at least a fortnight to repair.

The function of a slabbing mill is to reduce ingots weighing up to 20 tons to a slab section suitable for passing on to strip rolling mills, where the thickness is further reduced to plates and sheet steel. Such a mill is therefore an essential link in a chain of operations, and a failure means that the whole sequence of production is eventually brought to a stand-still.

The ingots are fed to the huge rollers driven by 27 000 h.p. motors on a roller conveyor system. The mill itself is massive in construction and its bulk prevents an operator positioned on one side from seeing the ingot once it has passed through the main rollers to the other side.

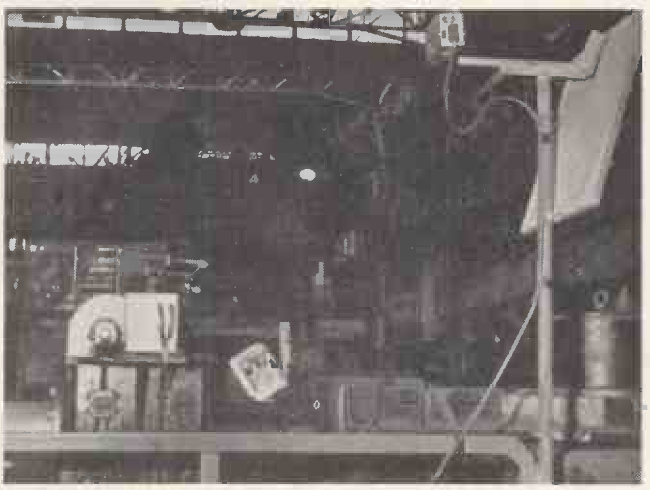
Under normal circumstances progress is controlled by two operators, one of whom is in charge of an equipment known as the manipulator. Once through the rollers, the ingot is stopped and is turned over by special mechanisms before being reversed through the rollers for further passes. This reversing process is repeated until the requisite section is reached.

Previously a similar breakdown on the front side of the mill and within the direct vision of the operator had been overcome by the use of a special hook, manipulated by an overhead crane. This time however it was necessary to use the same method on the outgoing side of the mill and this prevented clear sighting by the operators. It was therefore decided to provide an "eye" behind the mill by the use of industrial television, with a monitor screen at the manipulator operator's position. To accomplish this the Marconi Company hastened delivery of their equipment. The operator was then able to see the ingot's position exactly and the operation of affixing the crane-hook became a matter of routine.

The effect of the breakdown, was, by this means limited to such an extent that an average of 1 600 tons per shift was still being produced.

Interior of crane operator's cabin, with television screen displaying image of ingot.

Ingot in slabbing mill, with camera positioned top right.



Measurement of the Self-Capacitance of an Inductor at High Frequencies

By J. P. Newsome*, M.Sc., A.M.I.E.E.

The article defines and briefly discusses self capacitance. A detailed survey of methods of measuring this quantity is then undertaken. Certain of the methods described are not well known, while one is novel and can give a result of improved accuracy.

THE self-capacitance of an inductor is made up of a distributed capacitance due to turns and screening and a lumped capacitance due to the terminating assembly.

It is customary to consider the self-capacitance of a low loss inductor as a fixed value invariable with frequency in any lumped form of equivalent circuit. This procedure is substantially correct for frequencies below and removed from the self-resonant frequency of the inductor and is not without significance; it is not necessarily correct for frequencies approaching the self-resonant frequency.

To determine the precise behaviour of an inductor, a knowledge of the self-capacitance is required; usually this quantity must be measured, although empirical formulæ exist for its approximate calculation with inductors of regular form[1]. It is noteworthy that the mathematical calculation of self-capacitance even for inductors of regular form is a matter of very considerable complexity.

This article summarizes the methods of measuring self-capacitance. Certain of these methods are not well known, while one method is novel and can give a result of improved accuracy.

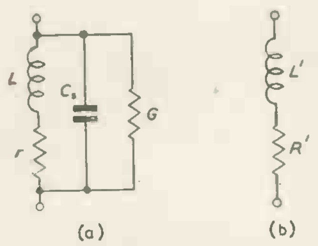


Fig. 1. Lumped equivalent circuits of an inductor

Equivalent Circuits of the Inductor[2]

It is intended to confine this article to low loss inductors operating at frequencies below the self-resonant frequency.

Experimental evidence shows that such an inductor may be satisfactorily represented by the lumped equivalent circuit shown in Fig. 1(a). At frequencies substantially removed from self-resonance, the inductance L and the self-capacitance C_s are found to be significantly constant over a wide frequency range (providing the permeability and permittivity of the magnetic and electric flux paths are not variable with frequency and flux density). This feature was commented upon by Howe in connexion with the calibration of wavemeters[3]; he later supported the experimental evidence with a fundamental analysis[4], which showed that the self-capacitance of a solenoidal inductor was primarily a function of coil dimensions, but independent of frequency.

In Fig. 1(a), r is a resistance representing the power loss in the inductor winding, the ferromagnetic core and the screen, G is a conductance representing the power loss in the dielectric of the capacitance C_s and in the insulation resistance. It is frequently possible to ignore G (a test for its significance is given by Legg[5]), which is usually only significant at frequencies approaching the self-resonant frequency: with this simplification, the equivalent circuit of Fig. 1(a) may be reduced to that of Fig. 1(b) using the simplified expressions:

$$R' = \frac{r}{(1 - \omega^2 LC_s)^2} \qquad (1)$$

$$L' = \frac{L}{(1 - \omega^2 LC_s)} \qquad (2)$$

Definition of Self-Capacitance and Self-Resonant Frequency of an Inductor

Moullin defines self-capacitance of an inductor as that capacitance C_s, which has to be subtracted from the capacitance required to resonate the coil at a given frequency and which is calculated using the low frequency inductance[6] [viz. the inductance L in Fig. 1(a)]. This definition implies that L is considered invariable with frequency though not necessarily C_s.

The self-resonant frequency of an inductor may be defined as that frequency at which the impedance of an inductor is wholly resistive or at which the impedance magnitude is a maximum. For a low loss inductor these conditions occur at substantially the same frequency.

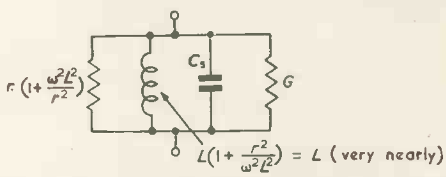


Fig. 2. Modified form of the equivalent circuit shown in Fig. 1(a)

Fig. 1(a) may be transformed to the equivalent circuit shown in Fig. 2.

Considering $r^2/\omega^2L^2 \ll 1$, which is the case for the low loss inductor, the self-resonant frequency would be given by:

$$f_{sr}^2 = \frac{1}{4\pi^2 \, . \, LC_s} \qquad (3)$$

This relation assumes that L and C_s are not variable with frequency, which is not always true, particularly for frequencies approaching f_{sr}.

Methods of Measurement of the Self-Capacitance of an Inductor

Methods Based on Resonating the Inductor

(a) The inductor is resonated with an external tuning capacitance C_T, at a number of frequencies as shown in simplified form in Fig. 3. Fig. 3(a) shows the shunt form of test circuit, Fig. 3(b) the series form, as in the Q meter. Resonance is indicated by the valve-voltmeter reading, which will be a minimum for the shunt circuit and a

* *University of Nottingham.*

maximum for the series circuit. In either case, for the low loss inductor, resonance is given by:

$$\frac{1}{4\pi^2 f^2 L} = (C_s + C_T) \quad \text{.............} \quad (4)$$

A plot is made of $1/f^2$ against C_T, C_s being obtained by extrapolating the curve, since when $1/f^2 = 0$, $C_s = -C_T$ [see Fig. 3(c)].

This method is well known; it is time consuming and will yield a result of poor accuracy unless carefully undertaken with an accurately calibrated tuning capacitor. Departures from linearity of the relationship of equation (4) indicate that either L or C_s must vary with frequency.

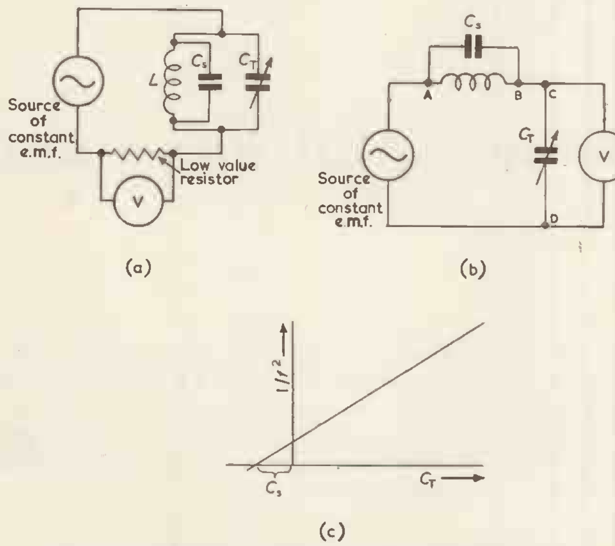


Fig. 3(a) and (b). Forms of test circuit for resonating an inductor (c). Graph of $1/f^2$ against C_T

(b) In a modification of this method used to obtain a rapid result, two readings are taken as indicated above. Consider the tuning capacitance to have values for resonance of C_{T1} and C_{T2} at frequencies respectively f_1 and f_2. C_s is given by:

$$C_s = \frac{C_{T1} - n^2 C_{T2}}{(n^2 - 1)} \quad \text{..............} \quad (5)$$

where $n = f_2/f_1$.

Again unless C_T is accurately calibrated, a result of low accuracy is obtained. Any change of L or C_s with frequency is not made apparent by this method.

Methods Based on the Determination of the Self-Resonant Frequency

If f_{sr} is known and also the capacitance C_T required to resonate the inductor at a substantially lower frequency f, then using equations (3) and (4), C_s is given by:

$$C_s = C_T \frac{f^2}{f_{sr}^2 - f^2} \quad \text{...............} \quad (6)$$

If it is considered that C_s and L may vary significantly as f_{sr} is approached, the procedure is modified as follows. A small known capacitance C_F is placed in shunt with the test inductor; this gives rise to an artificial self-resonant frequency f_{sr}' where:

$$f_{sr}'^2 = \frac{1}{4\pi^2 L(C_s + C_F)} \quad \text{...........} \quad (7)$$

Assume that the modified inductor now resonates with a tuning capacitance C_T'; equation (4) becomes:

$$\frac{1}{4\pi^2 f^2 L} = (C_s + C_T' + C_F) \quad \text{..........} \quad (8)$$

and C_s is now given by:

$$C_s = C_T' \cdot \frac{f_{sr}'^2 - f^2}{f^2} - C_F \quad \text{............} \quad (9)$$

This procedure may be carried out with increasing values of C_F and thus any variation in the value of C_s with frequency will be apparent.

Methods of Determining f_{sr} and f_{sr}'

(a) *The Q Meter Method*

[viz. using the circuit shown in basic form in Fig. 3(b)].

