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Commentary

THE Society of British Aircraft Constructors recently held their eighteenth Flying Display and Exhibition at Farnborough, the home of the Ministry of Supply's Royal Aircraft Establishment. The R.A.E. is the largest research and development establishment in the country and for more than fifty years has been the pioneer centre of British aeronautical experiments. It is the general adviser to the Services and to the aircraft industry on all technical problems of aeronautics and it collaborates closely with two sister establishments, the National Gas Turbine Establishment at Pyestock and the Royal Radar Establishment at Malvern. The work of the R.R.E. actually began at Woolwich Arsenal where, in 1878, the first official experiments with balloons were made. The establishment settled at Farnborough in 1905.

In the ensuing years the importance of electronic engineering, in its various forms, to aeronautics has increased steadily year by year; a fact which was most forcibly driven home at this year's S.B.A.C. display.

It is in fact true to say that there is no aircraft or guided missile that could operate without some form of electronic control. There is no airline in the world that could run with safety without the vital guidance of radio and radar while the air defence of Great Britain, now changing over to the guided missile, is utterly dependent on a network of electronics that extends from the initial detection of the enemy to the physical control of the missile itself, which may be travelling at many times the speed of sound.

For example, in one guided missile system, radar stations at strategic points around the coast give early warning of the approach of enemy aircraft, this information being passed automatically to the control post of the nearest battery of missiles. The control post automatically assesses when the target is within range and sends out signals which fire off as many missiles as is necessary. The system can differentiate between one aircraft and two flying in close formation, thus permitting each to be attacked separately. The missile is guided by a radar beam projected at the target, and reflected back to receiving equipment within the missile itself, the output from which acts upon the control surfaces of the missile.

On the civil side of aviation it is noteworthy that, although one pound of unnecessary weight on a trans-Atlantic airliner can cost £1 000 in a year of operation, the operator is, nevertheless, quite happy to carry in his aircraft several hundred pounds in weight of electronic equipment. The equipment can so improve the frequency, efficiency and safety of operation that its use can make the difference between commercial success and failure of the airline. The more crowded the airlines become, the more necessary are modern electronic aids.

The importance of electronic engineering is not, of course, confined to the operational side of aviation, it is equally important in the design, development and production stages. For, while the extremely rapid development of the flying machine which has taken place in the post war years could hardly have been accomplished at all without electronic aids, it could certainly not have been attained anywhere near so speedily.

Of the many contributions of the electronic engineer to the development of aeronautics, the analogue computer, or simulator, is, perhaps, of most consequence, for by its aid the effects of varying any of the many parameters determining the behaviour of an aerodynamic structure can be quickly and easily ascertained. In this way literally months of complex computation and prototype building can be saved.

At a later stage, when models are built for wind-tunnel testing, the electronic engineer again plays his part by providing most of the equipment for making the various measurements and for recording and analysing the mass of data that is obtained.

In the pre-production and production stages electronic methods of vibration testing and ultrasonic material testing techniques are used, while the resistance strain gauge is utilized for measuring the stresses on completed structures.

These are but a few of the uses which are being made of electronics but they serve to demonstrate the vital part which electronic engineering is playing in the rapid development of the aircraft industry.

That the aircraft and the electronic industries should co-operate so well is not surprising for they are both young industries. They are of much the same age, and have, so to speak, grown up together. That the acquisition of knowledge and the growth of the industries, both in size and importance, to the modern way of life, has been rapid, there is no denying. On looking at modern aircraft, guided missiles and the many complex products of the electronic industry it is indeed hard to realize that the development of both industries has taken place almost completely within the lifetime of over 25 per cent of the population of this country.

Applications of a

New Type of Cold Cathode Trigger Tube

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(Part 1)

The writing of this article has been prompted by the introduction of a new type of close tolerance cold-cathode trigger tube. Primarily, this tube has a trigger ignition voltage which is very stable over long periods; it is therefore suitable for use in accurate RC timers, overvoltage detectors, voltage regulators, relaxation oscillators and sensitive relay units.

The article describes the mode of operation of tubes of this type and outlines the design techniques needed for several applications.

IN the past most types of cold-cathode trigger tubes have shown a fairly wide spread in major characteristics, particularly trigger firing voltage, both from tube to tube and for any one tube during its operating life. This has meant that in relay or on-off applications such tubes could only be operated with large firing signals. Similarly, they have given rather poor service in applications such as *RC* timers and relaxation oscillators in which the accuracy of the device depends on the stability of the trigger firing voltage. Furthermore, trigger tubes have often possessed first order characteristics which were affected by ambient light and operation conditions. All these shortcomings have rather overshadowed the potential advantages of the tubes, e.g. long life, high reliability, instantaneous operation and negligible standby losses.

Over the past ten years, great advances have been made in the development of gas-filled stabilizers and reference tubes having very close tolerances, a high stability and long life. Recently attention has been given to applying the processes used in such reference tubes to the design of trigger tubes.

During the last few years tubes using these techniques, having a higher order of trigger ignition voltage stability and marked freedom from ambient light effects have been manufactured. These trigger tubes were, however, designed for specialized application and it is only in the last two years that comparable tubes, suitable for general purposes, have become available.

The earlier part of this article gives a description of the mechanism and operation of trigger tubes in general, together with an indication of the various limits which must be observed when designing circuits. This leads on to a brief description of the newer type of trigger tube.

The second section is devoted to a description of the applications for which the stable trigger tube is most suited.

To a number of readers, parts of the article will be common knowledge. In order that such portions may be ignored each section has been made as self-contained as possible, though at the cost of some repetition.

Principles of Trigger Tubes

General Properties

In many respects the general behaviour of the gas-filled cold-cathode trigger tube is similar to that of the hot-cathode gas-filled triode or thyratron. There are, however, a number of very basic differences in their characteristics and these must be taken into account when designing circuits around the two devices.

Both are made to conduct by the application of an appropriate signal to their respective control electrodes, and once anode to cathode conduction has commenced the control electrodes cease to have any significant effect on the discharge. It is a property of gas discharges that the voltage across the electrodes between which a discharge is flowing, is substantially constant and independent of the current. The current flowing in a gas discharge is, therefore, almost entirely dependent on the resistance in the circuit and also on the supply voltage.

To stop the current flow or extinguish a gas discharge it is necessary to lower the anode to cathode potential below the normal conducting potential.

The thyratron with its thermionic mode of cathode emission has a low anode to cathode voltage drop during conduction, say 10V. The trigger tube, however, relies on ionic bombardment of the cathode and has a drop or maintaining voltage V_m of the order of 100V.

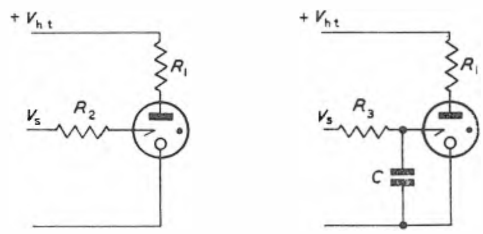


Fig. 1. (a) Current triggering circuit (b) Capacitor triggering circuit

For the same reason the process of initiating a discharge is different for the two devices. In normal operation* the thyratron control grid acts as a gate and, dependent on the voltages applied to the tube, either permits or prevents electrons emitted by the cathode from reaching the anode. In the trigger tube the input signal must establish a glow discharge between trigger and cathode which then causes an anode to cathode discharge. The input signal must therefore supply a quantity of energy to initiate an anode to cathode discharge. It should be stressed that the energy required is very small.

Fig. 1(a) shows a basic circuit for operating a trigger tube. V_{ht} is a voltage within the anode working range of the tube. If V_S is steadily raised to a voltage known as the trigger ignition voltage, $V_{t(ign)}$, a discharge will be formed between trigger and cathode of magnitude determined by R_2. If this discharge is of sufficient value (of the order of 100μA), it will initiate a discharge between anode and cathode (the main gap). The continuous main gap current, which may be of the order of 25mA is given by the expression $i_a = (V_{ht} - V_m)/R_1$.

* *Mullard Ltd.*

* *This refers to power control not pulse modulator service.*

In many applications, such as photocell relay duty, the firing signal for the tube may have a source of such high impedance that the trigger current necessary for transfer cannot be obtained directly. If a capacitor C is connected as shown in Fig. 1(b), then every time its potential is raised above $V_{t(ign)}$ it will partially discharge into the trigger cathode gap. Provided the capacitance of C is sufficient it can provide all the necessary trigger current to fire the main gap, R_3 can then be very high.

With the capacitor system of firing it is possible to obtain a current gain between input and output of the order of 10^5.

Principles of Operation

Mechanism of Glow Discharge

To make the rest of the article more easily understandable it will be useful to give a basic description of the mechanism of the normal two-electrode cold-cathode discharge. It should be added that the mechanism is complex and has been detailed elsewhere. The following notes offer a sufficient basis for the purpose of circuit design.

Consider a gas-filled diode, such as a simple stabilizer, which has a small positive voltage V_1 applied to its anode via a resistor R_a (Fig. 2). If the tube is in total darkness

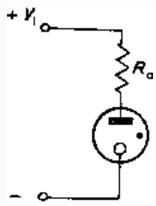


Fig. 2. Simple diode circuit

and can be shielded entirely from the cosmic radiation, traces of radioactive matter, X-rays, etc., it will act as a perfect insulator, within the voltage range normally considered. In practice, such shielding is almost impossible and occasionally one of these external agents will cause ionization of the gas. The electrons and positive ions which are thus liberated will be attracted to the anode and cathode respectively, producing a minute pulse of current in the external circuit.

If now the voltage V_1 is steadily increased, from say 10V, the ionization potential of the gas will be reached and above this point collisions between electrons and gas molecules can lead to the production of new positive ions and electrons. The positive ions formed will move towards the cathode under the influence of the anode-cathode potential gradient, which is still relatively uniform, and free secondary electrons when they strike it. These secondary electrons will be attracted towards the anode and will cause further ionization by collision. It should be stressed at this stage that the discharge is still dependent on a primary source of electrons and if this is removed the discharge will eventually cease. As the anode voltage is raised further a point is eventually reached at which, on the average, one electron freed from the cathode will lead, via the production of positive ions and their bombardment of the cathode, to another electron leaving the cathode. At this stage the discharge can be maintained without any auxiliary ionizing agent. This is the breakdown or ignition potential, V_{ign}, point B in Fig. 3.

Beyond this point the potential distribution changes drastically and the voltage across the tube falls to region CD. This is because slow moving positive ions in the discharge form a space charge or virtual anode near to the cathode. The result is that electrons liberated from the cathode enter a strong local field and obtain enough energy to cause ionization despite a fall in total tube voltage.

The tube is now operating in the glow discharge region and the discharge covers a small portion of the cathode. In this region the current density at the cathode is constant and if the tube current is raised within the region, the discharge will cover an increasing area of the cathode surface, without a great increase in V_m. When the whole area becomes covered a further current increase will raise the cathode current density and the discharge will start entering the abnormal glow región. This region is characterized by an increase in tube voltage drop and dissipation. Normally, therefore, it is only under peak pulse current conditions that trigger tubes are operated in the abnormal glow region.

In passing, it should be noted that the potentials and currents at which the above salient characteristics occur are determined by the gas pressure, composition of the gas-filling electrode geometry, and the surface material of the cathode. Changes in any of these parameters will alter the tube characteristics.

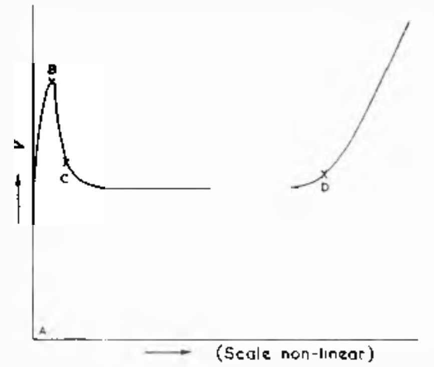


Fig. 3. Voltage/current characteristic of a cold-cathode diode

Forming the Discharge in Trigger Tubes

The process of establishing a discharge in a trigger tube may be divided into two stages:

(1) Establish a self-maintaining discharge between trigger and cathode.

(2) Increase the trigger discharge to a value at which it transfers to or initiates an anode-cathode discharge.

Consider for the present stage 1 in the sequence of events. From the foregoing discussion it will be clear that to establish a trigger discharge the trigger potential must be raised to the ignition voltage $V_{t(ign)}$ which will probably lie within the range 60 to 150V dependent on the form of the tube being employed. In addition it will be remembered, the presence of ionized gas or primary electrons must be ensured. The level of ionization produced by natural causes such as cosmic radiation, is small and of the order of several particles per minute. Unless an additional method of producing ionization is employed long delays ($\sim$ 1min) will occur between the application of the trigger ignition voltage and the formation of a discharge.

It is normal for cold cathode tubes to employ one or more of the three following mechanisms for priming the tube.

(1) Photo-electric emission from the cathode or other active surface.

(2) Stray ionization from an auxiliary priming discharge.

(3) Ionization due to an internal radioactive source.

The majority of the tubes with barium coated cathodes rely entirely on the first method of priming, but if such tubes are operated in complete darkness the priming source is removed and they will then exhibit long firing delays.

They may also fire spuriously if used in direct sunlight.

In tubes which have pure metal cathodes the photo-electric emission from the cathode is very low and one of methods 2 or 3 must be employed.

In tubes using the second form of priming a continuous self-maintained discharge of a few microamperes is normally sufficient. This can be obtained in several ways, for example, between a separate anode and the main cathode. The performance of tubes of this type is not altered by their use in either complete darkness or brilliant sunshine.

There are also an increasing number of cold-cathode trigger tubes which have a radioactive source inserted in some manner inside the tube. These tubes have similar properties to those with an auxiliary discharge priming.

The level of ionization considered up to this stage has been comparatively low and the effect of increasing it has only been to reduce the firing delays. The value of the trigger breakdown voltage is substantially constant over a wide range of ionization level provided sufficient time is allowed for the tube to trigger. To move on to the next stage in triggering a tube one must consider the effect of a self-maintaining discharge between trigger and cathode on the anode cathode breakdown voltage.

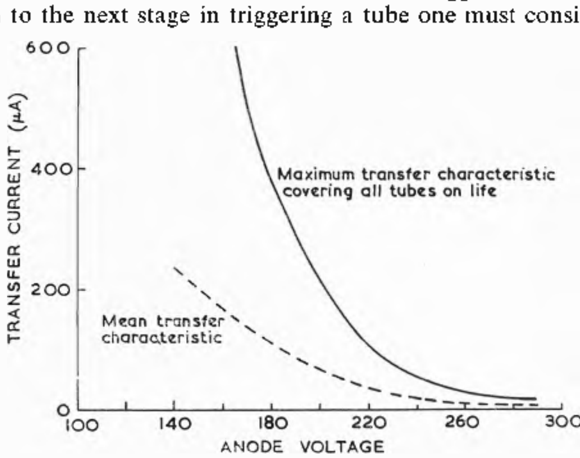


Fig. 4. Transfer characteristic of Z803U

A normal trigger discharge increases the ionization present by several orders of magnitude and under these conditions the breakdown voltage between other electrodes is reduced. The dotted line in Fig. 4 shows the typical reduction in anode cathode breakdown voltage as a function of the trigger current for the Z803U. (An example of a trigger tube with a priming discharge.) The mechanism described is often regarded as a transfer of the trigger discharge to the anode cathode gap.

In practice, the curve shown in Fig. 4 is used in another manner, the anode potential is set at some fixed value below $V_{a(ign)}$ ($\sim$ 290V) and the curves then show the magnitude of the trigger to cathode current necessary for the discharge to initiate an anode cathode discharge. For example, if the anode potential is $V_a = 250$V, a trigger current $I_t = 40\mu$A will be required. (The values obtained from the full curve include the performance of all tubes over life).

The curves cover the anode voltage range 170 to 290V. As already stated, the minimum anode cathode breakdown voltage in the absence of a trigger discharge is 290V, and if this voltage is exceeded spurious anode conduction may occur. At voltages below 170V this tube can still be triggered by a large trigger current, but a minimum voltage of 170V is necessary to establish a priming discharge.

Extinction

It was stated earlier that to extinguish a gas discharge the potential between the electrodes must be reduced below the maintaining potential. Once the voltage is reduced emission from the cathode will fall below the level required to maintain the discharge and further ionization of the gas by collision will also be reduced.

There are two basic methods of extinguishing a trigger tube, one is to raise the cathode potential and the other is to lower both the anode and trigger potentials. It is essential that the trigger discharge is extinguished, otherwise the tube will be retriggered immediately the full anode potential is re-applied. In the case of a tube with a priming anode, such as the Z803U, it is preferable not to extinguish the tube by raising the cathode potential since this will also extinguish the priming discharge. However, if the latter method is employed sufficient time must elapse for the priming discharge to re-establish before attempting to re-trigger the tube.

The actual methods by which the anode voltage is reduced will be discussed when describing applications. A self-extinguishing circuit employing capacitors across the trigger cathode and anode-cathode gaps will also be described later.

Some Characteristics of Trigger Tubes

Ionization Time

Considering the normal definition of ionization time, i.e. the interval between the application of the trigger firing voltage and the onset of anode conduction, this time may be split into three main portions.

(1) The delay before an ionized particle is drawn into, or liberated in the trigger/cathode gap. This time depends on the type of initial ionizing source and on the magnitude of the trigger over-voltage (i.e. the level of the trigger voltage above the static breakdown value).

In a tube in which a priming system provides a minimum level of continuous ionization the fluctuations in this time are relatively small.

(2) The formative delay, or time required to establish a trigger/cathode discharge after the appearance of the initial ionization. This time depends on the magnitude of the trigger over-voltage.

(3) The time taken for a trigger/cathode discharge to be transferred to the anode/cathode gap. For a given anode voltage this transfer time is inversely proportional to the trigger current, which in turn depends on the trigger voltage and circuit impedance.

In general, the last delay is of the order of 10μsec, it is impracticable to give separate values for the ionization and formative delay times because of the difficulty of differentiating between them.

For design purposes the total ionization time is the important parameter and it will be noticed from the above points that this time will be mainly dependent on the instantaneous trigger over-voltage.

Fig. 5 shows the typical relationship between over-voltage and total ionization time for the Z803U, measured with a trigger impedance of 5·6kΩ. It will be seen that this time ranges from 20μsec for an over-voltage of 10V, up to 2msec for an overvoltage of 0·5V.

Recovery Time

The recovery time or deionization time of a gas-filled tube is the time between the extinction of the discharge and the instant when the residual positive ions have been neutralized to an extent that allows normal anode and trigger pre-firing voltages to be applied, without false firing. This parameter sets a limit to the maximum frequency at which the tube can be used. The recovery time depends considerably on the circuit conditions and for circuit design it is necessary to appreciate the main variables involved and to be able to extrapolate from measurement under quoted conditions.

In a cold-cathode trigger tube the main variables are:

(1) Anode current prior to extinction.

(2) Potential of trigger and anode during deionization (V_L).

(3) Re-applied anode voltage (V_A).

In general, the greater the working anode current the longer will be the recovery time. For the Z803U this is shown in Fig. 6. It will be noticed that for instantaneous

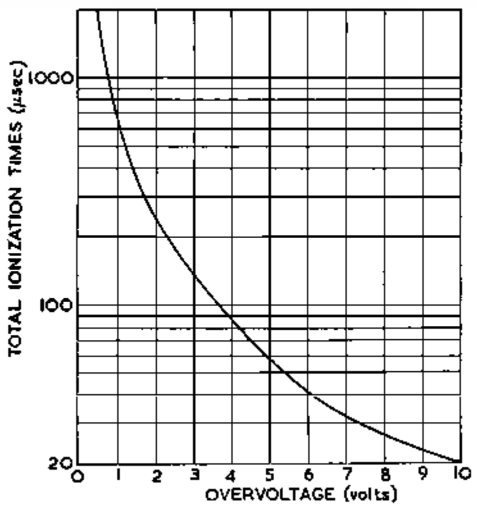


Fig. 5. Ionization times against trigger over-voltage for Z803U

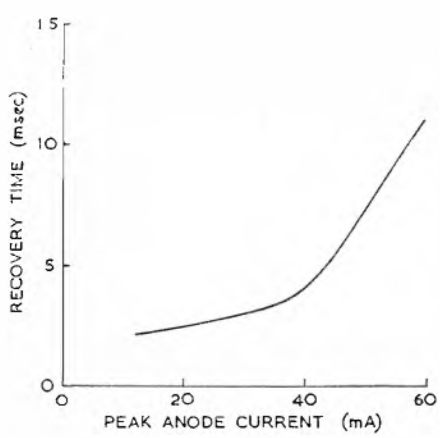


Fig. 6. Z803U recovery time against peak anode current $V_a = 220V$ $V_L = 0V$

anode currents above 20mA the recovery time rises very rapidly from the minimum level of about 2msec to 10msec at ≃ 60mA.

If, when extinguishing the tube, the trigger and anode potentials are only taken just below the maintaining value, ionization will still take place and the recovery time will, therefore, be prolonged. Alternatively, if the trigger and anode are returned to cathode potential, the residual positive ions will drift slowly to the various neutralizing surfaces and again the recovery time will be long.

To obtain the shortest ionization time, therefore, the anode potential should be sufficiently high to sweep charged particles rapidly out of the discharge path without causing further ionization.

This effect is shown in Fig. 7, though in this particular type the effect of the electrode potentials during deionization, is much less marked than in most cold-cathode tubes. From Figs. 6 and 7 it can be shown that raising the re-applied anode voltage from 220V to 290V increases the recovery time from about 2·0 to 3·0msec.

In the above discussion on recovery time it has been assumed that the anode voltage is re-applied as a step function. Normally this is not the case, the anode voltage rises slowly and the maximum operating speed of the circuit is therefore greater than the recovery time figures would suggest.

The results shown in Figs. 6 and 7 are based on a re-applied trigger voltage which is well below the critical value. If the trigger is biased near to the critical value another mechanism comes into play which can have the same circuit effect as long deionization time, though its

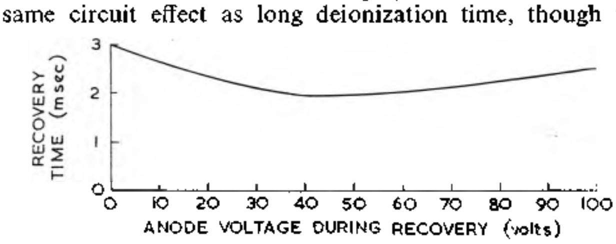


Fig. 7. Z803U recovery time against anode voltage during deionization, $V_a = 290V$ $I_a = 10mA$

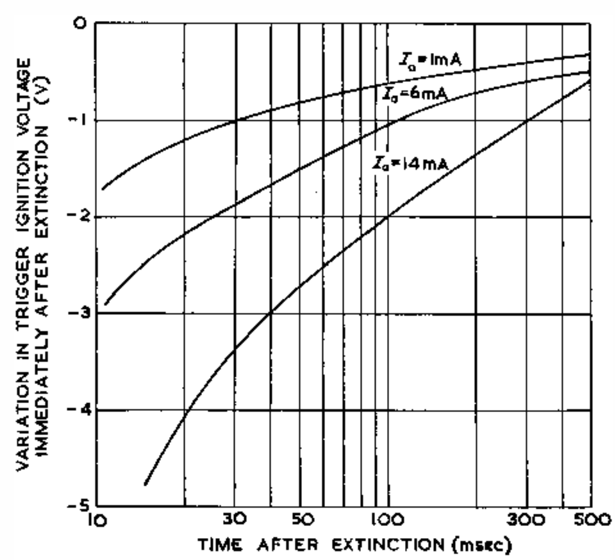


Fig. 8. Z803U trigger hysterisis.
Duration of conduction prior to extinction = 10msec

origins are very different. This so called hysterisis is shown in Fig. 8. It will be seen that the trigger ignition voltage of a tube is lowered after conduction. It only regains its true pre-conduction value after a period of about 100msec. The magnitude of the reduction of trigger ignition voltage at any instant prior to this is dependent on the current during conduction and the period of conduction. The reason for this behaviour is not fully understood, though it occurs in all trigger tubes. Work is now in progress to try and overcome this defect.

Trigger Effects

In designing circuits for cold-cathode trigger tubes the problem rapidly reduces to one of designing the trigger circuit. Apart from allowing for the tolerance in tube to tube spread of $V_{t(ign)}$ and ensuring a large enough discharge to cause anode cathode transfer, a number of other effects to be described, must be taken into account.

When the trigger potential is raised towards its critical value a small but increasing current will flow into the trigger cathode gap. This current corresponds to the portion AB of the general gas conduction curve, Fig. 3. The variation of this pre-ignition current with trigger potential for the Z803U is shown in Fig. 9. It will be seen that when the trigger is 1V below the critical value ($I_p = 10\mu A$), the current is 6×10^{-10}A. Reducing the priming current also reduces the pre-ignition current.

This pre-ignition current will obviously place a limit on the maximum trigger resistance in the capacitor firing circuit, shown in Fig. 1(b), since the voltage drop caused by the current flow in this resistor must be smaller than the difference between the trigger signal voltage and the critical trigger voltage of the tube. Fig. 10 shows a plot of the maximum trigger resistance as a function of the trigger supply voltage V_S.

A second factor which may limit the upper value of R_t under capacitor firing conditions is the total electrical leakage between electrodes, both internal and external to the tube. The trigger cathode leakage resistance of the

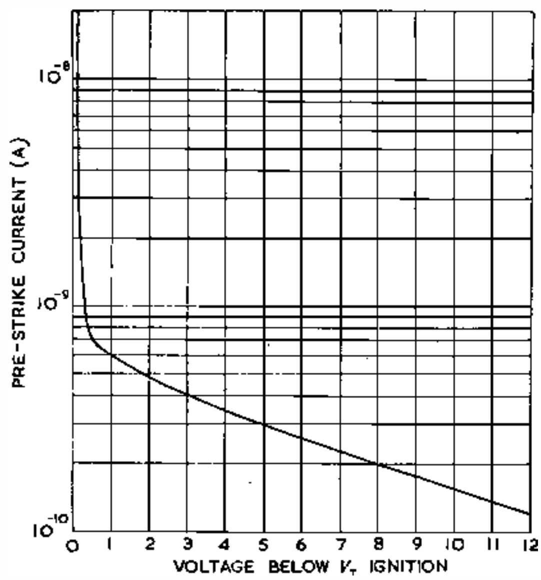


Fig. 9. Z803U pre-ignition current for different trigger voltages, priming current 10μA

Z803U is initially about $10^{11}\Omega$; leakage external to the tube will depend on atmospheric conditions and on the type of socket.

Up to the present, considerations necessary to ensure that the valve is triggered at the correct instant have been described but no attention has as yet been given to the effects which occur at the trigger once the discharge has been established.

Fig. 11 shows the potential distribution in an anode-cathode gap of a simple gas diode during a discharge. It will be seen that most of the tube is at a nearly constant potential V_p and that there is a rapid rise in potential near the cathode from 0 to V_p and a rather smaller one near the anode from V_p to V_m. If an electrode such as the trigger is inserted in this equipotential region it will tend to take up the potential V_p. If the trigger is taken via R_t to a potential above V_p normal positive trigger current will flow. Similarly, if it is taken below V_p negative trigger current will flow. The latter condition can be harmful to tubes such as Z803U unless the trigger resistance R_t is large, since the characteristics of the trigger surface are changed if it is called upon to act as a cathode.

The case in which the tube trigger current is either zero or positive is the normal condition and for the purpose of description can be divided into two sections, (1) $R_t < 1M\Omega$, and (2) $R_t < 1M\Omega$. Although a division has been made at this point it will be understood that there is a gradual change from condition (1) to condition (2).

In condition (1) the trigger current is comparatively high and the tube will behave as if it were two separate diodes in the same envelopes: trigger-cathode and anode-cathode. The trigger potential will therefore be greater than V_p.

If the trigger resistance is increased the current falls and the trigger potential, instead of rising as is normal in a gas diode, falls towards V_p, the condition in case (2).

The relative potential distribution outlined above will always be maintained. Thus if the cathode is raised in potential by say 20V all the other electrodes will tend to follow the same course. The currents in various circuits will

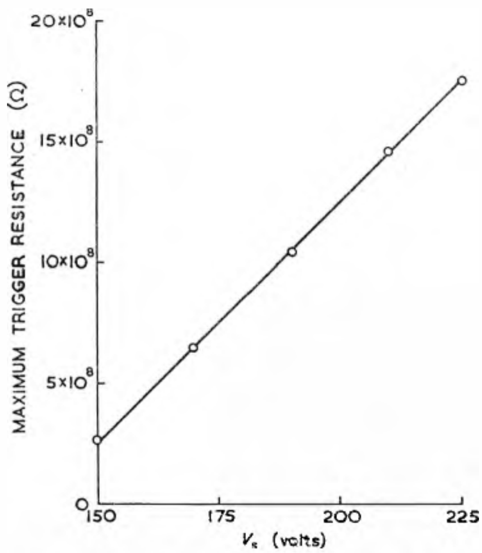


Fig. 10. Maximum trigger resistance/input voltage

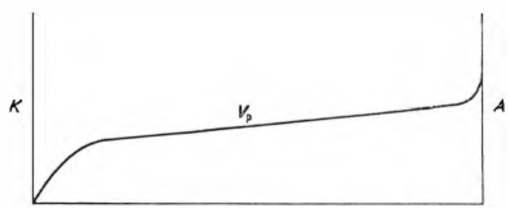


Fig. 11. Potential distribution in a glow discharge

adjust themselves accordingly. This fact is of importance in certain forms of self-extinguishing circuits to be discussed later.

CONSTRUCTION

There are numerous methods of construction of trigger tubes. Until recently most types have been made along conventional 'receiving' valve techniques, i.e. with the electrodes mounted between mica disks. However, with the newer sputtered envelope techniques, the use of mica is undesirable due to insulation and other problems. To overcome this difficulty all electrodes have to be mounted directly on to the electrode supports in the glass foot of the tube. The Z803U, which is an example of such a design, is shown in open view in Fig. 12.

The anode A consists of a stiff nickel wire, this wire is partially surrounded by a glass tube which increases the anode breakdown voltage of the tube.

T, the trigger electrode, is in the shape of a small hook whose position relative to the cathode K is closely controlled during assembly. The cathode is the largest electrode in the tube and consists of a plate, made of pure molybdenum which is bent in a U shape. It has a tripod type support.

It should be noted that the priming anode is on that side of the cathode remote from the other electrodes. This ensures minimum interaction between the priming discharge and the main discharge.

A very small quantity of uranium oxide is painted on the inside of the envelope. This is an emitter, whose external radiation is effectively attenuated by the glass envelope, but which can initiate ionization inside the tube at intervals of about 100msec. This ensures that a priming discharge, of about 10μA, can be readily established between the priming anode P, and the main cathode K.

The use of a pure metal cathode, as opposed to one coated with one of the usual barium compounds, results in a tube with relatively high operating voltages. For most industrial control applications this disadvantage is outweighed by the better life performance of the pure metal cathode tube.

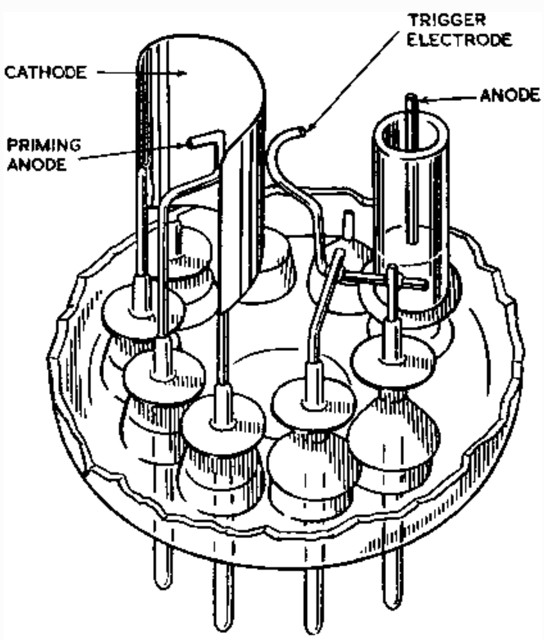


Fig. 12. Construction of the Z803U

The mechanical strength of the valve assembly is good, since all the electrodes, except the cathode, have a low inertia. The three cathode supports provide rigidity and the resonant frequency is above the range normally encountered.

During the process of manufacture a thin film of metal is evaporated or sputtered from the cathode on to the inside of the glass envelope. This has three main purposes:

(1) To establish a clean cathode surface.

(2) To remove or clean up any foreign active gases.

(3) To prevent any gas left in the glass from entering the tube.

Realization of the first aim also means that the tube is largely impervious to the effects of normal cathode bombardment, since such bombardment is simply a slight extension of the processing technique. However, this tube has been designed for operation with positive trigger and anode potentials and the cathode surface can be ruined if the tube is run even momentarily in the reverse direction.

The purity of the gas filling will only be impaired if the tube is run outside its ratings, since the high temperatures so produced can overcome the effects of the sputtered film.

The design of the tube, therefore, leads to the following properties, both initially and during life.

(1) Consistent electrode positioning.

(2) Pure and homogeneous cathode material.

(3) Pure and defined gas content.

These properties account for the stability of characteristics displayed by this tube over very long periods.

Reliability and Life

In general a tube fails in an application when one of its characteristics goes outside the limits tolerated by the particular circuit.

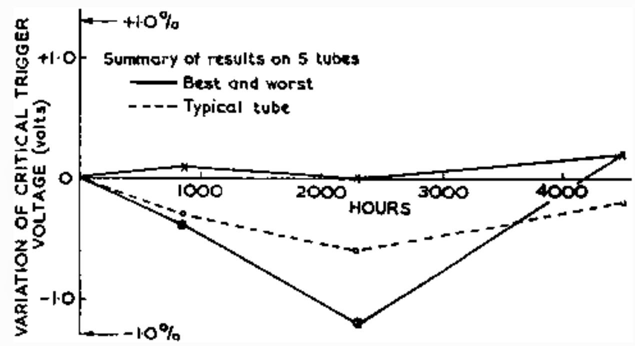


Fig. 13. Z803U variation of trigger ignition voltage on shelf

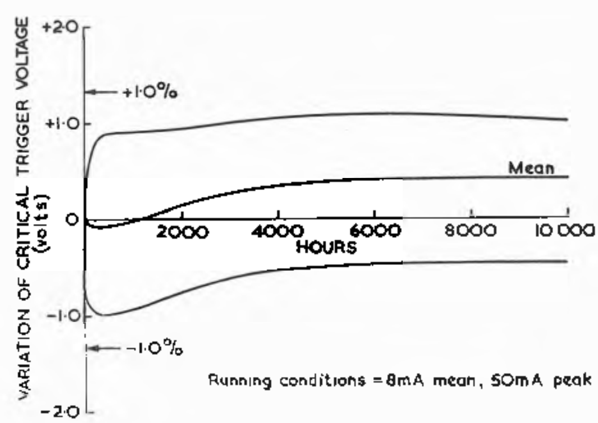


Fig. 14. Z803U trigger voltage variation

In a cold-cathode tube the changes in characteristic are likely to be either gradual ones, resulting from variations in gas purity, cathode surface composition, etc., or sudden changes due to mechanical shocks leading to movement of the electrode structure.