A tuned circuit is set up in the Q meter by introducing a suitable inductor between the terminals AB and resonated at a frequency estimated to be close to f_{sr}. The test inductor is then connected in shunt with C_T. If the frequency is f_{sr}, the impedance presented to C_T will be resistive and the tuned circuit in the Q meter will be damped, but not detuned; if the frequency is not equal to f_{sr}, then C_T will have to be adjusted to restore the resonant condition. The frequency is altered until the required condition is attained. The capacitance C_F is conveniently introduced by suitably increasing the tuning capacitance to a value C_T'' immediately before the test inductor is introduced into the circuit; then:

$$C_F = C_T'' - C_T \quad \text{..............} \quad (10)$$

If the coil is resonated at a substantially lower frequency than f_{sr} (or f_{sr}') by introducing it between the terminals AB, giving rise to an expression similar to equation (4), then using equations (4) and (7), it may be written that:

$$C_s = C_T \frac{f^2}{f_{sr}'^2 - f^2} - C_F \cdot \frac{f_{sr}'^2}{f_{sr}'^2 - f^2} \quad \text{.......} \quad (11)$$

(b) *The Maximum Impedance Method*

The frequency f_{sr} (or f_{sr}') is determined from the condition of maximum impedance magnitude using a circuit of the type shown in Fig. 3(a) with C_T removed. The frequency of the constant e.m.f. source is varied and the required condition is indicated by a minimum valve voltmeter reading. It is important that the source be free of harmonics as their presence can give rise to an erroneous minimum, the self-resonating inductor and low value resistor acting as a rejection filter to the fundamental. For this reason, this method must be regarded as approximate.

(c) *The Absorption Method* (for unscreened inductors)

The test inductor is placed in proximity with the tuned circuit of a feedback oscillator, whose output signal is monitored. The oscillator frequency is varied and a sharp drop in output is observed as the self-resonant frequency of the test inductor is approached. It is stated by Terman[7] that values of self-capacitance obtained using this method are low valued due to the distributed nature of the circuit and induced e.m.f. This method must therefore be regarded as approximate.

The Immersion Method

This approximate method is due to Meissner[8] and is intended for use with inductors having a minimum of solid matter associated with the inductor turns. The inductor is resonated as previously described at a given frequency using a tuning capacitance C_{T3}; the inductor is then immersed in a low loss oil of known relative permittivity ϵ_r, resonance being restored with a reduced tuning capacitance C_{T4}. C_s is extracted from the expressions:

$$f^2 = \frac{1}{4\pi^2 L(C_s + C_{T3})} = \frac{1}{4\pi^2 L(\epsilon_r C_s + C_{T4})} \cdot \quad (12)$$

Thus

$$C_s = \frac{C_{T3} - C_{T4}}{(\epsilon_r - 1)}$$

Note that this method does not require an accurate knowledge of frequency. The method suffers in most cases from the disadvantage that it is not possible for the oil to penetrate the entire volume through which the electric flux passes. Further, in order to repeat the measurement it is necessary to carefully remove the oil adhering to the coil turns, which is a tedious procedure.

Conclusion

The established methods of measuring C_s described are reliable, but tend to give a result of poor accuracy. Methods based on the determination of self-resonant frequency must be used with caution for the reasons stated. However, the Q meter method used in association with the principle of artificial self-resonance provides a method which has advantages, particularly for small values of C_s.

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Electronic Methods of Analogue Multiplication

By Z. Czajkowski*, B.Sc.

(Part 2)

Servo Multiplier

This is not in a strict sense an electronic multiplier, but is being considered here because it is the simplest instrument representing the feedback principle on which many completely electronic devices are based.

The simplest form of servo multiplier (Fig. 9) consists of a motor driving, through suitable gearing, the sliders of two linear potentiometers. Constant voltage E_b is applied

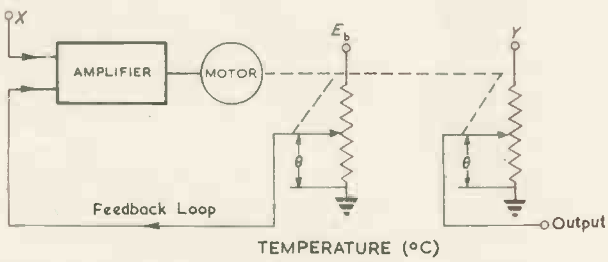


Fig. 9. Servo multiplier

to the potentiometer 1 and therefore the voltage at the slider will be:

$$V_t = \theta E_b$$

where θ is the angle represented as a fraction of the total sweep of the potentiometer 1. This voltage is then fed to the amplifier which compares it with the input voltage X and drives the motor in such a direction as to make the difference between X and θE_b tend to zero. Therefore:

$$X = \theta E_b$$

$$\theta = (X/E_b)$$

A voltage proportional to the second variable Y is fed to the potentiometer 2. The output from the slider of this potentiometer is:

$$V_{out} = Y\theta = XY/E_b$$

The accuracy of the multiplication so performed depends on two factors: gain of the amplifier and the similarity between the two potentiometers. It is important to note here that the accuracy does not depend on the linearity of the potentiometers as long as they are identical. As for the gain of the amplifier, it is related to the stability of the feedback loop, which in turn is governed by the phase lags (usually of a mechanical nature) of the system.

In practice, the accuracy may be of the order of 0·01 per cent, the time response 0·1sec. The system can be arranged to handle inputs of any polarity.

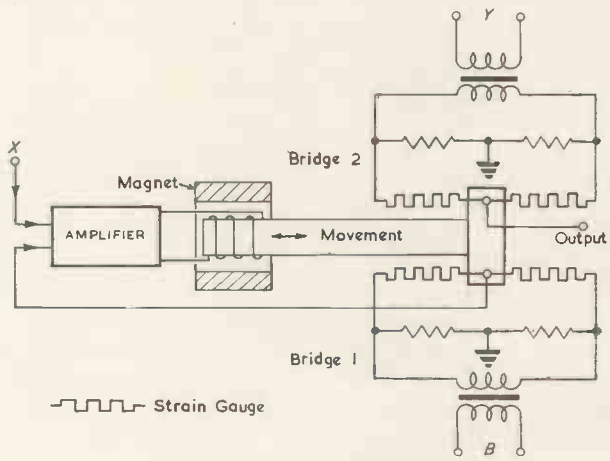


Fig. 10. Strain gauge multiplier

Strain Gauge Multiplier

This is the most direct development of the servo principle[17]. The system (Fig. 10) consists of four strain gauges arranged in two bridge circuits, as shown. The outer ends of the gauges are fixed, but the centre points are attached to a movable armature. This armature is displaced from the position of equilibrium by a moving-coil placed in a strong magnetic field. The output of the bridge 1, which is supplied with the reference voltage of constant amplitude, is compared with the input X, the difference is then amplified and after rectification fed into the moving-coil actuator. The input Y is applied to the bridge 2 which is identical with the bridge 1. As both pairs of strain gauges are displaced by the same amount, the output is propor-

* *Battersea Polytechnic and Winston Electronics Ltd.*

tional to XY, by the same reasoning that applies to the servo multiplier.

The system uses a mechanical link in its feedback loop, therefore its speed of response is somewhat limited. The difficult part of the system lies in finding a set of strain gauges with exactly the same characteristics.

Thermistor Multiplier

A thermistor is a non-linear element which changes its resistance with the change of temperature. This change of resistance is usually caused either by passing current through the element itself or changing the outside temperature. The indirectly heated thermistor has the form of a minute glass bead in which both the element and the heater are enclosed. The whole is placed in a vacuum envelope. The unit which was developed for the purpose of computation[18] contains one heater and two elements. The use of this unit for multiplication is illustrated in Fig. 11. The principle of operation is the same as that of previously described feedback units once it is noted that the coupling agent between the two attenuators, which in the case of the servo multiplier was the angular rotation of the shaft and in the case of the strain gauge multiplier a coil movement, here takes the form of the temperature of the glass bead.

The system is very simple though it lacks speed of

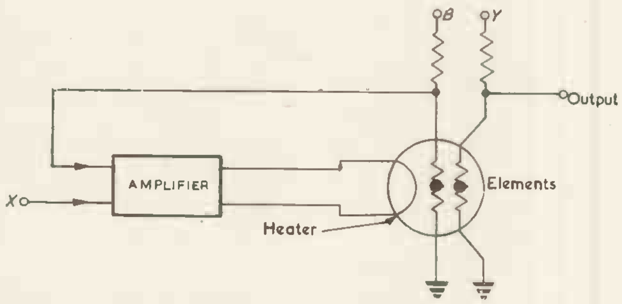


Fig. 11. Thermistor multiplier

response. The accuracy of the system depends entirely on the design and manufacture of the bead.

The Step Multiplier

This is a complex system which, although analogue in form, derives a great deal from the digital computer techniques[19]. Like the digital computers it is accurate, stable and relatively complicated. Fig. 12 gives a block diagram of the multiplier.

The variable attenuators each consist of 11 T-type resistance networks. The values of these networks are arranged in powers of two, i.e. 2, 4, . . . 1 024. These networks are switched on by each stage of a binary counter. The total conductance of these circuits corresponds, therefore to a count stored in the counter at any time. The counter is controlled by a gate circuit which can add or subtract according to the value of the input voltage e. The voltage e_1 is the sum of the voltages; the variable X, C and $-K$ which are fed into the amplifier A_1 via the conductances g, G_1 as shown in the diagram. Therefore, for the balanced condition of the amplifier and counter:

$$G_1 = \frac{(X + C)g}{K}$$

The second variable Y is applied to the amplifier A_2 and after its inversion fed into the amplifier A_3 via the variable conductance G_2. Therefore, the output of the amplifier A_3

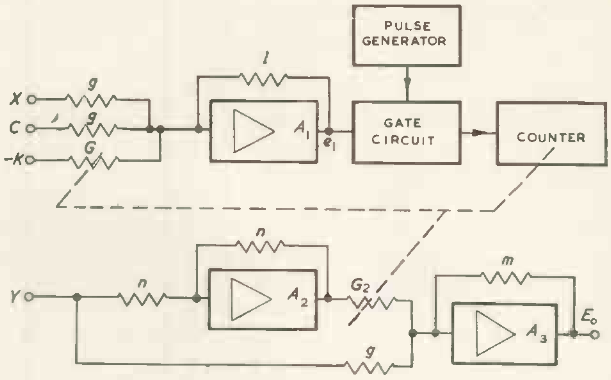


Fig. 12. Step multiplier

is equal to:

$$E_o = YX\,(g/K)$$

if G_1 is equal to G_2. This arrangement of the multiplier permits both input voltages to be of either polarity. The accuracy is of the order 1/2 000, the time of computation is about one second. About a hundred valves are used, in addition to twenty-two special high speed relays.

Pulse Feedback Multiplier

This type of multiplier is a development of the simple pulse multiplier described in the first part of this article. The two attenuators which are common to all types of feedback multipliers here take the form of two electronic switches which are both operated by the same pulse generator (see Fig. 13). The coefficient of attenuation K is defined as before:

$$K = (t/T)$$

where t = time during which the switch is closed.

T = time between two pulses.

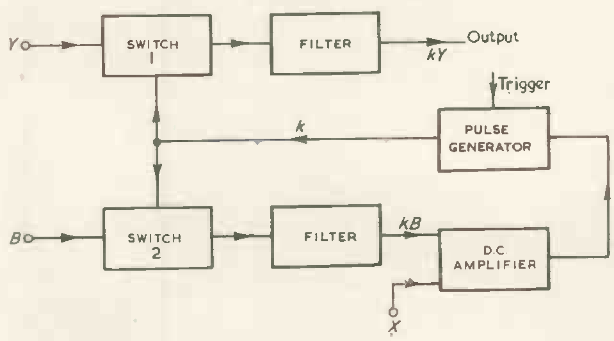
The pulse generator is triggered at a constant frequency and produces pulses of variable length, depending on the value of the controlling voltage E. A constant voltage B is applied to the switch 2 and after being attenuated and smoothed, is fed to the differential d.c. amplifier to which a voltage corresponding to one of the variables X is also applied. The output voltage from this multiplier controls the pulse length produced by the pulse generator and hence the coefficient K. Assuming the gain of the amplifier to be high enough, we have for the balanced condition:

$$BK = X$$

$$\text{i.e. } K = (X/B)$$

The variable Y is fed to the second switch, the output of

Fig. 13. Feedback pulse multiplier



which is, therefore:

$$V_{out} = KY$$
$$\text{i.e.} \quad = (1/B)XY$$

it must be noted here that, contrary to the simple pulse multiplier, the pulse generator is not required to have linear characteristics in relation to the controlling voltage E. In fact, it often takes the form of a simple delayed multivibrator[20] which is only approximately linear. The accuracy of the system depends mostly on the accuracy and stability of the electronic switches. Various types of such switches have been designed and the reader may refer to numerous works on the subject such as refs. 1, 2. For relatively slow speed of operation, high speed relays may be used. Another difficulty of the system is the design of the smoothing filters which are the main factor slowing down the speed of response of the multiplier. The best response is obtained at the highest possible pulse repetition frequency. The limitation of the simple system described is the difficulty with which the output can approach the zero value. This is due to two factors:

(1) The necessity of making the controlling pulse very short.

(2) The error which exists in all electronic switches when the input voltage is very small.

To overcome these difficulties a more complicated system can be built which is capable of four quadrant working. The block diagram of such a system[11] is similar to Fig. 12 except in so far that the variable conductances are substituted by switches and the counter which controls them by a pulse generator.

Single Attenuator Feedback Multiplier

The main difficulty of all the feedback multipliers discussed so far is that of finding two attenuators of exactly the same characteristics. This difficulty may be solved by the use of the same controlled attenuator to perform the function of both attenuators in the circuits previously described. As an example of this method of approach we may consider a double carrier frequency multiplier[1], the block diagram of which is given in Fig. 14. The principle of operation is as follows: two oscillators of frequency f_1 and f_2 respectively, are controlled by d.c. voltages in such a way that the amplitude of the generated carrier waveform is proportional to the d.c. voltage applied. The two carrier voltages of amplitudes B and X are fed then to the same variable gain amplifier which consists of a variable-mu pentode. The output of this amplifier will have two components of frequency f_1 and f_2 and of amplitudes KB and KX, where K is the gain of the amplifier. These are then applied to two filters tuned to the frequencies f_1 and f_2 respectively. The output of the filter 1 is then rectified and applied to a differential amplifier where it is subtracted from a voltage Y. The output of this amplifier is fed as the d.c. bias voltage to the variable gain amplifier.

The output voltage is obtained from the second filter and rectifier. By analogy with the previously described method it follows that this voltage:

$$V_{out} = KX$$
$$\text{i.e.} \quad = XY/B$$

Time Division Method

Another method by which the principle of a single variable attenuator can be used is that of time division (see Fig. 15). In this system the input of the variable gain amplifier is switched in turn to two voltages X and B which in this case can be of the same frequency or even d.c. Another switch running in synchronism with the first switches the output of the amplifier between the feedback loop and the output network.

It may be noticed that this method may be applied to any of the previously described feedback methods. The difficulty of the system is to find suitable switches. The speed of response is limited mostly by the frequency of operation of the switches.

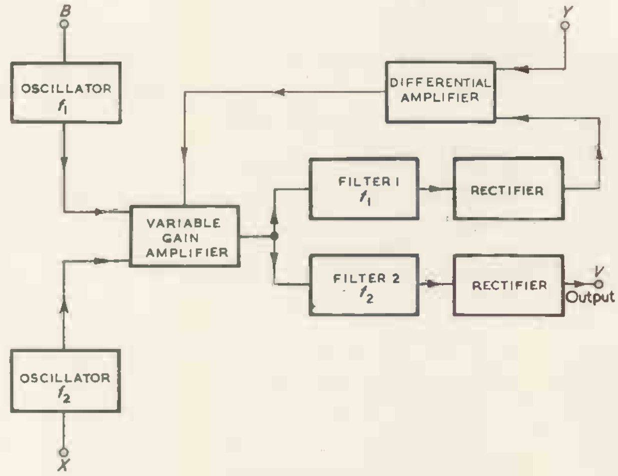


Fig. 14. Double carrier frequency multiplier

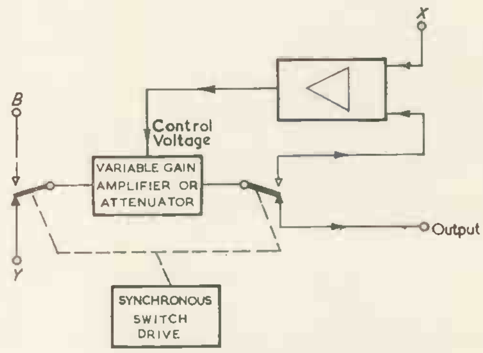


Fig. 15. Single attenuator—switching principle

Conclusions

Various methods which can be used for multiplication have been outlined, but the development work in this interesting field is still going on and many new ideas will, no doubt, appear.

Most of the methods described have advantages which make them suitable for a particular project. Table 1 gives a brief list of properties of the systems described. It is intended to give some idea of the possibilities of the systems and is only roughly accurate. In most cases the ideas may be improved by further development along the lines indicated.

From the accuracy point of view, it may be noticed that the feedback multipliers achieve accuracy above that obtained with direct methods; with the exception of the time modulation method where considerable accuracy is obtained by careful circuit design. Fortunately, the methods developed for radar measurements are easily adaptable for time modulation. It must be noticed, too, that all feedback methods will perform a division in addition to multiplica-

TABLE 1

Properties of the systems described

TYPE OF MULTIPLIER	TYPICAL ACCURACY (PER CENT)	SPEED OF RESPONSE	NUMBER OF QUADRANTS IN THE SIMPLEST FORM	COMPLEXITY OF APPARATUS
Time Modulation	0·5–1	200c/s	1	2–6 valves, no special components.
Frequency and Amplitude Modulation	1	Up to 30kc/s	2	5–10 valves, careful design and construction necessary if good accuracy is to be obtained.
Coincidence Multipliers	1–2	500c/s	1	About 7 valves. No special components. No critical adjustment.
Squaring Multiplier	1–2	200kc/s	1	About 6–10 valves. Either non-linear elements (specially adjusted valves are used) or another 20 diodes are needed.
Logarithmic Multiplier	3	200kc/s	1	Special non-linear elements are needed.
C.R.T. Multiplier	1–2	5–100kc/s	4	Special tube and auxiliary gear of about 10 valves.
Valve Characteristic Multiplier	5	200kc/s	1	Few valves but pre-selection often needed. Must be reset rather frequently.
Strain Gauge Multiplier	2	300c/s	4	Specially constructed electromagnetic device.
Step Multiplier	0·1	2c/s	4	About 100 valves, special relays.
Thermistor Amplifier	2	(A) 1·5c/s (B) 1Mc/s	2	One simple amplifier. Special thermistor unit.
Time Modulation Feedback Multiplier	0·1–0·5	200c/s	1	About 7 valves for the simplest instrument.
2 Carrier Frequencies Multiplier	1	20kc/s	2	About 10–15 valves.
Time Division Multiplier	0·5	50c/s	2	About 10 valves.

tion by simply applying a varying voltage instead of the constant voltage *B*.

The speed of the direct multipliers is usually of the same order for both channels. The feedback amplifiers on the other hand, have the speed of response limited for one channel because of the slow response of the feedback loop, whereas the second channel is very fast, except in the case of pulsed attenuators. This variation can be used with advantage in many applications. The polarity of the inputs (number of quadrants in the working range) is given only for the simplest system outlined. Usually the systems can be modified to accommodate both inputs of any polarity. This modification, however, adds expense, complexity and produces inaccuracy in the zero signal region.

It may be noticed that the multiplying apparatus need not necessarily be very complex, but some of the simplest systems involve the use of special components which are not easily obtainable.

Of the devices using non-linear elements, only the approximation to curves by means of biased-off diodes is accurate and stable enough to render the method useful for general analogue computer practice. In the opinion of the author, however, the application of non-linear elements has a great future in the feedback multipliers. Whereas it is very difficult to produce commercially a non-linear element of accurately determined and predictable characteristics, it is much easier to produce such elements in matched pairs. The only requirement here is the similarity of the characteristics of such a pair. Apart from the thermistor elements which were mentioned, pairs of non-linear resistors such as thyrite could be used with the same d.c. flowing through them to produce attenuation of a.c. computing signals. Pairs of current controlled magnetic transductors were also used for multiplication. Another interesting possibility is the use of voltage controlled capacitors employing a variable-*K* dielectric.

Acknowledgment

The author wishes to thank Professor d'Ombrain, Head of the Electrical Engineering Department at Battersea Polytechnic, for his suggestions and guidance in the research work on which this article is based.

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The Production of a Short Duration Pulse of High Velocity Electrons

By D. H. Le Croissette*, M.Sc., A.Inst.P.

This apparatus produces a pulse of electrons of duration 0·2μsec *suitable for initiating ionization in gas-filled systems. The pulse is repeated 50 times per second occurring during the time-base sweep of an oscillograph, so enabling the simultaneous display of the ionizing pulse and the resulting discharge build-up.*

In order to measure time delays and electron transit times in gas discharge systems it is usual to produce a short burst of ionizing particles or radiation and display the resulting discharge build-up on an oscillograph, the time-base of which is initiated by the ionizing pulse. Measurements of the time between the initial pulse and any portion of the waveform of the discharge have to allow for any delay occurring before the time-base commences operating and also for any non-linearity at the beginning of time-base sweep. Both these possible sources of error may be eliminated by producing a pulse of ionizing particles after the time-base has been initiated, so that the initial pulse and the discharge build-up may be displayed on the same sweep. One method of achieving this is to trigger the time-base of the oscillograph at a convenient repetition frequency and synchronize the rest of the apparatus from the time-base output. This method has been used here to produce a pulse of electrons of about 40keV energy repeated 50 times per second, the duration of each pulse being about 0·2μsec. The position of the pulse relative to the beginning of the time-base sweep may readily be adjusted.

Apparatus

A schematic diagram of the equipment is shown in Fig. 1. The oscillograph used is a Nagard type L103, having a maximum time-base sweep speed of 2in/μsec. The time-base is triggered by a multivibrator locked to the 50c/s mains supply. This provides a convenient operating frequency and has the further advantage of minimizing mains hum.

The electrons are produced from a tungsten filament operated in a continuously evacuated system. A triode gun is used with magnetic focusing and deflexion, the anode being at earth potential. The filament is heated by a silver-nickel 1·5V cell contained in an anti-corona mounting at 60kV. This potential is supplied from a radio frequency e.h.t. generator employing a voltage tripler rectifying circuit. The deflexion system is arranged so that the electron beam is deflected in unison with the spot on the cathode-ray tube of the oscillograph. The pulse is produced by allowing the electron beam to pass through a small aperture in the steel end face of the electron tube. In this way, exact synchronization between the time-base sweep and the pulse is achieved.