In the Z803U the predominant characteristic is the trigger ignition voltage, it is essential for many applications that this voltage remain stable over very long periods.

In many cases, such as safety applications, the tube will be used under standby or zero current conditions with only very occasional firing, and consequently, the 'shelf' behaviour of the tube is then of prime importance. Fig. 13 shows the average and extreme trends of trigger ignition voltage out of a group of five tubes during the first 5 000 hours of shelf life. It will then be seen that on these samples the trigger strike fell by an average of 0·5 per cent over the first 2 500 hours and then rose gradually: it is too early to state if this direction of trend is typical. The worst tube showed a maximum fall of $-$ 1·0 per cent.

A somewhat rarer duty is that of continuous operation. Fig. 14 shows the average, and worst trends, on a group of 13 early production tubes operated at 8mA mean, 50mA peak, from a half wave rectified 50c/s supply.

In both the above life tests the changes in other characteristics were small.

Tests at higher currents indicate a sudden falling off in life stability as the mean current is raised. It is likely that the maximum continuous current rating is about 25mA. At higher currents the dissipation is excessive and gross characteristic changes occur.

These results, which were obtained on tubes of early manufacture, suggest that the electrical reliability is good. In common with the experience on pure metal reference tubes, there are no signs to suggest any limit to the length of normal operating life.

There are also signs that under continuous current duty the greatest changes in trigger ignition voltage occur during the first few hundred hours of life. This also is in common with the behaviour of reference tubes.

At first sight it might be thought that this type of structure would be mechanically weak compared with that of a mica supported valve. Previous experience, however, has shown that this is not the case and in order to obtain some yardstick of this particular design a small batch of tubes have been subjected to the severe fatigue and shock tests laid down in the British Services specification for 'reliable' valves. (K1001 appendix 9).

Tubes were subjected to the fatigue test at 170c/s, 6*g*, over a duration of 100 hours. The average change (arithmetic) in trigger ignition was 2·5V.

Another batch of tubes was subjected to 20 impacts at 500*g*. The average change after this test was only 0·6 per cent. The results confirm that the mechanical strength of the tube is good.

(To be continued)

The Calibration of Vibration Pick-ups

By K. C. Foster*

Basic principles for the measurement of amplitudes of vibration are outlined and are suitable in varying degrees for applications within the audio frequency range. The methods described can also be adapted for the measurement of any form of vibration provided that due account is taken of the waveform to be measured.

SEVERAL methods are available for the calibration of vibration pick-ups, and choice of method will depend on the requirements of frequency and accuracy, and whether the calibration is for laboratory use or for production work.

For laboratory calibrations involving a relatively high degree of accuracy, measurements are made by optical means, while for production calibration comparison methods are employed, using a pre-calibrated standard pick-up and direct-reading meter.

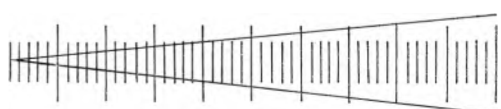
For all methods, a well designed vibration generator, (preferably of the electromagnetic type for audio frequency work or magneto-striction type for supersonic work), a stable oscillator and a reliable frequency meter are required, and the equipment used should be capable of covering the full range of frequency and amplitude required.

To obtain the best degree of accuracy, it is essential that the output of the vibration generator shall be truly sinusoidal and linear. The former quality is normally determined by the fidelity of the oscillator and its associated amplifier and the latter is determined by the mass balance of the vibration generator moving-coil system and anything which may be mounted on it. Any resonances will upset both conditions.

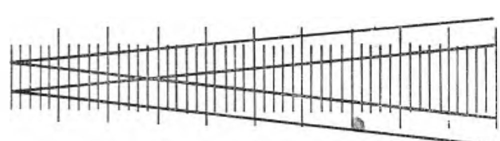
The vibration pick-up should be mounted on the vibration generator as simply and as rigidly as possible.

Method 1. Direct Optical Measurements (*Foppl's Triangles*)

The use of Foppl's triangles provides a simple method of measuring the amplitude of vibration but is limited to the lower audio frequencies and is also limited in accuracy by (a) the accuracy of production of the triangle and (b) normal reading error. Fig. 1(a) shows an example of the Foppl's triangle and Fig. 1(b) is an impression of what is seen during operation. To make such a triangle a large scale drawing is made as accurately as possible and is then photographically reduced to a size suitable for mounting directly on to the vibration pick-up. The triangle is mounted on the pick-up so that its horizontal axis is at right-angles to the line of vibration. When viewed under vibration two triangles will be visible, one being at the top of the movement and the other being at the bottom of the movement. The two triangles will cross over at a



(a)



(b)

Fig. 1. (a) Foppl's triangle
(b) Foppl's triangle subjected to vibration of $\pm 0{\cdot}140$in amplitude

certain point and this point referred to the vertical lines gives a direct indication of the peak to peak amplitude. For example, if the lines cross at the 3rd vertical and the verticals are of 1mm calibration the amplitude from peak-to-peak will be 3mm. In drawing up the triangle the angle

* *The Plessey Co. Ltd.*

between the converging lines should be as small as possible and the lines themselves should be as thin as possible consistent with visibility, minimum and maximum amplitudes required and the maximum permissible dimensions. The spacing of the vertical calibration lines will of course be determined by the angle between the converging lines. It is normally arranged that the spacing of the vertical lines is ten times the vibration amplitude. Under vibration, due to the thickness of the converging lines, a dark diamond is formed which often covers two or three calibration verticals. The centre of this diamond should be taken as the point of crossover.

In making the calibration the oscillator is tuned in turn to each of a number of selected frequencies within the range to be covered, and the amplitude of vibration is adjusted so that the crossover point of the triangle occurs at a pre-determined level. The output of the pick-up is then measured by means of a valve-voltmeter or oscilloscope. The amplitude reading, A, can be converted to velocity (v) or acceleration (a) by one of the two formulae:

$$v = 2\pi f(A/2) \qquad a = \frac{4\pi^2 f^2 (A/2)}{32{\cdot}2(\text{in}) \text{ or } 981(\text{cm})}$$

and output against velocity or acceleration can be plotted. Normally this will be a straight line within the working range of the pick-up.

An alternative to the Foppl's triangle is to use two parallel lines which are spaced apart by a known distance (say ·002in) the amplitude of vibration is adjusted until the white space between the lines just disappears, but this method is not to be recommended as two observers with differing eyesight will get different results.

Method No. 2. Cathetometer or Vernier Microscope

In this method a cathetometer or vernier microscope is used to measure the amplitude of vibration. With a good instrument this is perhaps the most accurate method for normal purposes since it is fairly easy to measure amplitudes of 0·01mm. Choice of instrument will be determined by local circumstances. If the vibration generator is a small one the vernier microscope is most suitable, but if the vibration generator is a large one the microscope cannot usually be mounted near enough to the pick-up to ensure adequate focus and the cathetometer is better. Whichever instrument is used however it is essential that it is mounted on a steady surface, and the vibration generator should be supported on sponge rubber or other suitable anti-vibration support. Vibration of the body of the vibration generator will not affect the calibration of the pick-up.

The microscope is focused on a portion of the body of the pick-up from which a strong light is reflected. Examination of this surface will show a number of needle points of light of varying size. A very small sharply defined point of light should be selected and lined up close to the vertical of the cross-wires of the microscope or cathetometer telescope. When vibration is applied to the pick-up the point of light will become a line with bright ends, as shown in Fig. 2. The horizontal of the cross-wire is lined up with either end of the line of light, the vertical of the crosswire being trimmed to be parallel with the line of light. The reading of the vernier scale is then taken. The microscope or cathetometer is then adjusted to bring the horizontal cross-wire to the opposite end of the line of light and a further reading is taken. The difference between the two readings gives the amplitude. Similar readings are taken throughout the required range of frequency. In this case the amplitude of vibration is so controlled that the output of the pick-up remains constant at all frequencies. This eliminates any error of calibration of the voltmeter or other device which is used for monitoring the output of the pick-up.

It should be noted that any non-linearity in the movement of the vibration generator will cause the point of light to form an elipse or even a circle and accurate measurements cannot be made under such conditions.

Method No. 3. Cathetometer or Vernier Microscope with Stroboflash

This method is similar to method No. 2 except that instead of using a steady source of light the light from a stroboflash is used.

The stroboflash is coupled to the oscillator or amplifier in such a way that the light is energized at the positive or negative peak of each cycle of the signal and arrangements are made to switch in a phase reversal circuit. In this way a single spot of light is observed at one extremity of movement of the pick up and when the phase reversal switch is operated the point of light is seen at the opposite end of the movement. The difference between measurements taken at each position of the point of light indicate

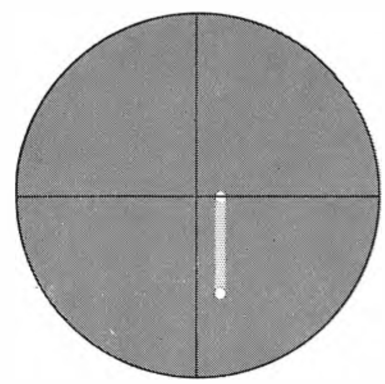


Fig. 2. Line of light viewed by microscope or cathetometer

the amplitude of vibration. This method involves more complex apparatus than that of method No. 2 but is somewhat quicker and more accurate.

Method No. 4. Reflected Light

Fig. 3(a) shows the set-up for this method, which is suitable for the measurement of small amplitudes at high frequencies. A small curved mirror or a prism with a curved reflecting surface (Fig. 3(b)) is mounted firmly in a convenient position on the pick-up and a narrow pencil of light is directed on it. When the vibration generator is at rest the light impinges on the centre of the mirror and is reflected on to the centre zero of a suitable scale mounted as far away as practicable. Calibration of the scale will be determined by the curvature of the mirror or prism. When vibration is applied the moving mirror projects a beam of light in an arc on to the calibrated scale, the line of light produced on the scale being proportional to the amplitude of vibration. By using all available space for the reflected light beam considerable amplification of the vibration amplitude can be achieved. Accuracy, however, is dependent on the accuracy of curvature of the mirror or prism and also on the accuracy of setting up of the reflecting medium. Any vibration of the light source will of course create considerable errors and this is perhaps the most difficult factor to cope with.

An alternative arrangement is shown in Fig. 3(c) which may be more convenient in certain circumstances. In this case a plane mirror is mounted at an angle on the pick-up

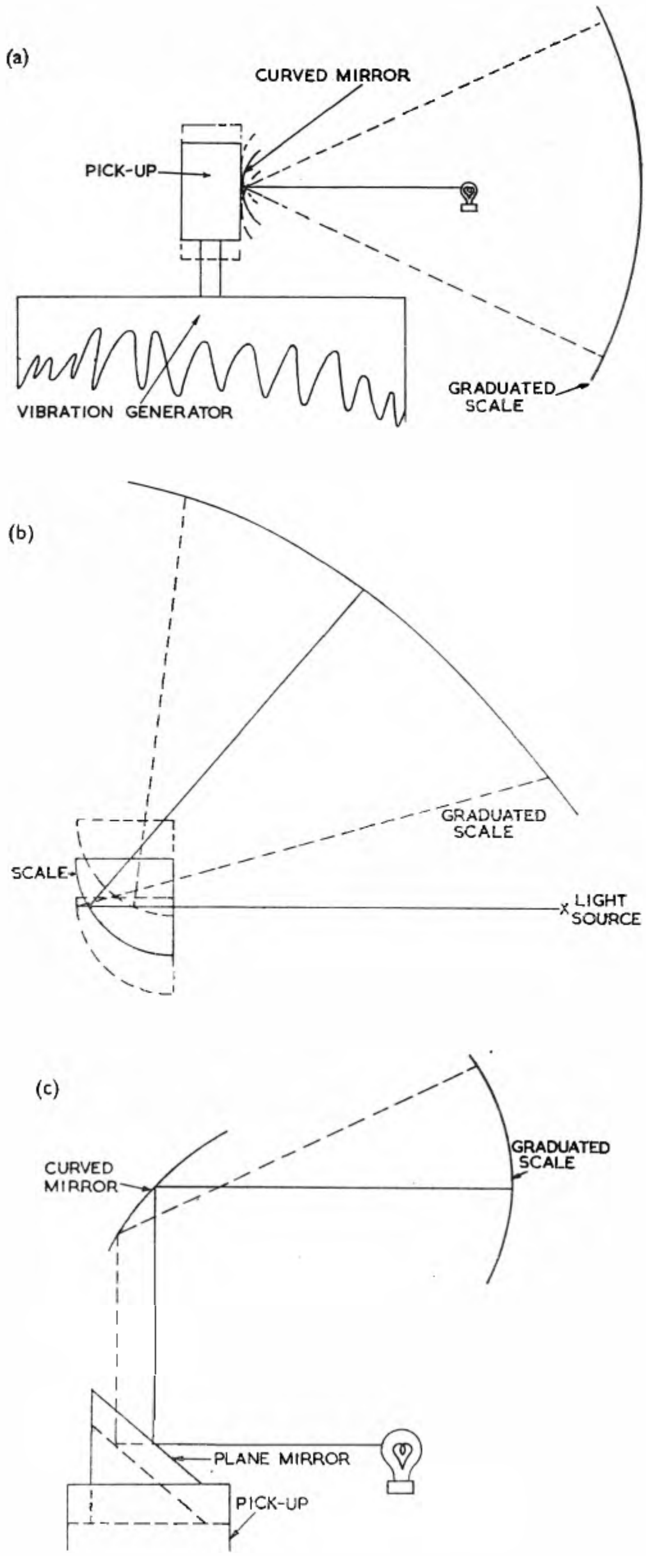


Fig. 3. (a) Method 4
(b) Prism with curved reflecting surface
(c) Double reflected beam

and this reflects the light on to a curved mirror which, in turn, transmits the light to the calibrated scale.

Method No. 5. Resistance Strain Gauge

If a thin flat spring is supported between a rigid block and a close-fitting slot located as closely as possible to the vibration pick-up (Fig. 4) the spring will be flexed by the movement of the vibration generator and pick-up. The degree of flexing, and consequently the amplitude of vibration, can be measured by means of a resistance strain gauge fixed to one surface of the spring. It is important that the slot shall be a very close fit on the spring and it is also important to ensure that no fundamental or harmonic resonance occurs in the spring at the frequencies for which measurement is required. This method is suitable for small amplitudes and, where the amplitude is low enough, the spring may be anchored rigidly to the pick-up or its mounting, to eliminate compliance in the slot.

Method No. 6. Matrix and Counter

For this method it is necessary to have a matrix which comprises a number of parallel lines of exactly equal thickness, spaced apart by a distance exactly equal to the thickness of a line. The matrix can be constructed of a number

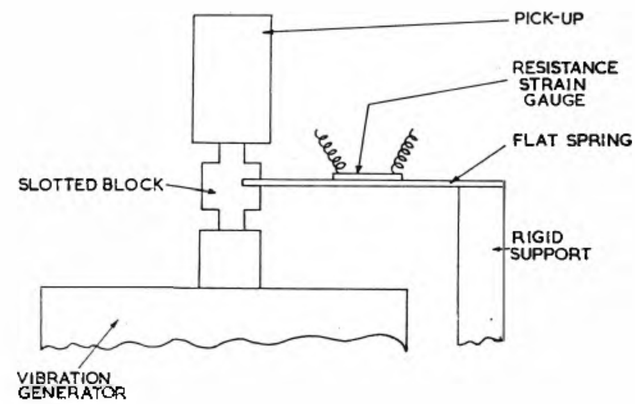


Fig. 4. Resistance strain gauge

of fine wires stretched across a frame or, better still, it can be a photographic plate which is a reduced photograph of a large drawing. The matrix is fitted rigidly on top of the pick-up and a plate bearing a slit is independently and rigidly supported as close to it as possible, so that the slit is parallel with the lines of the matrix. The width of the slit is exactly equal or slightly less than the width of a line or space on the matrix. A parallel beam of light passing through the slit will alternately be blocked or allowed to pass by the lines and spaces of the matrix as the pick-up is being vibrated. By means of a photo-electric cell and electronic counter the light pulses produced are counted. The amplitude of vibration can be calculated by the formulae:

$$A = \frac{N \times 2T}{f}$$

where N = Number of counts per second
T = Thickness of line or space
f = Frequency of vibration.

by using two matrices in juxtaposition, one being fixed and the other one moving, a considerable increase of illumination of the photocell is obtained, with consequent improvement in counting reliability. This method is limited to amplitudes of the order of ·001in upwards due to the effects of parallax, fringe effects etc.

Method No. 7. Interference Fringe Effect

This method is suitable for the measurement of very small amplitudes and entails the use of a synchronized monochromatic stroboflash. The interference fringe effect is well known and is often used for the measurement of very small distances down to the wavelength of the monochromatic light. Fig. 5 shows the principle of the fringe effect in which a monochromatic light (preferably produced

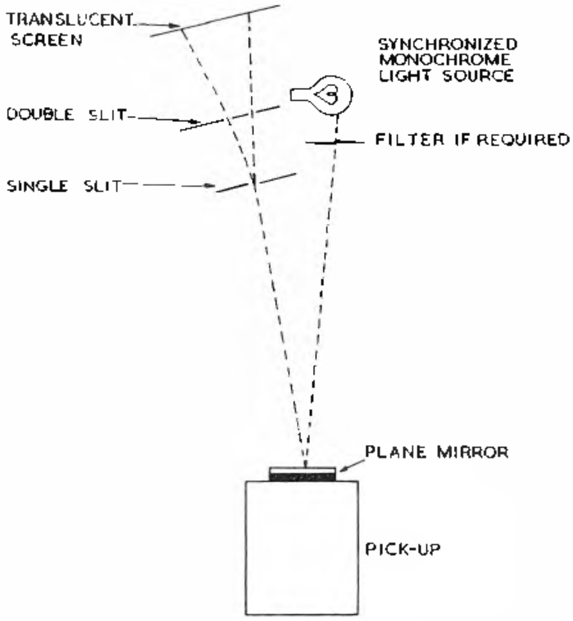


Fig. 5. Interference fringe effect

by a sodium lamp or a mercury vapour lamp fitted with a suitable filter) passes through a single narrow slit and then through two narrow slits. Interference fringe patterns are produced on the translucent screen. The position of the fringe lines will depend on the distance of the light source from the single slit.

For vibration measurements a suitable reflector, such as a small optically flat mirror is mounted on top of the vibration pick-up and the monochrome stroboflash light is focused on the mirror, the light being synchronized to be energized at the positive peak of each cycle. reflected light passing through the slits produces interference fringe lines of which the position is noted. The stroboflash is then synchronized to operate on the negative peak of each cycle and the displacement of the fringe lines will enable the amplitude of vibration to be calculated.

General Comments

The methods outlined in this article are all practicable methods but the results obtained will depend for accuracy on the skill and technical knowledge of the operator in addition to the accuracy of the various pieces of equipment involved.

In general, for amplitudes greater than 0·02mm Method No. 2 is the most accurate and reliable. It is however possible to combine some of these methods to obtain greater accuracy. For example, in Method No. 1 the Foppl's triangle used with normal vision is limited to amplitudes greater than 0·1mm. By viewing a much smaller triangle through a microscope or cathetometer amplitudes of 0·01mm can be measured without difficulty, and, in extreme cases it is possible to measure amplitudes as small as ·001mm. This latter case however, involves an extremely high degree of photographic skill in making an accurate reduction of 1 000:1 from a master triangle measuring 20in in length with calibration divisions marked at 1mm intervals, the tangent of the angle at the apex being 0·1.

In carrying out measurements of amplitude it is of the utmost importance that the motion produced by the vibration generator shall be truly in line with the normal motional axis of the generator and that measurements shall be made as close to this axis as possible.

A New Radio Telescope

A radio telescope with a 45ft parabolic reflector has recently been commissioned by the Royal Radar Establishment at Malvern for use in radio astronomy research.

Designed for operation over a wide range of wavelengths, it will be used initially for a study of the distribution of heavy hydrogen in the galaxy.

As a result of careful planning and improvisation, the cost of this equipment has been of the order of £10 000 only. The work has been completed from the time of the initial conception of the overall requirement in under 18 months. As an example of the improvisation to keep the cost low, the support for the aerials at the focus of the telescope is a steel telegraph pole bought from the G.P.O. while the mount for the reflector is an old radar turning mechanism which has been raised on a tower.

The principal item of expenditure has been the 45ft diameter parabolic reflector which has been constructed of aluminium.

The operating conditions for the reflector were rigid and required that the reflector maintains its true paraboloid shape to within a tolerance of $\pm 1/4$in under the following conditions: —

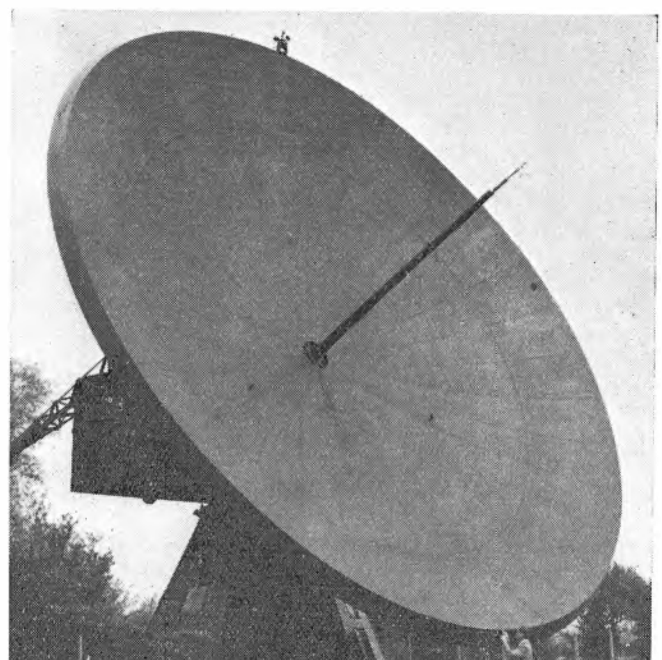
(1) The reflectors to be capable of turning through 360° in the horizontal axis and 80° in the vertical.

(2) Accuracy to be maintained in any position; at ambient temperatures from 4°C to 25°C; with temperature differentials across the surface, due to part shadowing, of up to 35°C; at relative humidities up to 100 per cent; and in steady winds up to 30 mile/h.

(3) The reflector to be capable of withstanding steady winds of up to 60 mile/h from any direction, with gusts of 10sec duration up to 110 mile/h; deposition of up to 2in of ice or 12in of snow on all upward-facing surfaces; without damage, vibrations, resonances, permanent deformation, or failure of any kind.

The reflector was designed jointly by the Engineering Department of the Establishment, by Precision Metal Spinnings (Stratford-on-Avon) Ltd., and by Messrs. Jenkins and Potter who acted as consultants; general advice on riveting by Dr. Whiteman, of the Military Engineering Experimental Establishment, Christchurch. The reflector was fabricated by Precision Metal Spinnings.

British Standard aluminium alloy 410-WP was used and was supplied by British Aluminium Ltd.



Error-Predicting D.C.-Restoring Circuits for Television Signals

By E. L. C. White*, M.A., Ph.D., M.I.E.E.

The performance of d.c.-restoring circuits of direct and negative feedback types is analysed. It is shown that, by a combination of techniques, the low frequency cut-off of the circuit prior to d.c.-restoring may be raised by a factor of 11 *compared with that required with methods commonly employed, for a specified degree of distortion.*

These techniques are particularly useful for long transmission circuits.

The new techniques have been used experimentally.

It has become common practice in the transmission of television signals to restore lost 'd.c.' and low frequency components by utilizing the fact that certain periodic datum portions of the waveform, for example the tips of synchronizing pulses[1], or portions at black level[2] immediately following these, should be at a constant potential, and any deviation from this can be measured and compensated.

The low frequency components become attenuated as a result of passage through any high-pass elements in the circuit, such as transformer or capacitive couplings. There may be appreciable phase errors introduced, even when there is not much attenuation, for example in transmission through a cable link on a carrier system using a vestigial sideband.

D.C.-restoring circuits working on this principle, in particular the 'black level clamp' are fairly satisfactory provided the cut-off frequency of the channel, considered as a high-pass filter, is sufficiently low.

It is believed that this principle has not been exploited so as to realize its full capabilities, particularly with regard to transmission through long circuits. For example, in the 2 500km circuit being considered by the C.C.I.R. and C.C.I.F. as a reference standard for the long-distance relaying of television signals by radio or cable, the cumulative loss of low frequencies renders it difficult to provide an overall performance such that a simple black-level clamp will effectively remove the signal distortion. Arising out of this, there are also difficulties in deciding how to specify the permissible low frequency performance of the link.

The aim of this article is to contribute to the solution of these problems by discussing the best possible performance of black-level clamps and related circuits, including, in particular, 'negative feedback clamps' in which the datum level is observed and compared with a reference potential, the difference being suitably processed to predict subsequent errors in the incoming datum levels and to make corresponding corrections.

The Nature of the Problem

Before proceeding to discuss the effectiveness of various circuits, it is useful to consider more closely the basic problem.

The signal after transmission through the high-pass circuits may be regarded as being the original undistorted signal together with a distortion component which comprises (reversed in phase) the missing d.c. and low frequency components, and the datum portions provide periodic samples of this distortion component. The problem is then to observe these periodic samples and reconstruct from them as accurate a replica as possible of the distortion component.

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Analysis of the problem can be carried out using either time or frequency as the variable. Steady state analysis reveals some interesting aspects, but is not in general so well suited to the problem as transient analysis, and for reasons of space the latter will be mainly used, with interpretation of some of the results in terms of frequency.

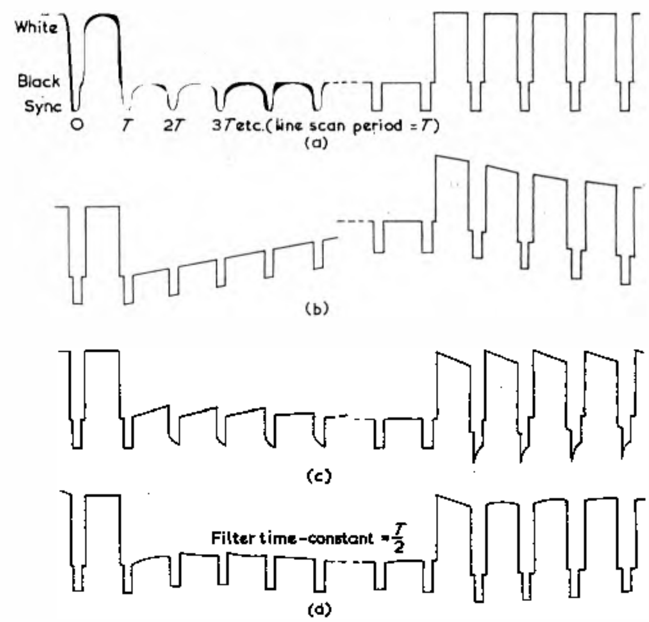

Fig. 1. Signal waveforms before and after d.c. restoration by some simple circuits

(a) Signal with extreme steps of mean level. (b) Effect of passing (a) through CR coupling with CR = 5T. (c) D.C.-restoration of (b) by simple clamp (Fig. 4c). (d) D.C.-restoration of (b) by simple clamp with smoothing (Fig. 4d) or (approx.) by feedback clamp with first order prediction.

In Fig. 1(a) is shown a signal waveform with sudden steps of mean level from an all white picture to an all black picture and vice versa. This is a suitable signal for transient analysis, as the lower frequencies, which are of most interest for this problem, are weighted most heavily. Fig. 1(b) shows the effect of passing this through a simple high-pass network consisting of a capacitive coupling into a resistive load, with a time-constant of $5T$ chosen for the illustration, where T is one line scan period. Fig. 1(b) may be regarded as the original signal of 1(a) with superimposed on it a distortion or error component as in Fig. 2(a).

The datum portions in 1(b), either the tips of the sync pulses or the short period of black level following sync pulses, are periodic samples of 2(a). These are observed by a sampling switch or gate, operated by pulses derived from the synchronizing pulses[2].

In the well-known diode d.c.-restoring circuit shown in Fig. 4(b), which is the simplest arrangement of all, the switch is a single diode, the sync pulses in the signal applied to the diode being themselves the switching pulses, and the datum portions observed being the tips of sync pulses. This circuit is not further discussed here, as its operation is asymmetrical and is not a suitable basis for the high performance systems considered.

The switch assumed in all the circuits considered, is a bi-directional one[3], such as a two-diode or four-diode bridge, operated by externally supplied pulses.

The output from the sampling switch may be applied directly to cancel the distortion component in the signal, or the output signal itself may be sampled, and the switch output used in a feedback circuit[4].

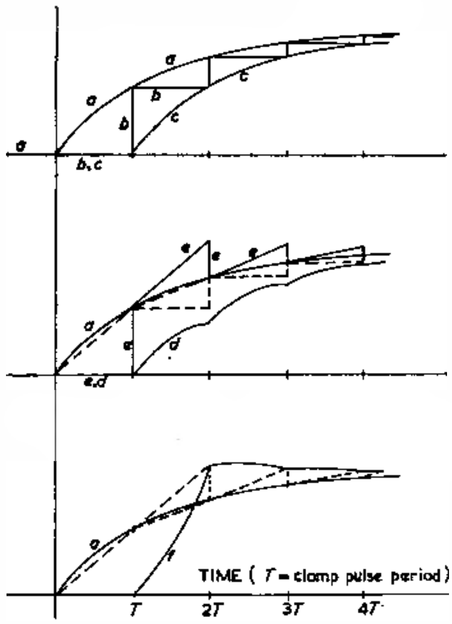

Fig. 2. Error and correction waveforms

(a) Incident distortion, for unit step input. (b) Zero order prediction (simple or feedback clamp). (c) First order prediction (feedback clamp). (d) Simple clamp with smoothing. (e) Modified first order prediction. (f) Second order prediction.

Basic Circuits

The basic circuits are shown in Figs. 4(a) and 5(a). E_i is an input signal, such as Fig. 1(a). A represents the transmission network up to the point where d.c.-restoration is to be applied, and F adds to the distorted signal a correcting signal to give the final output E_o.

A clamp pulse generator (C.P.G.) closes the sampling switch S periodically for short times during the selected datum portions of the signal.

For direct operation, S samples the input signal, and the sampled signal is fed to F, through a low-pass filter B.

For feedback operation, S samples the output signal, and the sampled signal is fed to F through a filter or processing amplifier P. P (or F) must incorporate phase reversal effective at very low frequencies and d.c.

Theoretical Limit of Performance

Sampling theory[5] shows that periodic samples at intervals T can carry information over a frequency range 0 to $1/2T$. To realize this requires the sampled signal to be passed through a low-pass filter, such as B in Fig. 4(a), with an infinitely sharp cut-off at $1/2T$.

Such a filter would have an infinite time delay, so in practice some compromise is essential. The attenuation at frequencies $\geqslant (1/2T)$ must be high, say >40dB, and the useful range for re-insertion is from zero to the frequency at which attenuation is only just noticeable.

Thus from one consideration the aim is a filter with as steep a cut-off slope as practicable. This will have a correspondingly large delay, and an equal delay must be provided[6] in the main signal channel from A to F if the restored low frequency components are to be properly timed with relation to the higher frequency components.

Since such a compensating delay must have no attenuation up to the top end of the video-frequency band, a large delay is difficult to achieve. Ultrasonic delays, such as mercury lines or quartz polygons offer possibilities, but are too expensive for normal use.

In the following discussion such delay compensation in the main signal channel is not considered.

Direct D.C.-Restoration

The circuit considered is Fig. 4(a), described above.

Let the output of A with unit amplitude step input be $A(t)$. The sampling switch only observes differences between the undistorted signal and the ouput of A, i.e., $A(t) - 1$, and by its operation it produces carrier frequencies $1/T$, $2/T$, $3/T$, etc., each with side bands due to the modulating signal $\left\{ A(t) - 1 \right\}$ and also passes the modulating signal itself. It is only the latter which is required, but in order to reject the unwanted sampling frequency, with its harmonics and sidebands, it is necessary for the filter characteristic $B(j\omega)$ to attenuate heavily (e.g., more than 30dB) at frequencies $1/T$, $2/T$, $3/T$, etc., and their sidebands. It is also necessary, in order to distinguish between lower sidebands of the carrier $1/T$ and the modulating signal, that $A(t) - 1$ should contain no substantial components at frequencies above $1/2T$.

Simple Clamp

The simplest example of direct d.c.-restoration, other than the asymmetrical diode circuit of Fig. 4(b), is the widely used 'clamp' circuit[3,7] of Fig. 4(c). The switch S is in series with a storage capacitor C_s, and it is assumed that (a) the time-constant of C_s with the resistance of the source and the switch in the conducting state is such that C_s reaches substantially the datum portion potential during the switching pulse, and (b) there is no appreciable load on the output, e.g., it is connected only to the grid of a valve. Here the filter B is merely the storage capacitor C_s, and the sampled potential held in this is directly subtracted from the input signal since, except during the actual sampling times, it is simply in series with the input.

Such a filter provides infinite attenuation at the clamp frequency $1/T$ and its harmonics, as required, but does not entirely reject the sidebands.

At this point it is opportune to consider the operation on a waveform basis. Fig. 1(b) shows the typical effect $A(t)$ on a step waveform entering A, due to the loss of low frequencies in A. Fig. 2(a) shows $1 - A(t)$, which can be expressed as

$$\left\{ 1 - A(t) \right\}_{t > 0} = a\ (t/T) + b\ (t/T)^2 + c\ (t/T)^3 + \ldots \quad (1)$$

The abrupt change of slope a/T at $t = 0$ is the most objectionable feature.