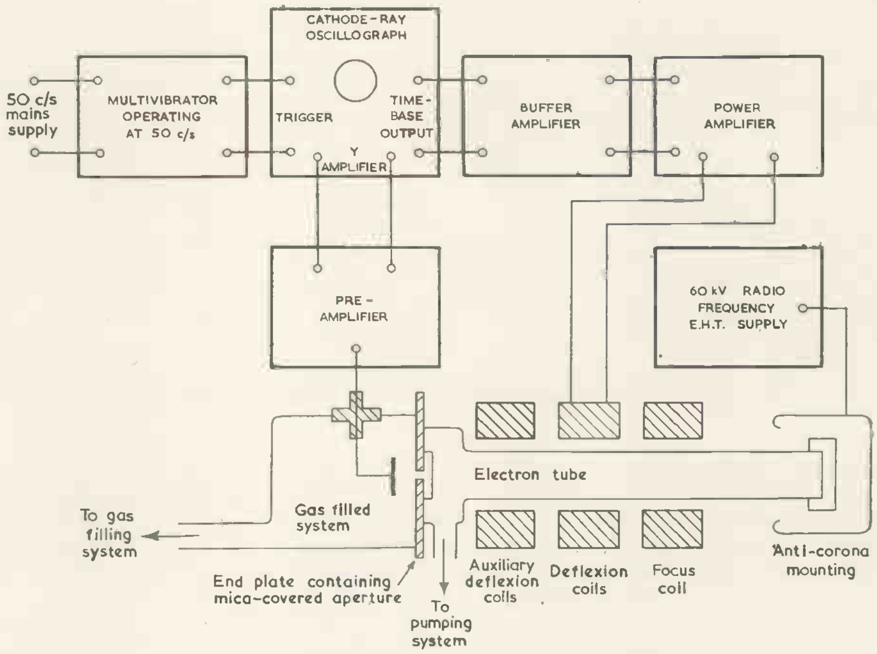


Fig. 1. Schematic diagram of the apparatus

Operation of the Circuit

The negative-going sawtooth voltage from one of the X plates of the oscillograph is fed to the input of a two-valve buffer circuit shown in Fig. 2. Valve V_1 is operated with a bias of 6V. As soon as the time-base commences to operate, the grid of V_1 goes negative and the valve is cut off in about 10 per cent of the sweep time. The anode voltage of this valve rises by about 100V to the h.t. potential in the same time, remaining at this value for the duration of the sweep. The cathode-follower output stage delivers a voltage of the form shown in Fig. 3(b) across a cathode load of 200Ω.

Standard television deflexion coils are used to deflect the electron beam. A rate of change of current of 0·1A/μsec through the deflecting coils is required to produce a deflexion rate of 1cm/μsec past the aperture. This current is supplied by the two valves V_3 and V_4 (Fig. 2) used as electronic switches. These two valves are operated in parallel and are biased to pass a total current of about 10mA in the absence of any input signal. The rapidly rising input pulse sends the grids of these valves positive and permits a large anode current to flow. The rate of rise of the total anode current (di/dt) is limited by the voltage

* Physics Department, The University of Southampton.

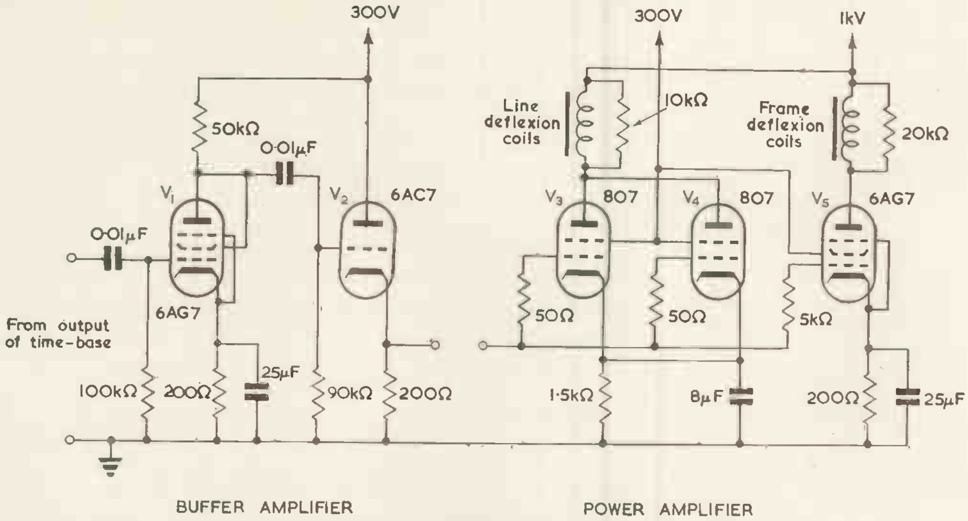


Fig. 2. Circuit of buffer and power amplifiers

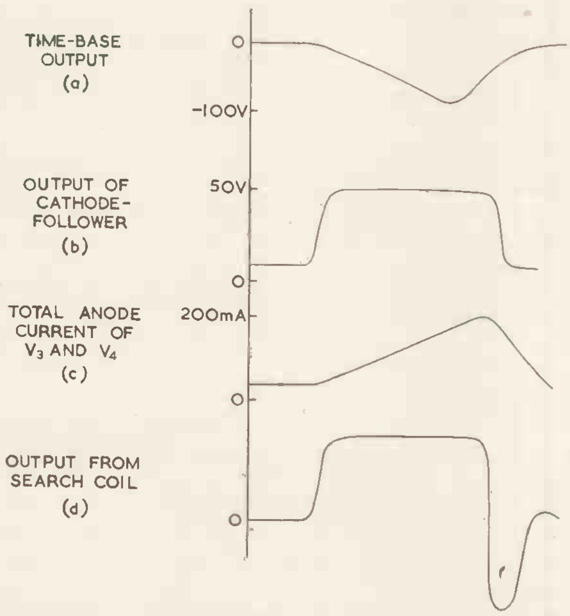


Fig. 3. Circuit waveforms when the time-base is operating with a sweep time of 2μsec

applied to the valves (1kV) and the inductance (L) of the line deflexion coils (9mH). As 100V across the valves is needed to bring the operating point above the knee of their $I_a - V_a$ characteristic, there is 900V available to offset the back e.m.f. across the coils.

Hence:

$$E = L \,.\, di/dt$$

where E = 900V, giving:

$$di/dt = (E/L) = 10^5 \text{A/sec.}$$

Fig. 3(c) shows the variation in the total anode current of valves V_3 and V_4 when the time-base sweep has a duration of 2μsec or less. The rate of change of anode current remains constant over this period of time without an excessive current being taken from the valves. For longer sweep durations, however, the rate of change of anode current attains this value only at the beginning of the sweep and then drops to a lower value. The duration of the electron pulse is proportional to the rate of change of deflexion of the electron beam past the aperture. The deflexion of the electron beam is proportional to the magnetic flux produced by the deflexion coils, which is in turn proportional to the anode current of the output valves. In order to produce a pulse of the shortest possible duration, therefore, the beam must pass the aperture when it is being deflected at its greatest rate. A knowledge of the rate of change of anode current during the time-base sweep enables this to be achieved. This information may be obtained by placing a search coil in the magnetic flux of the deflexion coils. The e.m.f. developed across the search coil is given by:

$$e = n \,.\, A \,.\, dB/dt \text{ volts}$$

where A is the area of the search coil in square metres,

n is the number of turns on the search coil,

B is the flux density in webers per square metre.

The output from the search coil may be displayed on the oscillograph screen as shown in Fig. 3(d). For a time-base sweep of up to 2μsec duration, the output e.m.f. of the search coil is nearly constant. Since the flux density is proportional to the current through the deflexion coils, the e.m.f. developed across the search coil at any time is proportional to the rate of deflexion of the electron beam. This e.m.f. is inversely proportional to the duration of the pulse, if the beam is crossing the aperture at that time.

Focusing and Collection of the Electron Pulse

The electron beam is focused by a combination of a permanent magnet ring and an electromagnetic focusing coil. Adjustment of the beam relative to the aperture is obtained by using auxiliary deflexion coils arranged at right-angles and fed from a variable d.c. source. The location of the pulse along the time-base sweep can be varied by these controls. To avoid a second pulse being produced on the flyback stroke of the time-base, the frame coils of the television deflexion coil system are fed from valve V_5 as shown in Fig. 2. The different time-constants of the circuits feeding the frame and line coils produce an elliptical beam path, ensuring that the aperture is traversed only once every cycle of operations.

The electron pulse has been used to operate a gas filled counter system. For this reason, a thin mica window is placed in front of the aperture in the steel end plate separating the electron tube from the counter. The weight of the mica is about 1mg/cm² and thus the energy loss in penetrating the window is about 20keV. The joint between the mica and the steel plate is made vacuum tight by an O-ring seal as shown in Fig. 4. The aperture in the steel plate is 0·5mm diameter. Care is needed when evacuating the systems on either side of the window to avoid fracturing the mica. Providing the air is not pumped out or admitted too quickly, the systems may be operated with a pressure difference of one atmosphere.

A beam current of up to 100μA has been used. The pulse duration is then about 0·2μsec. This is determined by the rate of deflexion of the electron beam past the aperture and the sum of the diameter of the aperture and the

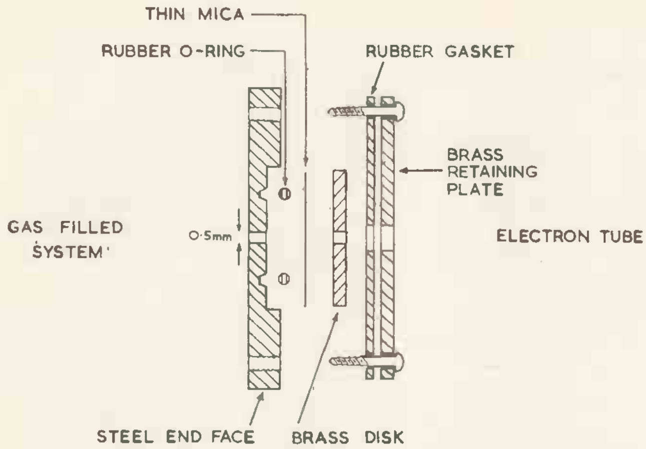
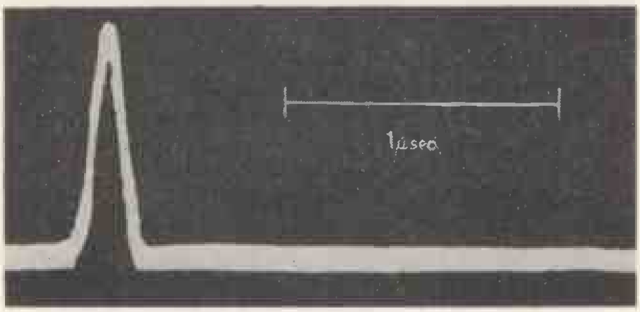


Fig. 4. Exploded view of the seal between the electron tube and the gas-filled system

Fig. 5. Output pulse as viewed on the oscillograph screen



diameter of the focused beam. No attempt has been made to reduce this time since the rise time of the electronic detecting equipment is about 0·09μsec.

The pulse is detected on a small collecting electrode connected via a cathode-follower to a pre-amplifier and the main oscillograph amplifier, the total gain of the system being 5×10^4. The time-constants of the input circuit are made smaller than the rise time of the pulse. Hence the voltage variation of the collecting electrode as displayed on the oscillograph is proportional to the current flow during the pulse. The rate of rise and fall of the pulse as viewed on the screen is determined by the profile of the focused beam and the rise and fall times of the amplifying system.

Application

This apparatus is being used for measurements on Geiger-Müller counters. The electron pulse is arranged to ionize in a region close to the cathode of the g.m. tube. A small electrode collects a fraction of the incident pulse which is amplified and displayed on the oscillograph. After an interval determined by the transit time of the counter, the pulse obtained from the anode wire of the g.m. tube occurs. Thus the transit time can be obtained from the calibrated time-base sweep. By measurement of the relative height of the two pulses a calculation of the gas amplification can be made when the counter is operated in the proportional region.

Acknowledgments

The author wishes to thank Professor A. M. Taylor for his encouragement of this work. The electron tube system was kindly supplied by Mr. G. A. R. Tomes, Managing Director of 20th Century Electronics Ltd. who also gave valuable advice.

A Flutter Simulator

A Mathematical Services Group comprising nine senior mathematicians was formed last September by Bristol Aircraft Ltd.

Among the equipment with which the Group has been supplied are a DEUCE* computer and a flutter simulator.

The DEUCE computer is being employed principally on problems relating to performance data of the Bristol Britannia on prescribed routes and to the computation of trajectories of guided missiles.

The Flutter Simulator, which was designed by Bristol Aircraft Ltd, is an analogue machine for the determination of critical flutter speed in aircraft.

With the aid of the simulator, the problem of flutter is solved by expressing as a list of equations the various forces acting on the aircraft tending to displace it. The displacement is complete but it can be simplified by assuming it to contain no more than six degrees of freedom.

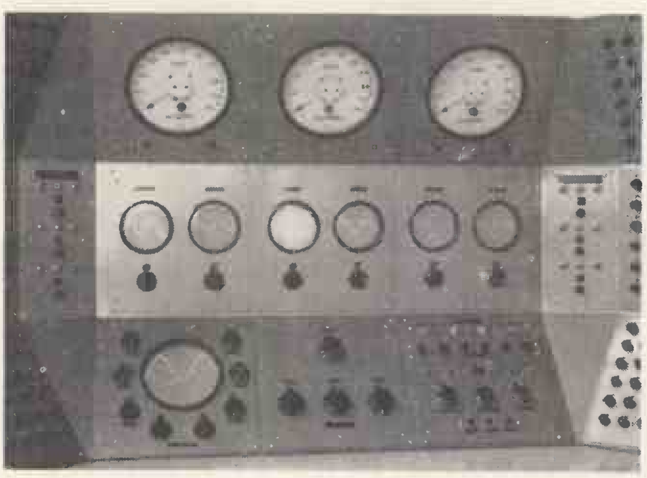
The six equations and the numerical values of the coefficients representing the forces on the aircraft are set up on the simulator and the results are displayed on six cathode-ray monitor tubes.

The control representing the forward speed of the aircraft can be varied until flutter conditions are indicated by the waveforms on the monitor tubes and the actual flutter speed is thus readily obtained.

The photograph shows the console of the flutter simulator with a cathode-ray tube monitor for each of the six degrees of freedom. Any two of these waveforms may be simultaneously displayed on the large cathode-ray tube shown on the lower left-hand side of the console.

The value of the simulator lies in the fact that it helps the designer to ensure, while his aircraft is still in the drawing board stage, that it will be completely free from flutter when it flies. If any modification is needed to a component as it is first designed, the simulator will tell him and the modification can be effected merely by altering a drawing.

The control desk of the flutter simulator



* DEUCE (A New Digital Computer). *Electronic Engng.* 27, 179 (1955).

Short News Items

The President and members of the Council of The Institution of Electrical Engineers recently received over 2 000 members and guests at a conversazione held at the Royal Festival Hall, London. More than one hundred exhibits and demonstrations of advances in electrical science and engineering and their applications were on show. Demonstrations of closed-circuit television applications, of the use of electro-magnetic pumps for pumping liquid metals, as is required in certain forms of nuclear reactors, were given, and the Science Museum, in collaboration with the Post Office, presented a display showing the history of transatlantic cable communication.

The Royal Military College of Science at Shrivenham recently held an open day when the work of teaching and research into the sciences and technologies appropriate to the equipment of the Army was on display.

Pye Marine Ltd have been awarded a contract by the G.P.O. to supply and install an f.m. v.h.f. coastal radio telephone station. The first of its kind to be set up in Britain, this station will be in operation on the Clyde in the autumn and should be of great assistance to shipping and trade. The new system will enable ships to be in swift and direct contact with harbour masters and dock superintendents, and vice versa. Clyde shipbuilders conducting trial runs will be able to save hours and even days in evaluating results.

The board of Electric and Musical Industries Ltd announce the appointment of Mr. Clifford Metcalfe to the post of Managing Director of E.M.I. Electronics Ltd. This company, on behalf of the E.M.I. Group, will take over the full range of research, development and manufacture on Government contracts and commercial electronics.

Mr. R. F. Holtz has been appointed General Manager of Laboratories RCA Ltd, Zurich. Mr. C. G. Mayer, who established the Swiss company, has become Managing Director and Chairman of the board of RCA Great Britain Ltd, with headquarters in England. He will continue to be available as Consultant to the Laboratories.

Plannair Ltd announce a new licence agreement for the manufacture of their high efficiency axial flow blowers in France and this will result in the distribution of these products becoming virtually world-wide. The agreement is with Societé Air Equipement, 18 Rue Basly, Asnières (Seine), France.

Armstrong Whitworth Aircraft Ltd, of Coventry, have embarked on an extensive expansion and development programme to complete the establishment of a fully equipped aircraft and guided missile research and design centre on the site of their plant at Whitley, near Coventry. The centre will include design offices, laboratories and experimental shops. Equipment on order includes additional electronic brains and pressure chamber facilities. A new mechanical test laboratory will have twice the space and equipment as the existing facility at Baginton.

Marconi's Wireless Telegraph Co Ltd announce the appointment of Mr. V. J. Cooper to the newly created post of Chief Television Engineer. In addition, three distinct Television Development Groups and one Audio Development Group have been formed, under Mr. E. Davies, Mr. N. N. Parker-Smith, Mr. J. E. Nixon and Mr. S. J. Gooderham, each responsible to the Chief Television Engineer.

At the Damascus International Fair to be held from 1-30 September, seven leading British radio and television manufacturers are taking part in a composite stand under the auspices of the British Radio Equipment Manufacturers' Association. Other manufacturers of radio and electronic equipment are taking part in the Fair independently of the BREMA composite stand.

Hilger & Watts announce that the divisional offices at 48 Addington Square, London, S.E.5, are now closed and the sales and administrative departments of the company are now combined at Camden Town, the address being 90 St. Pancras Way, Camden Road, London, N.W.1. Additional telephone lines have been provided.

The South East London Technical College, Department of Electrical Engineering, announce various courses, beginning in October, on High Voltage Engineering, Electrical Engineering Economics, Communication Engineering Economics, Vector Analysis and Fundamental Electromagnetic Theory, Operational Calculus with Applications to Electric Circuit Theory, Optical Instruments, Electric Power Transmission Calculations. Further details may be obtained from the Head of the Department of Electrical Engineering and Applied Physics, L.C.C. South East London Technical College, Lewisham Way, London, S.E.4.

The Council of the Royal Society has appointed Professor E. N. da C. Andrade, F.R.S., to deliver the Rutherford Memorial Lecture for 1957 in Australia, under the terms of the scheme to commemorate the late Lord Rutherford of Nelson.

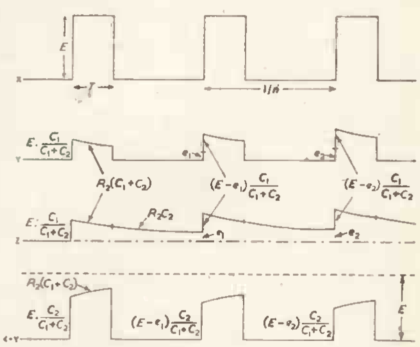
Racal Engineering Ltd have been awarded a contract by the National Physical Laboratory for the design and supply of a special purpose electronic timing and recording equipment. This equipment is to be used by the National Physical Laboratory with their shock tube for aerodynamic research into flight at very high speeds and is to give precise measurements of the speed of propagation of the shock wave at various points in the tube.

Radiovisor Parent Ltd, manufacturers of photo-electric and electronic controls, have recently moved their complete organization from 1 Stanhope Street, London, N.W.1, to the new factory at Stanhope Works, High Path, London, S.W.19.

Aero Research Ltd, of Duxford, Cambridge, have changed their telephone number to Sawston 2121. The Telex number is 10-101.

Craven Electronics Ltd is the new name for the Craven Electronic Instrument Co. The registered office is at 184/5 Swan Arcade, Bradford 1.

Errata. In the article "A Diode Pump Integrator" by J. B. Earnshaw (January issue, p. 26), the waveform *X-Y* in Fig. 7 was drawn and calculated wrongly. The correct version is printed below.



In the Savage Transformers Ltd advertisements in the May and July issues (pp. 116 and 120) the telephone number was given as Devizes 93. This should be Devizes 932.

LETTERS TO THE EDITOR

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

High Impedance Current Generators

Dear Sir,—A rather different application of high impedance current generators, similar to those described by McGuire[1] and Clarke[2], is in high voltage power supply stabilizers, particularly

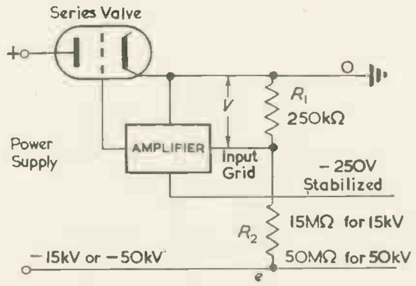


Fig. 1.

$$V = e R_1$$
$$V = R_1 + R_2$$

where the positive line is at earth potential.

In such a power supply, series stabilization of the output voltage is preferably done near earth to avoid insulation difficulties in the associated amplifier. Using a resistance divider network as in Fig. 1, there is a large attenuation to signal inputs. Taking as typical, the values in the diagram, it is 60 and 200 (voltage ratio) for 15kV and 50kV respectively.

Replacing R_1 by a circuit such as Fig. 2, an effective impedance of 10MΩ may be obtained for the same voltage drop as across R_1. This reduces the attenuation to less than 3 for 15kV and 6 in the 50kV case. Using the circuit in Fig. 3, an effective impedance of well over 100MΩ may be obtained.

Yet another application which appears interesting is that of using a circuit of this type as the anode load of a pentode,

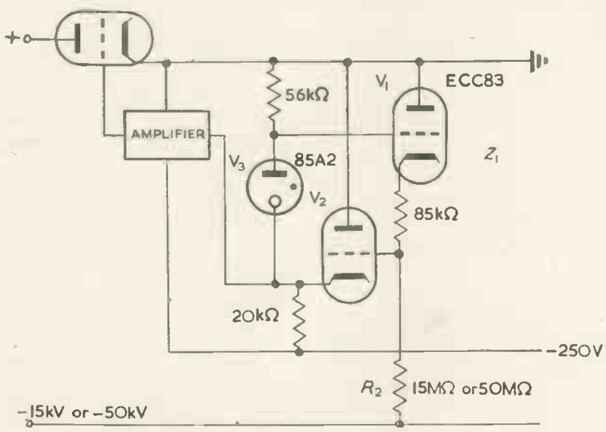


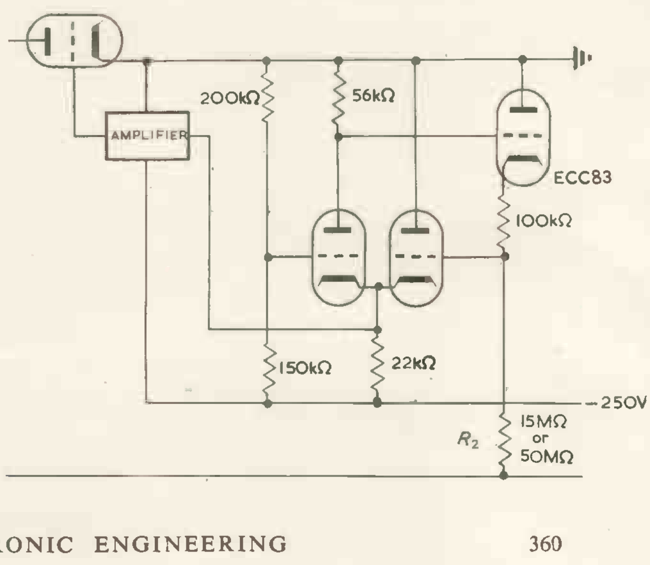
Fig. 2.

$$Z_1 = (\mu G + 1) R_k$$

G = *gain from grid* V_2 *to grid* V_1

Fig. 3.

$$G = 50$$
$$Z_1 = (\mu G + 1) R_k$$
$$= (100 \times 50 + 1)\ 100\text{k}\Omega$$
$$= 500\text{M}\Omega$$



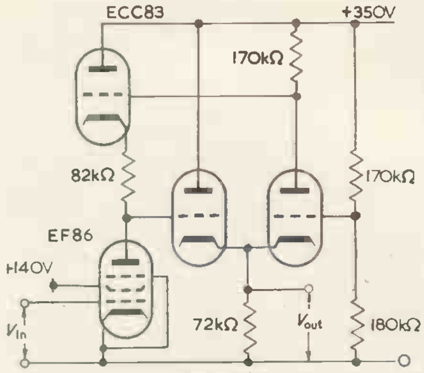


Fig. 4.

$$\mu \text{ of } EF86 = g_m r_a = 1{\cdot}8 \times 10^{-3} \times 2{\cdot}5 \times 10^{6} = 4500$$

$$\frac{V_{out}}{V_{in}} = \frac{\mu Z}{r_a + Z} \quad Z \gg r_a$$

$$\therefore \simeq \mu. \qquad g_m = 1{\cdot}8\text{mA/V} \quad r_a = 2{\cdot}5\text{M}\Omega$$

as shown in Fig. 4. Here we can utilize nearly the full μ of a pentode, which may be as high as 5 000; the output impedance is low, being that of the cathode circuit of the cathode-coupled amplifier. A limitation of this circuit, however, is the anode capacitance of the pentode which restricts the frequency response.