Fig. 2(b) shows the result of sampling the waveform $1 - A(t)$ and storing in C_s. It will be seen that the residual error is the difference between curves 2(a) and 2(b), and is of the sawtooth shape shown in 3(b), having a maximum amplitude a, from equation (1).

Consider for example a high-pass network A consisting of five stages of CR coupling, buffer amplifiers being assumed to prevent each stage loading the previous one. This filter has been used as an example because its indicial response $\dfrac{p^5}{(p + (1/CR))^5}$ is readily calculable, but the nature of the response of any multi-stage high-pass filter or television link will be broadly similar. The time scale has been normalized to units equal to the time-constant CR of one stage.

Fig. 8(a) shows the response to a unit step. The initial slope is $5/CR$, and equating this to a/T in equation (1) gives $a = 5(T/CR)$. Thus if the residual error amplitude a is to be, say 3 per cent, then $5T/CR = (3/100)$, or $CR > 167\,T$.

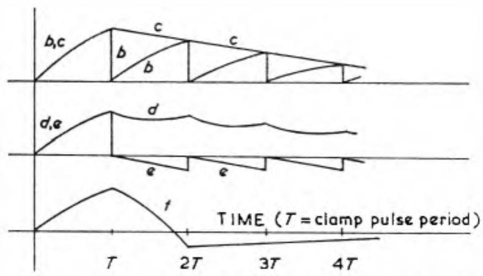

Fig. 3. Residual error waveforms lettered to correspond with Fig. 2

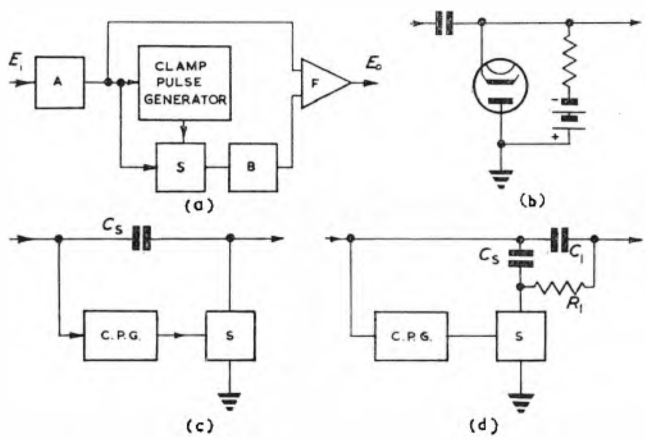

Fig. 4. Direct d.c. restoration circuits

(a) Generalized block schematic. (b) Diode d.c. restoring circuit. (c) Simple clamp d.c. restoring circuit. (d) Simple clamp with smoothing.

This can be compared with the theoretical limit by finding, the frequency at which $A(j\omega)$ gives say 0·3dB attenuation, which is approximately $\omega = (8/CR)$ or $\omega/2\pi = (1{\cdot}26/CR)$. But for limit performance this would be $1/2T$, or $CR = 2{\cdot}5T$. Thus the simple clamp falls short of the limit performance by a factor of $(167/2{\cdot}5) = 67$.

Simple Clamp with Smoothing[2,5]

The small additions shown in Fig. 4(d) reduce the residual error considerably except immediately following the step wave. The correcting signal of Fig. 2(b), developed across C_s, is smoothed by R_1 and C_1 to give Fig. 2(d), the residual error being as in Figs. 3(d) and 1(d). Since there is charge-sharing between C_s and C_1, C_1 should not exceed about $1/5\ C_s$. The optimum value of C_1R_1 is about $T/2$.

Feedback D.C.-Restoration

Instead of acting on the signal directly, or via a passive filter, as in the circuit of Fig. 4, d.c.-restoration may be carried out by observing the error existing during successive datum portions of the signal, and feeding back a derived waveform to an earlier point in the signal amplifier[4]. From this point onwards, the amplifier must have finite gain at zero frequency, and preferably have a substantially constant gain at all frequencies in the video band.

The system is shown in Fig. 5(a). It should be noted that the sampling circuit has no direct effect on the signal output, but only via the the feedback path. If the output impedance of F is not negligibly low compared with that presented by S when conducting, it is desirable to interpose a buffer amplifier, e.g. a cathode-follower, in front of S, but usually this is not necessary.

In general the design of the signal amplifier is easier with feedback d.c.-restoration, since the observation of error and insertion of correction is done in side channels which do not have to handle the highest signal frequencies.

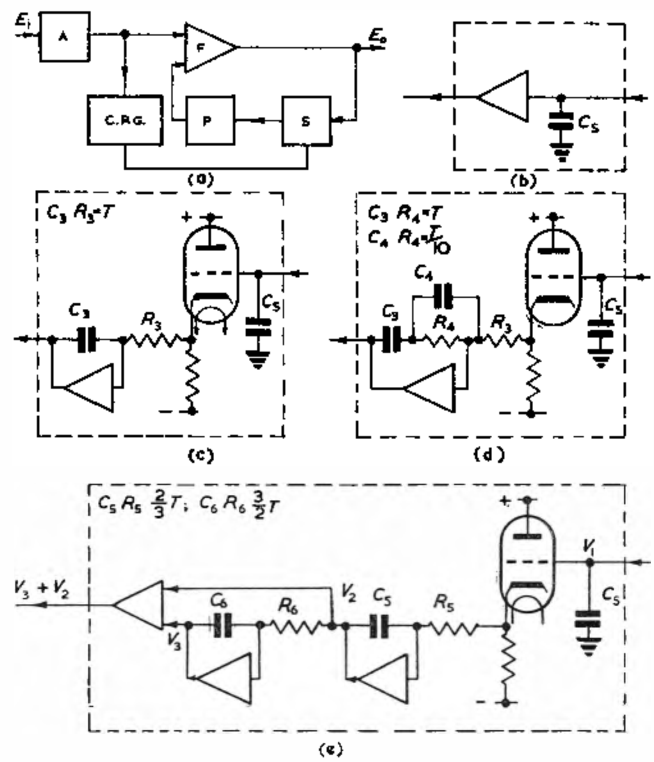

Fig. 5. Feed-back d.c. restoring circuits

(a) Generalized block schematic. (b) Form of 'P' for zero order prediction. (c) Form of 'P' for first order prediction. (d) Form of 'P' for modified first order prediction. (e) Form of 'P' for second order prediction.

There are other general advantages over direct d.c.-restoration. For example, the signal may be observed and effectively corrected at the final output of the amplifier, whereas direct clamping needs a low-impedance output to drive it and must often be followed by a cathode-follower or equivalent circuit to give a low impedance final output. Again, the feedback method restores d.c. in all stages of the amplifier between the feedback insertion point and the error observation point, and as this usually includes those stages working at the highest signal level, it is particularly useful if there is any degree of non-linearity, as it prevents the sync pulse amplitude being modulated by mean picture level.

Zero Order Prediction

The simplest arrangement for P is shown in Fig. 5(b) and consists of a storage capacitor C_s followed by a high-gain amplifier with a response time less than the duration of a sampling pulse. In this case, d.c.-restoration is exact at each pulse, and the correction is maintained constant till the next pulse, i.e., the operation is indistinguishable from that provided by simple direct d.c.-restoration. The circuit may be regarded as predicting that the correction required at the next sampling pulse will be unchanged from that required at the previous one, and hence may be described as a 'zero order prediction' arrangement.

First Order Prediction

By incorporating one integrating circuit in P, so proportioned that the output for an error a in the storage capacitor is $1/T\int a\,dt$, a correcting waveform as in Fig. 2(c) will be developed, with residual error proportional to the slope of $1 - A(t)$, as in Fig. 3(c).

Since no discontinuities are introduced, even though the average residual error is greater, this is an improvement over zero order prediction. Fig. 5(c) shows a practical arrangement using a virtual earth or 'Miller' integrator.

One way of removing the constant residual error, except for a transient caused by the abrupt change of slope, is to arrange two first order prediction circuits in tandem, the performance being very similar to that described under 'Second Order Prediction'.

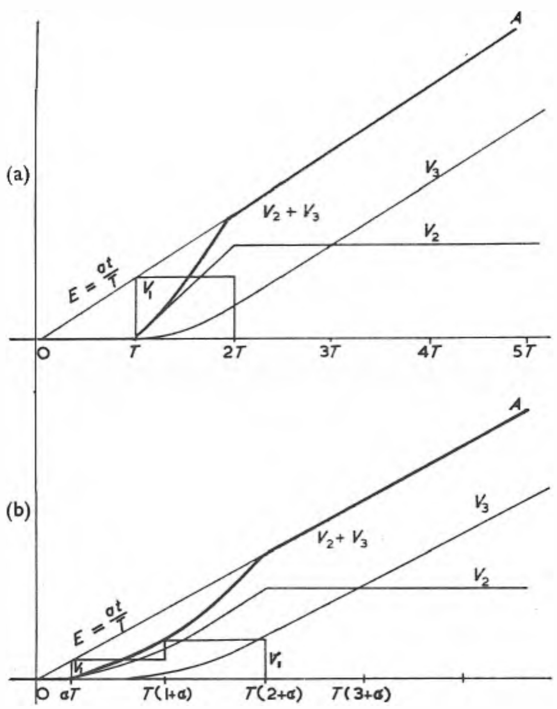

Fig. 6. Operation of Fig. 5(d)

(a) Step in input waveform coincident with sampling pulse. (b) Step in input waveform at time T before next sampling pulse.

Modified First Order Prediction

The residual error of first order prediction can be eliminated for a constant slope of input error signal, at the cost of re-introducing a discontinuity, by combining zero and first order prediction[9].

Fig. 5(d) shows one way of doing this, by adding R_4 shunted by C_4 in series with C_3 in Fig. 5(c). By making $R_4 = R_3$, the correcting waveform develops an immediate correction equal to the observed error, and C_3 causes the correction to grow as though extrapolated from the last two observations, as shown in Figs. 2(e) and 3(e). The correction must be delayed substantially for the duration of the clamp pulse, by C_4 shunting R_4, or else the full initial error will not be retained in C_3.

Second Order Prediction

To obtain zero residual error in a single d.c.-restoration circuit when compensating for a constant slope distortion without the discontinuity introduced by the previous system *P* requires two integrators in tandem. A steady slope in the compensating signal output of the second integrator requires a constant input, which can be supplied from the first integrator even though the residual error input to the latter is quickly reduced to, and remains at, zero.

For stability it is necessary to add the outputs of the two integrators, as in Fig. 5(e).

Let the residual error in the storage capacitor, C_s be V_1, and the output of the first integrator be $V_2 = (1/T_1)\int V_1 dt$, where T_1 is a parameter of the integrator. In particular, if the integrator is of the virtual earth type, with the see-saw arms consisting of resistance R_5 and capacitance C_5, then $T_1 = R_5C_5$. Similarly, the output of the second integrator is $V_3 = (1/T_2)\int V_2 dt$.

The appropriate values of T_1 and T_2, for the case of $F = 1$, to give exact compensation after, at most, two sampling pulses following a change of slope of the input distortion, are derived in the Appendix. They are $T_1 = (2/3)\,T$, $T_2 = (3/2)T$, and the relevant waveforms are shown in Fig. 6.

The circuit in effect predicts the correction required at the next sampling pulse by linear extrapolation from the two previous samples, with smooth curves joining the sampling points, Thus it is simple to derive graphically its performance with an input of smoothly changing slope, as in Fig. 2(f).

This type of circuit is well-known in auto-tracking radars, which are essentially feedback arrangements controlled by a sampled error signal[10,11]. 'Second order prediction' corresponds to 'zero velocity lag' in the radar case.

Reduction of Interference

The graphical construction mentioned above is readily applied to determine the reduction of interference, e.g., from power lines, picked up by the transmission circuit.

Consider an interfering signal of frequency $\omega/2\pi$ and unit amplitude. Its maximum slope is $\pm\omega$, and its maximum rate of change of slope is $\pm\omega^2$.

Thus the peak residual signal from zero-order prediction or first-order prediction is $\pm\omega T$, and from second-order prediction is $\pm(\omega T)^2$.

For example, 50c/s pick-up on a 405 line 50 field/sec system is attenuated 30dB by the simpler circuits and 60dB by the second-order circuit.

Experimental Verification

The circuit of Fig. 7 was set up to check the performance of second order correction. The circuit is divided into blocks to correspond with Fig. 5. No particular care was taken over high frequency responses or linearity of the signal channel, as the object was not to produce an engineering design, but to test a principle.

The signal amplifier F was designed to have a gain of 6 from either input, and has a cathode-follower output stage. A suitable output signal level is 10V overall, and minimum external load resistance about 1kΩ. The switch S is a diode bridge as commonly used in clamp circuits. Each integrator in P is a virtual earth circuit as in Fig. 5(e), each high gain amplifier comprising a triode amplifier and cathode-follower. Since the first integrator loads the input with a resistor R_5, the storage capacitor C_s is followed by a cathode-follower to feed R_5. A biasing resistor (shown as 2·7kΩ) is included in the cathode circuit in order to allow the signal output from F to be at approximately earth potential during the observed black level datum portions, while the grid at the junction of C_5R_5 is necessarily at about −3V.

Summing V_2 and V_3 is performed by inserting an additional resistor R_6 directly in series with C_6, so that the output of the second integrator repeats V_2 as well as providing V_3, the integral of V_2. The output $V_2 + V_3$ is tapped down the cathode-follower load resistor to give a suitable bias potential of about −250V for the amplifier F, and the resulting attenuation of 1/6 is made up by the gain of F. Many other detailed arrangements are, of course, possible.

Practical Results

The two time-constants T_1 and T_2 were made variable by the capacitors C_5 and C_6, and adjusted by trial to obtain the least distortion in the output when a heavily distorted signal of the type of Fig. 1(b) was fed in. A 625-line signal was used, so that $T = 64\mu$sec. The distorting network mostly used comprised five stages of a CR ladder network, the time-constant of each stage being 1msec and the impedance level increasing by a factor of 3 at each stage. It may be shown that the initial slope a of the distortion following a unit step is $0{\cdot}43/T$.

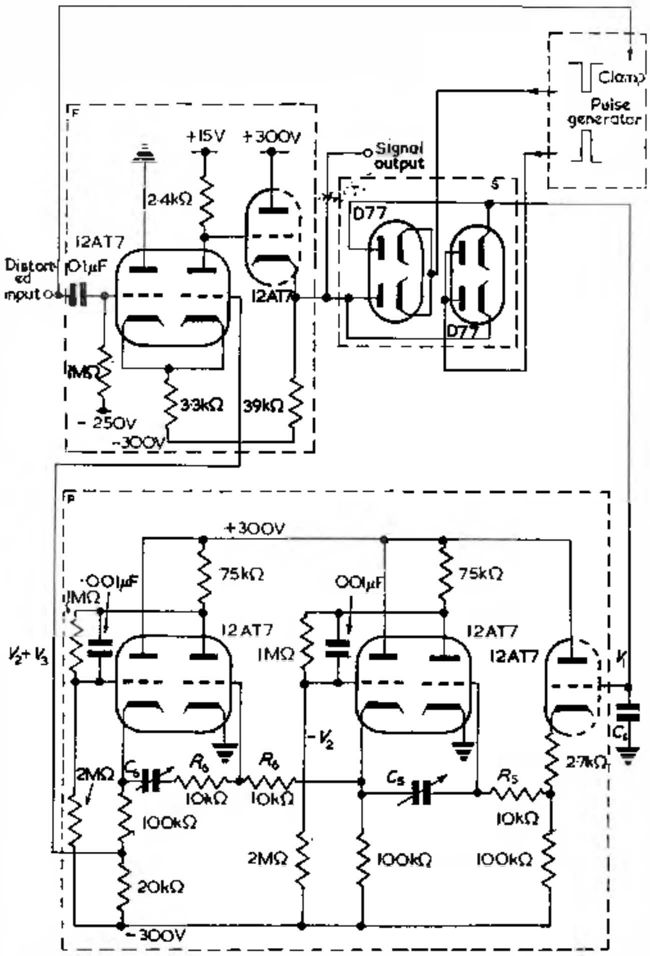

Fig. 7. Experimental second order prediction circuit

From a previous section it is seen that the theoretical values of T_1 and T_2 are 43μsec and 96μsec respectively. The best values of C_5 and C_6 found by trial were $0{\cdot}0018\mu$F and $0{\cdot}01\mu$F respectively. The resulting value for T_2 is very close to the theoretical, since $R_6 = 10$kΩ, but T_1 agrees less well. Allowing for the attenuation in the cathode-follower feeding R_5, which gives an effective value of $1{\cdot}42R_5 = 14{\cdot}2k\Omega$, T_1 is 26μsec instead of 43μsec. More careful measurement of the gain of F gave this as 4·9 instead of 6, which would make the expected value of T_1 35μsec. The remaining discrepancy is probably due partly to the see-saw amplifiers each having a gain of 23 rather than the assumed value of infinity, and partly to the finite length of the sampling pulse.

With the optimum adjustment, the residual error in the output waveform corresponded with Fig. 3(f).

Compensation of Slope Discontinuity

Figs. 2 and 3 show that increasing orders of prediction, though giving smoother outputs and reduced residual error for smooth input distortion waveforms, do not reduce the amplitude of the transient error due to a discontinuous change of input slope.

The distortion following a unit step has been expressed above as $\left\{1 - A(t)\right\} = a\,(t/T) + b\,(t/T)^2 + c\,(t/T)^3 + \ldots$

A considerable improvement results if the discontinuity in slope can be removed prior to d.c.-restoring, by compensating the first term $a\,(t/T)$. This may be described as first order slope equalization, and requires an integrator to give a constant slope $a(t/T)$ from a unit step input. Perfect integration is unnecessary, as the second and higher order terms are not being corrected, and it is sufficient if integration is fairly accurate for a time T following the step, by which time the d.c.-restoring circuit will take over.

Taking again as an example the distorting network A of Fig. 8, Fig. 8(b) shows the result of adding to Fig. 8(a) a suitable proportion of its integral, the total being $\dfrac{p^5 + 5p^4}{(p + (1/CR))^5}$.

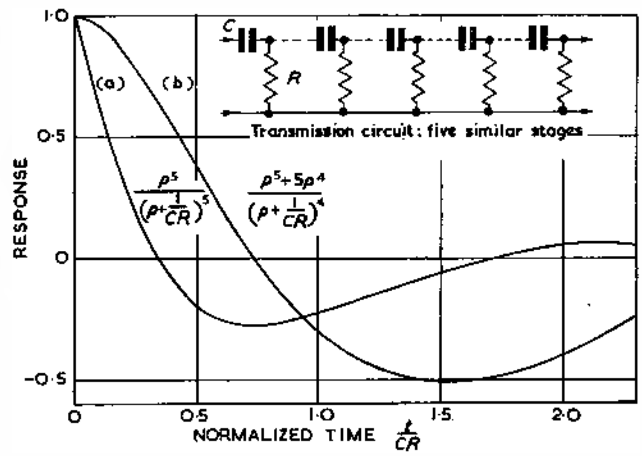

Fig. 8. Response to unit step of high-pass transmission circuit
(a) Before, and (b) after, slope equalization.

Consider the effect of sampling these waveforms at intervals $T = (CR/n)$, and suppose that a maximum residual distortion amplitude of 3 per cent is tolerable. Acting on the original curve (a), it was shown previously that either a first or second order prediction circuit would require n to be 167 and that this corresponds to a frequency for $-0{\cdot}3$dB response before d.c.-restoration of $1/133T$. On the slope equalized curve (b), a first order circuit would be three times better, i.e., $n = 55$, since the maximum slope is approximately 1/3 of that of curve (a).

The combination of curve (b) and a second order prediction circuit, however, reduces the required value of n to 15, as may be seen by taking second differences of the ordinates of curve (b) at abscissa intervals of 1/15. This is an improvement of 11 times in CR, or low frequency cut-off, and only falls short of the theoretical limit of $1/2T$ by a factor of 6.

Slope Equalizing Circuits

The simplest practical arrangement is to insert an adjustable capacitor C_7 in series with the anode load resistor R_7 of an earlier amplifier stage, as in Fig. 9(a), the feed current being supplied through a resistor R_8 such that C_7R_8 is several times T. C_7 is adjusted by observing the output distortion of the d.c.-restoring circuit, and will require to be approximately such that $C_7R_7 = (T/a)$. It will be seen that this circuit merely needs suitable choice of the commonly used decoupling components.

Another very satisfactory method, not requiring access to an earlier amplifier stage, is to take an adjustable portion of the signal output of the d.c.-restoring circuit of Fig. 5(e), integrate it, and add it to F. In the practical circuit of Fig. 7 this is conveniently done by interchanging the two integrators, so that the second one now does a pure integration, and giving an additional input to it through a resistor from an adjustable tap on the signal output, as shown in Fig. 9(b).

By either of these methods, a signal distorted as described can be restored so as to be almost indistinguishable from the original on a waveform monitor, and certainly indistinguishable on a picture monitor.

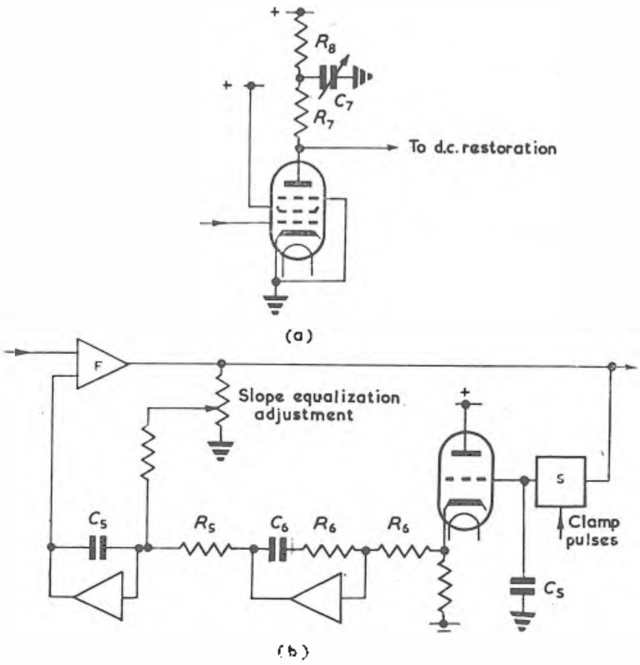

Fig. 9. Simple slope equalizing circuit

(a) Simple slope equalizing circuit. (b) Use of integrator in 'P' for slope equalization.

The above suggestions apply to existing equipment, but new equipment would preferably be designed so that each stage or unit gives no slope discontinuity.

Other Possibilities of Improvement

With all the improvements so far described, performance is still limited fundamentally, by the average delay of $T/2$ before an error is observed. This leads to the correction waveform developed being extrapolated from the two previous observations.

If the video signals are delayed by T before adding the correction waveform, it becomes possible to interpolate instead[12], which reduces the residual error eight times in the case of second order prediction applied to a parabolic distortion. Alternatively, the cut-off frequency can be raised 2·8 times for the same distortion, thus coming within a factor of about 2·2 of the theoretical limit.

Delaying video signals by T can probably be done, using an acoustic delay line, but considerably greater refinement of technique would be necessary than for the acoustic delay lines now used for microsecond pulse storage in electronic computers.

Conclusions

If a simple clamp d.c.-restoration follows a high-pass transmission circuit without low frequency phase corrections, the nominal cut-off ($-$ 0·3dB) frequency of the transmission circuit must not be more than about 1/133 of the line scan frequency if waveform distortion is not to exceed 3 per cent.

By a combination of slope discontinuity compensation of the input signals, or 'slope equalization', and 'second order prediction', it is possible to reduce this ratio to about 1/12.

Negative feedback clamping is recommended in preference to direct clamping. It separates all the error observation and processing from the video circuits, facilitates more sophisticated processing, and allows all the high-level stages in the video amplifier to operate with the d.c. component present.

Acknowledgments

Thanks are due to E.M.I. Electronics Ltd for permission to publish this work.

Appendix

Design of Second Order Prediction Circuit

To derive the appropriate values of T_1 and T_2, consider Fig. 6(a), in which OA represents a distortion component suddenly changing slope from zero to a/T volts/sec. The instant of change is taken to coincide with one of a series of sampling instants at intervals of T, other relative timings being considered later. At the next sample ($t = T$) an error $V = a$ volts is recorded and stored, so that $V_1 = a$. Hence

$V_2 = (1/T_1)\int_T^t a \; dt = \dfrac{a(t - T)}{T_1}$, for $T < t < 2T$, since V_2 is zero initially. Then $V_3 = (1/T_2)\int_T^t \dfrac{a(t - T)}{T_1}\, dt = \dfrac{a(t - T)^2}{2T_1T_2}$ also for $T < t < 2T$.

If the parameters are correctly chosen, the error V is reduced to zero at the second sample, and remains so subsequently. Hence for $t > 2T$, $V_2 = a\,(T/T_1)$ and $V_3 = (1/T_2)\int a\,(T/T_1)\, dt$. But the value of V_3 at $t = 2T$ is $aT^2/2T_1T_2$, hence $V_3 = (aT^2/2T_1T_2) + \dfrac{aT(t - 2T)}{T_1T_2}$.

For exact compensation, $V_2 + V_3 = (at/T)$ for $t > 2T$.

$$\therefore\; a\,(T/T_1) + (aT^2/2T_1T_2) + \frac{aT(t - 2T)}{T_1T_2} = (at/T)$$

This can be satisfied if $T_1T_3 = T^2$ and $(T_2/T) = 3/2$, i.e., $T_1 = (2/3)\,T$; $T_2 = (3/2)\,T$.

The waveforms of V_1, V_2, V_3, and $V_2 + V_3$ are shown in Fig. 6(a), and it is seen that only the sample at $t = T$ is in error. If the change of slope occurs part way through an interval T, full correction takes two sampling periods after the next sample, as shown in Fig. 6(b), which is drawn for the same parameters as Fig. 6(a). For the more general case of a rate of change of slope following the initial sudden change, Fig. 2(f) shows the way in which the compensating waveform follows. An additional integrator in P would correct for the second differential after three errors have been observed, giving third order prediction, and so on.

REFERENCES

1. Willans, P. W. British Patent Specification No. 422,906.
2. Blumlein, A. D. British Patent Specification No. 458,585.
3. Browne, C. O. et al. British Patent Specification No. 449,242.
4. Willans, P. W. British Patent Specification No. 464,828.
5. Oliver, Pierce, Shannon. Philosophy of P.C.M. *Proc. Inst. Radio Engrs. 36*, 1324 (1948).
6. White, E. L. C. British Patent Specification No. 459,853.
7. Wendt, K. R. T.V. D.C. Component. *RCA Rev.* (March, 1948).
8. White, E. L. C. British Patent Specification No. 693,220.
9. Nuttall, T. C. *J. Televis. Soc. 5*, 257 (1949).
10. Porter, A., Stoneman, F. W. A New Approach to the Design of Pulse-Monitored Servomechanisms. *Proc. Inst. Elect. Engrs. 97*, 597 (1950).
11. Lees, A. B. Interpolation and Extrapolation of Sampled Data. *Inst. Radio Engrs. Trans. on Information Theory, 11* (March 1956).
12. Collard, J. British Patent Specification No. 521,744.

Circuit Techniques Associated with
Transistor Broadcast Receivers

(Part 2)

By J. N. Barry*, M.Sc.

OSCILLATOR AND MIXER CIRCUITS

Assuming a standard i.f. of 470kc/s is being used, the oscillator must be capable of operating with sufficient amplitude and stability up to a frequency of 2 070kc/s.

The maximum possible frequency of oscillation for a transistor, f_{max}, will be the maximum frequency at which the transistor, in any circuit configuration, can give a power gain. (Note: it can be shown theoretically that this frequency is the same for all configurations). Hence, putting G_{max} equal to unity in equation (8), one obtains:

$$1 = (1/8\pi) \, . \, (1/f_{max}^2) \, . \, \frac{f_{\alpha}}{r_{bb'} \, . \, C_c}$$

or:

$$f_{max} = \sqrt{\left[\frac{f_{\alpha}}{8\pi \, . \, r_{bb'} C_c}\right]} \qquad \ldots\ldots\ldots\ldots (16)$$

Hence the maximum frequency of oscillation is proportional to the square root of the merit figure (for an alloy type junction transistor).

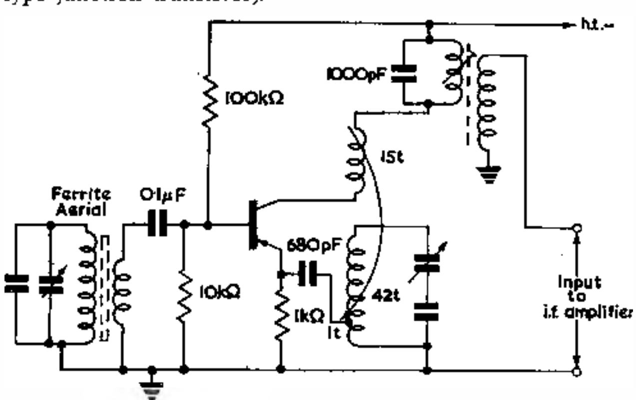

Fig. 13. An experimental oscillating-mixer stage for use with relatively low grade transistors

The use of a self-oscillating mixer circuit will only be feasible if f_{max} in equation (16) is appreciably in excess of 2 070kc/s. For transistors which are likely to give borderline operation it is desirable to design the circuit to give optimum performance as an oscillator. This probably requires the use of phasing circuits in the current feedback path to compensate for phase changes in the transistor current transfer function as the oscillation frequency increases, and may result in some loss of conversion gain.

However, if such a circuit is used, an example being shown in Fig. 13, it is found that the relationship:

$$f_{max} \propto \sqrt{M}$$

is borne out fairly well in practice. The experimental results of maximum oscillator frequency versus $\sqrt{M}$ using a number of transistors in the circuit are plotted in Fig. 14.

When using junction transistors having higher values of 'α' cut-off frequency and merit figure some of the above problems become much less difficult, and completely satisfactory oscillating-mixer circuits having nearly constant conversion gain over the band are possible. Such circuits can be designed to have oscillator frequency shifts of not more than about 1kc/s for a 2:1 decrease in h.t. voltage when using transistors having $f_{\alpha} > 5$Mc/s with typical conversion gains of the order of 22 to 26dB.

R.F. AND AERIAL CIRCUITS

Due to the fact that one of the main potential advantages of a transistor portable receiver is small size, it is customary to employ some form of ferrite rod aerial as the signal pick-up device.

The design of the aerial loop represents a compromise between conditions of best signal pick-up and best selectivity. For the former the coil should be wound about the centre of the rod; for the latter at one end of the rod. The variation of unloaded Q with the position of the coil on the rod may be seen from the curves of Fig. 15.

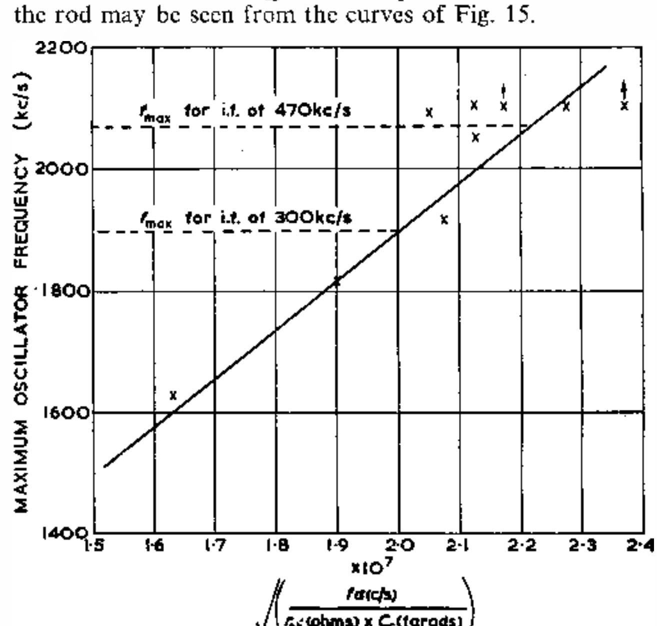

Fig. 14. f_{max} versus $\sqrt{M}$ for type GET3 transistors

If h = effective height of aerial loop,

Q_o = unloaded Q value,

it can be shown that for a given rod the optimum coil design is when the product hQ_o has a maximum value. This usually occurs with the coil wound about a mean point situated at approximately one quarter along the length of the rod. As will be seen from Fig. 15, this also provides the minimum variation of Q_o over the frequency band.

It can also be shown[11] that the signal pick-up will increase as the ratio l/D increases, where

l = length of rod.

D = mean diameter of rod.

The possible improvement here is usually limited by space considerations in the receiver.

It is not possible to match the aerial circuit impedance to the input impedance of the transistor mixer stage over the whole of the band, due to variation of the latter with

* Research Laboratories of The General Electric Co. Ltd., Wembley.

frequency. This may be seen by reference to the hybrid-π common emitter equivalent circuit[13] of Fig. 16(a), which represents the mixer stage from the signal aspect. From Fig. 16(b), the input impedance is given to a good approximation by:

$$Z_{in} \simeq r_{bb'} + \frac{r_{b'e}}{1 + j\omega C_{b'e} r_{b'e}} \quad \ldots\ldots\ldots\ldots (17)$$

or, since $r_{b'e} = (r_e/(1 - \alpha_0))$
and $C_{b'e} = (A/r_e\omega_\alpha)$,
the above equation (17) can be written

$$Z_{in} \simeq r_{bb'} + \frac{(1 - \alpha_0)r_e}{A^2}(\omega_\alpha/\omega)^2 - j\,.\,(r_e/A)\,.\,(\omega_\alpha/\omega) \quad (18)$$

Equation (18) is valid provided:

$$A^2\,(\omega/\omega_\alpha)^2 \,.\, \frac{1}{(1 - \alpha_0)^2} \gg 1 \quad \ldots\ldots\ldots (19)$$

where A = a constant = 1·22.