Yours faithfully,

B. G. L. Braun,
Great Malvern,
Worcs.

REFERENCES

1. McGuire, J. H. Some High Impedance Current Generating Circuits. *Electronic Engng. 27*, 529 (1955).
2. Clarke, L. N. Letter to the Editor. *Electronic Engng. 28*, 223 (1956).

A Modified Twin-T Oscillator

Dear Sir,—Your correspondent, Mr. H. L. Armstrong describes an interesting low frequency oscillator with electronic frequency control (April 1956 issue page 179), based on the use of suppressor grid modulation of a pentode to vary the amount of cathode current which reaches the anode circuit. He obtains a frequency swing of a few per cent and suggests the principle may have other applications.

He may be interested to know that K. C. Johnson[1] has described a single valve frequency modulated oscillator using a tuned *LC* circuit with the same principle of modulation. Johnson claimed ± 15 per cent frequency swing at central frequencies up to 10Mc/s using an EF50.

Yours faithfully,

W. H. P. Leslie,
Department of Scientific and Industrial Research,
Mechanical Engineering Research Laboratory.

REFERENCE

1. Johnson, K. C. Single Valve Frequency Modulated Oscillators. *Wireless World, 55*, 121 and 168, (1949).

The correspondent replies:

Dear Sir,—When my comment on the frequency-modulated twin-T oscillator was written, I was not aware of Mr. Johnson's article. I appreciate Mr.

Leslie's bringing it to my attention, for the circuit described seems to be more generally useful, especially in that it goes to higher frequencies.

Yours faithfully,

H. L. Armstrong,
National Research Council, of Canada,
Ottawa.

Multivibrator Circuits Using Junction Transistors

Dear Sir,—Mr. A. E. Jackets in the introduction to his article "Multivibrator Circuits Using Junction Transistors" (May, 1956) states: "It is better when designing the circuits to approach the problem from an understanding of the mode of operation of the transistors." Yet in the article only a passing reference is made to the collector leakage current I_{co} which plays an important part in the design of these circuits.

Due to change of I_{co} with temperature the frequency of oscillation of the circuit in Fig. 7 changes from 760c/s at 20°C to 865c/s at 45°C.

The expression $t = CR \ln\left(\dfrac{V_B + E}{V_B}\right)$

can be modified to include I_{co}, since I_{co} flows in the base circuit of the transistor during the timing period; the modified expression being

$$t = CR \ln\left(\frac{V_B + E + I_{co}R}{V_B + I_{co}R}\right)$$

Table 1 shows some practical results using the circuit of Fig. 7 together with the results predicted by the above equation, making the assumption that $V_B = E = V$ supply.

TABLE 1

		TRANSISTOR 1	TRANSISTOR 2	COMPLETE MULTIVIBRATOR
20°C	I_{co}	2·5µA	2·0µA	
	calculated t	678µsec	681µsec	1.359msec
	,, f			735c/s
	observed f			760c/s
45°C	I_{co}	20	16	
	calculated t	588	604	1·192msec
	,, f			840c/s
	observed f			865c/s

These results show that the variation of frequency with temperature can be predicted. The differences between the predicted and observed values are due to assuming that $E = V_B$ which neglects the bottoming voltage and the base-emitter voltage of the transistor.

Also the discrepancies above 10kc/s which Mr. Jackets describes as not fully understood can in fact be accounted for by the voltage drop across the small cross coupling capacitors needed at these speeds. The situation is made worse by the small capacitors reducing the loop gain and hence reducing the switching speed.

Yours faithfully,

L. P. Morgan,
Mullard Research Laboratories,
Surrey.

The Author replies:

Dear Sir,—In reply to Mr. Morgan I agree that I_{co} has an effect on frequency and that it can be included in the expression for the timing period in the way he suggests. Although I did not include this in detail in my article, I pointed out that the effect of I_{co} was most pronounced at high values of R. The value of R should, therefore, be kept as low as possible for high temperature operation.

It is interesting to consider what happens to the frequency drift of a circuit operating at a nominal frequency of 735c/s when the value of R is reduced. The value of C must of course be changed as well; in practice it is also desirable that the value of R_L be reduced in proportion, in order to maintain a good waveform. Assuming that the two transistors have an I_{co} of 2·5µA at 20°C, and that this rises to 20µA at 45°C, the following results are obtained for the frequency drift of the circuit of Fig. 7, if $E = V_B = 8$V.

(1) With $R = 100\text{k}\Omega$ the change of frequency with temperature due to the increase of I_{co} is approximately 15 per cent.

(2) With $R = 22\text{k}\Omega$ this change is reduced to about 3 per cent.

Another interesting point is the effect of V_B on the stability of frequency with temperature. If V_B is made much greater than the value of $I_{co} R$ at the maximum temperature, the effect of temperature can be made negligible. Thus frequency stability can always be improved by increasing the base supply voltage.

The explanation by Mr. Morgan of the discrepancies above 10kc/s appears rather vague. I think these discrepancies can be largely attributed to a discharge of the coupling capacitor through the initially low input resistance of the transistor, at the instant of switching.

A. E. Jackets,
Research Laboratories,
The General Electric Co. Ltd.,
Wembley.

Dear Sir,—In his article on Junction Transistor Multivibrator Circuits (May 1956), Mr. Jackets is rather misleading about the effects of the temperature sensitive element, I_{co}. In his circuit, shown in Fig. A, after having reached the conditions where X_2 is non-conducting and X_1 is bottomed, he defines the maximum voltage to which C_2 can charge by the equation:

$$E = V_c - R_L I_{co} \quad \text{............(1)}$$

and thence proceeds to derive the expression for frequency:

$$f = \frac{1}{2R_L C \ln\left[\dfrac{V_B + E}{V_B}\right]} \quad \text{............(2)}$$

which contains the term E (a function of I_{co}) and is thus extremely temperature sensitive.

In fact, when X_1 is bottomed, a large positive voltage is applied to the base of X_2 with respect to its emitter so that the current I_{co} is backed off. Equation (1)

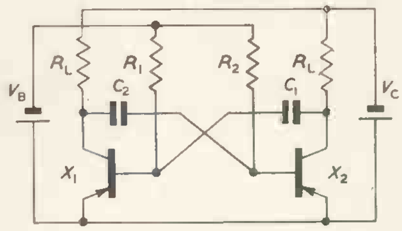


Fig. A.

thus becomes simply:

$$E = V_c$$

and the expression for the frequency

$$f = \frac{1}{2RC \ln\left[\dfrac{V_B + V_C}{V_B}\right]} \quad \text{............(3)}$$

As Mr. Jackets states, the ratio V_c/V_B is usually unity in practice and equation (3) may be written:

$$f = \frac{1}{2RC \ln 2} \simeq \frac{1}{1{\cdot}4RC}$$

This discussion indicates the inherent temperature stability of the type of transistor circuit employing transistors in the switched condition.

Yours faithfully,

A. McKeon,
D.H. Propellers,
Hatfield.

The Author replies:

Dear Sir,—Mr. McKeon is incorrect when he states that the leakage current I_{co} of a junction transistor may be "backed off" by making the base positive with respect to the emitter, for in a transistor type EW53 (and also in similar units) the value of I_{co} is almost independent of the d.c. voltage applied across the collector junction over the range 0 to 15V. Thus even if the base is positive with respect to the emitter, a small leakage current I_{co} (a few microamps, at 20°C) will flow.

So in practice the value of $R_L I_{co}$ is very small compared with V_c even at high temperatures, and the effect of temperature upon E is usually negligible.

I wonder if Mr. McKeon has not confused I_{co} with another parameter known as I_{co}' which is sometimes quoted in transistor literature. I_{co} is the leakage current flowing between collector and base when the collector is biased negatively and when the emitter is open circuit. I_{co}' is the leakage current flowing between collector and emitter when the base is open-circuit; this latter condition is seldom of importance in circuits of this type. I_{co}' is related approximately to I_{co} by the expression.

$$I_{co}' \simeq [1/(1-\alpha)]I_{co}$$

The effect of I_{co} and hence of temperature on frequency may be seen more fully from the letter written by Mr. L. P. Morgan and my reply.

Yours faithfully,

A. E. Jackets.

BOOK REVIEWS

The Services Textbook of Radio, Volume 3 Electronics

By J. Thomson. 225 pp., 60 figs. Demy 8vo. H.M. Stationery Office. 1955. Price 12s. 6d.

THE Services Textbook of Radio is a series of seven volumes which has been written to provide Service readers with an introduction to current radio techniques. The present volume "Electronics" (which happens to be published first) is the third volume of the series. The volume deals with the properties of valves, cathode-ray tubes and semi-conducting devices. The treatment is at two levels, at an elementary level in the main text and at a more advanced level in the various "starred" parts of the text and in the appendices.

The first two chapters deal with the behaviour of electrons in metals leading on to thermionic, photo-electric, secondary and high-field emission. These chapters form an excellent introduction to the physics of the subject but one or two practical points require correction. For instance, on page 23 it is stated that cathode efficiency is given by "emission per unit area for a given heating power input". This should read "emission per unit power input" as substantially all the heat is lost by radiation so that, at a given temperature, there is a direct relation between emission and power input. On the same page is the statement "It is rare for a small valve to be seriously affected as to emission by a change of voltage of ± 5 per cent". As the previous text has been concerned with *saturated* emission, which of course depends very considerably on heater voltage, and the effect of space charge has not been mentioned, this statement appears somewhat incongruous.

Chapters III and IV deal with the structure and characteristics of diode, triode and multi-grid valves. The significance of g_m, r_a and μ are shown with reference to a set of triode characteristics restricted to the central linear region. As the three-halves power law is given in the main text it would have been helpful to have derived them by differentiation. This has the advantage that the dependance of g_m and r_a on current is immediately apparent.

Chapter V deals with the valve as a distorter but as the whole of the discussion is concerned with a sinusoidal input, one presumes that the behaviour of valves in pulse and shaping circuits will be dealt with separately in the forthcoming "Circuitry" volume of the series. Frequency Changing, Modulation and Demodulation are considered in Chapters VI and VII.

Chapter IX on High Power Valves and Chapter X on High Frequency Effects, both contain material not easily available without consulting original papers, and to a lesser extent this is also true of Chapter XI on V.M. Valves and Magnetrons. A large part of Chapter XIV on Semi-conductors is "starred" as being of an advanced standard. It is virtually impossible to deal with this subject in simple terms and perhaps it would have been better to have restricted the coverage to the characteristics of actual devices. In view of the great (although largely potential) importance of semiconductors to Service users, there is a good case for a separate volume in the series devoted entirely to the subject.

Chapters XV and XVI deal with Electron Optics and the Cathode Ray Tube. Here the proofs for the deflexion sensitivities of both e.m. and e.s. tubes are somewhat untidy and in the case of the e.m. tube the sensitivity is given in an implicit form from which the effects of the various parameters are not easily appreciated. A further difficulty is that the meaning of the various symbols is hidden in the text and does not appear on the diagrams. The final chapter is on Photo-Electric Devices and includes notes on the Photo-Multiplier and on Television Camera Tubes.

Altogether the book is well worth its modest price and probably well adapted to the needs of the three Fighting Services; the treatment is too abbreviated and non-mathematical to make it of value to those whose current aim is to pass examinations.

V. H. ATTREE.

The Principles of Technical Electricity

By M. Nelkon. 250 pp., 30 figs. Demy 8vo. Blackie & Son Ltd. 1955. Price 10s 6d.

The book is written as a textbook of electrotechnology for students up to the standard of the C and G telecommunications examinations, Principles, I & II. The subject is well taught, with adequate quantitative instruction, although largely from the physicist's point of view, as will be seen from some of the examples. An excellent feature is the number of worked examples, with other examples in each chapter for the student to work out: answers are given at the end of the book. Basic instruction in electrical effects and elementary circuits, d.c. and a.c. theory, etc, is well brought out, with an introduction to the telecommunications aspect, including diode and triode characteristics and the cathode-ray tube. The copper-oxide rectifier is mentioned, but not the selenium. Some calculus is brought in, mainly supplementary to the text, which is quite 'readable'. M.K.S. units are briefly mentioned in an Appendix but are not used in the text.

Terminology and symbols are generally very good, but there are some old terms such as condenser. An unpractical example is the academic teaching of shunting a copper instrument coil with a shunt which in practice has a low temperature-coefficient. As described the resulting temperature error would be prohibitive. While the calculation is of use as a basis, it should not be left at that to mislead the student. What is the International Electrotechnical Committee of Electrical Engineers, mentioned on page 240? The book is excellent value at half a guinea.

E. H. W. BANNER.

Electronic Transformers and Circuits

By Reuben Lee. 360 pp. 293 figs. Demy 8vo. 2nd Edition. John Wiley & Sons, Inc., New York, Chapman & Hall Ltd., London. 1955. Price $7.50 or 60s.

NOT a very large number of people earn their livelihood by designing electronic transformers. Although a considerably larger number are concerned with using them, the majority of these unfortunately regard the transformer as an entity they wish to purchase from a competent manufacturer, whose business it is to know how to make it. For this reason, the market for a book on electronic transformer design is essentially limited, which undoubtedly explains why there are so few authoritative works on this subject.

This book follows the same outline as its first edition: two chapters on transformer construction in general; then two on transformers and reactors for rectifier and supply circuits; followed by two on transformers in amplifiers. The remaining chapters cover transformers working at r.f. and for wide-band operation, components for electronic control circuits, magnetic amplifiers, and pulse and video applications. It is noted that several of the chapters are arranged in pairs, the titles of which suggest that the first one of the pair deals with the transformers or reactors for that application, while the second deals with the associated circuits. However, the subject matter does not lend itself too readily to this kind of division, so we find that both aspects of the section are distributed throughout both chapters.

It is evident from this book that the author is an engineer whose training and main experience has been with equipment for supply circuits: this section is extremely well covered, with an abundance of charts and practical data, even to a couple of design forms, the use of which is discussed at length. In the other areas, an attempt has certainly been made to provide similar coverage. For example, the section on amplifiers contains about as many charts and tables as that on rectifiers. But the coverage is very incomplete: the performance of audio transformers is covered only for a very limited combination of source and load impedances, quite inadequate for practical application. Information about filter design is very scant: without previous knowledge of this subject it would be incomprehensible, while any previous knowledge would find it inadequate.

The chapters to which new material has been added, on magnetic amplifiers, pulse circuits, r.f. power supplies, are rather too qualitative in their nature. While much useful data is presented, and not a few useful formulæ, it is doubtful how far an engineer confronted with a problem in these fields could apply them without his own private fund of data. This reviewer has the impression that the progress made in this field during the eight years since the first edition was released considerably exceeds the extent by which the book has been brought up to date.

However, the book is one of the few good ones available in its field, and the situation described at the outset means we are unlikely to see many competitors. The reviewer is in agreement with the author that users of electronic transformers should know more about them, instead of expecting the manufacturer to take the entire responsibility, because the design of a transformer and of the circuit with which it will be used are inseparably connected. A working knowledge of transformer design, which this book gives in quite a compact compass, is essential to effective design of circuits to go with them.

N. H. CROWHURST.

Working Metals by Electro-Sparking

66 pp. 23 figs. Demy 8vo. Her Majesty's Stationery Office (for Department of Scientific and Industrial Research). 1956. Price 5s.

THIS is an abridged translation of an original Russian work. The book was prepared as a practical manual for engineers, technicians, foremen and all those concerned with the operation of electro-spark units.

Electro-spark machining is one of the newer techniques for working metals, particularly metals and alloys of great hardness. In this technique the erosive effect of repeated electrical discharges between tool and workpiece is used instead of conventional metal-working tools to carry out the work required. The tool itself does no actual abrading but acts only as an electrode.

The publication covers the physical laws governing the technique, the technology requirements for using the process and the results which can be obtained. A great deal of operating experience has already been obtained in Russian factories and the translation includes an outline of workshop procedure based on operations carried out in the workshops of industrial co-operative societies.

Fixed Resistors

By G. W. A. Dummer. 180 pp. 50 figs. Demy 8vo. Sir Isaac Pitman & Sons. 1956. Price 28s.

THIS book is the first of four volumes of a proposed series written on the subject of components with a bias towards Service requirements.

Consisting of nine chapters, this 180 page book embodies a great deal of information relating to a component which for many years has only provided material for numerous articles and papers, much relevant data on industrial techniques being witheld by manufacturers.

No printing errors were observed but on page 38 the author, in referring to temperature coefficients states that—"for most *pure* metals the value of α is 0·0038..." and on page 71 "The temperature coefficient of *pure* metals is positive and ranges from 0·3 to 0·6 per cent per degree centigrade". Two somewhat contradictory statements.

The author, however, has gathered together and co-related a great many facts, figures and manufacturing methods and very ably presented them in a logical sequence commencing with generalities and Service specifications, progressing through measurements, rating, stability, performance, manufacturing processes and techniques, special resistors and experimental types to future developments, ending with 24 pages of comparison charts setting out details of mechanical, electrical and climatic characteristics in both grade 1 and grade 2 categories of most British manufacturers from $\frac{1}{8}$ watt to 100 watt types together with pictorial representations of their construction.

An extensive bibliography containing some 350 references is intended to form the nucleus of a much-needed general bibliography on components and which, in its comprehensiveness, signposts the way to orderliness from the existing chaotic welter of published material.

As a book for the industrial electronic design engineer it is not too technical to be comfortably readable, generalizing as it does rather than particularizing, extremes of detail being taken care of by references in the bibliography. Nevertheless, it is librally sprinkled with useful data charts and diagrams.

Many manufacturing processes are described in considerable detail and information is given for the most modern applications such as high-frequency performance, especially where used as radar waveguide loads, thermistors, printed circuits, subminiaturization and resistors potted in and extruded with ethoxylene resins.

"*Fixed Resistors*" undoubtedly fills a gap in modern electronic literature and no designer or engineering library should be without a copy of this comprehensive volume.

R. H. MAPPLEBECK

ELECTRONIC EQUIPMENT

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

GONIOGRAPH

(Illustrated below)

Standard Telephones and Cables Ltd, Oakleigh Road, New Southgate, London, N.11

The A2209 Goniograph has been developed for the automatic recording of radio direction finding bearings during site testing of d.f. stations.

It is primarily intended to present, in graphical form, a continuous plot of bearings determined by the direction finder in comparison with those obtained visually of the aircraft under observation.

The imperfections of the site in the immediate vicinity of the d.f. station, and also the topography of the surrounding country inevitably affect the accuracy of d.f. bearings. Previously, this has been difficult to assess quantitatively, since manual recording is often too slow and may miss rapid fluctuations.

The Goniograph enables a speedy evaluation of a d.f. site to be made. A permanent location for a direction finder can be selected with efficiency and economy by using the A2209 in conjunction with mobile d.f. equipment.

The equipment can be used with almost any direction finder whose bearing information is produced in the form of balanced d.c. voltages, $V\sin\theta$ and $V\cos\theta$ (where θ is the d.f. bearing angle and $V > 30$V). A switched attenuator permits direct connexion to v.h.f. and u.h.f. direction finders.

The machine records indicated bearings at the rate of about five per second, and plots them against time in cartesian form. True bearings are measured by tracking the aircraft visually with a theodolite connected to the machine by a servo link and plotted on the same graph. Calibration lines at appropriate angular and time intervals and an indication of signal strength (derived from the d.f. receiver a.g.c.) are also printed automatically. They appear on 12in (30·48cm) × 17in (43·18cm) sheets of Teledeltos recording paper.

ELECTRONIC PHOTOGRAPHIC PRINTER

Cinema-Television Ltd, Worsley Bridge Road, London, S.E.26

The "Cintel" Electronic Photographic Printer has been designed principally to overcome the high density variations existing in negatives produced by aerial photography.

Due to the inability of printing paper faithfully to reproduce the full range of densities present in the negative, the amount of information obtained from the print is severely restricted unless steps are taken to equalize these variations.

This new equipment is basically a contact printer, but the light source used is a cathode-ray tube on which is produced a single spot of light which is made to scan out a rectangular raster in a similar manner to that of a television picture.

The amount of light transmitted through the negative and printing paper, which are in contact with one another will vary in accordance with the density variations of the negative and this quantity of light is collected by a photo-electric cell positioned within an integrating cube.

The output from the photocell is amplified and fed back to control the brightness of the spot of light on the cathode-ray tube such that it decreases in intensity as the less dense areas of the negative are scanned.

The effects of negative density variations can thus be reduced to any desired value by control of the degree of feedback employed.

The complete equipment is constructed in console form and employs a vertically mounted c.r.t. on which is reproduced a 1½in square raster. This is projected by a suitable lens to fill a 10in × 10in printing plane at a working height of approximately 3ft.

PRE-SET POTENTIOMETER PANELS

(Illustrated below)

Atlas Process Components Ltd, Alexandra Road, Ponders End, Middlesex

These pre-set panels which have been designed mainly for use in radio and television receivers can contribute greatly to reducing the cost involved in mounting 9 or 10 individual carbon track potentiometers on one panel.

The basic unit consists of a panel 1in wide on which the track is mounted. Individual manufacturers' requirements regarding the size of the fixing panel and distance between and position of fixing holes vary greatly, so the design has been made flexible enough to include a second panel which can be cut exactly to manufacturers' drawings. This second fixing panel also does much to improve the unit as a whole as it gives protection to the actual track against the adjusting screwdriver as well as adding to the rigidity of the assembly.

DIRECT READING MAGNETOMETER

(Illustrated below)

Newport Instruments (Scientific & Mobile) Ltd, Newport Pagnell, Buckinghamshire

The magnetometer is designed to measure the variations in the magnitude of the field of a television focus magnet. It is also suitable for the investigation of permanent magnets and the measurement of stray fields of transformers and electromagnets and also for examining the fields of the large focusing electromagnets associated with particle accelerators.