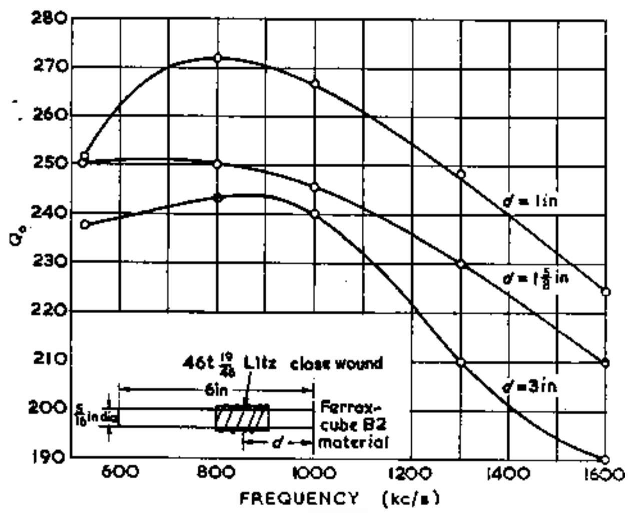

Fig. 15. Variation of unloaded Q of aerial coil with position on ferrite rod

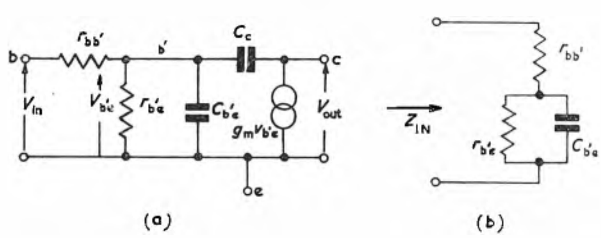

Fig. 16. Hybrid-π equivalent circuit at high frequencies

Thus it may be seen that as the frequency increases the magnitudes of both the resistive and reactive components of the input impedance decrease.

Measured values of input impedance for a mixer circuit similar to that shown in Fig. 13, obtained by using a suitable impedance bridge, confirm the above theory. Results are shown in Table 1, the standing emitter current being 0·35mA. Impedances are given in terms of an equivalent series circuit, i.e.

$$Z_{in} = r_s - jx_s \quad \ldots\ldots\ldots\ldots\ldots (20)$$

In practice the step-down ratio from aerial to mixer is usually chosen to reduce the unloaded Q value of the aerial coil by approximately 2 : 1 at the high frequency end of the band, where the performance of the ferrite aerial is beginning to fall off. Under these conditions loaded Q values of the order of 50 to 75 are usually achieved. This provides image rejections of the order of 35dB at the high frequency end. The loaded Q at the low frequency end is usually appreciably more (> 100), and better image rejection figures are then achieved.

TABLE 1.

FREQUENCY	TRANSISTOR 1; f_α=2·3Mc/s		TRANSISTOR 2; f_α=3·9Mc/s		TRANSISTOR 3; f_α=7·8Mc/s	
	r_s	$-jx_s$	r_s	$-jx_s$	r_s	$-jx_s$
550kc/s	340	−j150	399	−j658	373	−j833
750kc/s	220	−j130	280	−j424	343	−j580
1 000kc/s	200	−j75	240	−j344	347	−j390
1 600kc/s	120	−j35	225	−j192	250	−j275

Although values of this order are usually acceptable for small portable receivers, they are not really adequate for car radios. Hence, in this case, it is usually considered necessary to incorporate an r.f. stage. In this way image rejection figures in excess of 60dB may be achieved, though the use of such a stage can introduce certain other disadvantages. These include variation of the r.f. stage gain over the pass-band, which to a first order will be according to the curve of Fig. 9 (unless a transistor having an α cut-off frequency greater than about 15Mc/s is used). In addition it is necessary to provide more complex tuning arrangements.

Some early experimental car receivers operated the r.f. stage transistor under low collector voltage conditions with the object of reducing its noise factor. Recently, however, junction transistors have been manufactured having noise factors which are both lower in magnitude and also less dependent on collector voltage. Special operating conditions on the r.f. stage are hence likely to be of less importance in the future.

It can be shown[13] that the noise performance of a car receiver using a capacitive type aerial and an r.f. input stage will be slightly inferior to that of an equivalent all-valve receiver, approximately by a factor equal to $\sqrt{N_F}$, where N_F is the noise factor of the r.f. transistor.

For portable receivers using a mixer input stage the noise performance of valve and transistor equipments should be fairly comparable, unless the mixer transistor has a noise factor in excess of 6dB.

A.G.C. Circuits

Due to the limited amount of gain which can be controlled in a transistor i.f. amplifier, particularly in the last i.f. stage where overloading troubles must be avoided, the range of a.g.c.* in a transistor receiver is usually somewhat limited. Thus although the values required for a portable receiver (say 30 to 40dB) can usually be met, the range required for a car receiver (say 60dB) may present difficulty, although in this case some additional control can be obtained by applying a.g.c. to the r.f. stage.

Two methods of a.g.c. are possible, namely emitter current control or collector voltage control. From the point of view of range of control there is not much to choose between them, but technically emitter current control has some advantages and is the method usually adopted. This is principally because variations in collector voltage produce variations in effective transistor output capacitance which can cause a shift in tuning, but also because the voltage method requires the provision of decoupled resistance in the collector lead across which the control voltage is developed.

In order to reduce to a minimum the control power required for either of the above methods, it is usual to provide a.g.c. action by varying the base current of the controlled stage. The transistor then acts as its own d.c. amplifier and produces a correspondingly larger variation in emitter (and consequently collector) current. For an emitter current controlled stage using a pnp transistor a.g.c. is initiated by applying a positive-going potential

* *The control range is considered to be the change of input signal needed to produce a 6dB change of output.*

to the base. The reduction in gain is then principally due to changes in the resistance at the input terminals for either common base or common emitter connected stages. This may be seen qualitatively from the following relationships which apply at low frequencies, and are hence valid for a.g.c. considerations.

Common base:

$$(r_{in})_b \simeq r_e + r_b(1 - \alpha_0) \quad \ldots\ldots\ldots\ldots (21)$$

now

$$r_e = \frac{kT}{q(I_e + I_{cbo})} \quad \ldots\ldots\ldots\ldots (22)$$

where k = Boltzmann's constant,
q = electronic charge,
T = absolute temperature,
I_e = controlled component of emitter current.
I_{cbo} = residual emitter current with base-emitter junction reverse biased.
α_0 = low frequency value of α,
r_b = total base resistance.

Substituting for r_e in equation (21):

$$(r_{in})_b \simeq \frac{kT}{q(I_e + I_{cbo})} + r_b(1 - \alpha_0)$$

or

$$(r_{in})_b \simeq \frac{A_1}{(I_e + I_{cbo})} + r_b(1 - \alpha_0) \quad \ldots\ldots\ldots (23)$$

where A_1 = a constant at a given temperature.

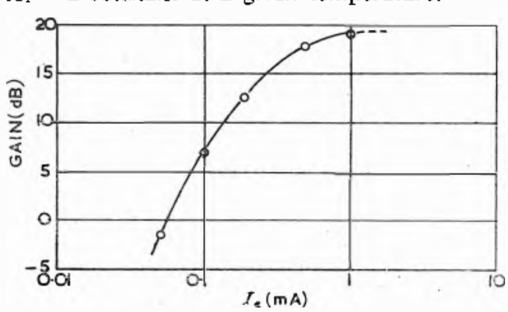

Fig. 17. Gain variation with emitter current—common emitter stage

Common emitter:

In this case one has, to a first approximation:

$$(r_{in})_e \simeq (r_{in})_b \times \frac{1}{1 - \alpha_0}$$

i.e.

$$(r_{in})_e \simeq \frac{A_1}{(1 - \alpha_0)[I_e + I_{cbo}]} + r_b \quad \ldots\ldots (24)$$

From equations (23) and (24) it may be seen that in each case the first term on the right-hand side tends to be inversely proportional to I_e for values of $I_e \gg I_{cbo}$. It is also evident that the maximum value of r_{in} in equation (24) is given by:

$$[(r_{in})_e]_{max} \simeq \frac{A_1}{(1 - \alpha_0)\, I_{cbo}} + r_b \quad \ldots\ldots\ldots (25)$$

When transistors having still higher α cut-off frequencies are available for use in the controlled stages, so that the working frequency is appreciably below the range $0{\cdot}1 < f/f_\alpha < 1$, it should be remembered that the stage gain then becomes more dependent on the value of α or α_{cb} for common base and common emitter stages respectively, and the variation of these two parameters with emitter current will also have to be taken into consideration.

The amount of gain variation which is achieved in practice is similar for either of the two circuit arrangements, although common emitter stages show a slightly faster rate of change at low values of I_e. For a common emitter amplifier operating at 470kc/s, practical measurements of gain for varying emitter currents gave the curve shown in Fig. 17. The transistor used had an α cut-off frequency of about 2·3Mc/s, and was designed to provide maximum gain at an emitter current of 1mA. It will be seen from this curve that the most rapid a.g.c. action takes place for low values of I_e, (I_e < 0·5mA), as would be anticipated from equation (24). A similar state of affairs exists in the case of the common base circuit, as shown by equation (23). It is hence customary to use a normal standing emitter bias of about 0·5mA for the controlled stages.

For common emitter stages a.g.c. action also produces changes in the reactive component of the input impedance and can consequently cause some variation in the tuning. However, practical measurements have shown that such variation is small, usually about 1 to 1·5kc/s in an upward direction for a reduction in emitter current to about 0·05mA. This is due to the fact that the value of the effective input capacitance is reducing (see equation (18)), and its effect on the tuning of the preceding collector circuit quickly becomes small, particularly if a large value of tuning capacitor is used.

For common base circuits variation in the tuning with a.g.c. is extremely small. This is due to the fact that the reactive component of the input impedance is to a first

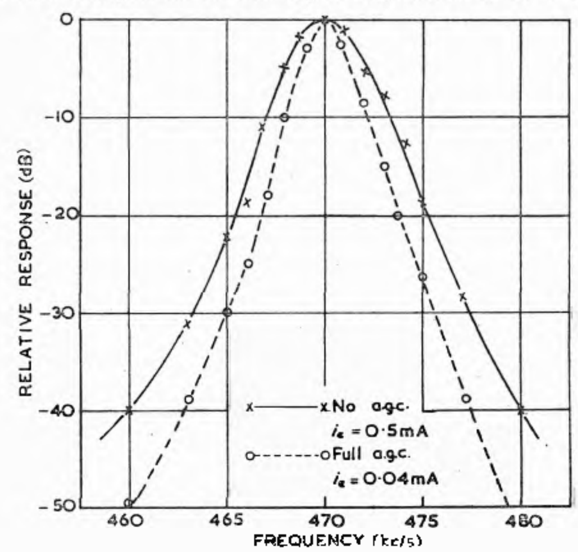

Fig. 18. Variation in selectivity with a.g.c. action—common base stages

order unaffected by emitter current changes. This may be seen by substituting the complex value of α given by:

$$\alpha = \frac{\alpha_0}{1 - jf/f_\alpha} \quad \ldots\ldots\ldots\ldots\ldots\ldots (26)$$

for α_0 in the expression for $(r_{in})_b$ in equation (23), and writing the l.h.s. as an impedance. This gives the relation:

$$(Z_{in})_b \simeq \frac{A_1}{[I_e + I_{cbo}]} + r_b\left[1 - \frac{\alpha_0}{1 + (f/f_\alpha)^2}\right] + j\,\frac{r_b\alpha_0 \,.\, f/f_\alpha}{1 + (f/f_\alpha)^2} \quad \ldots\ldots\ldots (27)$$

For either type of circuit the result of a.g.c. action is to increase the selectivity. This may be seen in the curves of Fig. 18, which show two i.f. selectivity curves for a car receiver having three i.f. stages. The stages were common base connected, and the effective control range for a change in emitter current from 0·5mA to about 0·04mA was approaching 45dB for 6dB change in output. It may be seen that practically no change in tuning has occurred.

If gain control is required primarily for the purpose of extending the range of input signals for which the receiver can be used, it may be possible to accept a reduced range of a.g.c. in conjunction with some modified form of volume control. The outline of such a circuit[14] is shown in Fig. 19. This method can be of use for portable receivers, as advantage can be taken of increasing the effective bandwidth under strong input signal conditions, thus aiding reproduction quality when most usable. In addition such a circuit can obviate variation in the a.c. load presented to the detector, which can occur with the operation of a conventional volume control. It also allows this load to be increased in value, thus improving threshold sensitivity.

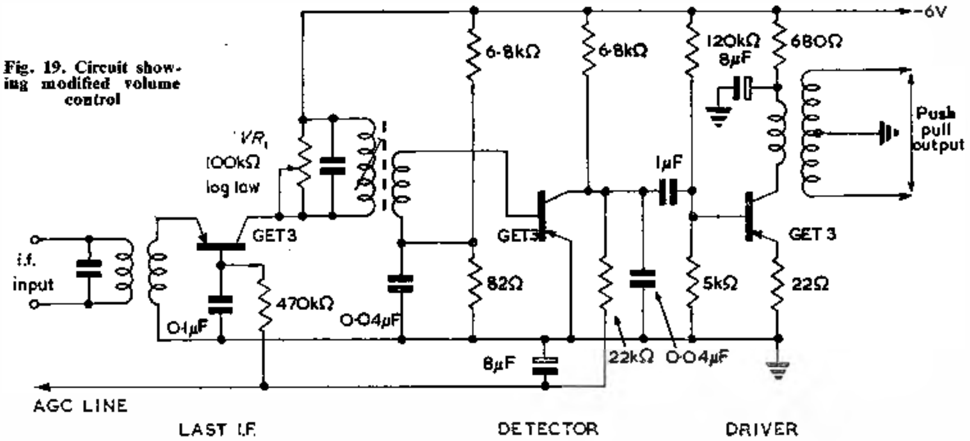

Fig. 19. Circuit showing modified volume control

Receiver Performance and Economics

Electrical Performance

It is firstly of interest to make a brief comparison between the performance requirements for portable and car receivers. The main differences lie in the facts that the car receiver is usually required to have better sensitivity and selectivity, better a.g.c., and larger power output than the portable. In addition, as it is required to operate satisfactorily over a larger range of ambient temperatures (at least 0°C to 45°C), the car receiver supplies a more rigorous problem as regards technical requirements.

Table 2 has been compiled to compare the relative performance figures of several transistor battery receivers, a transistor car receiver and a typical thermionic valve portable receiver, and indicates the type of electrical characteristics achieved. The following explanatory notes may be useful.

The sensitivity figures are quoted for a 1Mc/s input signal modulated 30 per cent at 400c/s, and refer to 50mW output level (except for 'pocket' receiver No. 4 and the car receiver where they refer to maximum output).

The a.g.c. range is the approximate change in input signal level for 6dB change in output, and is measured over a range of input levels prior to that at which the second detector begins to overload.

The mean current consumption refers to average broadcast reception conditions with the volume control adjusted to give maximum undistorted output.

The first three transistor receivers (Nos. 1-3) are small battery portables, the fourth (No. 4) is a 'pocket' type receiver.

Transistor portables Nos. 1 and 2 are experimental receivers using British transistors only, the other transistor receivers are American.

From Table 2 it will be seen that if a given transistor receiver excels in one characteristic e.g. selectivity, it usually fails in another e.g. bandwidth. This trend is due to the fact that it is very difficult to achieve all the desirable electrical characteristics simultaneously with an all-transistor receiver which is reasonably economical to produce.

Economic Considerations

In order to assess the most important economic factors involved, block diagrams of a transistor car radio and battery portable receiver are shown in Figs. 20(a) and 20(b) respectively. In these diagrams the figures in brackets refer to the number of transistors employed in each stage. Although these figures do not correspond to any particular receiver design instanced in Table 2, they are considered to represent the most recent design trends using the best transistors currently available. On the above assumptions the electrical performance obtained in the case of the portable receiver should be reasonably comparable to that of a four-valve battery receiver.

TABLE 2.

Characteristic	Transistor Portable 1	Transistor Portable 2	Transistor Portable 3	Transistor 'Pocket' Portable 4	Valve Portable 5	Transistor Car Radio 6
No of valves or transistors	10 transistors	7 transistors	7 transistors 1 crystal diode	4 transistors 1 crystal diode	4 valves	9 transistors
I.F.	470kc/s	470kc/s	455kc/s	262kc/s	470kc/s	455kc/s
Sensitivity	600μV/m	1mV/m	400μV/m	>1mV/m	500μV/m	4μV at aerial
Image rejection at 1Mc/s	31dB	42dB	43dB	35dB	35dB	57dB
Selectivity (attenuation at 10kc/s) off tune	20dB	20dB	37dB	8dB	30dB	40dB
A.G.C. range ..	34dB	20dB	34dB	32dB	48dB	63dB
Electrical response at −3dB points	70c/s to 4kc/s	300c/s to 2·8kc/s	200c/s to 1·2kc/s	300c/s to 5kc/s	50c/s to 2kc/s	150c/s to 2kc/s
Maximum power output	425mW	140mW	75mW	35mW	200mW	2W
Operating voltage ..	15	6	6	22½	90 (H.T. only)	6
Mean current consumption	22mA	14mA	22mA	4mA	13·3mA (H.T. only)	300mA
Estimated useful battery life	150 hours (layer type battery)	600 hours (U.2 cells)	500 hours* (U.2 cells)	30 hours (special midget battery)	L.T. 25 hours H.T. 60 hours	—

* Manufacturer's claim.

Since the prices of transistors to receiver manufacturers are appreciably higher than valves, it can be said that the initial cost of the transistor version is likely to be appreciably higher. This is based on the fact that the total complement of ancillary components is not likely to be very different in the two cases, particularly if a transistor receiver incorporating two wave bands is being considered. It is however possible that the provision of special components, e.g. centre tapped speakers, may ease the situation a little in favour of the transistor version.

On the other hand the transistor portable receiver will show an appreciable economy in operating costs, of the order of five or ten to one, due to its greatly reduced battery consumption; and the battery itself is much cheaper to replace. Although at the moment sufficient data has not really been obtained, it is likely that improved reliability would also result.

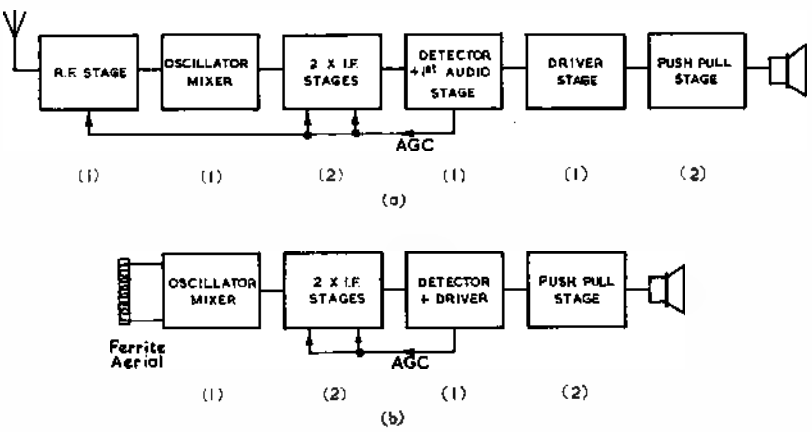

Fig. 20. Typical transistor car and battery portable receivers

In the case of car receivers the elimination of vibrator supplies and probably reduced installation costs should prove advantageous. The latter economy results principally from the fact that a much more compact receiver can be constructed using transistors, particularly if the speaker is treated as a separate item and mounted outside the main receiver.

The adverse circuit economy of transistor receivers tends to be reflected in the fact that in general their electrical performance is usually a little inferior to that of their thermionic valve counterparts, as shown by Table 2. Should their performance become strictly comparable in all respects, it would be found that the initial cost factor was tilted even more in an unfavourable direction. However, it should be remembered that the above arguments only refer to present conditions; for example should the initial cost of transistors be reduced appreciably the economic emphasis could be quite different.

Future Trends

Some of the economic difficulties outlined in the previous section have led to alternative solutions being adopted in certain commercial applications. One such has been the investigation of 'hybrid' receivers containing both transistors and thermionic valves[17]. In this way the better high frequency performance of valves is retained, while at the same time the advantages of the use of transistors in the audio stages are still utilized. Much work has been performed recently in America on the production of a valve having good amplification characteristics when operated from an h.t. supply of only 12V. It should, therefore, soon be possible to use such a combination of valves and transistors in car radio applications without the need for a vibrator unit, and it appears quite possible that in the future such a receiver will be preferred to either the all-valve or all-transistor receiver.

If the use of 12V valves is not considered desirable, a hybrid car receiver can still be constructed without the need for a vibrator by using a transistor d.c. convertor. The design of such a unit has been discussed at some length in a paper by Light and Hooker[15], who show that at low power levels (up to a few watts) such circuits are likely to have the technical advantages of higher working efficiency, long life and good reliability, at the expense (at the present time) of higher initial cost.

In the near future it is likely that improved transistors will be available in respect of both frequency performance and power handling capability. Experimental transistors of the diffused base type are already making their appearance on the American market. Such devices are capable of amplifying in the 100Mc/s region but are likely to be relatively expensive to begin with. They do however open up the technical possibility of an all-transistor broadcast receiver operating in the v.h.f., f.m. band.

Considering the i.f. requirements of such a receiver, it is usual to employ a frequency of about 10·7Mc/s. Assuming that the gain-frequency relationship of a diffused base transistor follows the same type of curve as shown in Fig. 9 it appears that a gain of about 24dB per stage might be possible at frequencies of the order of 10Mc/s. This estimate is based on published data on such transistors which quote α cut-off frequencies of about 200Mc/s, and a maximum available power gain of 10dB minimum at 100Mc/s. Some constructional details of such transistors have already been published[16].

The above gain may be compared with that from a suitable battery valve at the same frequency, which would probably be of the order of 30dB.

In such v.h.f receivers it is usual to employ an r.f. stage, principally to reduce oscillator radiation from the front end. In this case the valve circuit would have a greater advantage with something like 26dB of gain against 10dB in the transistor case.

Transistors using similar techniques to those just described are at present only in the experimental stage in this country. It hence appears that practical transistor v.h.f. broadcast receivers lie a few years into the future, particularly as the economic disadvantages of such receivers would probably be found more severe than in the cases previously discussed.

Other transistor types being investigated are the pnip and drift transistors. Both of these are likely to represent advances in two directions over the existing alloy junction types. These consist firstly in higher α cut-off frequencies, and secondly in reduced collector-base capacitances. It may well be that the gain performance of such types in a.m. receivers will not show a significant improvement over the best alloy junction transistors, but the reduction in capacitance should aid performance in other directions, particularly with varying operating conditions.

The variation of collector capacitance with voltage for these two types is much smaller than in the case of conventional alloy transistors provided they are operated above a certain lower voltage limit (usually in the region of 9V).

Drift type transistors having α cut-off frequencies of about 40Mc/s are now available on the American market. The advent of such devices means that the function of an oscillating mixer for the short wave broadcast bands (assumed to cover approximately 4Mc/s to 22Mc/s) should also soon become possible. In such a case the type of aerial used in a short wave battery receiver, and the matching between it and the first transistor stage, will probably need careful attention.

On the power side, devices are just becoming available on the British market which are capable of delivering powers of the order of 10W or more from a push-pull pair operating from a 12V supply, though such transistors have been available in America for some time. The diffused base type of construction might open up the possibility of a high power, high frequency transistor. However, such a device, if practical, must again lie well into the future.

In addition to these possible improvements in semiconductor devices, there are also likely to be some advances in the components field. Such components as miniature tuning capacitors, shallow speakers and small i.f. transformer assemblies are already beginning to make their appearance. Another field where considerable work is being undertaken is that of batteries. Small batteries having good storage capacity can be provided by mercury cells. They are, however, relatively expensive compared with the conventional Leclanché cell, and are likely therefore to have only a limited use in commercial receivers.

At the present time there is a fair amount of evidence to indicate that circuit techniques are keeping well abreast of device development. It therefore appears that the time elapsing before transistor receivers are introduced on the commercial market on a large scale will depend on the time interval before devices become available in large numbers at appreciably lower prices. In the meanwhile it is possible that special types of receiver, aimed to reduce to a minimum the economic disadvantages previously mentioned, may make their appearance. A particularly attractive receiver in this category could be a combined home radio and car receiver, which could operate either from the car battery, or from an ordinary dry battery when used outside the car or in the home.

Appendix

Thermal Stability Considerations in Common Emitter Amplifiers

List of Symbols Used

α_0	=	Low frequency value of α.
F	=	Feedback factor.
I_{cbo}	=	Collector cut-off current (with emitter open circuited).
I_{ceo}	=	Collector cut-off current (with base open circuited).
r_e	=	Transistor emitter resistance.
R_E	=	External resistance in emitter circuit.
r_b	=	Transistor base resistance.
R_b	=	External resistance in base circuit.
T_0	=	Arbitrary reference temperature.
T	=	Temperature in excess of T_0.
ΔT_j	=	Junction temperature rise due to reverse collector current.
V_c	=	Collector voltage.
Θ	=	Thermal resistance of transistor.

For a transistor connected in the common base arrangement, the rise in junction temperature for a rise of $T°$ in ambient is given by:

$$\Delta(T_j)_b = \Theta \,.\, V_c \,.\, [I_{cbo}]_{T_0 + T} \quad \ldots\ldots\ldots (1)$$

In the common emitter connexion, the maximum value of the equivalent junction temperature rise will be greater and is given by:

$$\Delta(T_j)_e = \Theta.V_c.\,[I_{ceo}]_{T_0 + T}$$

$$\text{i.e.} \quad \Delta(T_j)_e = \Theta.V_c.\frac{[I_{cbo}]_{T_0 + T}}{1 - \alpha_0} \quad \ldots\ldots\ldots (2)$$

assuming α_0 remains constant.

The variation of I_{cbo} with temperature is given by:

$$[I_{cbo}]_{T_0 + T} = [I_{cbo}]_{T_0} \,.\, e^{BT} \quad \ldots\ldots\ldots (3)$$

where B = a constant.

For thermal stability:

$$\frac{d\,[I_{cbo}]_{T_0 + T}}{dT} \leqslant 1/\Theta V_c$$

$$\text{i.e.} \quad B\,[I_{cbo}]_{T_0}\, e^{BT} \leqslant 1/\Theta V_c$$

$$\text{or} \quad [I_{cbo}]_{T_0 + T} \leqslant 1/\Theta V_c B \quad \ldots\ldots\ldots (4)$$

Comparing equations (1) and (4), it is seen that:

$$\Delta(T_j)_{max} \leqslant 1/B \quad \ldots\ldots\ldots (5)$$

For germanium transistors B has a mean value of about 0·07°C^{-1}, and hence:

$$\Delta(T_j)_{max} \leqslant 1/0{\cdot}07 \leqslant 14°\text{C} \quad \ldots\ldots\ldots (6)$$

Allowing a safety factor of 3:1, it is usually advisable to ensure that:

$$\Delta(T_j)_{max} \leqslant 4{\cdot}5°\text{C} \quad \ldots\ldots\ldots (7)$$

Inserting the latter value in equation (2):

$$4{\cdot}5°\text{C} \geqslant \Theta.V_c.\frac{[I_{cbo}]_{T_0 + T}}{1 - \alpha_0} \quad \ldots\ldots\ldots (8)$$

If the common emitter amplifier has d.c. negative feedback applied, equation (8) becomes:

$$4{\cdot}5 \geqslant \Theta.V_c.\frac{[I_{cbo}]_{T_0 + T}}{1 - \alpha_0\,(1 - F)} \quad \ldots\ldots\ldots (9)$$

where F = the feedback factor (which is less than unity).

If the feedback is produced by adding external emitter resistance R_e, F is given to a good approximation by:

$$F \simeq \frac{R_e + r_e}{R_e + r_e + R_b + r_b} \quad \ldots\ldots\ldots (10)$$

[For equation (10) to be strictly true, the base bias voltage should remain constant.]

As an example suppose it is required to operate a common emitter stage up to a maximum temperature of 55°C.

If it is assumed that:

$(I_{cbo})_{25°C}$ = 12μA (a fairly high value), then, from equation (3) with the value of B given:

$(I_{cbo})_{55°C}$ = 100μA.

Let: Θ = 0·2°C/mW
V_c = 10V
α_0 = 0·97.

Then from equation (9), $F \simeq 0{\cdot}016$.

If R_b constitutes the source resistance and is of value 2kΩ, then from equation (10), if r_e = 12Ω, and r_b = 100Ω (typical values for i_e = 2mA):

$$R_e = 21{\cdot}2\Omega.$$

Hence a resistance of 22Ω should be added to the emitter circuit.

It will be found that should the collector voltage be increased, then, all else being equal, the amount of negative feedback required will increase rapidly i.e. a considerably larger value of R_e will be necessary. In the above example, if V_c was 20V, the value of R_e required would be 120Ω.

References

11. Everden, W. A. Ferrite Rod Aerials. *Wireless World.* *60*, 440 (Sept. 1954).
12. Giacoletto, L. J. A Study of P-N-P Alloy Junction Transistors from D.C. through Medium Frequencies. *RCA Rev.* *15*, 506 (1954).
13. Freedman, L. A. Design Considerations in the First Stage of Transistor Receivers. R.C.A. Report. No. LB 1032.
14. British patent application No. 36574/56.
15. Light, L. H., Hooker, P. M. Transistor D.C. Converters. *Proc. Instn Elect. Engrs.* *102* No. 6 (Nov. 1955).
16. Lee, C. A. A High Frequency Diffused Base Germanium Transistor. *B.S.T.J.* *35*. 23 (1956).
17. Funkschau. Die Erste Reisesuper mit Transistoren. No. 5. 172 (March, 1956).

The Magnetic Tape Store for Pegasus

By T. G. H. Braunholtz* and D. Hogg*

The design of the magnetic tape store attached to Pegasus[1] was started two years ago, and the first model has now been in operation for about six months. A general description of the tape store is given first, followed by a brief description of some of the principal anticipated applications. This is followed by more detailed description and discussion, first of the engineering design and then of system design.

THE general arrangement of the tape store is quite standard and is shown in Fig. 1.

Information to be transferred from the computing store of Pegasus[1] is first of all deposited in a buffer store, eight words at a time, by means of a special order. This order is called the 'interchange order' and interchanges the contents of any block in the computing store with any block in the buffer store. From the buffer store the information is written on tape by the tape control, and Pegasus can, in the meantime, continue with computation.

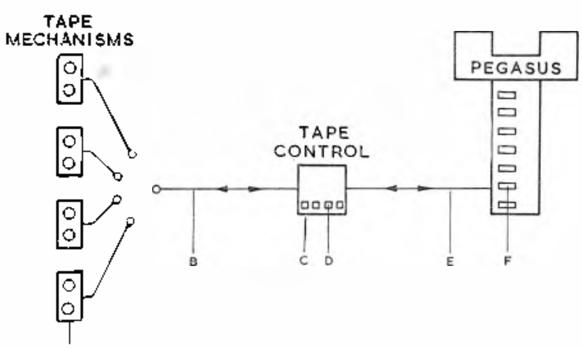

Fig. 1. Schematic diagram of Pegasus with magnetic tape

- A *Four tape mechanisms can be attached. A switch on each mechanism selects 16 or 32 word working*
- B *Each tape order can select any one of the mechanisms*
- C *Each tape order is carried out by the tape control, autonomously, only one tape order at a time can be carried out by the tape control*
- D *Four blocks of eight words in the buffer store*
- E *Interchange the contents of any block in the buffer store with any block in the computing store*
- F *Computing store of 55 words*

The buffer store contains four blocks of eight words, making 32 words in all. When reading and writing on tape the buffer can either be used in one 32-word section or two 16-word sections.

Four tape mechanisms may be connected to a tape control, and two tape controls with four mechanisms each could be connected to Pegasus.

A tape control can carry out four orders. The specification of an order is contained in one word, which states:

(1) Which tape order is to be carried out (search, read, write, rewind).

(2) Which mechanism is to be used.

(3) The address of the section of tape referred to.

(4) Which half of the buffer is to be used.

To set the tape control into action this word is sent to register 20 where it is stored for use by the tape control. The act of sending a word to register 20 starts the tape control.

The four control orders are:

SEARCH: the tape in the selected mechanism is moved (forward or backward) into position for reading or writing on the specified section.

READ: read the specified section from the selected mechanism into the buffer store. The tape control will search for the correct section before carrying out the read order.

WRITE: write on to the specified section of the selected mechanism from the buffer store. Again, the tape control will search for the correct section before carrying out the write order.

REWIND: set the selected mechanism into Rewind. This will take a few milliseconds, after which the tape control becomes free to carry out other tape orders while the mechanism is rewinding.

If an error in carrying out a tape order is detected by the checking circuits an indicator is set. The setting of this may be sensed by the programme through register 21, which will be all ones if an error was detected and will be zero otherwise. The order may then be repeated. Whenever an error is detected a buzzer on the tape control sounds, in addition to register 21 being set.

The interchange order, the tape control word sent to register 20, and the register 21 staticizor are the only controls over the tape store the computer requires.

The reader will have gathered that tapes are addressed for 16 or 32-word sections. Each tape, of course, is either all 16-word sections or all 32-word sections, and the tape control carries out a 16 or 32-word transfer depending on a switch setting on the selected mechanism, which must be set by the operator to correspond to the tape in the mechanism.

There is a special feature for checking that writing has been carried out correctly: each mechanism has two heads, and writing is carried out by the forward head and is checked by reading back with the trailing head as part of the write order. The two heads are at present mounted one inch apart.

The following is the specification for the tape store:

(1) Tape is ½in wide, and can be supplied in lengths of multiples of 600ft (up to 3 600ft) with the tape tested and addressed.

(2) The tape speed is 75in/sec.

(3) The time to stop or start between rest and full speed is less than 7msec for either stopping or starting.

(4) Writing (and reading) speed: 9 250 characters per second.

(5) Transfer times per section:

32 words: 53msec

16 words: 41msec

assuming that the tape is in position for the quickest possible transfer.

* *Ferranti Ltd.*

(6) Rewind time for a 3 600ft tape: 6min (average speed of 120in/sec).

(7) *Property (3 600ft tape)*	*16-word section*	*32-word section*
Number of sections per tape	13 000 or more	10 000 or more
Number of words per tape	208 000 or more	320 000 or more
Length of section	2·84in	3·75in
Length of address	0·23in	0·23in
Length of stop-start space	0·7in	0·7in
Length of head spacing gap	1·0in	1·0in
Length of information	0·91in	1·82in

(8) Search time per section:

32 word: 53msec

16 words: 41msec.

This is the time it takes to run over one section at full speed and is also the time per section for continuous reading or writing. It is actually the same time as in (5) above, which is the time from rest to completion of the order.

The Application of the Tape Store

The designer of a computer for scientific and commercial use, and for use by organizations with very different amounts of money available, is faced with a difficult choice in deciding how expensive a tape system he should plan. The principal variables at his disposal are the size of sections of information on tape (which may be variable) and the degree of autonomy to be given to the tape control.

In scientific and engineering computation Pegasus will use magnetic tape for two principal purposes. First there are the problems using more data than can be stored on the drum. Second it will be found very valuable to reserve one mechanism for a tape containing the library of subroutines, and also perhaps individual's private routines and intermediate results; all this saving a great deal of programme assembly and paper tape input time, and allowing more efficient use of machine time.

These uses of the tape store account for some aspects of the design chosen. Both will involve dividing the tape up into parts for storing different sets of data, and so on, as distinct from using a tape as one long file. If the tape is not supplied marked into sections in some way, but is a continuous medium for recording, there is a problem in overwriting one portion of the tape while leaving adjacent portions unaffected: for instance variations in tape speed cause difficulty. Therefore it was thought desirable to mark out sections, either with or without addresses. If addresses are not recorded on tape the programmer will either have to keep count of where he is on each tape in use, or, when he wishes to go to a particular section, he will have to run back to the beginning of the tape and start counting from the beginning. This situation, together with some considerations on tape life and quality, determined the use of addressed tapes.

The applications of magnetic tape to commercial problems also divides into categories, which will be taken as sorting, file processing and programme aid.

The programme aid category covers the second use mentioned for scientific applications.

Sorting usually involves four tape mechanisms, A, B, C, D. The basic cycle as it affects magnetic tape is as follows. Either mechanism A or B is selected, more or less randomly, and the next section forward on the tape is read into the computer. One section of information is then written on the next section forward on tape C, D being unused, and the cycle is repeated. If two tape transfers can be carried out simultaneously (i.e. if there are two tape controls connected to the computer) then the write on to C takes place concurrently with the read from A or B. But more quick access storage is then required in the computer for information in transit between A or B and C, in order that it may be examined by the computer, while the tape orders are in progress, to determine which of A or B is to be used next. In either case the transfer time to and from tape is probably the limiting factor on the speed of the sorting operation. On the usual methods for estimating cost, sorting on the computer comes out more expensive than using a punched card sorter, but nevertheless its greater real speed and completely automatic functioning make it an important activity of commercial computers.

File processing involves running a master file (for instance of insurance policies, or stock control information in a factory or warehouse) through a mechanism, and probably writing out an amended copy of it, while taking in or putting out a number of other smaller files of information from the computer. The proportion of items in the master file which are affected will vary between one in one and one in twenty or less. The number of input and output chanels required varies a great deal, and may be either punched cards or paper tape, or magnetic tape for subsequent printing or conversion to cards. If the amount of information to the output channels is small and there are many of them it may be preferable to put them all on one tape first and separate them after the main file has been processed, when more tape mechanisms will be free. If there are any alterations to the master file a completely new copy of the main file will usually be produced, for reason of security of information, which means that all the unaffected records will be copied too. Main files will often occupy a number of tape spools, and for very large files hundreds, and in this case rewinding the tape and changing the spool becomes a major time consuming operation. This time can be saved by using two mechanisms alternately for the successive tapes of a file, and thus rewinding and loading one mechanism while the other is in use.

Tape Mechanisms

It is considered that ½in tape is a useful minimum width for computer use since 8 channels may readily be accommodated. A separate address track is used, and since a clock track is necessary for identification of '1's and '0's on the information tracks ½in tape can accommodate 6 information channels. ¼in tape would accommodate only 2 information tracks assuming 4 tracks in all. It is not true to say, however, that ½in tape carries 3 times the information of ¼in tape since the start-stop gap between blocks on tape is proportionally greater on ½in tape than on ¼in tape for a given size of information block. Given then that ½in tape be used, as specified, a minimum length of tape of 2 400ft is necessary for a large number of commercial applications. Moreover a high tape speed is necessary on a tape system for use with a computer so as to obtain the highest frequency of signal from tape, consistent with a packing density of digits on tape which does not introduce unduly stringent limitations on tape guiding, skew of tape across recording heads, etc., and unreliability due to digits being small compared with blemishes in the tape surface. Finally a rapid start-stop time is necessary so as to reduce to a minimum the tape wasted between blocks of information on tape. For both programming and engineering reasons fixed block lengths are used on Pegasus and in the case of 16-word blocks these occupy 0·91in of tape.

In view of the time that would have been involved in developing a suitable tape unit it was decided to purchase tape units made by ElectroData Division of Burroughs Corporation of Pasadena, California. These tape units were chosen as being considered the best available. As supplied the units have a tape speed of 75in/sec ± 1 per cent and a start-stop time of 7msec. Standard $10\frac{1}{2}$in NARTB spools are used which in fact accommodate up to 3 600ft of the thin polyester base tape.

Reference to the photograph, Fig. 2, will give a general idea of the unit. The tape passes over two contra-rotating capstans from which it is drawn into two tape vacuum

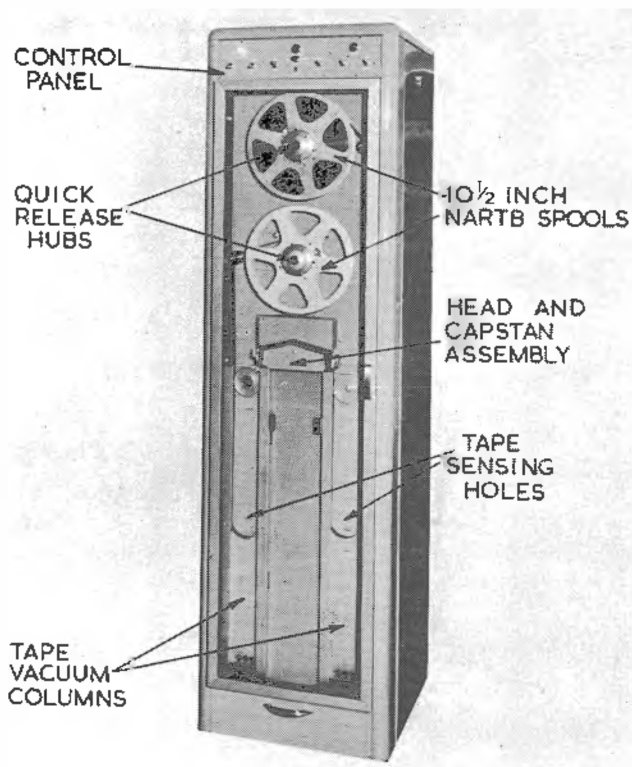

Fig. 2. Electrodata magnetic tape transport unit

columns. With the unit at rest the tape in each column takes up a position between two rows of holes. Each row of holes communicates with a pressure chamber which operate magnetic blowout microswitches. The tape is moved into the forward or reverse direction by selecting a solenoid which presses the tape against the appropriate capstan by means of a pinch roller. As the tape moves into one vacuum column it passes the bottom row of holes. This operates the microswitch which directly switches the tape spooling motor armature current so that the tape spool takes up the surplus tape from the column. Similarly as tape is drawn from the other column so it passes the top row of holes in that column. This operates the microswitch so that the other tape spooling motor delivers tape to the vacuum column. In this manner loops of tape are held ready to permit the rapid start-stop time required as well as to ensure a continous drive of tape as required. A high-speed rewind operation is provided in which neither pinch roller is actuated. The bottom spooling motor drives continuously at its maximum speed with the tape in the left-hand column running over the pulley wheel at the top. The right-hand column maintains the flow of tape from the top spool as required, by the normal servo action. N.B., the left-hand column with the bottom spooling motor and the right-hand column with the top spooling motor form two independent servo systems.

A further interesting feature of the tape unit is that single-edge guiding of the tape is used. This results in reduced modulation of tape signals as compared with the more common double flange guiding.

On each tape unit the tape chamber itself, i.e., that part of the unit containing the tape spools, capstans, tape vacuum columns, etc., is connected in a closed circuit air conditioning system which maintains the air at a temperature between 65° and 85°F, and 40 to 60 per cent relative humidity. This is necessary to ensure reliable operation of the tape irrespective of conditions in the computer room. Excessive temperature rise may lead to static electricity causing the tape to stick in the vacuum columns and moist conditions lead to oxide sticking upon the heads and tape guides. The pressure in the tape chamber is slightly greater than atmospheric, preventing the entry of dust, and an electrostatic filter is built into the air conditioning system to maintain a high order of cleanliness in the tape chamber. The air conditioning equipment is mounted on the rear door of each unit. Each tape unit is on wheels and immediate access may be had to all working parts.

Magnetic Tape

When work was started on this tape store the only tapes available from European suppliers were standard audio quality. Such tapes all contain, depending upon the makes, greater or less quantities of impurities in the oxide, blemishes in the base and splices per given length. All these factors prevent the correct recording of digits on tape and so lead to unsatisfactory and unreliable operation of a tape store. It is therefore necessary to arrange that no recording takes place on bad regions of the tape. This is done by setting aside one track of the available eight as a block marker track (in fact addresses are used on Pegasus, as explained previously) in which the block markers or addresses are so positioned as to avoid the bad regions of tape. Having decided to do this it is better to choose a tape for its mechanical properties, specifically, mechanical strength and bonding of oxide to base, than for comparative freedom from blemishes. If bonding is poor the recording heads will will get dirty quickly and loose oxide will accumulate in the tape unit and cause reading errors. It will have been gathered that the magnetic properties of the tape are comparatively unimportant. It so happens that the tape finally chosen is not only guaranteed splice free in lengths of 3 600ft, mechanically strong since it has a polyester base, but is also among the best from the point of view of purity of oxide and absence of blemishes. Blemish free tape is still not available in Europe.

Tapes are checked and addressed on a special piece of equipment—not on the computer. In this way it is possible to check and address a tape in two passes, by using a series of heads as follows. The first head records throughout all tracks at approximately double the working frequency, the second head is a reading head and checks for drop-outs, i.e., loss of signal amplitude. The third head erases 7 tracks and writes addresses at the correct spacing (for 16 or 32-word operation) on the address track. The fourth head checks that every address is correctly written. Evidently the distance between the second and third head is dependent upon the required block length on tape. Basically the principle of operation is that no address is written until a portion of error-free tape of the required length has passed the second head. It is found that about 93 to 95 per cent of the tape used is acceptable to the tape addresser. Each address is approximately 0·23in in length.

Layout of Information on Tape

In addition to a general requirement for a very high order of reliability of the tape store it was considered essential to be certain that every tape write order had been carried out successfully before proceeding with the next tape order. If a single read-write head were used this would involve reversing over the block of information and then reading it off in the forward direction and checking that it was correctly written. This would result in every write order taking about three times as long as a read order. The alternative scheme, which was adopted, is to mount read and write heads side by side, as close together as possible. There is a distance of 1in $\pm 1/64$in between the lines of the gaps of the two heads. A simplification in switching is also then ensured since one head only is ever connected to the reading circuits and the other to the writing circuits.

A tape write order causes the selected tape unit to search for the specified address. When this is found in the forward direction of tape as read by the reading head the writing head starts to write the block of information on the tape. There is, therefore, a gap of 1in between the address on tape and the block of information to which it refers. At 75in/sec tape speed this represents an increase of 13·3msec per transfer but this is felt to be more than justified for the reasons given above. The tape continues forward and the block is read immediately it reaches the reading head. A check is then made on the written information. It is generally true to say that once a block of information has been written on tape and read successfully then it will always subsequently be read correctly. The general arrangement of the heads and layout of information on tape is shown in Fig. 3.

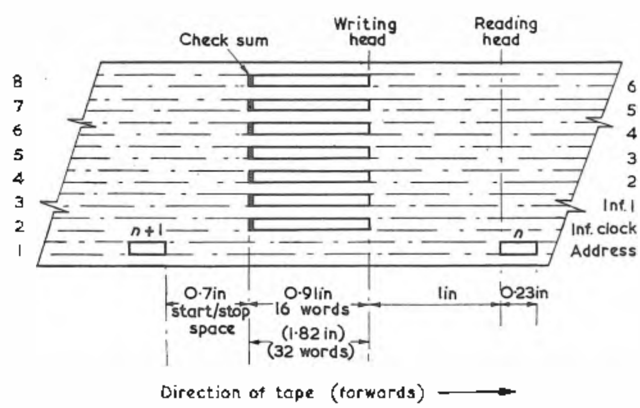

Fig. 3. Layout of information on magnetic tape

It is necessary to specify the head spacing to 1in $\pm 1/64$in so that tapes written on one unit may be over-written on any other unit. In point of fact the writing head does not start writing immediately the required address is found but switches to a d.c. erase condition for a period of time sufficient to ensure that any digits of a block written on a tape unit whose head spacing is less, within the limits specified, than that on the unit currently used are erased. In this way interchangeability of tapes from unit to unit is readily ensured.

Information is written on the tape at the rate of one 6-bit character every 108μsec, i.e., at a frequency of 9·225kc/s. A clock pulse is written in the clock channel with every character. The clock track is not a parity track. At a tape speed of 75in/sec a packing density of 9 225/75 = 123 bits/in is obtained. Thus each digit on tape has a cell length of about 0·008in. It would appear from our experience that packing densities of 250 to 300 bits/in are quite feasible.

Reading and Writing Heads

Reading and writing heads are identical and interchangeable. Each head has 8 tracks 0·045in in width separated from the adjacent tracks by 0·015in. The individual heads are made from mumetal laminations in a simple ring form. There are 200 turns of 46 s.w.g. wire per limb. The gap width is ·0005in and the inductance of each track is approximately 3·0mH. A direct current of 15mA will completely saturate all medium coercivity tapes. The method of recording is the return system[8]. It is felt that the recording techniques are sufficiently well established to warrant no further description in this article. The read signal from the reading head is approximately 15mV peak-to-peak. Such a large signal from a low impedance head means that head amplifiers on each tape unit are unnecessary provided earth return loops are avoided and adequate screening provided.

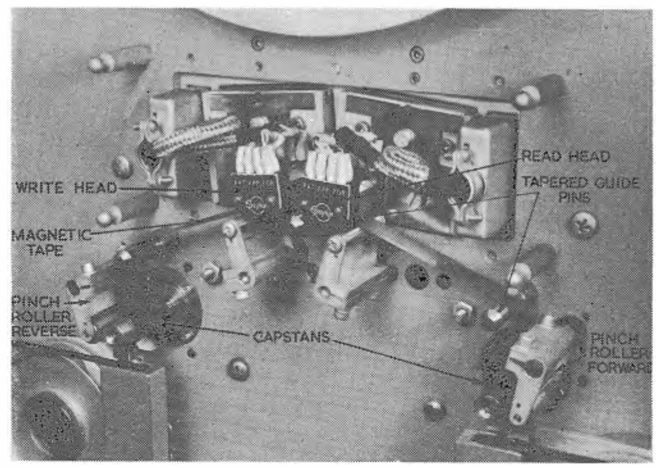

Fig. 4. Head and capstan assembly

Each head is mounted on an individual head mounting plate and may be adjusted so that the tracks on the head register with the channels on a specially prepared master tape, so that the head face is parallel to the tape and so that the line of the head gaps is perpendicular to the direction of movement of the tape. In addition each head may be moved up and down in a vertical direction so that the amount of wrap-around may be adjusted. The use of pressure pads is not, in our view, practicable in a high speed tape unit. Our experience is that the wrap-around should be the least possible consistent with reliable operation. Excessive wrap-around leads to oxide sticking to the heads. With the adjustments described it is possible to adjust, and to lock into position, the two heads on each tape unit to exactly the same relative position with respect to the tape by means of a double beam oscilloscope and the specially prepared master tape referred to above. It is evidently necessary for interchangeability of tapes that all heads should be in the same relative position, as well as that the heads on each unit should be 1in $\pm 1/64$in apart, as described previously. It will be appreciated that with fine screw adjustments of the heads and a tape recorded at 20kc/s, which means at 75in/sec that one cycle corresponds to about 0·0038in, head adjustment is both easy and precise. The photograph, Fig. 4, shows the disposition of the heads, tape and tape guides.

Control Desk, Reading and Writing Circuits

There is one set only of reading and writing circuits for all four tape units. These circuits are built on Pegasus packages and are mounted in the control desk attached to the side of the cabinet of equipment which controls all tape operations. Relay switching is employed to connect the heads of the selected unit to the reading and writing circuits. The drive signal to the tape units is selected electronically and since the relays operate in about half the time taken for the tape to get up to full speed the heads on the selected tape unit are connected to the reading and writing circuits before signals are reached. In this way relay switching is not a limiting factor on speed of changeover between tape units.

Separate 25-way cables are used for reading and writing head leads. The read cable carries only 8 pairs of leads to the 8 amplifiers. The write lead carries, in addition to writing currents, all control signals to and from the tape unit, i.e., drive forward, and reverse, rewind and rewind hold, 16 or 32-word operation, etc. Associated with each tape unit there is mounted on the control desk a 'write inhibit' switch so that a master file may be safely placed on a given tape unit with the 'write inhibit' switch on in the certain knowledge that no errors in programming or faults in the equipment can cause erasure of important records.

Length of Sections on Tape and Size of the Buffer Store

Variable length sections on tape were excluded by the decision to use addressed sections. It is desirable to use long sections on tape both in order to pack more information on a tape and to obtain higher effective tape speeds. On the other hand the buffer store to hold the long sections is expensive, since in Pegasus it is made of single-word nickel delay lines. A section length of 16 words is desirable on the ground that it contains 92 alphanumeric (6-bit) characters, and therefore one 80 column punched card comfortably. (A card to tape, tape to card, and tape to printer convertor unit has been designed and built as part of the Pegasus System, and will be described in a later article.) Also, for uses in which the tape speed is adequate, 16-word sections have an advantage in being often simpler to handle by programme than 32-word sections, since twice the number of such sections can be kept in the quick access store at one time as of 32-word sections. But with 16-word tape only 40 per cent of the tape length is actually used to hold information. Therefore an alternative longer section was introduced. A length of 32 words was chosen, with about 60 per cent tape utilization. A longer section of, say, 64 words would be too large to be contained in the computing store, and would have to be transferred to the Pegasus drum in two halves, wasting about 20msec, which would cancel out the gain in effective tape speed.

The buffer store can be used to hold either one 32-word section or two 16-word sections. When it is used for 16-word sections it is possible to use one half of the buffer for transferring to or from tape and the other half for eight-word transfers to and from the computing store of Pegasus.

Reading and Writing Backwards

In some large commercial problems it would be useful to be able to read and write on tape in both directions. But it would be very expensive to do this, and appeared to be one of the least worthwhile of possible refinements.

Inter-Section Gap

The distance required along the tape for stopping from full speed and reaching full speed again (called the stop-start space), with safety tolerances added, is 0·7in. The problem arises whether or not to make this gap much shorter and expect the computer either to be ready for the next section when it arrives, while running the tape at full speed between sections, or to waste time in stepping back to the beginning of the section. But $6\frac{1}{2}$msec gap time must be allowed in any case, for 32-word transfers, for testing the error register, clearing the buffer store and initiating the next order and during this time the tape moves nearly 0·5in. Thus with a 0·5in stop/start space the question of inter-gap wastage no longer arises, and with 0·7in is not a significant problem. If there were two buffer stores used alternately, however, the possibility of saving the stop-start space would arise again.

Method of Control

The division of the buffer store into blocks of eight words for transfer to and from the computing store is obvious, since transfers between the drum and computing store are in blocks of eight words. The interchange order exchanges the contents of a block of the buffer store with a block of the computing store; an alternative arrangement would be to have two orders, one transferring a block from the computing store to the buffer and the other transferring a block from the buffer to the computing store; the choice was one of convenience of design.

The method of specifying a tape order and initiating it is based on the assumption that each order would contain the address of the section to be read. The alternative would be to have tape orders of the form 'Read the next section forward' and 'Step back one section', and have only the search order specify an address. This method has been adopted on the Ferranti Perseus computer, but requires a little more equipment. With the Pegasus system the information required to specify a tape order will not fit into one computer order, so one Pegasus word is used, and given this control word the form of initiation of a tape order is as simple to use as any.

A method of repeating orders that have been incorrectly carried out alternative to that adopted in Pegasus would be to have, say, up to three repetitions carried out by the tape control automatically. But the Pegasus system leaves complete freedom to the programmer to take what action he wishes. The order to be repeated is still held in the tape control so that the programme does not have to try to unravel which was the last tape order to be initiated. By leaving action following an error at the discretion of the programmer, the way is left open for changes in after-error action, and possibly for large scale record-keeping by the computer on the frequency and nature of tape reading and recording errors.

Reliability and Built-in Checks

It is difficult to develop a rational approach to the problem of how much checking should be built into a computer. What appears to occur in practice is that each design group adopts a different attitude to the matter, and goes ahead with the design of its machine. When the machine is finally installed and tried out in practice an assessment can begin to be made on the degree of checking that must be built into the programmes and into the organization external to the machine.

The less reliable the machine, the greater the trouble and money put into the external checking, but however the balance lies, reliable computers by present standards can be successfully applied to commercial problems. Pegasus has shown a high standard of reliability, and the criterion adopted for the tape store was that this standard should be more than maintained.

We are cautious about any claims of perfect reliability in any machine which is not checked throughout by duplication of components or equivalent devices. A claim that no undetected error has been known to occur is different from a claim that it is known that no undetected error has occurred. Errors undetected (in the sense of not detected until after the results have been removed from the computer) should be so rare that there is always doubt whether it was the computer's fault or not. In the case of Pegasus, as with most commercial systems, some programmed limit and sum check tests on the numerical part of commercial output may be required to achieve such a standard of accuracy.

There is one feature which magnetic tape has in common with some other input/output media, and which is not common to the internal part of a computer: errors in recording or reading off will from time to time occur, and are not to be treated as indicating that the system is in a faulty or marginal state and requires maintenance, nor is any doubt thrown upon the preceding calculations. Therefore far more errors will actually occur in transferring information to and from the tape than internally, and thus far more opportunity for compensating errors passing the checking system. In addition, the storage of tape has a very low usage rate compared with the interior of a computer, and so it is not subjected to continuous testing by use. For these reasons a higher standard of checking is required on tape than, for instance, on internal storage.

The checking arrangements built into the Pegasus tape store will now be described.

(1) In write orders, as explained above, the information is read back by a reading head one inch behind the write head, and check sums are compared. This checks almost all of the reading circuits and writing circuits against each other. In this comparison three check sums are compared: the check sum formed in the logic as the information ran out of the buffer store to be written on tape; the check sum formed while reading the section off tape; and the check sum read off tape. Any difference sets the error staticizor. This comparison of three check-sums makes sure that writing has in fact occurred, so that, for instance, failure of the h.t. supply to the write circuits will be detected at once.

(2) The parity check digit in each word of the computing store is made full use of. It is retained on transfer to the buffer store, from the buffer store to tape, from tape to the buffer store and from the buffer store to the computing store. When being written on tape each word has its parity digit checked, as also when being transferred from the buffer store to the computing store, and a failure stops the machine.

(3) The check-sum mentioned in (1) above is formed as follows: There is a check-sum register, which is empty at the start of a write order. As each 6-bit character is written on tape it is added to the check-sum character with end around carry. (End around carry means that carry off the most significant bit position is added into the least significant bit position.) A check-sum has the advantage over a parity check that two errors in the same sense on one track cause a compensating error on a parity system, but not on a check-sum system. The end around carry of the check-sum is necessary to avoid the most significant channel having only its parity checked. The frequency of errors obtained so far is in the region of one per 3 000 to 15 000 sections. Without having recorded any figures, there is nevertheless good theoretical reason to suppose that errors tend to be in the same sense at any one time. Also, errors which do occur consist usually only of one or a few digits. Thus the probability of a compensating error on the check-sum seems to be pretty remote.

(4) Another form of fault which would not affect the check-sum is the case when, due to distortion of tape or misalignment of heads, digits are read as being in the wrong character, either in front of or behind where they should have been. A check has been put in to prevent this happening. The device detects any signal arriving on

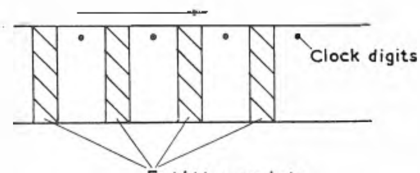

Fig. 5. Detection of faults due to distortion of tape a misalignment of heads

the information channels during a 30μsec gap between characters (Fig. 5).

(5) A count is kept of the number of information clock digits received during reading. If there should be a gap of 0·5msec during which no clock digit arrives before the full number of clock digits have been received, the error staticizor is set. This will detect drop-out on clock digits.

(6) On the address track a parity count is kept of the address itself and a count of the clock digits is kept also. The form of addresses on tape is as shown,

$$1\ \delta\ 1\ \delta\ 1\ \delta\ 1\ \ldots\ 1\ \delta\ 1\ p\ 1$$

where the '1's are clocking digits and the δ the address digits (of which there are 14) and the p is the address parity digit. The parity on addresses would not appear to be so safe as the check-sum on information, but the following facts render the degree of safety great. Several successive drop-outs would include a clock digit, in which case the clock count would detect a fault; an error cannot do any harm unless the error turns the address into the address being searched for; in large scale working the next section forward will almost always be the section referred to, and a compensating error could do no harm.

(7) A buzzer on the tape control will ring for half a second whenever an error is detected. This will keep the operator and maintenance engineer informed of the state of the tape mechanisms; otherwise they would have very little knowledge of when the performance of a tape or mechanism was deteriorating.

Acknowledgments

Our thanks are due to ElectroData, Division of Burroughs Corporation for the ElectroData tape mechanisms and advice on their use; to Epsylon Sound Accessories Ltd, for the manufacture of the special heads used; to the Minnesota Mining & Manufacturing Co. Ltd (whose tape we use), for their assistance and co-operation.

This work would have been impossible without the knowledge and experience gained from the development of the 35mm tape store for the Ferranti Mk.I computer.

Our thanks are also due to Mr. J. R. Travis, Dr. J. M. Bennett, Mr. J. F. Davison, Mr. G. E. Felton and Mr. P. M. Hunt for their invaluable contributions to the project.

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REFERENCES

1. Braunholtz, T. G. H., Emery, G. The Design of the Pegasus Computer. *Electronic Engng.* 29, 358 & 420 (1957).
2. Robinson, A. A., McAulay, F., Banks, A. H., Hogg, D. A Magnetic-Tape Digital-Recording Equipment. *Proc. Instn. Elect. Engrs.* 103B, Suppl. No. 2, 337 (1956).

The Recording of Digital Information on Magnetic Drums

By D. G. N. Hunter*, M.A., and D. S. Ridler*, A.M.I.E.E.

Methods of representing binary digital information on magnetic drums are discussed with regard to their reliability, cost and technical merits.

THE designer of a computer or a computer-like equipment who has decided upon the use of a magnetic drum for a digital store has the choice of several known methods of representing binary digital information on the magnetic surface. These methods differ in reliability, in cost, and in the facilities they offer.

This article is written to assist the designer in making the best choice possible from among the known methods, which are described and compared.

An Outline of Known Recording Methods

PULSE OR DIPOLE METHOD

With this method binary information is written on the magnetic surface of the drum by passing short pulses of current through the recording head[1]. The polarity of a pulse indicates the binary condition, one sense indicating a '1' and the other sense a '0'. The pulse amplitude is such that the recording medium lying in the neighbourhood of the gap at the time of recording is magnetized to saturation one way or the other, irrespective of the previous state of the medium. In this way a train of pulses representing '1's and '0's can be made to print a series of cells on the magnetic surface.

The pulse method has several variations, the first two of which concern the initial state of the track which can be either unmagnetized or else magnetized uniformly in one direction.

(a) *Track Initially Unmagnetized*

In this case the writing current pulses (Fig. 1(a)) produce magnetic dipoles with unmagnetized medium between them (Fig. 1(b)). When the head scans the dipoles during reading, an e.m.f. is induced proportional to the rate of change of flux with distance in the medium (Fig. 1(c)). A binary '1' is indicated by a positive followed by a negative swing, and a '0' is indicated by a negative followed by a positive swing; these e.m.f.'s are equal and opposite in theory.

A similar result can be obtained by initially magnetizing the track to saturation transversely to the direction of motion. However, this appears to offer no advantage over the unmagnetized background and there is no record of its use in practice.

(b) *Track Initially Magnetized*[2]

When the track is initially magnetized in the direction of motion to the state corresponding to '0', the signal read for a '1' is twice as big as in the previous case since the available flux change is doubled in the same distance (Fig. 1(e)). However, the signal for a '0' is very much reduced, or even eliminated, since there is no substantial flux change.

Given the signals produced by reading information which has been recorded by either of the above variations in the initial state of the track, there are alternative ways of treating the signals in order to recover the digital information:—

(1) By examining the recovered signal at the time-instant just before (or just after) the time-instant at which the recording was made.

(2) By differentiating the signal and examining the resulting waveform at the time-instant corresponding to the usual recording time.

Whatever the method of recovery, when the track is initially magnetized the signals must be decoded by measuring them against a finite threshold voltage since '0' corresponds either to the absence of a signal or to only a very small signal of opposite polarity to that for a '1'. The signal is therefore taken to indicate a '0' or a '1', dependent on whether it is above or below the threshold. In contrast, when the track is initially unmagnetized no such threshold is necessary since the waveforms for '1' or '0' are equal but opposite in sign so that only the polarity has to be identified at the instant of examination.

The pulse method therefore has four variants according to the initial state of the track and according to the way in which the signals are recovered. A unique feature of the method of recovery not involving differentiation is the ability to recover the stored information a short time before the recording time-instant is reached. This feature makes it possible to read a stored digit and, if desired, to change it in a single passage of the head over the digit cell, thus avoiding the necessity for external storage in certain applications.

In the pulse method it is possible to change isolated digits on the magnetic surface without affecting adjacent cells, and the method is therefore suitable for use in systems which may require only a part of a track to be modified at one time without the necessity for re-recording the whole track.

TELEGRAPH OR NON-RETURN-TO-ZERO METHOD, AND POULIART'S VARIATION

The principle of the telegraph method is to pass a current through the recording head for the entire digit period (Fig. 1(f)), driving the track into either positive saturation for '1's or negative saturation for '0's. The magnetization pattern has the form shown in Fig. 1(g). In contrast to any other method of recording, the maximum possible fundamental frequency in the signal from the track is only half the digit repetition rate, so it would seem to be possible to record twice as many digits per inch of track. Unfortunately the practical difficulties of dealing with the recovered waveform considerably reduce this advantage.

Referring to Fig. 1(h), it can be seen that a negative-going signal is induced for each transition from '1' to '0', while a positive-going signal is induced for each transition from '0' to '1'. It is therefore convenient to establish two thresholds, one at half the height of a positive-going peak and the other at half the height of a negative-going peak. The signal is then measured at a suitable time-instant and declared to lie in one of three distinct zones:—

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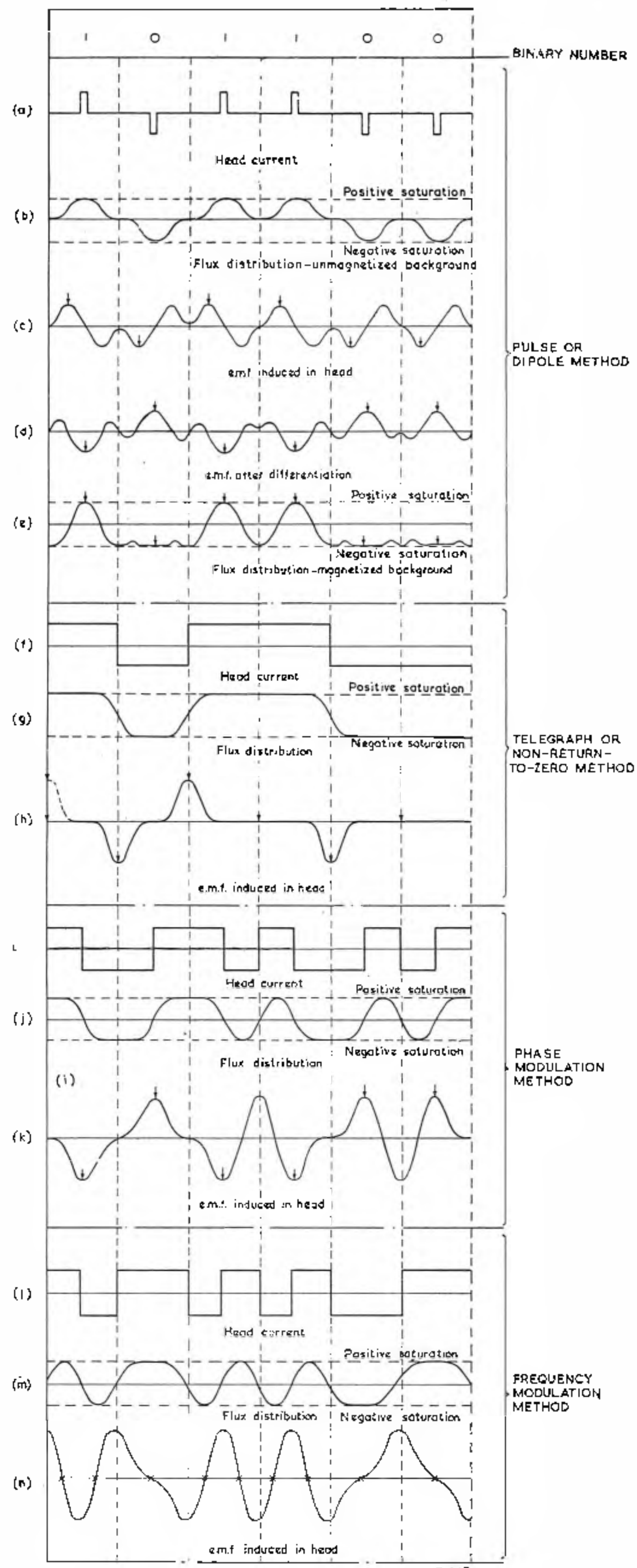

Fig. 1. Comparison of methods of recording on a magnetic drum

(a) Below the lower threshold, indicating a '0'.

(b) Between the two thresholds, indicating no change from the digit read previously.

(c) Above the upper threshold, indicating a '1'.

The first recorded element in a train of elements serves only as a 'start indication' and cannot be used to store information.

A variation of the telegraph method, due to Pouliart, consists of a telegraph system as just described, but preceded by a circuit for converting telegraph signals from one code to another according to the following rules:—

(1) The first element in an arbitrary train of elements is passed through unchanged, i.e. a '1' remains a '1' and a '0' a '0'.

(2) For all succeeding elements, if a '1' is received the output remains at its previous value, but if a '0' is received the output changes its state. For instance, a group of '1's causes the output to remain steady at positive or negative saturation, and a group of '0's causes the output to change its state on every digit.

Thus, as in the original telegraph method, the highest frequency in the recovered signal is again half the digit repetition rate—it is only the interpretation of the patterns which is different. In the Pouliart variation also three states must be recognized in the received signal:

(a) Above the upper threshold, indicating a '0'.