The normal ranges of the instrument are from 0 to 5 oersteds, 0 to 50 oersteds, and 0 to 500 oersteds. All these ranges are obtainable with the general purpose probe supplied with the instrument and may be selected at will by means of a front panel switch. The ranges may be extended by the use of a special probe. An increase in sensitivity of ten times may be obtained on all these ranges by a control on the front panel.

The principle of operation of this equipment is the gyromagnetic coupling in ferrite materials. A ferrite toroid is excited to saturation by a toroidal winding, a solenoid being wound on the core at right-angles with this winding. In the presence of an external magnetic field a voltage appears across the solenoid proportional to the value of the field and at a frequency $2f$ (where f is the exciting frequency). The voltage derived from this solenoid is fed into a stabilized gain, tuned amplifier, the output of which is measured by a valve voltmeter. The indicator is a 4in circular scale meter located on the front of the instrument. An electrical output is provided so that slowly varying fields may be displayed on an oscilloscope or pen recorder.

As the short term accuracy of the instrument is better than ± 1 per cent, a standard calibrating solenoid is available to enable the absolute value of a field to be determined to this accuracy.

TELEVISION AERIAL ATTENUATOR

(*Illustrated below*)

Spencer-West Ltd, North Quay, Great Yarmouth, Norfolk

Type 60 is continuously variable from 6dB to at least 36dB. Full matching is

maintained at all settings and a lockable control is fitted. This attenuator is based on a new circuit arrangement which ensures that no frequency selective circuits likely to spoil the picture quality are present.

TELEVISION LINE SELECTOR

(*Illustrated below*)

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

The Television Line Selector L.196 enables any normally triggered oscilloscope to give a jitter-free display of one or more lines from a television video signal. It can be used on transmitter and receiver circuits to display depth of modulation, d.c. levels, bandwidth, blanking, synchronizing pulses and for the examination of similar waveforms in detail.

Any desired line may be selected from either odd or even frames and both line and frame synchronizing pulses can be selected for display.

The line selector has high and low impedance input circuits and can be used either to terminate or bridge a video link.

The output p.r.f. can be switched to 50c/s, enabling both frame waveforms to be superimposed provided that a 50c/s square or sine wave input is available.

The internal mains can be switched to provide this input with a mains locked television system.

The output is a sharp front 30V pulse, which is delayed with respect to the frame pulse by up to 25msec. The delay is "stepped" in synchronism with the line pulse and coarse and fine delay controls make it simple for the operator to select the required display.

VIDEO NOISE LEVEL METER

(*Illustrated below*)

The Wayne Kerr Laboratories Ltd, Sycamore Grove, New Malden, Surrey

This instrument deals with the measurement of the type of noise similar in nature to thermal noise, but of higher amplitude, caused by transmission systems and which can limit the use of signals and cause television link picture definition to suffer. Formerly this has been assessed by an oscilloscope method which depends on the skill of the operator and such factors as trace brightness and speed.

Here the noise is fed through two attenuators switching in 10 and 2dB steps, to the input of a video amplifier. The output of the amplifier is fed through a cathode-follower to the mean square detector which in turn feeds a moving-coil meter calibrated from +1 to −10dB.

The meter is switched to monitor the output as a built-in calibration oscillator,

which can be switched into the input of the instrument as well as the output from the detector.

In this way the instrument is calibrated internally for each measurement made, and provision is made by switching the output of the calibration oscillator for setting up the amplifier gain, and checking the law of the system.

Measuring range 20 log signal (r.m.s.)/noise (r.m.s.)=0 to −60dB.

Video bandwidth 10kc/s to 1·5, 3, 6, or 10Mc/s, switchable.

The response is flat from 10kc/s to the stated frequencies: above these frequencies the cut-off rate is extremely sharp.

Input impedance 75Ω

Calibration level 0·353V r.m.s. sine wave.

Accuracy ± 1dB.

The instrument has been developed to meet the requirements of the Planning and Installation Department of the BBC for the measurement of signal-to-noise ratio on their television link circuits.

SILVER RIVET CONTACTS

Johnson, Matthey & Co. Ltd, 73-83 Hatton Garden, E.C.1

The use of composite rivet contacts with silver facings has hitherto been severely limited, despite the economy afforded by the reduction in weight of silver, by the expense of the additional manufacturing operations involved. This limitation has been overcome by novel manufacturing techniques that enable silver-faced composite rivets to be produced at lower cost, size for size, than the larger types of solid rivets.

The copper shank, tubular in section to simplify riveting, is electrically and thermally efficient, while the thickness of the silver face—a minimum of one third of the total head thickness—ensures long life. The rivets are given a light silver electro-deposit so that good contact can be made with the contact member on assembly.

To enable the new tubular contact to be manufactured and marketed as economically as possible a range of four head sizes from $\frac{1}{8}$in to $\frac{1}{4}$in in diameter has been set up in two types, TF (flat contact face) and TD (domed contact face).

HIGH RANGE ELECTROSTATIC VOLTMETER

(*Illustrated above*)

W. G. Pye & Co. Ltd, Cambridge

Catalogue No. 11314 has been developed to provide an instrument capable of measuring the higher voltages now being encountered in e.h.t. work on cathode-ray tubes and similar devices. The suspension and optical systems are similar to those of the other "Scalamp" instruments.

The range is direct reading up to 40kV d.c. or 40kV r.m.s., a.c. The overall accuracy is at least 2 per cent of full-scale deflexion and is not affected by reversal of polarity.

A specially designed vane is held in the proximity of an insulated electrode by a taut suspension of the galvanometer type and the movement is damped by a magnetic damping device. The "Scalamp" optical system projects a bright vertical light band with a fine hair line on to an engraved translucent scale 15cm long.

The dust-proof cast aluminium case provides screening from external electric fields, encloses all the components of the instrument and is provided with three rubber feet, the instrument requiring no special levelling.

Notes from ——————

NORTH AMERICA

Second I.R.E. Instrumentation Conference

The Professional Group on Instrumentation and the Atlanta Section of the Institute of Radio Engineers will sponsor the Second Instrumentation Conference at the Biltmore Hotel in Atlanta on 5-7 December 1956. Dr. B. J. Dasher, Director of the School of Electrical Engineering, Georgia Institute of Technology, will again serve as Conference Chairman and Sessions will be devoted to the folowing subjects:—

Industrial Nuclear Instrumentation and Instrumentation for Radiological Safety; Industrial Application of Instrumentation; Guided-Missile Range Instrumentation and Wind-Tunnel Instrumentation; Solid-State Devices and Other Components; Laboratory Instrumentation and Test Equipment; Recorders.

Colour Television in Plutonium Plant

The Hanford plutonium plant near Richland, Wash., operated by General Electric for the Atomic Energy Commission, now uses a closed circuit colour television system to enable crane operators to manipulate crane hooks in radioactive areas. Colour television was chosen primarily because the remote manipulation problem involves colour coded objects and also because it gives added depth perception and better object resolution. The rotating colour wheel principle is used, for good colour rendering and definition and low maintenance cost.

The two cameras, mounted in the radiation area and attached to the crane bridge, give the operators a clear, wide-angled view of the chemical equipment used in the extraction of plutonium from fissioned uranium 80ft below the crane. The crane cab is windowless, has thick steel walls and is located behind a concrete parapet, complete remote control of the cameras including all electronic adjustments, pan and tilt.

The two receivers in the cab, connected to the cameras by wires, have 7in screens, magnified to give a 10in effect.

High Precision A.C. Motor-Tachometer

A 400c/s miniature a.c. motor-tachometer, manufactured by Precision Components, Inc., Westbury, L.I., N.Y., has been designed to give a high level of accuracy and reliability for application in air-borne integrating and rate servos.

Tachometer performance features 0·05 per cent linearity error over a 0 to 4 000 rev/min range, 0·05 per cent scale error over an ambient temperature range of -55 to $+75$°C and the 400c/s null voltage is reduced to less than 1mV in phase and quadrature. The output voltage at 3 600rev/min is 3·6V ± 0·05 per cent and is in phase with the excitation voltage within 0·05°. The tachometer has a built-in heater and temperature sensing winding and the power to the heater is controlled by an externally located magnetic amplifier.

The 4 pole servo motor has a no load speed of 9 500rev/min with a stalled torque of 0·9oz. in at 7W per phase input. The motor-tachometer unit is flange mounted and measures 1·437in in diameter by 4¼in long.

New Standard Diode Pulse Recovery Test Set

(*Illustrated below*)

The Hauman Instruments Company, Model ND-1 has been designed for the purpose of measuring the inverse pulse recovery of silicon and germanium diodes, and is to be allocated Armed Services specification JAN-256.

The unit features an accurate and correlatable method of determining true or absolute diode recovery characteristics due to its inherently low load time-constant, which allows the recovery time of the test set itself to be much lower than that of most available diodes, accomplished by circuit design eliminating excessive capacitance caused by blocking diodes, measuring probes and the like.

When used with a suitable square wave generator and oscilloscope the test set will operate over the following ranges:—

Forward bias current	2-40mA
Inverse voltage	5-50V
Load resistance	500-3 000Ω
Load capacitance	10-100pF
Inverse enhancement current	5mA-50μA
Measurement time interval	0·1-10·0μsec

The test set is built in a standard panel front cabinet with test clips affording easy test diode insertion and contains an integrated power supply as well as all associated test circuits for complete operation.

General Electric Company's New Divisions

The General Electric Company has announced the re-organization of its electronics business into three separate Divisions as a result of "rapid expansion and future growth prospects".

They are the Industrial Electronics Division with temporary headquarters at New York; the Electronic Components Division with temporary headquarters at Owensboro, Ky, and the Defense Electronics Division with headquarters at Syracuse, the respective general managers being Messrs. Harold A. Strickland, Jr., L. Berkley Davis and George L. Haller.

It is also announced by the General Electric Research Laboratory at Schenectady, N.Y., that Dr. A. M. Gurewitsch has been appointed one of the Laboratory's European scientific representatives, operating from the new GE research office in Zurich, Switzerland.

Right-Angle Valve Sockets for Printed Wiring

Mounting valves parallel to printed-wiring boards by means of right-angle sockets, is the latest space-saving contribution of the Aerovox Corporation.

These sockets reduce the height and depth of printed-wiring assemblies and are equally adaptable to hand or machine-insertion methods. The terminals are of adequate length to slip into printed-wiring holes and be dip soldered, the silver-plated contacts being engineered to provide non-fatiguing contact pressure with suitable insertion and withdrawal pressures. Metal parts and mounting hardware are plated to meet salt-spray and MIL specifications.

These components are available in 7 and 9-pin sockets and in four different versions:— type A for general-purpose applications where extra rigidity and resistance to vibration and shock are not important factors; type AX for special applications requiring extra rigidity; type B, as A, but with valve shield added; type BX, as AX but with valve shield added.

Ribbon Cable Assemblies

"Tempbraid", a product of Hitemp Wires, Inc., Mineola, N.Y. was designed to combine from 2 to 20 conductors ranging in size from 12 to 32 AWG) or pairs of parallel conductors, with a maximum width of 1¾in, in a single ribbon assembly. This flat construction is claimed to effect saving in space, time and conductor footage over conventional twisted cable assemblies and to take a 90° bend, connexion-making being simplified by almost unlimited colour coding.

The 100 per cent "Teflon" construction is said to offer:— Low loss factor, low dielectric constant, high volume resistivity, low coefficient of friction, to be non-inflammable, unaffected by moisture, inert to all known commercial solvents and to have a working temperature range of -250° to $+500$°F.