(b) Between the two thresholds, indicating a '1'.

(c) Below the lower threshold, indicating a '0'.

Although the same waveform does not represent the same code in the telegraph method as in the Pouliart variation, the two codes may be discussed together from the point of view of the reliability of the magnetic recording process.

The information in the telegraph method is theoretically recovered at the same time as the recording is due to start, but it is not possible to read and record information in a single passage of the head over the digit cell since there are delays in the recording and reading circuits.

In the telegraph method it is theoretically possible, but probably difficult, to change isolated digits in a track satisfactorily without recording the entire train. Thus this method is rather unsuitable for use in systems requiring the partial modification of the track without a complete re-recording. Pouliart's variation is even less suitable since the change of any one digit demands a series of changes along the remainder of the track.

In the telegraph method it is necessary for the reading circuit to have a knowledge of the previous digit since a pulse may not be induced in the head if the read digit is not the same as the previous one. This difficulty is avoided in the Pouliart variation by transferring the problem to the recording circuit which has to have a knowledge of what was previously recorded. The first recorded element serves again as a start indication.

Phase Modulation Method[3,4]

This method derives its name from the fact that the current waveform in the recording head is in phase for a '1' and out of phase for a '0', with one cycle of a square wave repeating at digit frequency (Fig. 1(i)). At the centre of the digit period the current in the head reverses, the direction of the reversal being characteristic of the state of the cell ('1' or '0'). On the other hand, at the beginning of the cell there may or may not be a reversal, depending on whether the two digits are the same or different.

Bearing in mind that a change in head current during recording results in a peak in the waveform recovered during reading, it is evident that the optimum time-instant for reading is half way across the cell (Fig. 1(k)). This implies that information sent into the recording head is best recovered at a time at least half a digit period later than the time-instant at which recording commenced. Only the polarity of the signal needs to be detected at the reading time-instant. The signal is twice as big as that of the pulse method using an unmagnetized background.

It is apparently possible to change one digit at a time on the magnetic surface when using the phase modulation method, but the authors have had only slight experience of this particular aspect.

Frequency Modulation Method

In this method the same polarity of current is passed through the recording head for the entire digit period to represent a '0', whereas the polarity of the current is reversed at the centre of an element to represent a '1'. There is always a reversal at the beginning and end of a cell (Fig. 1(e)).

The most convenient way of recovering the information is to count the number of transitions during a digit period, one transition indicating a stored '0' and two transitions indicating a stored '1' (Fig. 1(n)). Alternatively, the signal polarity may be strobed at the end of each cell and the result of the strobing compared with that at the end of the previous cell; similar results indicate a stored '1', and dissimilar results a stored '0'. The first element can be used as an information-storing element, provided that the track is magnetized in a standard sense initially and that the first half-cell to be recorded is magnetized in the opposite sense. In this manner the first strobing may be compared with a known starting polarity.

This method is clearly related to the phase modulation method and possesses most of its properties, particularly when noise is considered. One exception, however, is that if one digit in a group is changed then it is necessary to change the rest, as in Pouliart's variation on the telegraph method.

Reading Methods, and the Effect of Noise

Sources of Noise

In any method of representation the signal from the track must compete with noise originating from external sources and from imperfections in the magnetic track. In this section the extent and nature of this noise will be discussed in relation to the method of recording. The reading method using a strobe pulse will be the only one considered; other known reading methods seem to give no fundamental advantage, and generally their use appears to be restricted.

The most significant source of noise is the presence of imperfections in the magnetic track that generally produce 'splurges' lasting from one to two or two to three digit periods. This interference seldom has a duration much shorter than one digit period since the digit-packing density is usually near the upper limit for the medium, and in consequence the frequency spectrum of the track noise extends downwards from the digit frequency rather than upwards. In other words, it is more common to find a splurge lasting for two or three digits than a 'spike' lasting for much less than a digit simply because such a spike is incapable of being reproduced.

The cause of splurges is not yet fully understood. In the case of nickel, experience shows that immediately after deposition in the plating bath the track noise is quite small but that, if at any time the surface is struck so hard as to stress the nickel or underlying metal beyond its elastic limit, then a permanent splurge is produced, possibly by some magnetostriction effect. Further, the blow apparently produces a net magnetization which cannot be removed by the usual demagnetization techniques. To stress the material less than its elastic limit is quite innocuous, but it should be remembered that even a light blow with a hard object produces a high local stress.

Track noise can be reduced in comparison with the signal by arranging that the frequency response of the amplifier falls away at frequencies below the lowest signal frequency to be recorded.

For drum storage systems working at 10kc/s and above, all other sources of noise can be made small compared with track noise. The valve noise caused by shot effect and flicker effect is quite unimportant; microphony is more serious but can be made considerably smaller than the average track noise by suitable choice of valve and mounting and by arranging that the input circuit impedance is as high as practicable in order to have a large signal at the input; hum should never be a serious trouble.

Stray pick-up from electrical machinery can be eliminated by making the earthing of the input circuit, which may include a selection network, a coaxial shield around the wiring; this shield should be earthed at only one point.

In order to achieve the best signal-to-noise ratio, F. C. Williams advocates the use of an input transformer, tuned broadly to the digit frequency, for three main reasons:

(a) It raises the input signal level slightly.

(b) It reduces the high frequency external noise and, more important, the low frequency track noise compared with the signal.

(c) It allows the phase of the recovered waveform to be adjusted slightly to improve timing margins. This is also useful in the case where each head of a group has an individual transformer of this type and for one reason or another the phases of the individual head signals are not identical.

This type of transformer is very desirable, provided that no crosstalk is experienced between a recording and a reading head. When this is expected to occur it is thought to be best to extend the pass-band of the amplifier above and below the signal frequency and to rely on the difference in timing at which reading and writing take place to minimise the effects of the interference.

Reading Balanced Waveforms

The waveforms induced in the reading head of a magnetic drum are said to be balanced when the signals corresponding to '1' and '0' are equal and opposite and when there are equal areas above and below zero during each digit period. Balanced waveforms are produced by the pulse method which uses an unmagnetized background and by phase and frequency modulation. The waveforms produced by the other methods do not fall within the definition and are said to be unbalanced.

Firstly, it is necessary to amplify to a convenient level a balanced signal induced in a reading head and then to pass the signal to a threshold stage. Ideally, the output of this stage should generate a waveform having only two levels, one when the input is above zero (i.e. positive) and the other when the input is below zero (i.e. negative). However, in a practical circuit this ideal cannot be achieved since there must always be a lower limit of input signal below which the zero threshold stage will not respond. In practice, therefore, there are two voltage levels, very close to each other in a well-designed circuit, that establish the following conditions when the threshold stage output is examined at the optimum time for reading:

(a) Signal exceeding upper level, indicating a '1'.

(b) Signal less than lower level, indicating a '0'.

(c) Signal between the levels, giving rise to an indefinite condition.

The importance of maintaining only a small difference between the two levels will be understood with reference to Fig. 2 in which:

E_s stands for the minimum signal voltage,

E_n stands for the maximum noise voltage, and

E_t stands for half the difference between the two levels.

It will be seen that one must have for successful recovery of information:

$E_s > (E_n + E_t)$, regardless of sign.

It is evident that E_t merely adds to the effective noise voltage, and therefore in the design of the amplifier and recording circuit it is essential to arrange that the gain of the pre-amplifier and/or the sensitivity of the squaring stage is such that:

$E_t \ll E_n$

In other words, track noise alone should drive through the threshold stage, so that the only condition for correct reading is that the signal voltage shall exceed the track noise. Similar considerations apply to the design of a threshold stage which establishes a pair of levels, close to each other, but not at zero voltage; in particular it can be shown that the

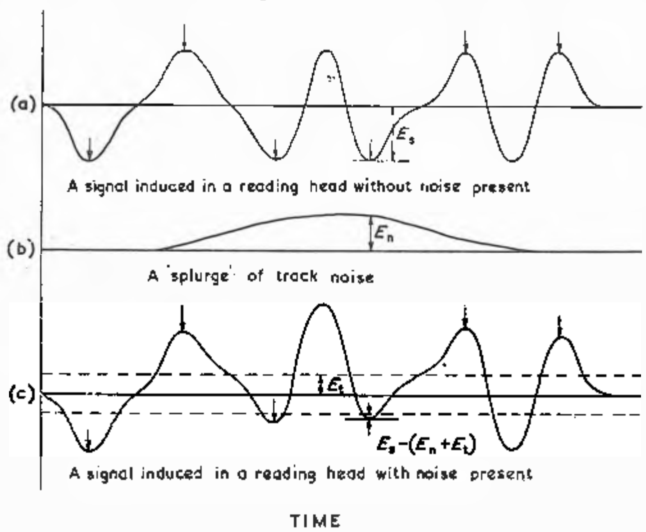

Fig. 2. The effect of track noise on a balanced waveform

difference between the levels adds effectively to the track noise.

It will also be clear that the threshold stage must be capable of changing rapidly from one output state to the other, or else time margins will be lost.

The reading process is not complete until the stored digit has been transferred to an output trigger circuit. This is usually achieved by examining the polarity at the output of the threshold circuit when the signal-to-noise conditions at the input are optimum, which is normally when the original signal is at a maximum. The results of the examination are then applied to a trigger circuit which provides the useful output of information from the track circuit.

It will be seen that this technique excludes all effects of noise not occurring at the instant of examination.

Summing up, the method of reading a balanced waveform is to sense its polarity at a particular instant, and the result will be correct provided only that the signal amplitude exceeds the noise amplitude at that instant. This therefore forms an 'ideal' method of reading in the sense that no other reading method could be better.

Reading Unbalanced Waveforms

It has already been pointed out that the unbalanced waveform recovered from a recording made on an initially magnetized track must be decoded by measuring it against a non-zero threshold voltage. It is necessary to take into account the following factors when deciding the value of this threshold voltage:

(a) If the drum is at all eccentric, then the clearance from the head to the recording surface changes within each revolution, so altering the amplitude of the recovered signals. The importance of this was first recognized by R. L. Wallace[6], who showed that the reduction in signal strength due to the introduction of a clearance was:

$$S = 20(\ln 2\pi)d/\lambda \text{dB}$$
$$= 54{\cdot}6\ (d/\lambda)\text{dB}$$

where d = space between head and drum, and

λ = recorded wavelength in the same units as d.

Substitution in this formula shows, for instance, that a variation of 0·0004in at a nominal clearance of 0·001in produces a difference in signal of 3dB for a packing density of 70 to the inch. This assumes that the track has been recorded equally strongly all the way round. In practice the recording will be strongest where the head is closest to the track, and so the variation in signal may be somewhat more than 3dB. It is perhaps not unreasonable to assume a maximum variation during life, due to all causes, of 9dB (or about 3:1 in the linear scale) although the mechanical designer should aim at a higher standard.

It will be apparent that the threshold voltage must be chosen with regard to the minimum rather than the maximum signal within one revolution. In practice one would probably arrange that the threshold lay at half the minimum signal height, in which case track noise would have more chance of breaking through the threshold in the region where the head was closest to the drum; the effect of drum eccentricity on the unbalanced waveform is therefore to worsen the signal-to-noise ratio by an anticipated factor of 3:1 or, in the general case, by the ratio of the maximum to minimum peak signal amplitudes within one revolution.

(b) In order to maintain the chosen threshold voltage at its optimum value one must hold the following five quantities constant:

(1) The speed of the drum
(2) The recording current
(3) The average clearance between head and drum
(4) The threshold voltage
(5) The gain of the amplifier.

It may be that items (1) (2) and (3) will actually be held constant for other reasons as well. Nevertheless, any variations in these quantities will cause corresponding variations in the signal-to-noise ratio.

(c) The difficulties of using one piece of reading equipment with several heads are increased by the need to adjust the output of all heads to be equal. This always means making each of them equal to the worst in the group.

It may be observed that none of these difficulties is present when the waveform is balanced because the zero voltage level for decoding can be reproduced without any trouble. For reading unbalanced waveforms one can copy the system commonly used in television circuits and arrange that there is a floating threshold derived from the local signal, but this means allocating special digital positions to hold '1's permanently all round the track because otherwise a signal could not be guaranteed. Nor is an a.g.c. system always practical because the result of recording on the head from which one is also reading would produce some undesirable changes in gain and output level.

From the foregoing remarks it will be clear that the requirements for reading an unbalanced waveform do not allow the most to be made of the signal-to-noise characteristics of the medium.

Other Factors Affecting Performance

Effective Signal

It may be noted from Fig. 1 that, for a given thickness of medium, etc., the different methods of representation do not produce the same amplitude of read signal since the flux change with distance along the track is different. For example, the signal induced in the case of the telegraph and phase modulation methods is twice that for the pulse method using an unmagnetized background. At first sight the doubling of the voltage excursion would seem to result in the doubling of the signal-to-noise ratio since the track noise is substantially constant. Unfortunately this is not so when a non-zero threshold is involved since the noise has only to exceed the threshold to cause a fault.

The value of a threshold is best set at half the peak signal when the waveform for a particular digit is considered since a burst of noise can equally add to the absence of signal

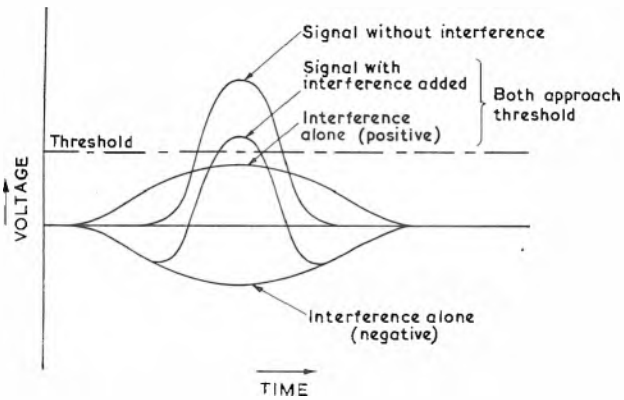

Fig. 3. The effect of positive and negative interference on threshold setting

or subtract from a signal. This point is illustrated in Fig. 3.

Probably the best method of comparing the influence of voltage excursion and threshold in the various methods is on the basis of effective signal (see note below), and consideration shows that in the phase modulation method the effective signal is twice as large as in the telegraph method or in either of the pulse methods not involving differentiation, i.e. phase modulation offers a 2:1 advantage over other methods in signal-to-noise ratio from this point of view.

For the pulse methods which involve differentiation it is not clear how the effective signal compares with that of other methods. It is observed however that the signal delivered by the head is steeper than a sine wave having the same amplitude and frequency in the neighbourhood of the time at which recording took place. This means that a sinusoidal source of noise having the same amplitude and practically the same frequency will not cause an error since, after differentiation, the steep parts of the signal will yield larger outputs than any part of the noise. Therefore, in Table 1 the effective signal of the two pulse methods using differentiation has been rated as $> \frac{1}{2}$.

Meaning of 'Effective Signal'

Imagine a drum storage system working at some digit repetition rate F, and a signal generator connected in series with the reading head set to deliver sine waves at a frequency which can be anywhere in the range from zero up to but not including F in order to provide a controlled source of noise. The output of the signal generator is increased until the reading circuits just fail; in this survey the 'effective signal' will be taken to mean the amplitude of the sine wave under the conditions just mentioned. This definition is aimed at removing the confusion that arises when systems using zero-level threshold, one half-level threshold and two half-level thresholds are compared.

The Stability of an Unmagnetized Medium

A further important factor in assessing the value of a particular method of recording is the stability of the magnetic medium in the particular state, or states, used to represent a binary digit. For example, the pulse method using an unmagnetized background depends on three primary states, namely saturated positively, saturated negatively, or unmagnetized (Fig. 1(b)). Experience in the laboratories has shown that a so-called unmagnetized medium is rather difficult to reproduce and that it is not always possible to maintain equal and opposite conditions for '0' and '1'. This is probably the reason why other workers have abandoned this method in favour of one using a magnetized background. However, this is never said directly, but instead one's attention is usually drawn to the doubling of the output signal, a ruse which, as we have seen above, is quite misleading because the effective signal is unchanged. On the other hand, the two saturation states have been found to be both reproducible and stable in most media.

There is a further hazard to the unmagnetized background because the head has no control over the unmagnetized region between digit cells. For example, a replacement head may have its gap located in a slightly different position so that regions influenced by the original will not be influenced by the replacement. A similar effect can result from a variety of fault conditions. In these circumstances it is not sufficient to pass a succession of '0's into the head in order to clear the track to the '0' state. It must be carefully demagnetized, with gradually diminishing a.c., and then a set of '0's must be recorded.

In general the use of the pulse method with an unmagnetized background is not to be recommended unless the requirement for reading and recording in one pass is absolutely essential. However, this does not mean that better results can necessarily be expected from the use of the pulse method with magnetized background which is unsatisfactory from other points of view.

Power Requirements

The pulse method requires that the recording current generator delivers larger peak power but lower mean power than is the case with the other methods. This larger peak power makes it difficult to use currently available junction transistors for switching writing head currents using a pulse method.

Circuit Convenience

The recording current waveform in the telegraph method contains frequencies extending from zero to at least several tens of thousands of cycles. This range precludes the use of a transformer in the recording circuit, even if a transition can be expected every few digit periods in a particular application, a condition which only slightly alleviates the problem. At one time the necessity for direct coupling when using this method was considered a drawback since it implied that a head had to be wound with many turns of fine wire to match into the valve impedance and that its winding would be at h.t. potential, furthermore, most circuits which used a centre-tapped recording head only used half the winding at a time which aggravated the winding problem. With the introduction of transistors, and the consequent lower operating impedance and two-way current flow, this drawback is no longer such a serious objection.

Equipment Convenience

It cannot be seriously maintained that the equipment requirements are very different for any of the methods. For example, it would appear at first sight that the phase modulation method uses more equipment for writing than the pulse method because of the need to convert to phase modulation before recording. However, experience with the pulse method indicates that the output pulse from a gate is totally unsuitable for passing directly into a recording valve; instead it has been found necessary to include equipment to generate much larger pulses upon the arrival of the small pulses and to apply these large ones to the recording valves. This restores the balance.

Changing One Digit Only at a Time

In any particular method the ease with which any isolated digit on the magnetic surface can be changed depends upon the form of the flux distribution and upon the effect any local change may have on the waveform induced in the reading head. For instance, it will be apparent that the flux distribution in Fig. 1(b) is suited to the change of an isolated digit by a single pulse in the reading head since there is an unmagnetized region separating cells. On the other hand, a change is more difficult in the case of phase modulation since the recording waveform for a single digit must start late and end early in order to cancel the effect of flux spread under the head. In the telegraph method it is considered to be even more difficult to change only one digit due to the nature of the flux distribution and the use of the threshold for reading which tends to emphasize slight defects in the flux distribution

While there remains the necessity for more experimental data on the changing of isolated digits, there are indications that there is likely to be a significant loss of signal-to-noise ratio due to the writing head field demagnetizing adjacent digits. It is therefore desirable to isolate elements, or groups of elements to be changed at one time, by short lengths of track. This isolation is desirable even when the pulse method is used, although perhaps not essential.

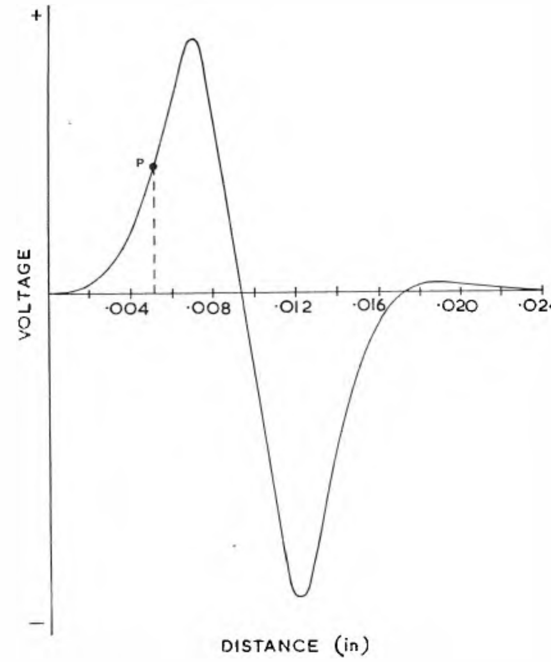

Fig. 4. E.M.F. induced in reading head after recording by pulse method on neutral background

Reading and Recording in One Pass

It has been pointed out that the advance information given by the pulse method makes it possible to read a digit and to change it conditionally during one pass of the head over the digit cell, thus eliminating the need for external

TABLE 1.

		TYPE OF SIGNAL READ BACK If unbalanced, a non-zero threshold must be used and signal-to-noise is worsened by ratio of max. to min. signal amplitude	EFFECTIVE SIGNAL (Phase modulation =1)	READING TIME IN RELATION TO RECORDING TIME If in advance, it is possible to read and record in one pass.	Is it possible to change isolated digits on the magnetic track independently of the others?	Is the Magnetic Track condition used stable?	THEORETICAL DIGIT PACKING DENSITY (Telegraph =1)	RECORDING POWER REQUIREMENTS
Pulse method unmagnetized background	Without differentiation	Balanced	½	¼ digit period in advance or in retard	Yes	Not very	½	High peak Low mean
	With differentiation	Balanced	>½	Equal	Yes	Not very	½	
Pulse method magnetized background	Without differentiation	Unbalanced	½	¼ digit period in advance or in retard	Yes	Yes	½	
	With differentiation	Unbalanced	>½	Equal	Yes	Yes	½	
Telegraph Method		Unbalanced	⅓	Equal	Unlikely	Yes	1	Low peak High mean
Pouliart Method		Unbalanced	½	Equal	No	Yes	1	
Phase Modulation		Balanced	1	½ digit period later	Difficult	Yes	½	
Frequency Modulation		Balanced	1	1 digit period later	No	Yes	½	

storage. However, there are important limitations to this technique which are best understood by referring to an example.

Fig. 4 shows a typical waveform of the e.m.f. induced in a reading head after recording by the pulse method on a neutral background. The examination time position must precede the time at which the waveform crosses the time axis because of various circuit delays (i.e. in the condition (or logic) circuit, in the recording circuit, in the current rise on recording, in the current rise on reading and in the amplifier) and because of the finite width of the pulse. In practice this cumulative delay is about 4 to 6μsec. It follows that, if the waveform is examined not earlier than the typical point P, equal to half the maximum amplitude, then the time taken between P and the point of intersection of the time axis must be 4 to 6μsec or greater. Moreover, an examination will show that this time is approximately 2/11 of the total waveform interval, which must therefore be made at least $(11/2) \times 6 = 33\mu$sec for satisfactory operation (to assume the higher delay figure). Hence, it is to be expected that the practical upper frequency limit for single-pass operation with this medium will lie in the region of 30kc/s or less and that above this frequency operation will become progressively less reliable. The medium in this example is a stressed nickel plate, and the equivalent packing density is 45/inch.

It is necessary to take a further factor into account to allow for the possibility of adjacent digits having to be changed. Because the recording power level is very much higher than the reading level there is a danger that the reading amplifier will overload and not recover in time to read the following digit. In the example, if a p.r.f. of 30kc/s were adopted, about 25μsec would be available between the conclusion of recording and the point of reading. Recovery times of this order demand a specially designed amplifier.

Yet a further factor concerns amplifier gain. The relatively low level of point P demands an amplifier with a high gain, and stability is not always easily achieved when the recovery time must be short.

Differentiation of the Recovered Signal in the Pulse Method

Differentiation improves the signal-to-noise ratio upon recovery for two reasons:

(a) Splurges are reduced in amplitude since the overall response is modified to discriminate against the lower frequencies.

(b) The recovered waveform is examined where the head had most control over the medium when recording took place. The slope of the signal as it crosses the zero level implies the presence of two peaks in the signal; both peaks therefore play a part in this reading process whereas only one is involved when differentiation is not employed.

In practice more noise is tolerated by a pulse system using differentiation than by one which does not. Naturally, the delay involved by differentiation precludes the possibility of reading and recording in one pass.

Conclusions

There is no one method of representing digital information on a drum surface which can be universally adopted since each method offers a different range of facilities. These are tabulated in Table 1.

Generally speaking, if an application does not demand reading and recording in one pass of the head over the digit cell it is technically very desirable to use the phase modulation method since it can be expected to give the most reliable operation throughout the life of the drum. If a less satisfactory standard of reliability can be accepted, and if it is necessary to read and record in one pass, then an appropriate pulse method may be adopted provided care is taken in its use.

The telegraph method appears to be less satisfactory than the phase modulation method and has not been widely adopted largely due to the inconvenient nature of the recovered waveform. However, advances in components and in circuit techniques are reducing this disadvantage.

Acknowledgments

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REFERENCES

1. Booth, A. D. A magnetic digital storage system. *Electronic Engng. 21* ,234. (1949).
2. Hill, J. L. Design features of a magnetic drum information storage system. Paper delivered to the association for computing Machinery Conference, March 28-29, 1950, at Rutgers University, U.S.A.
3. Thomas, G. E. A magnetic storage system for use with an electronic digital computer. M.Sc. Thesis, University of Manchester, 1951.
4. University Mathematical Laboratory, Cambridge. Report of a Conference on High Speed Calculating Machines, 22-25 June, 1949.
5. Williams, F. C., Kilburn, T., Thomas, G. E. Universal High Speed Digital Computers: A Magnetic Store. *Proc. Instn. Elect. Engrs. 99, Pt.* 2, 94. (1952).
6. Wallace, R. L. The Reproduction of Magnetically Recorded Signals. *Bell Syst. Tech. J.* October 1951.
7. Harvard University Computation Laboratory. A History of the development of magnetic drums at Harvard is given in the following Progress Reports: Nos. 1, 2, 3, 4, 5, 6, 7, 8, 9, 10, 13, 16, 17, 18, 19, 21, 22, 23, 25, 28.

A Simple Physiological Stimulator, Using a Transistor Oscillating Circuit

By W. T. Catton*, L. Molyneux* and B. Schofield*

The instrument described is designed to replace the conventional induction coil in experiments in which a simple stimulator is required. A single point-contact transistor is used as a relaxation oscillator and operation may be single pulse or repetitive.

THE instrument to be described was designed to replace the conventional induction coil in all class experiments in which a simple stimulator is required. It has proved satisfactory over a period of use by large classes for one year.

It offers the following facilities:

(a) Continuous control of frequency of pulse output over a range from about 1 to 120p/s.

(b) Continuous control of output voltage in 2 ranges, respectively 0 to 2·5 and 0 to 25V.

(c) Alternative repetitive or single pulse output.

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(d) Provision for triggering of single pulse by kymograph contacts, or other external switch.

The pulse-width is about 1msec, and is not adjustable. The output impedance is about 2kΩ, but is partly dependent on the setting of the voltage control. Power is supplied by a 45V layer-type dry battery, and since the current drain is only from 3 to 5mA, battery life is long. The instrument is entirely self-contained, and is now in production by the Newcastle Scientific Apparatus Co.

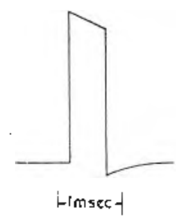


Fig. 1. The output pulse

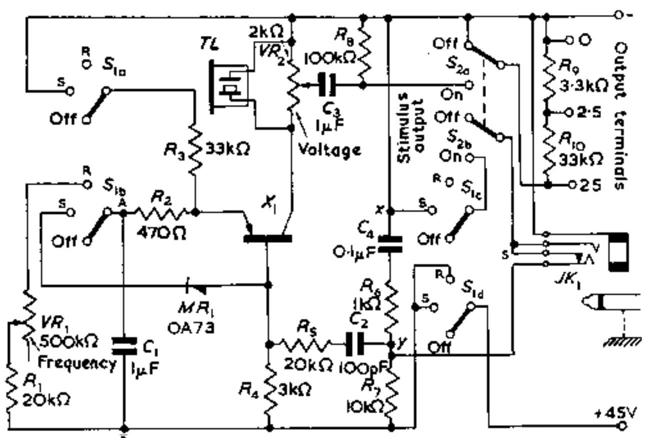


Fig. 2. The complete circuit

VR_1 = 500kΩ, inverse log law; S_1 = single-wafer, 4-pole 3-way Yaxley switch; S_2 = post-office key switch (one contact (S_{2a}) slightly bent, so as to make before the other). X_1 = Brimar TP1 point-contact transistor. TL = crystal microphone from deaf-aid equipment. Supply voltage must not exceed 45V; reversal of polarity may destroy the transistor.

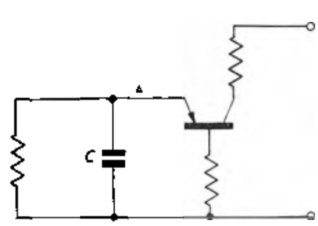


Fig. 3. The fundamental oscillatory circuit

A single point-contact transistor (Brimar TP1) is used as a relaxation oscillator (e.g. Lo[1]), and simple stimulators using a similar principle have already been described[2,3]. Such a circuit generates an output pulse which is nearly rectangular (Fig. 1) and is consistently reproducible whether the circuit is used as a free-running oscillator or is triggered from a quiescent state in order to produce single pulses only.

Technical Description

The circuit diagram (Fig. 2) may for convenience of explanation be divided into two parts, (1) the fundamental oscillatory circuit and (2) the switching necessary for simple operations.

(1) The Transistor Relaxation Oscillator

The basic circuit of the free-running oscillator is given in Fig. 3. The action of this circuit has been described by Lo[1] and others.

The repetitive rate is determined by the rate at which C discharges after the pulse has been formed and is adjusted by varying R. The pulse length is determined by the value of C and the constants of the transistor. It is not affected by the value of R, for any but very small values of R; thus the repetition rate may be varied without changing the pulse length. In order to give an approximately 'linear' frequency scale, a logarithmic control (VR_1) is used in the practical circuit.

Single Pulse Operation

If R is removed (Fig. 3) and point A is maintained a little negative with respect to the base, the circuit will not oscillate. It can be made to go through one cycle of oscillation by driving the base negative with respect to the emitter, when a pulse is formed. By making the trigger pulse short with respect to the output pulse, the shape of the latter is unaffected by variations of the former. The derivation of this trigger pulse is described below.

(2) The Switching Arrangements (Fig. 2)

S_1 is the main control switch, which selects the type of operation required. It has three positions, of which the first is OFF, in which the battery positive supply is disconnected (S_{1d}). In position 2 (S = single stimuli) negative bias is applied to the emitter via R_3, sufficient to prevent free oscillation. The lower contact of S_{2b} is connected to the negative line and earth via S_{1c}, and battery positive is connected via S_{1d}. The production of a single pulse now awaits the depression of S_2, which applies a short-circuit via S_{2b} across points x and y, the jack socket auxiliary contact S being normally made when the jack is not inserted. Capacitor C_4 now rapidly discharges through R_6, and point y is carried negative. A negative pulse so derived passes via C_2 (whose capacitance is small) and R_5 to the transistor base. This negative pulse moves the base bias sufficiently to initiate a single cycle of oscillation. Even though the key S_2 is held down indefinitely, no further oscillations occur, since C_2 remains charged to its new potential. When S_2 is released, C_4 recharges through R_7 and R_6, and the potential at y undergoes a relatively slow positive change, which does not of course produce a discharge from the transistor. Just before S_{2b} applies the short-circuit across x and y initiating the output pulse, S_{2a} connects the output from the collector load potentiometer VR_2 via C_3 to the output attenuator consisting of R_9 and R_{10}, the output terminals being short-circuited in the OFF (spring-loaded) position of S_2. S_2 thus serves to give single stimuli in this condition (main switch S_1 to S, no jack inserted). When the jack is inserted, however, the operation differs, essentially because the auxiliary socket contact S is now broken. If the jack leads are not shorted, operation of S_2 now has no effect. The jack leads are normally connected to the simple make and break switch on the kymograph, although any simple shorting switch will serve. When this switch makes, the single stimulus is produced in exactly the same manner as described above, so long as S_2 is also depressed. It is seen that S_2 now decides when the operation of the external switch shall be effective in producing a stimulus, and it is intended to be used in this manner.

When S_1 is turned to the third position (R = repetition), the emitter is disconnected from the negative line and the diode and is connected instead to the variable resistor VR_1 in series with R_1. Now $(VR_1 + R_1) = R$ in Fig. 3, and its total value determines the discharging time of C_1, and hence the interval between successive pulses; VR_1 now acts as a frequency control, and R_1 is a ballast resistor which sets the upper frequency limit. (The lower frequency limit is chiefly set by the value of C_1). Depression of S_2 now causes the transistor output, developed across the collector load potentiometer VR_2 to be applied to the output attenuator. VR_2 bears a linear voltage calibration which serves for both output ranges of 0 to 2·5 and 0 to 25V. The crystal microphone element TL across VR_2 gives an audible signal of each pulse.

The following notes complete the description of the circuit. R_2 limits the maximum emitter current flow. R_8 serves to hold the lower contact of S_{2a} at d.c. earth, to prevent a spurious output pulse on depressing S_2 due to accumulated charge on C_3. The jack frame is connected to negative line and the jack body is connected to the earth terminal of the kymograph switch. This prevents the appearance of spurious pulses on operating S_2, due to the capacitance to earth of the stimulator, and provides a single earth point arrangement. Resistor R_7 serves to prevent short-circuiting the battery when S_2 is depressed. Diode MR_1 is included in order to re-establish the potential of point A (Fig. 3) as rapidly as possible after the completion of one pulse. This is necessary in order to provide for the firing of pulses in rapid succession when the kymograph switch arms are set very close together. The minimum interval within which two discrete pulses can be generated may be termed the 'recovery time' of the stimulator. In order that it shall be as short as possible, the recovery period of the stimulator should also be made short, which is the limiting factor. Diode MR_1 presents a low-resistance charging path for C_1, by which the recovery period is much reduced. With the practical circuit values the recovery time is about 5msec. It can be made less by using a lower value of C_1, but this will shorten the pulse-width. It can be greatly increased by inserting a high-value resistor (1 to 2MΩ) between point y and R_7. This may be a useful modification when only widely-spaced single stimuli are required and where the external contacts used are not of good quality, since this addition will prevent multiple firing with the rapid intermittent making and breaking of such contacts. The rate of stimulation may thus be compulsorily limited to not more than one every second or so, according to the value of resistance inserted.

The Output Pulse

As explained above, the output pulses are consistent in shape and size, and conform closely to rectangular waves (Fig. 1). A pulse width of about 1msec was chosen; it is dependent, as pointed out above, on the value of C_1 and the electrical properties of the individual transistor. In the thirty-six models produced so far there has only been found a small variation from one transistor to another. When changing over from the Mullard OC51 used in early models to the Brimar TP1 transistor, it was found that the pulse width was scarcely affected, but that the low frequency end of the repetition range was somewhat extended, and the high frequency end a little restricted. Thus the value of R_1 had to be reduced to restore the high frequency upper limit; this did not affect the lower limit. No other modifications proved necessary.

Power Supply

The experience gained so far with the use of the instrument has been mainly limited to models supplied individually by 45V dry batteries of the layer type. The current used is from 3 to 5mA, and with intermittent use the batteries have been found to last for a year or more. Exhaustion of the battery generally causes failure of single pulse output before repetitive. It is necessary to insist that the instrument be switched off after use, and if this is not done the battery is exhausted within a day or so. As an alternative to battery supply, a low-power individual mains supply unit using a metal rectifier in a half-wave circuit has been found satisfactory.

Operating Instructions (see Fig. 4)

(1) Connect stimulating electrodes to output terminals according to range required.

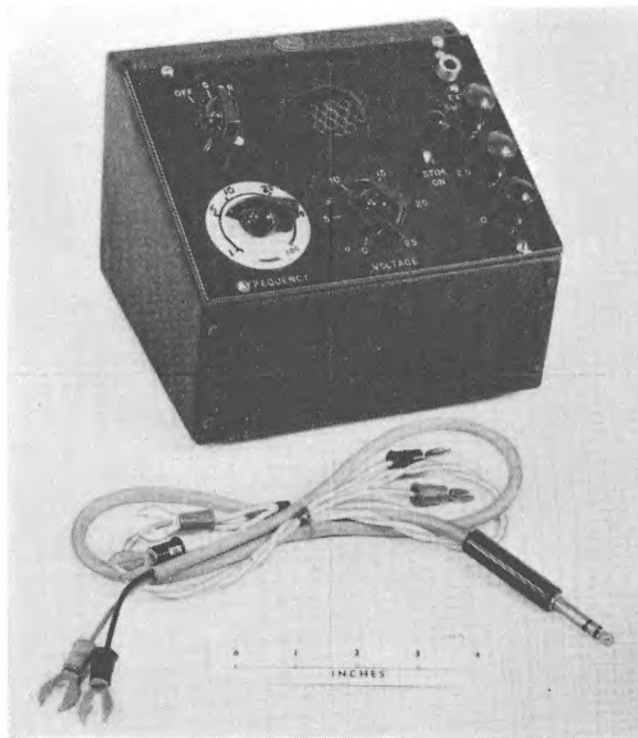


Fig. 4. Transistor stimulator and connecting leads

(2) For single stimuli, manually produced, turn main switch to position S (jack plug not inserted). For each depression of the key switch ('stimulus on') a single stimulus will be produced.

(3) For single stimuli, synchronized with kymograph spindle rotation; main switch at S, and insert jack plug into 'ext. switch' socket. Connect jack leads to kymograph switch contacts (black lead to earth). Depression of stimulus key will not now produce a stimulus (if the kymograph contacts are not made), but so long as the key is held down, a stimulus will occur each time the kymograph contacts operate. With the kymograph contacts separated to various angles, double stimuli will be produced until the interval is less than 5msec (the resolving time of the stimulator) when only one will appear.

(4) For repetitive stimulation. Turn main switch to R and depress stimulus key switch. Control the voltage and frequency on the appropriate dials.

(5) When not in use turn main switch to OFF, to conserve the battery.

REFERENCES

1. Lo, A. W. Transistor Trigger Circuits. *Proc. Inst. Radio Engrs. 40*, 1531 (1952).
2. Molyneux, L., Pask, E. A. A General-purpose Stimulator. *Brit. J. Anaesth. 26*, 284 (1954).
3. Marshall, C., Goodwin, G. W. Miniature Stimulators. *J. Neurosurg. 12*, 651 (1955).
4. Catton, W. T., Molyneux, L. A New Stimulator Using a Transistor Relaxation Oscillator. *J. Physiol. 128*, 27 (1955).

V.H.F. LINE MEASUREMENTS

(A New Method)

By F. J. Charman*

The device described gives a measure of the impedance or admittance components of the unknown load in a form suitable for graphical solution, but not directly applicable to a Smith Chart. It is compared with the impedometer, which gives a measure of the reflection coefficient in a form directly applicable to a Smith Chart. The new method offers certain advantages which tend to make it more accurate than the impedometer.

A NEW method of making transmission line impedance measurements is described. The method employs a single loop-type directional coupler which is normally used in a fixed position for the observation of the reflection coefficient, but which makes use of the variation of output which results when the coupler is rotated. The new type of instrument, which measures impedance, is compared with the reflectometer which measures reflection coefficient. Because of its relative simplicity it offers the possibility of better accuracy than does the impedometer.

In the v.h.f. range it is not practicable to use conventional instruments such as self-contained bridges for the measurement of impedance or reflection coefficient. This is mainly because of the difficulty of defining the terminals or point of connexion at which the unknown is measured. This difficulty becomes serious at frequencies above, say, 200Mc/s. It is therefore necessary to resort to instruments with which the observations can be made inside a uniform transmission line which is continuous with the circuit to be examined.

The most popular form of instrument is the slotted line from which, by means of a travelling detector probe, the standing wave pattern in the line can be delineated, and the impedance or reflection coefficient of the unknown can be calculated or plotted directly on a Smith Chart[1]. The 'terminal plane'† can now be chosen anywhere in the transmission line system, and the corresponding impedance or reflection coefficient found by simple angular rotation on the chart corresponding to the electrical distance between the required terminal plane and the observation points. The slotted line has the advantage that it is easy to understand and to use, and gives a visible picture of the wave inside the line, but it is relatively slow in use, and it is difficult and expensive to construct a line suitable for precision measurements, particularly at the lower frequencies where the measuring line is several feet long.

Another method which leads to a more compact instrument is the 'Impedometer' described by Parzen[2]. This uses two directional coupler loops in the wall of the line adjusted to separate the incident and reflected travelling wave components which together make up the standing wave pattern on the line. The ratio of the outputs of the two couplers gives a direct measure of the ratio of the two wave amplitudes, i.e. of the magnitude $|\rho|$ of the reflection coefficient. In order to determine the phase-angle of the reflection coefficient it is necessary to insert two other devices into the line, at least one of which is a voltage probe giving an output proportional to the line voltage at its point of insertion, i.e. proportional to $|1 + \rho|$. These two amplitudes $|\rho|$ and $|1 + \rho|$ can be plotted directly on the Smith Chart to give the complete ρ or the corresponding normalized impedance z, which can be transferred back to the required terminal plane by rotation of the chart, as with slotted line observations.

The Impedometer, like the slotted line, will work over a wide range of frequencies, and has the advantage that compact instruments can be made to work at any frequency where concentric lines would normally be used. The directional couplers can be made nearly reflectionless, and on this account, the instrument can be more precise than a slotted line.

The cost of precision lies more in the design and the skill employed than in accuracy of construction. The chief limitation is the disturbance introduced by the additional detector required for determining phase-angle; unless the instrument is very insensitive this introduces an error of a few per cent which cannot be compensated.

The new method (discovered by M. Borchert of Telefunken G.M.B.H.) uses a directional coupler loop which can be rotated about its axis (normal to the line) to give outputs at various positions of rotation which can be used to determine the real and imaginary parts of the impedance, or of the reflection coefficient. A brief description of the loop coupler, as used in reflectometers and impedometers, will help to make the new method clear.

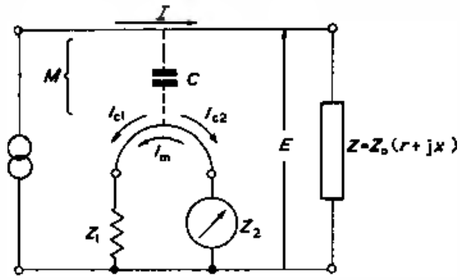


Fig. 1. Equivalent circuit of directional coupler loop

The Directional Coupler

Fig. 1 shows, diagrammatically, a loop coupler. It is inserted into a sleeve in the side wall of the main transmission line, and arranged so that it can be rotated in order to adjust its discrimination between the incident and reflected waves. One end of the loop is terminated in a resistance which has roughly the same value as the characteristic impedance of the loop—say 50 to 100Ω. (This resistance is not necessarily equal to the main line impedance). The other end of the loop is taken either to a detector mounted in the coupler unit itself, or via concentric cable to a receiver or other r.f. indicator.

The equations for the directional coupler will be discussed with reference to Fig. 1. Let the forward wave direction be represented by the arrow indicating the direction of the current I, and the voltage from the line conductor to earth, by E. E/I defines the impedance of the load as seen through the line from the coupler position. The forward and

* *E.M.I. Electronics Ltd.*

† *A convenient cross-sectional reference position in the line to which all measured impedances can be referred.*

reflected waves, considered separately, are pure travelling waves and therefore for each of these, the ratio E/I is equal to the characteristic impedance Z_o. Hence one can solve the network for a travelling wave, considering only one direction at a time.

The loop is coupled to the line by mutual inductance M and capacitance C. The inductive coupling induces an e.m.f. in the loop in the direction shown for the loop current I_m. The current through the capacitance branches either way in the loop, as indicated by I_{c1} and I_{c2}.

Assuming that $1/\omega C \gg Z_1$ or Z_2, one finds that

$$I_{c2} = j\omega CE \frac{Z_1}{Z_1 + Z_2} \quad (1)$$

$$I_m = \frac{j\omega MI}{Z_1 + Z_2} \quad (2)$$

Now $I_2 = I_{c2} - I_m$ and this will be zero if

$CEZ_1 = MI$, that is

$$M/CZ_1 = E/I = Z_o \quad (3)$$

which is the required condition for a detector in branch 2 to read zero for the wave travelling in the direction of the arrow I.

In the other branch, I_{c1} and I_m are additive so that, with the above balance adjustment, the total current is

$$I_1 = \frac{j\omega(CEZ_2 + MI)}{Z_1 + Z_2} \quad (4)$$

and using equation (2) one finds that, if $Z_1 = Z_2$, the current in branch 1 is

$$I_1 = j\omega CE = j\omega MI/Z_1 \quad (5)$$

The assumption that $1/\omega C \gg Z$ is not fulfilled exactly. Equation (1) is based on the assumption that the admittance to I_c is $j\omega C$. The finite resistance in series with C gives I_c a phase-angle which is slightly different from 90°, and therefore the balance represented by equation (3) is never complete. This error is partly compensated by the leakage reactance of the loop and any residual inductance in the resistor, and may be reduced to zero by means of additional capacitance across the resistor or from the loop to earth.

When there is a standing wave on the line, owing to reflection, the current and voltage on the line are, by definition of the reflector coefficient.

$$I = I_o(1 - \rho)$$

$$E = E_o(1 + \rho)$$

where the subscript $_o$ refers to the forward wave and ρ is the complex reflection coefficient at the position of the coupler loop in the line. The current in the detector branch can now be written

$$I_2 = j\omega \frac{CE_oZ_1(1 + \rho) - MI_o(1 - \rho)}{Z_1 + Z_2} \quad (6)$$

If the balance condition (3) is included, the current in a reflected-wave detector is given by

$$I_2 = \frac{2j\omega CE_o\rho}{Z_1 + Z_2} \quad (7)$$

Looking the other way, the sign of M is changed in equation (6) resulting in

$$I_2 = \frac{2j\omega CE_o}{Z_1 + Z_2} \quad (8)$$

and this corresponds to the current in a forward-wave detector. The ratio of the two, for the same values of M and C, is $|\rho|$.

The important results of these equations are first that the balance condition is independent of the detector impedance or the frequency, and secondly that identical couplers facing either way will give outputs in a ratio equal to the magnitude of the reflection coefficient. Changes in frequency or Z_2 only affect the sensitivity of the coupler.

In construction, it is not practicable to design the loop to give a perfect balance: fine adjustments are necessary. For this purpose the coupler unit is made to rotate and is given an excess of mutual inductance over capacitance coupling by suitable choice of conductor dimensions, and is fitted with a small trimmer capacitance (about 1pF) between loop and earth. With the line terminated, a rotation of the loop, together with adjustment of the trimmer, will bring about a balance. The discrimination, or balance between waves in the two directions can usually be set to over 50dB at any frequency.

A pair of such couplers will give $|\rho|$ only, and this independently of their positions in the line. In order to resolve the phase-angle of ρ, a third detector is used, in the form of a plain capacitance voltmeter probe in the line. If this is set to the same normal sensitivity as the loop couplers, it will give a relative output proportional to E, that is, to the value $|1 + \rho|$ in which ρ has the phase-angle at the voltmeter position in the line. These two values $|\rho|$ and $|1 + \rho|$ can be plotted directly on a Smith Chart, $|\rho|$ being measured from the centre of the chart proportional to its radius, and $|1 + \rho|$ to the same scale from the point $Z = 0$, the intersection of these vectors giving ρ or z. There are two intersections possible, and thus it is necessary to make a further observation to determine whether the required value ρ, or its corresponding normalized impedance, is on the capacitive or inductive half of the chart. Parzen added a plunger which could momentarily place a small extra capacitance across the line—the writer has used a second voltmeter to give a check reading at a different point on the line.

Rotating Directional Coupler

The new method is based on the variation of output which occurs when the loop is rotated, and thereby disposes of the second loop coupler and voltage probe which are needed in the impedometer. From readings taken at particular setting angles, it is possible to determine ρ and to separate the real and imaginary components of the unknown load. As with the impedometer, the sign of the phase-angles is indeterminate, and an extra observation is necessary.

Putting the normalized load impedance $z = Z/Z_o = r + jx$, and the angle between the loop and the line cross section as α, and replacing M by $M_o \sin \alpha$, the current in the detector branch becomes:

$$I_2 = \frac{j\omega}{Z_1 + Z_2}(CRE + IM_o \sin \alpha) \quad (9)$$

Putting $E = IZ_oz$, and transposing,

$$\frac{I_2(Z_1 + Z_2)}{IZ_oj\omega} = (CRZ + m_o \sin \alpha) \quad (10)$$

where $m_o = M_o/Z_o$

The effect of rotation of the head is given by the variation of $\sin \alpha$, and the output is proportional to the right-hand side of this equation, all terms except I_2 on the left-hand side being constant against variation of α.

If now, α is adjusted to the angle α_o giving maximum discrimination against a correct termination, (i.e. z = unity), then, $CR = -m_o \sin \alpha_o$. Both these terms can be regarded as constants for the instrument, and therefore one can divide out equation (10) and transfer CR to the l.h.s., so that

$$I_2 \simeq (z - (\sin \alpha/\sin \alpha_o)) \quad (11)$$

This is the equation for the output variation against rotation of the loop.

The first result is obtained when α is varied for minimum output. Since $z = r + jx$ and the sine ratio is real, this minimum occurs when the two real parts cancel:

$$r = \sin \alpha / \sin \alpha_0 \quad \ldots\ldots\ldots\ldots\ldots (12)$$

while the residue output must be proportional to

$$I_{2(\min)} \simeq |x| \quad \ldots\ldots\ldots\ldots\ldots (13)$$

The next step is to set the loop at the two angles $\pm\alpha_0$ corresponding to maximum directivity of the coupler for either direction of the line, giving

$$I_2 \simeq |z \pm 1| \quad \ldots\ldots\ldots\ldots\ldots (14)$$

and it should be noted that the ratio of these readings gives

$$|\rho| = \left|\frac{z-1}{z+1}\right| \quad \ldots\ldots\ldots\ldots\ldots (15)$$

A further reading can be taken when $\alpha = 0$, giving

$$I_2 \simeq |z| \quad \ldots\ldots\ldots\ldots\ldots (16)$$

There are more readings here than are necessary, so that redundant readings provide a check on accuracy of measurement. They are summarized in Table 1.

TABLE 1

	SETTING	OUTPUT
A	$\alpha = 0$	$I_2 \simeq \lvert z \rvert$
B	$\alpha = \alpha_0$	$I_2 \simeq \lvert z - 1 \rvert$
C	$\alpha = \alpha_0$	$I_2 \simeq \lvert z + 1 \rvert$
D	$\alpha = \alpha_{\min}$	$I_2 \simeq \lvert x \rvert$
E	$\dfrac{\sin\alpha_{\min}}{\sin\alpha_0}$	$= r$

Determination of z

The instrument line is terminated accurately by means of a suitable load (e.g. a long length of cable if nothing better is available) and the coupler is set up for zero output by rotation and adjustment of its trimmer capacitance. This determines the angle α_0. When the unknown is connected, two other readings, such as A and D will resolve $z\angle\phi$. The angle $\alpha_{\min}$ of D gives r absolutely while the ratio

$$D/A = |x/z| = \sin\phi$$

from which

$$|x| = r\tan\phi = \frac{r}{\sqrt{[(A^2/D^2) - 1]}} \quad \ldots\ldots\ldots (17)$$

Alternatively, if B or C is taken instead of A, one has

$$D/C = \left|\frac{x}{1 + r + jx}\right|$$

giving

$$|x| = \frac{1 + r}{\sqrt{(C^2/D^2 - 1)}} \quad \ldots\ldots\ldots\ldots (18)$$

and a corresponding equation in B/C and $(1 - r)$.

Readings B and C are more suitable for finding ρ.

The graphical solution is simple. In the $r + jx$ plane (Fig. 2) the ordinate kx is marked proportional to reading D. Reading A, or kz must then complete the right angled triangle ODA, after which the triangle is scaled down to fit the known value of r, and to give x. It will be seen that variation of α simply moves the output vector along the line BC. Readings B and C for $z \pm 1$ should therefore also fall on the line DA, and it may be preferable to use these readings and check that A falls in the centre of BC. The diagram is shown for the case where $r < 1$. When r is greater than unity, B must fall to the positive side of D. Thus r, k and ϕ are found, except that the sign of x is still indeterminate.

Measurement of ρ

The magnitude of the reflection coefficient is given by equation (15) and is the ratio of readings B and C of the Table 1. Since the phase-angle of the ratio of two vectors is the difference between their individual angles, the phase-angle of ρ must be given by

$$\arg(\rho) = \arg(z - 1) - \arg(z + 1)$$

$$= \tan^{-1}\frac{x}{r-1} - \tan^{-1}\frac{x}{r+1} \quad \ldots\ldots (19)$$

These two angles are the angles of vectors OB and OC to the positive real axis of Fig. 2 and hence

$$\text{angle of } |\rho| = BOC$$

Sign of Phase-Angle

Because this method, like the impedometer method, only measures magnitudes, and does not give the signs of phase angles, an additional facility is necessary. The simplest arrangement is to alter the reactance term of z artificially, and the method used by Parzen is applicable, namely a spring-loaded plunger opposite the coupler loop, which normally rests flush with the wall of the line, but when depressed places a small capacitance across the line. In the case of Fig. 2 which is drawn for a positive reactance, pressing the plunger reduces the total reactance and therefore the readings B, C, or D will fall. On the other hand, if operation of the plunger increases the outputs, x is negative and the angle of ρ or z, taken from the diagram of Fig. 2, must be changed in sign.

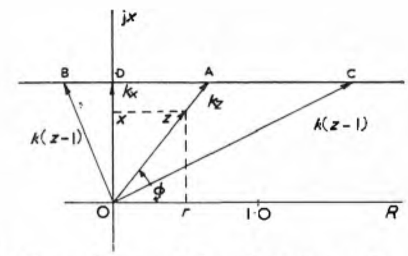


Fig. 2. Vector diagram solution of rotating coupler outputs

Impedance Range

The range of the instrument is limited. If r is greater than $\sin 90°/\sin\alpha_0$, i.e. $r > \operatorname{cosec}\alpha_0$, the minimum output will occur at $\alpha = 90°$, but will not give the correct result. For high values of r, the loop coupler must be designed to balance at a low value of α, i.e. it must be given a large excess of mutual inductance or a small value of resistance. This is undesirable as it leads to a coupler in which the setting α_0 varies rapidly with frequency.

When r is much lower than unity, a high value of α_0 is desirable for best accuracy: the loop should be arranged to balance at about $\alpha_0 = 60°$, by suitable adjustment of its terminating resistance. The case $r > 1$ can always be dealt with by increasing the line length between the coupler and the unknown impedance by $\lambda/4$: a handicap which also affects the impedometer.

Construction and Application

The balance equation (3) for the loop assumed that the Z_1 arm of the loop is purely resistive. There are two sources of residual phase error, mentioned above, which are compensated by means of a small trimmer capacitance. It can be shown that when this adjustment is made, the loop places an almost reflectionless load on the line.

A suitable width of loop is about one quarter of the overall line diameter. This will give an output of the order of 20 to 30dB less than the main line power when the top of the loop is a little way inside the outer wall of the line.

In order that the capacitance coupling into the line shall not vary as the coupler is rotated, the top of the loop should take the form of a flat disk or ring of wire mounted on two wire pillars from the base (see Fig. 3). Such a loop would be suitable for use at frequencies of 200Mc/s upwards, but would be rather insensitive at 50Mc/s as the output is proportional to frequency. However, impedometers have been made with more than one turn in the loop, to work at quite low frequencies, and there is no

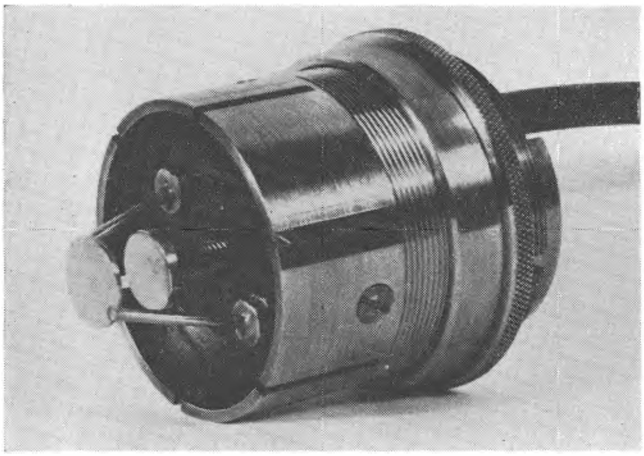


Fig. 3. Directional coupler

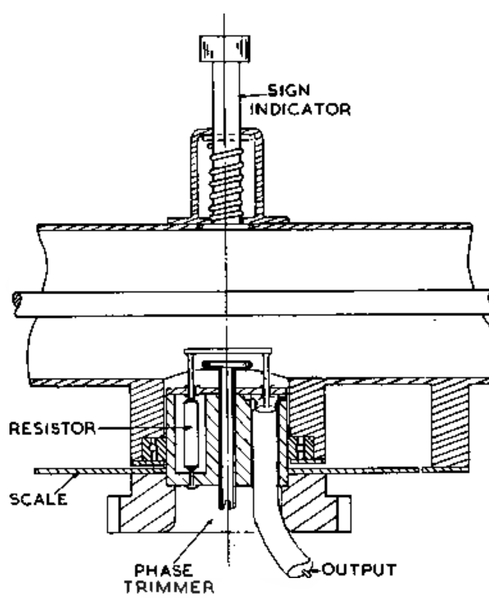


Fig. 4. Layout of rotating loop impedance meter

reason why the new instrument could not be modified in this way to work in the lower frequency range.

The instrument itself can be built into a short section of line of suitable diameter, with taper ends to the required coaxial connectors. Provided the diameter ratio is maintained throughout the taper, it introduces no tangible error. The coupler can be used in either direction of the line, and therefore the line can be reversed, so that if the coupler is set off-centre by one-eighth of the mean working wavelength, reversing the instrument will change the length of circuit between the coupler and the load by $\lambda/4$ and r can thus always be made less than unity. Fig. 4 shows a section of the coupler in the line.

The best method of working is to use a tuned receiver as an output indicator and, using a calibrated attenuator between the generator and the line, to obtain constant output for the various adjustments shown in Table 1. A stabilizing attenuator pad of say 10dB loss is desirable, if sensitivity is adequate, but is not essential. Modulation of the source, together with tuned audio amplification of the output, can be used to increase the sensitivity.

The measurements, of course, give the impedance of the unknown as observed from a point on the line opposite the coupler. This must be referred back to the required terminal plane, say the output terminal of the instrument, or the far end of a line. It is therefore necessary to know the electrical length of circuit between the coupler and the terminal plane. This is found by short-circuiting the terminal plane and measuring the resulting reactance at the coupler position. In these circumstances the minimum output (reading D) should be obtained at $\alpha = 0$, while readings B and C should be equal. The ratio B/A gives $|(jx - 1)/jx|$ whence

$$x = \frac{1}{\sqrt{[(B^2/A^2) - 1]}} = \tan\theta$$

where θ is the required electrical length $\pm(n\lambda/2.)$ The sign of x should be found. The reference of measured imped-

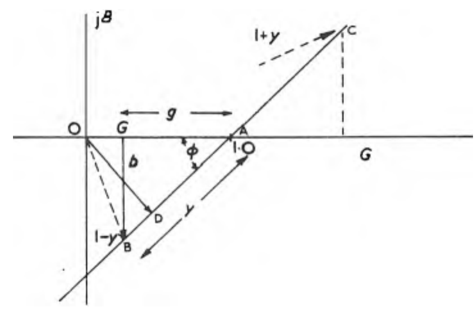


Fig. 5. Vector diagram of outputs reduced to admittance form

ances back to the load is best done with a Smith Chart; the distance from the intercept of the x curve with the edge of the chart to the point $z = 0$ is the required rotation, and this rotation should be applied to the measured z or ρ in the opposite direction.

Results Expressed as Admittance

For some purposes it may be more convenient to reduce the readings to the form $y = g + jb$. If one repeats the development of equation (10), but dividing by y, one has

$$I_2/E \,.\, \frac{Z_1 + Z_2}{j\omega C R m_0 \sin\alpha_0} = 1 - y\,\frac{\sin\alpha}{\sin\alpha_0}$$

It can be seen from this that readings A, B and C, with $\alpha = 0$ and $\pm\alpha_0$ give unity, and $1 \mp y$. $|\rho|$ remains equal to B/C.

But varying α now changes both g and b with reference to unit output A. The condition for minimum output is, thus, $g\,(\sin\alpha_{min}/\sin\alpha_0) = 1$ with a residue $b\,(\sin\alpha_{min}/\sin\alpha_0)$. The way in which this affects operations is best seen by a vector diagram of equation (10). In Fig. 5, the letters correspond to the readings of Table 1, e.g. DA is the output when $\alpha = 0$, and is unit reference. Hence $OG = 1 - g$ and $GB = -b$, so that OB is $1 - y$

Varying α now moves the output vector along AB. This will be a minimum when OD is normal to AB, and triangles AGB and ADO are similar, with the angle ϕ (the phase-angle of y) common to both.

Thus,

$$\frac{\sin\alpha_{min}}{\sin\alpha_0} = (AD/AB) = \frac{\cos\phi}{y} = (g/y^2)$$

The table then takes the form shown in Table 2.

TABLE 2.

	SETTING	OUTPUT
A	$\alpha = 0$	$I_2 \simeq$ unity
B	$\alpha - \alpha_0$	$I_2 \simeq 1 - y$
C	$\alpha = \alpha_0$	$I_2 \simeq 1 + y$
D	$\alpha - \alpha_{min}$	$I_2 \sim b/y = \sin\phi$
E	$\dfrac{\sin\alpha_{min}}{\sin\alpha_0}$	$= \dfrac{g}{y^2}\ (-r)$

Case E does, in fact, give r and the series-parallel transformation $r = (Rx^2/(R^2 + x^2))$ is equivalent to g/y^2. The phase angle of ρ is, as before, the angle BOC.

To construct the diagram, make OA equal to unit conductance and draw a circle on O of radius OD (the readings at α_{min}). Draw a tangent from A to the circle. Readings B and C will give the intercepts B and C, and hence y, g, and b, together with a check on the accuracy.

Comparison With Other Instruments

The nearest relation of this instrument is the impedometer already mentioned; this uses at least three detectors in the line which give ratios proportional to $|\rho|$ and $|1 + \rho|$ and these can be plotted directly on a Smith Chart. The new device gives impedance values which must be plotted on an $R + jx$ diagram. It has the possibility of increased accuracy, since only one detector is placed in the line. There is no essential difference in coupler design except the provision of an angular scale, and the same coupler could be used in either instrument. The extra arrangements which are necessary for determining the sign of the phase-angles, or for dealing with high impedances are common to both methods.

By comparison with the slotted line, both types of loop coupler system are more compact. They are most sensitive when the line is nearly matched, since the readings at α_0 or α_{min} approach a null balance. In the slotted line, the position of the minimum is most clearly defined when the matching is poor, i.e. the two methods are complementary.

There are two other rotating detector instruments which may be mentioned[3,4]. In each of these, a three-way right-angle line junction is used, one for the generator, one for the load and one for a normalized unit reactance. A rotating probe exploring the field at the junction gives an output which corresponds to that of a slotted line probe, one revolution representing one wavelength of line. The two types differ according as the probe is electric or magnetic. These give a slotted line result, but can be built compactly. Such instruments are very difficult to design, because of the problem of field distribution at the junction. By comparison, it should be noted that the new method does not give a slotted line delineation when the coupler is rotated, because it samples both current and voltage simultaneously.

Acknowledgments

Acknowledgment is made to E.M.I. Electronics Ltd. and Telefunken G.M.B.H. for permission to publish this article, the latter Company having discovered and communicated the method.

REFERENCES

1(a). SMITH, P. H. Transmission Line Calculator. *Electronics* (Jan. 1939). SMITH, P. H. An Improved Transmission Line Calculator. *Electronics 17*, 130 (Jan. 1944).

1(b). JACKSON, WILLIS. High Frequency Transmission Lines. (Methuen Co. Ltd., 1951).

2. PARZEN, B. Impedance Measurements with Directional Couplers and Supplementary Voltage Probe. *Proc. Inst. Radio Engrs. 37*, 1208 (1949).

3. EGGER, ALOIS. Widerstand Messung mit der Meßverzweigung. *Fernmeldetech. Z. 8*, 277 (1955).

4. The Polytechnic Research and Development Co. of New York manufacture a rotary s.w.r. indicator using electrostatic coupling.

Computers for British Railways

Two electronic computers have recently been installed by British Railways as part of its modernization scheme.

The first of these has been brought into operation in the Regional accountants' office of the Western Region at Swindon for commercial, industrial and public service accounting and industrial research computing.

The machine, a Powers-Samas Programme Controller computer will feed, process and punch up to 7 200 cards an hour. As it works directly with decimal and sterling numbers, conversion and reconversion to some other scale of notation is eliminated.

An unusually flexible method of programming is employed and each specific programme for a particular application is set up on special programme units which are inserted in the machine each time the particular type of work is to be undertaken.

The machine has a main store with a total capacity of 2 560 sterling or decimal digits, and in addition six immediate-access stores and storage for input, output and programme instructions where required. In addition the punched cards themselves provide very comprehensive information stores.

Various automatic checking safeguards are incorporated in the machine. These cover the checking of the accuracy of the input information; similar checking of the accuracy of the punched results, while the computations are fully checked by a method of alternative programming.

One of the principal uses to which this computer will be put is the calculation of wages for the 10 000 employees of the Swindon Works.

Each of the 10 000 wages grade employees may receive many varying hourly rates of pay during a week, and be entitled to flat time, time and a quarter, time and a third, time and a half, and double time in the same week.

The cycle of operations starts with the marking of the pay cards. These cards are already partly punched with the employees' numbers, pay rates, etc., by automatic process at the end of the previous week's operations.

The marking consists of pencil marks made on the cards in appropriate positions to represent the different kinds of hours worked. The cards are then fed through an electronic 'Mark Sensing Punch' which automatically reads the marks and punches corresponding holes in the cards, at over 5 000 cards an hour.

The cards are then ready for the computer which firstly calculates the uplift to be applied to the hours worked according to the nature of the overtime, multiplies the results by the hourly rates, adds in bonus and allowance, and so arrives at gross pay.

Then, without any need for reference to P.A.Y.E. tables, the computer works out and deducts the income tax due, subtracts various other deductions and punches net pay into the card which is then fed into a tabulator to print the payslip and payroll.

The whole computation from 'hours worked' to 'net pay' takes only two seconds per employee.

Another essential weekly task for the computer is to calculate each employee's piecework bonus. The bonus is earned by gangs, and has to be divided among the members of the gang in exact proportion to the hours worked by each man and his basic hourly rate.

This calculation is performed by the computer at the rate of one employee per second.

The second computer has been installed at the accountants office of the London Midland Region at Derby to provide information on inventory control of stores.

Inventory control, which is a major job at the works, covers the control of all stores holdings (by both quantity and value), ensuring that stores items are immediately available without the necessity of carrying large stocks. Derby Stores deals with 60 000 different items needed for the building and repair of locomotives, carriages and wagons.

There are about 3 000 stores movements daily at Derby, each one of which affects the inventory both in quantity and value. The machine, a Hollerith "550", will supply accurate and timely statistical information, performing as a matter of daily routine the calculations necessary for the adjustment of inventory records, resulting from stock issues and receipts.

THE S.B.A.C. EXHIBITION

A description, compiled from information supplied by the manufacturers, of a few of the electronic exhibits shown at the recent exhibition of the Society of British Aircraft Constructors at Farnborough.

PICTURE TRANSMISSION

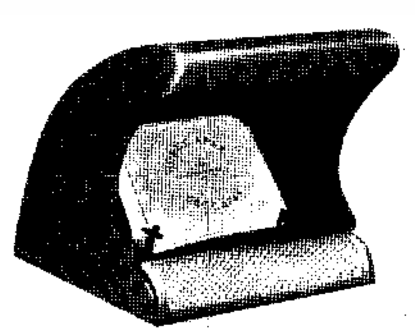
(Illustrated below)

Decca Radar Ltd, 1-3 Brixton Road, London, S.W.9

Deccafax is the name given to a new picture transmission system introduced by Decca Radar Ltd. This low cost system transmits television style pictures and sound between points inter-connected with co-axial cable at distances of up to 2 000 yards.

Based on new applications of television receiver principles, the Deccafax master unit can relay both sound and picture to modified standard television receiving sets for one way announcements etc., or to other master units for two-way speech/vision inter-communication in office, store or factory.

The most interesting feature of Deccafax is its versatility. The transmitted picture or message can take many different forms, ranging from complete 35mm cine-films to simple transparencies of a diagram or pro forma. As an example, announce-



ments could be routed from a single master unit to all or any of a number of television receivers at an airport terminal, these receivers being also used to receive the normal television programmes for entertainment purposes.

Pictures derived from slides or strip films represent an ideal medium for advertisement purposes and Deccafax can relay them to television receivers appropriately placed in large stores etc. Equipment to move the film on or change slides automatically is available.

An important application of the Deccafax system is to meet the requirement of information collation and filtering centres where complicated visual displays are necessary, such as at Air Traffic Control centres and Defence Operations Rooms. A number of remote master units can be inter-connected to feed different types of information to a central receiver for enlargement on to a viewing screen. Control can be exercised over the various inputs so that the main display shows only the specific information necessary at any one time. Filtering in this way makes it far easier for viewers to appreciate the significance of the events which concern them.

The Deccafax system has many other uses. For instance photographic negatives can be reversed and displayed as positives, several pictures can be superimposed on to a single viewing screen and private film shows can be given without using a cinema and without having the noise of a portable projector in the room (there would be no need for total darkness either).

The Deccafax is a low cost, versatile picture transmission system. The standard master unit is small, compact and suitable for table use. Installation is simple and systems can be provided to cover a very wide variety of applications.

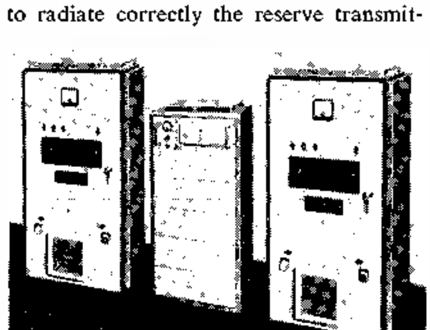
NAVIGATION AND COMMUNICATION EQUIPMENT

Redifon Ltd, Bromhill Road, Wandsworth, London, S.W.18

Aeronautical radio navigation equipment shown by Redifon Ltd, includes the new 350W medium frequency non-directional beacon type G175. The G175 is designed to reduce maintenance time to an absolute minimum; the front-door is opened by a car-type handle to reveal the interior of the transmitter and provide adequate access for all routine maintenance, also, sliding chassis ensure rapid access to all components. Some normally 'difficult-to-reach' components such as the fan motor are mounted in the front door of the G175. *(Illustrated below).*

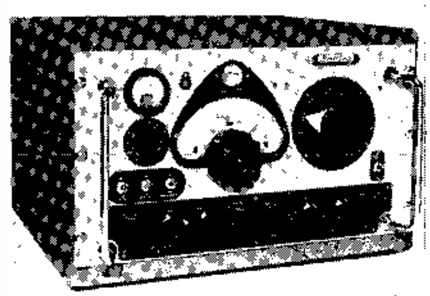
The Redifon G175 can be supplied with alternative modulators for either keyed tone identification or voice modulation and keyed tone operation; voice modulation is used where Meteorological reports are transmitted from the beacon. All service check points can be rapidly monitored by a large scale test meter, all components are of full tropical rating and the beacon can be switched for either local or remote operation.

For navigation beacons where absolute reliability is essential, Redifon offer a dual version of the G175 using two transmitters and an automatic changeover unit. If the transmitter in use fails to radiate correctly the reserve transmitter is automatically switched into circuit and the alarm system operated.



Among the new communication equipment recently developed by Redifon and displayed at Farnborough are the receivers R145 and R150, and the GK185 frequency shift key unit with complete radio teleprinter receiving terminals. The R145 receiver provides precision tuning over the frequency range 2 to 30Mc/s with accuracy and stability. Special features include a selectable sideband unit and an adaptor to extend the frequency range down to 15kc/s. The R150 provides both continuous tuning and crystal controlled spot frequency reception at a modest price. (R145 *Illustrated below*).

The GK185 frequency shift keying unit allows standard Redifon transmitters, and those of other manufacturers, to be converted for radio teletypewriter operation for the aeronautical fixed telecommunica-



tions network (AFTN) or similar communication requirements. The GK185 shown at Farnborough is the commercial version; a special Services version (GK185A) has been purchased in quantity by the Admiralty.

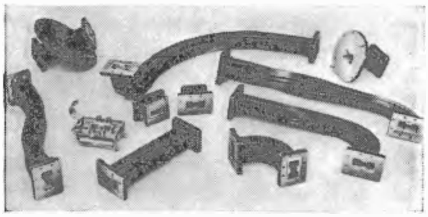
RIDGE WAVEGUIDE COMPONENTS

(Illustrated above right)

W. H. Sanders (Electronics) Ltd, Gunnels Wood Road, Stevenage, Hertfordshire

Due to existing differences of opinion concerning the optimum frequency for airborne weather penetration radar systems, leading manufacturers of such equipment have selected different operating frequencies for their designs, viz: 5·7cm (C Band) or 3·2cm ('X' Band).

An aircraft waveguide run from radio rack to radome should therefore be able to operate at either waveguide length without degrading the performance of the system, particularly where the run is to be installed permanently. To meet these requirements Airtron Inc., Linden, New Jersey, U.S.A., developed a double ridge waveguide of extremely wide frequency range complying with the specifications of characteristic No. 529 of Aeronauti-



cal Radio, Inc. (ARINC). Under licence agreement with Airtron Inc., Sanders are now manufacturing quite a wide range of these components.

Among these are pressurized bulkhead assemblies, straight sections, twists, transistions, 90° and 45° 'E' and 'H' plane bends, flexible ridgeguide, quick disconnects, plain and gasket flanges and r.f. pressure gaskets. Transistions and r.f. unit adapters are available to suit Bendix RDR-1 or RCA AVQ-10 etc. weather radar as required. The ridgeguide bends are based on 2·214in and 7in radii measured to the centre of the waveguide, and the 90° twists are incorporated over a length of 8in. The pressurized bulkheads are tested to 30lb/in and can therefore fully meet air regulation board requirements in this respect. By using these components an aircraft waveguide installation becomes fully adaptable to either 'C' Band or 'X' Band radars, and as a result commercial aircraft so equipped are usable for whichever radar the customer may choose.

Substantial savings in weight and space are also achieved with a ridge waveguide run as this is only slightly larger than the 1in × ½in o.d. guide size used in normal 'X' Band operation. The design of such double ridge waveguide components can thus be easily adapted to suit any aircraft configuration.

COMMUNICATIONS EQUIPMENT

Murphy Radio Ltd, Welwyn Garden City, Hertfordshire

V.H.F. Transmitter/Receiver MR370

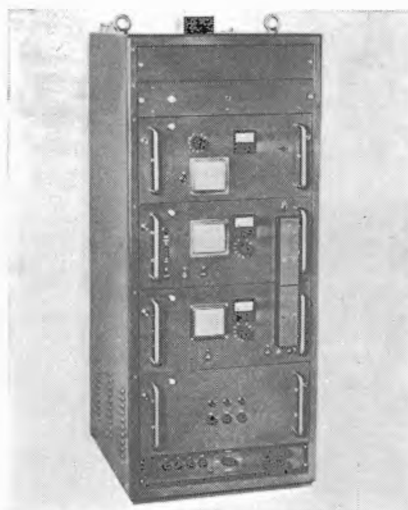
(Illustrated above right)

The equipment has been designed for use as the fixed station in ground-to-air and sea-to-air communications in the frequency band 100 to 156Mc/s. It offers a large number of channels (622 at 90kc/s spacing or 560 at 100kc/s spacing) or 24 preset channels, with rapid channel selection and great flexibility of control. High output power is combined with extremely low spurious radiation.

It comprises four main units, a transmitter/modulator unit TM371, a receiver unit RX372, an exciter unit EC373 and a power unit PU374, all of which fit into 19in racking. The main rack is divided into seven compartments, four of which each accommodate a main unit, and three smaller compartments into which may be fitted auxiliary control units. The main units are mounted on sliding rails which enables them to be withdrawn and tilted through 90° to facilitate servicing. The cabinet normally stands on four anti-vibration mounts, and two further similar mounts are fitted on the top of the cabinet to facilitate mounting in vehicles or ships.

The equipment may be remotely controlled in several ways, the method depending on the circumstances. In the land based version, control may be exercised for short distances over a multi-wire system, or over a two wire system for distances up to 3 miles. In either case all channels are available, or the 24 preset channels. It is also possible to effect simultaneous control of the transmitter and receiver, with either of these units situated at a remote location, from a common control box: in this case separate exciter units must be used.

Channel selection is by means of a decade switching system, and any channel may be selected in 4 to 5sec. Other features include 'ready' signalling on completion of channel selection combined with an alarm signal if the channel change has not been effected, carrier signalling from the receiver, and the possibility of monitoring the speech



under transmission. The exciter unit determines the frequency of operation, controls the overall tuning and contains the servo system controlling the channel selection. It contains the master oscillator and the associated crystal controlled and interpolation oscillators.

The master oscillator is a variable frequency oscillator tuned mechanically by a servo motor system. It has two functions, acting as the first oscillator for the receiver, and also providing a reference frequency with which the transmitter frequency may be compared and stabilized.

The receiver is a conventional double superheterodyne circuit. After r.f. amplification, the signal is mixed with the output from the master oscillator a 1st i.f. of 13Mc/s. This is mixed with the output from a crystal controlled oscillator at 11·3Mc/s, resulting in a final i.f. of 1·7Mc/s. This is followed by a detector, a.g.c. and peak noise limiter circuits, and an a.f. stage delivering 1·5W of audio. Connexion may be made to standard 600Ω lines and/or external loudspeaker if required.

V.H.F. Multi-channel Aircraft Equipment

(Illustrated below)

The MR300 v.h.f. equipment is a transmitter/receiver offering radio telephone facilities on forty-four spot frequencies anywhere in the band 118 to 132Mc/s. It meets all the current requirements for airborne radio installations and is suitable for all types of civil aircraft, being compact, light in weight, fully tropicalized, and capable of operation up to a height of 40 000ft.

The equipment comprises two main units and a control unit. The transmitter power unit, TP301, and the receiver unit, R302, are separately housed in cabinets designed for mounting in standard R.C.E.E.A. racking. A common back plate provides interconnexion between these units and the control unit. The control unit, VF305, is housed in a light alloy case and contains all the operating controls including channel selection switches.

Both the transmitter and the receiver are controlled from the same crystal oscillator and no re-tuning is required when switching from one channel to another. The control unit permits direct channel frequency reading, rapid selection of the required frequency being made by two rotary switches. The first switch is calibrated in 1Mc/s units from 118 to 131Mc/s, and the second switch is calibrated from 0 to 0·9 in units of 0·1Mc/s, thus providing the decimal selection. Rear illumination of the figures by transmitted red light renders the selected frequency easily readable.

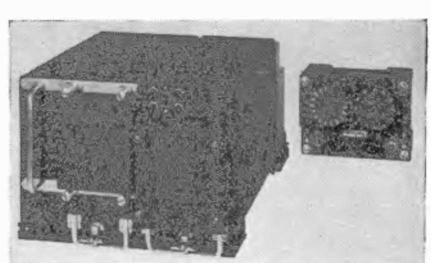
Intercommunication and side-tone facilities are provided and the receiver may feed up to three pairs of headphones.

The transmitter power unit TP301 contains the modulator circuits, and the rotary transformer power supply. The modulator comprises a microphone pre amplifier driving a push-pull modulator which provides series modulation of the anodes and screen gride of the transmitter power amplifier.

The output from the common crystal multiplier is mixed in the transmitting circuits with the output from a further crystal oscillator and passed via further amplifier/multiplier stages to the push-pull p.a. stage, which provides an output of 3/5W into 45Ω at the final frequency.

Power is supplied to the equipment by a rotary transformer fed from a 28V d.c. input. The rotary transformer delivers 300V d.c. at 200 mA.

The receiver is a double superheterodyne circuit employing one r.f. stage, 1st



and 2nd mixer stages, three i.f. stages, diode detector, noise limiter and a.g.c. circuits, and an output stage. The first i.f. is 27·5Mc/s which is converted to 1·95Mc/s by the second mixer stage using a crystal controlled local oscillator.

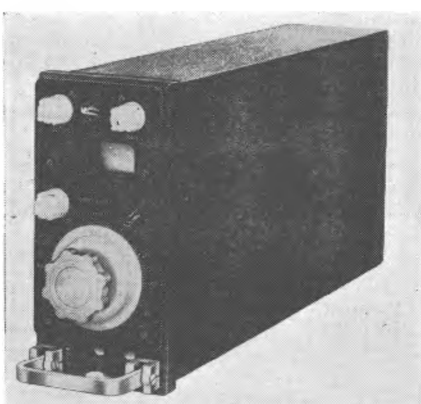
The control unit contains the operating controls including selector switches for forty-four channels. A volume control, on/off switch, dimmer, and an r.t./i.c. switch are fitted.

AIRBORNE SEARCH RADAR

Ekco Electronics Ltd, Southend-on-Sea, Essex

Three types of Search Radar equipment have been introduced, type E120 'X' band equipment as installed in Bristol 'Britannias', a new 'X' band equipment Type E160 intended for the de Havilland Comet IV, the Vickers Vanguard and the Boeing 707 and a prototype 'C' band equipment.

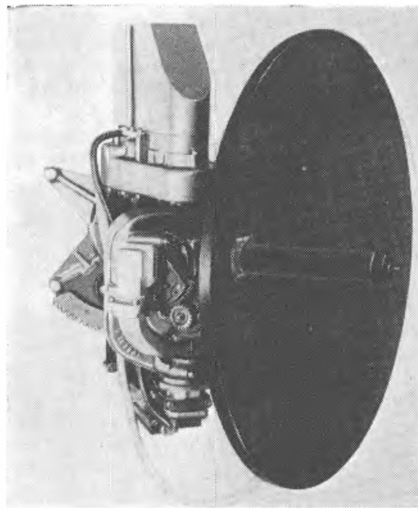
Ekco search radar scans a sector of space ahead of the aircraft, to a maximum range of 120 miles and provides warning on a p.p.i. display, of turbulence within cloud formations. The already established iso-echo contour



feature gives visual distinction on the screen between degrees of turbulence. The equipment also provides land map-painting facilities as an aid to navigation, if required, while the recently introduced Ecko doppler drift measurement unit *(illustrated above)* can be added to facilitate dead-reckoning navigation.

Type E120 3·2cm equipment employs full platform stabilization for scanning unit, with stabilization limits of + 18° to 22° in pitch and ± 45° in roll. Provision is made for tilting the radar beam within limits of + 7° and − 17° and the azimuth coverage of the scan is ± 75° about the aircraft heading.

In the new Type E160 3·2cm equipment, a simpler stabilization system is employed in which pitch stabilization and tilt are combined and cover ± 25°. Roll limits are ± 45° and the scanner sweep is ± 90° about the aircraft heading. Provision is made for switching the pencil beam to an equal-energy-distribution beam to give improved map-painting facilities. *(Scanner Unit illustrated above right).*



The prototype 'C' band equipment displayed, operates on a 5·5cm wavelength and consequently provides a modified response to certain kinds of weather conditions but in general it offers similar facilities to the Type E160 equipment.

ULTRASONIC GENERATOR

Dawe Instruments Ltd, 99 Uxbridge Road, Ealing, London, W.5

The Type 1151 equipment is a high power version of the type 1150 ultrasonic cleaning equipment. It operates at the same frequency and is capable of fully loading eight type 1161/B36 immersible transducers, having a total effective radiation surface of 350². It is suitable for ultrasonic cleaning in tanks of up to 25 gallons capacity.

The generator uses a pulsed oscillator with a relatively small number of heavy duty components, including only one electronic valve. The simple design and rugged construction ensure trouble-free operation under production conditions and simple maintenance.

Brief Specification

- Generator Frequency: Tunable 36 to 40kc/s pulsed at mains frequency.
- Output: High frequency power approx. 1kW average. 4kW peak.
- Mounting: Bench unit.
- Power Supply: 200-250V, 50-60c/s, 2·5kW approx.
- Dimensions: 25in, 17in, 19in high.
- Weight: 1 751lb.
- Cooling: Forced air.

DOUBLE-BEAM OSCILLOGRAPH

A. C. Cossor Ltd, Cossor House, Highbury Grove, London, N.5

The specification of the new oscillograph (model 1035Mk III) is based on that of its predecessor, model 1035Mk II, and a great improvement in performance is claimed.

The instrument is fitted with a new double-beam p.d.a. 4in tube. The A1 amplifier provides an output deflexion on all ranges of not less than 2cm to peak over the range 5c/s to 7Mc/s with a rise-time (in the 50V-150mV sensitivity range) better than 0·1μsec with less than 5 per cent overshoot. An additional terminal is provided so that input voltages of up to 1 500 may be displayed and measured. The maximum gain of the channel is 3 000. The stabilization system employed assures a high order of gain constancy. The A2 amplifier has a frequency response extending from 5c/s to 250kc/s and a switch allows the phase reversal of the display so that both channels may be directly compared. The gain of this channel is 30. The time-base is variable in eight steps from 100msec to 10μsec and a variable delay network is incorporated. It will fire repetitively or by trigger, the potentiometer adjustment for these two modes of operation allowing a very sensitive trigger facility. Provision has been made to trigger the time-base from either line or frame pulses in a composite video signal. Both Y channels are fitted with shift and calibration facilities with a measurement accuracy to ± 10 per cent, and measurement of time along the X axis are carried out in a similar manner to some accuracy. A test waveform of 10V peak-to-peak is provided.

SINGLE-SIDEBAND TRANSMITTER-RECEIVER

Mullard Ltd, Mullard House, Torrington Place, London, W.C.1

Important progress in air communications has resulted from developments in single-sideband systems. One of the most advanced equipments of this type is the Mullard airborne transmitter/receiver (Type X7443) which has been developed in conjunction with the Royal Aircraft Establishment, Ministry of Supply.

Designed to fulfil the need for a long range, pilot-operated r.t. aircraft set, flight tests have shown this equipment to have marked advantages over comparable d.s.b. systems.

The main improvements are in reduced fading and distortion providing increased intelligibility and range. The 300W transmitted power gives an effective range equal to that of a 2·5kW d.s.b. equipment.

Ease of operation is a particular feature and 12 pre-set crystal-controlled channels can be selected, the changeover period being about 15sec only.

The equipment comprises three main units, namely: power supply; power amplifier and s.s.b. unit. These can be mounted in either civil or service type racking and are associated with a number of plug-in sub-units.

The overall weight of the three main units rack mounted and complete with junction box and inter-connecting cables is only 85lb.

An alternative version of this equipment has been developed for fixed station services where medium or short distance point-to-point r.t. communication is required.

LETTERS TO THE EDITOR

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

Accurate Measurement of Rise Time

DEAR SIR,—The usual methods for measuring the rise-time of a transient waveform depend for their accuracy on the linearity of the oscilloscope time-base and upon the accuracy with which the time-base can be calibrated from the timing waveform.

The method to be described eliminates these and other possible errors and enables the rise-time of transient waveforms to be measured accurately.

The waveform to be measured is displayed on the oscilloscope which is triggered by an early trigger pulse from the circuit under test.

The early trigger pulse is also used to trigger a ringing h.f. oscillator whose output is loosely coupled to the X plates of the oscilloscope.

The oscilloscope time-base thus becomes velocity modulated by the oscillations from the ringing oscillator and the transient becomes the time-base for the oscillations. The number of cycles of the oscillation on the relevant portion of the transient waveform may then be counted and, knowing the frequency of the oscillation accurately, the time interval may be derived.

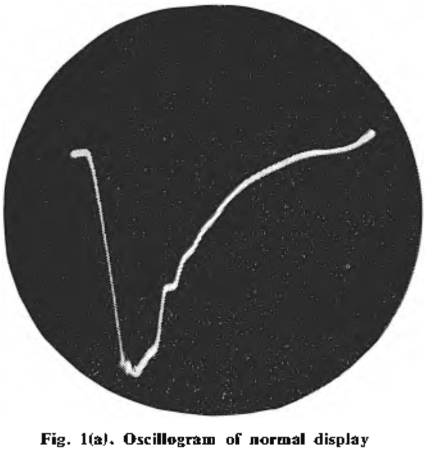


Fig. 1(a). Oscillogram of normal display

Fig. 1(b). Oscillogram of velocity modulated display

Fig. 1 (a) is a normal display of the voltage pulse applied to the cathode of a magnetron. Fig. 1(b) shows the same pulse with the oscilloscope time-base velocity modulated by the ringing oscillator. The frequency of oscillation is 15Mc/s.

It can be seen that the accuracy of measuring rise-time using this method is not dependent upon the linearity of the oscilloscope time-base. The only requirement of the time-base is that it is fast enough to separate the transient to be measured from the rest of the waveform.

When a very accurate measurement of rise-time is required the ringing oscillator may be calibrated against a crystal controlled oscillator. The reading accuracy of the measurement depends upon the frequency of the oscillations, the higher the frequency the greater will be the reading accuracy.

The method described above could be included in the circuits of high speed oscilloscopes as a useful facility. In this case the time-base waveform could be used to trigger the 'built in' ringing oscillator.

The method of measuring rise-time as described in this letter was originally proposed by Mr. C. A. Lea of The General Electric Co. Ltd.

Yours faithfully,

M. R. HARKNETT,
Portsmouth, Hants.

High Current Low-Tension Transistor Stabilizers

DEAR SIR,—In his letter in the July 1957 issue, Mr. van de Stadt seems to have missed a few points in regard to the circuit configurations used in his stabilized power supply.

(1) The use of an auxiliary power supply E_2 for the feedback amplifier is necessary not only to provide a higher gain but also to make the output resistance independent of R_i. Since variations in E_2 cause variations of output voltage, it is preferable to stabilize E_2 against variations in mains input voltage.

(2) One of the chief advantages of using the long-tailed pair as a d.c. amplifier is that the variation of 2·5mV/°C in the base-emitter voltage of the two transistors causes negligible variation in the voltage between the two bases. In addition, if the resistances of the base circuits are low, differences in the value of I_{co} of the two transistors will also cause little variation in the voltage appearing between the two bases.

(3) In considering the current sharing of transistors connected in parallel, the most important factor is the spread in base-emitted voltage for a given collector current. For this reason, to ensure adequate current sharing stabilizing resistors should be connected in series with the emitters. This is in fact what is stated in the reference quoted. For maximum effect, the voltages developed across the emitter resistors should be large compared with the spread in base-emitter voltage of the transistors at the required current.

Due to the high collector resistance of the transistors, the use of resistors in series with the collectors will have negligible effect, and the 1Ω resistors in series with the bases will be almost completely swamped by the base resistance of the transistors themselves.

Yours faithfully,

W. L. STEPHENSON,
Mullard Research Laboratories,
Salfords.

The Correspondent Replies:

DEAR SIR,—I would like to make the following comments regarding Mr. Stephenson's valued remarks:

(1) It is, of course, correct, that addition of a second voltage supply, E_2 does make the output resistance independent of R_i. The matter of stabilization of E_2 has been looked into, but as our specifications (power supply for a digital computer) did call for an output variation of 1 per cent for half load to full load, it was not found necessary to stabilize E_2 for a mains variation of 10 per cent. This is possibly also due to the 'common mode rejection' of the long tailed pair circuit.

(2) In the fourth paragraph of my letter of July 1957 the offset of effects of temperature, and more implicitly, of differences in I_{co}, are indicated.

(3) Equal current sharing of transistors in parallel can indeed be realized better in the manner indicated by Mr. Stephenson. However, then a larger signal voltage should be applied between the output line and the bases of the stabilizing transistors. In the case of the stabilizer of Fig. 3 delivering its rated 5A, each series transistor should nominally carry 1·25A. Then its emitter to collector voltage is about 2V, which results in a internal resistance for this transistor of about 1·6Ω. The resistors of 1·67Ω then do assure adequate current sharing in this application, the collector currents being equal within about 15 per cent for 4OC16's randomly selected.

The remark regarding the 1Ω base resistances is quite correct; in the ultimate version these have been omitted.

Yours faithfully,

W. VAN DE STADT
Computer Laboratory,
Mathematisch Centrum, Amsterdam.

BOOK REVIEWS

The Theory of Networks in Electrical Communication and Other Fields

By F. E. Rogers. 560 pp. 341 figs. Royal 8vo. Macdonald and Co. (Publishers) Ltd., London. 1957. Price 65s.

DIETZOLD has identified three distinct phases in the history of network theory: a decade of infancy, a decade of adolescence, and a period of maturity. The first phase began with the development of the discrete element analogue of the continuous line and came to an end when interest shifted from the network as a frequency selective device to the network as a dynamical system. At the end of the first phase the subject was encumbered with catalogued network responses and network theorems. In the second phase, network theory became more like mathematical physics. The image parameter approach of the first decade was found to be unsuitable for the study of more general systems and network theory was remodelled on an energy basis. Two major aspects emerged: analysis and synthesis. The first two phases were dominated by steady-state frequency response considerations but the time response of networks became important with the advent of electro-optical devices. Hitherto, the study of transients was regarded as a separate activity and belonged to circuit theory. The recognition of the relationship between the frequency and time responses led to a union of these two facets of network behaviour and network theory was unleashed from the Kennelly-Steinmetz sinusoidal stead-state and entered the phase of maturity.

This new book by Mr. F. E. Rogers of the Regent Street Polytechnic has as its purpose the presentation of network theory of the kind required by final-year engineering degree or diploma students. Accordingly, the book dwells largely in what Dietzold identified as the first decade. After an introductory chapter in which the needs of the heavy-current and the light-current engineer are compared there follows two chapters headed "The Foundations of Network Theory" and "The General Theory of Multi-mesh Networks". For this reviewer, the treatment of these vital topics is disappointing. The main foundation of modern network theory is the science of network configuration (network topology). A secure grasp of the principles of this aspect of network theory is essential if the student is to formulate his problems without doubt and uncertainty but the author disposes of this fundamental issue with little more than the vague remark that the " closed paths are known as meshes, and the periphery of a mesh is called the mesh contour ". With such loose definitions the difference between a mesh and its contour is difficult to see whereas, in fact, they are distinct basic ideas. A little tuition in the elementary notions of linear graph theory would reveal to the student the real significance of the electric network problem and equip him to approach network analysis with confidence and understanding.

The author proceeds next to the laws of Ohm and Kirchhoff and gives some simple examples of the Kirchhoff method of analysing resistance networks. The treatment is not as clear as it might be; for example, nothing explicit is said about linear independence of the current equations. Similarly, the frequent reference to the preliminary prediction of branch current directions is misleading; the determination of these directions is one of the objects of the analysis. The integro-differential equations of some simple circuits and the use of complex numbers (called vectors) in the solution of sinusoidal steady-state problems are considered next and the discussion concludes with an indication of the use of the Fourier integral.

The third chapter introduces the Maxwell circuited method of analysis. The topological relation giving the number of independent circuits in terms of the number of branches and nodes of a connected network is stated but no suggestion is made as to how an independent set of circuits can be found in a complicated configuration. Presumably, to borrow from Guillemin, it is taken for granted that the student will straighten this " obviously simple " matter out for himself. In the discussion of circuital currents the author states that, although the current directions are arbitrary, they must be the same throughout the network. This is certainly not so. How is the remark to be interpreted in the case of an intricate non-planar network? The use of the circuital method of deriving the integro-differential network equations is explained for the simple case of two circuits coupled magnetically but the extension to the general case is restricted to the single-frequency sinusoidal steady-state. The Maxwell nodal method of analysis is not mentioned.

The remainder of the book settles down to a thorough-going account of first decade steady-state network theory. The fourth chapter is devoted to network theorems. All the theorems are proved and adequately illustrated by specific examples including one proposed by the author and concerned with the superposition of powers. The following five chapters deal with conventional network structures, equivalent networks, two-terminal configurations, Foster networks, inverse networks, and networks with constant impedance. Transmission along uniform lines and cables and the classical theory of four-terminal networks in iterative and image connexion are considered in detail. During the discussion of the latter an analytical device invented by the author is introduced by which the image impedance of a four-terminal network can be changed without affecting its image transfer function. Insertion loss and impedance matching are treated in the tenth chapter and the eleventh chapter is concerned with the principles of wave-filter theory. The final chapter deals with steady-state network measuring techniques.

The book is well planned and bears evidence of careful preparation. Worked examples, many of which are taken from the London University B.Sc. examinations, append most of the chapters. The typography is excellent and the pages have a pleasant appearance. In recommending the book to anyone pursuing a course in line communication networks one cannot suppress the feeling that an opportunity to provide young engineering undergraduates with a sound introduction to modern network theory has been missed.

S. R. DEARDS

Plant and Process Dynamic Characteristics

Proceedings of a Conference sponsored by the Society of Instrument Technology at Cambridge. April 1956. 246 pp. Demy 8vo. Butterworth Scientific Publications, 1957. Price 50s.

IN April last year, the Society of Instrument Technology organized a conference on Plant and Process Dynamic Characteristics. This was a unique arrangement in so far as this country is concerned. The lecturers contributing to the conference came from distinguished industrial companies in this country, such as I.C.I. and British Petroleum, together with men of similar experience from Holland. In addition, Universities and Industrial Research Associations were able to provide experts to read papers. Following a conference which covered such an important field it is not surprising to find that there has been a demand for a book which recorded not only the lectures themselves, but all essential parts of the discussion which followed each paper.

Although fundamentally the lectures were biased towards chemical engineering, exceptional care has been taken to give adequate cross references in the bibliography so that electronic servo engineers, who are not necessarily familiar with the techniques of process engineering, will still be able to take an interest in these recorded proceedings of the conference. It must, however, be stressed to electronic engineers that the lecturers were not themselves electronic engineers, but were experts engaged in fundamental research on the dynamics of processes, and men very experienced in measurement and control of chemical plants actually in operation. Because this book is basically a collection of papers which these experts read, it must of course be realized that it does not provide, nor does it pretend to provide, a systematic survey of this very wide field.

There is, however, a paper on Data Reduction Equipment which is of great interest to electronic engineers. It not only discusses the functional requirements of such equipment, but also describes actual electronic and electromechanical devices which meet these needs. Another paper by three members of the Wool Industries Research Association gives an interesting application of electronic measuring and controlling equipment for a woollen carding engine.

J. R. BOUNDY

Transistors Handbook

By William D. Bevitt. 410 pp. 380 figs. Demy 8vo. Prentice-Hall, New York, Bailey Bros. & Swinfen, London. 1957. Price 63s.

THE production of a handbook on such a rapidly growing subject as transistor applications presents problems to the book publisher. If the author organizes himself an information service, to make sure the manuscript is up to date when submitted, and makes additions right up to the last possible date before actually going to press, the book will still be "out of date" when it appears. It will be even more out of date by the time the reviewer can scrutinize it, and the review appears in print. But such books are needed, despite these problems, to provide the best reference possible under the circumstances.

In this one, Mr. Bevitt has collated a vast amount of material from a wide range of publications covering his subject, and has presented a concise précis of these articles, well illustrated. He has arranged the book in a progressive sequence, from fundamental definitions and concepts, through the various types of transistor and methods of measuring them, to applications: audio and radio frequency amplifiers and oscillators, a.m. and f.m. detection (demodulation), radio and television receivers, relaxation oscillators, and finally computer applications.

As he has observed good proportion in his précis, the coverage is approximately proportional to the work done (at his dead-line) in each area. The book includes a couple of useful appendices, giving the standard terms (IRE) laid down for the industry (1954) and listing essential characteristics of a number of transistor types. The latter table only includes products by American manufacturers, and is by no mean up to date with them. But it has the useful feature of identifying the characteristics of transistors referred to throughout the text of the book.

Not the least valuable feature is the complete bibliography (again to the author's dead-line) presented at the end of each chapter.

NORMAN H. CROWHURST.

Digital Calculating Machines

By G. A. Montgomerie. 256 pp. 40 figs. Demy 8vo. Blackie & Sons Ltd. 1957. Price 30s.

THIS is the first book among four on computers, written about British and European machines, that the reviewer has had the pleasure of reading.

Unlike its American counterparts, which accent the large data processing electronic installations, this book describes all mechanical aids to computation from the abacus used by the Chinese since the 19th century, through key-response adding machines, to the present day Sequence Programmed electronic calculators.

The author has categorized the large range of machines into types and describes at length both the mechanical configuration and the arithmetical logic behind some of the machines in each category.

More than three quarters of the book is devoted to mechanical and electromechanical machines (i.e. desk and punched card) and only the latter quarter to, what the author calls, Group IV Electronic Machines.

Though allowances should be made for the reviewer being favourably biased because the book includes descriptions of electronic machines he designed, potential readers can rest assured that they will find in this book a great deal of useful and detailed information, clearly set out, well illustrated, and what is most important in books covering such a large range, ample references.

W. WOODS-HILL

Short News Items

The Council of the British Institution of Radio Engineers has announced the awards of premiums to authors of papers which were published in the Journal of the Institution during 1956. The senior award, *the Clerk Maxwell premium*, has been made to Dr. K. D. Froome for his paper "Microwave Determinations of the Velocity of Light". Dr. Froome is with the Metrology Division of the National Physical Laboratory. *The Heinrich Hertz premium*, which is awarded for the most outstanding paper dealing with the mathematical or physical aspect of radio, goes to Dr. A. G. Edwards for his paper "The Effects of Atmospherics on Tuned Circuits". Dr. Edwards is at present with the A.E.I. Research Laboratories. *The Brabazon premium* for the most outstanding contribution on radio and electronic devices for aircraft safety has been awarded to Mr. K. E. Harris, Director of Research at Cossor Radar Ltd., for his paper "Some Problems of Secondary Surveillance Radar Systems". *The Louis Sterling premium* for the most outstanding paper on television technique has been awarded to Dr. A. van Weel for his paper "Some Remarks on the Radio-frequency Phase and Amplitude Characteristics of Television Receivers". Dr. van Weel is with Philips Research Laboratories, Eindhoven, Holland. *The Marconi premium* for the most outstanding engineering paper has been awarded to Professor P. M. Honnell for his paper "Prescribed Function Vibration Generator". Professor Honnell is in the Department of Electrical Engineering, Washington University, St. Louis, U.S.A. *The Dr. Norman Patridge Memorial Award* dealing with improvements in the quality of sound reproduction will be made to Mr. H. J. Leak for his paper "High Fidelity Loudspeakers; the Performance of Moving Coil and Electrostatic Transducers".

Huddersfield Technical College, Department of Electrical Engineering, have arranged a series of ten lectures entitled Principles of Closed-Loop Control, to be held weekly from 9 October. Further details may be obtained from the Head of the Department of Electrical Engineering, Huddersfield Technical College, Huddersfield.

British Railways North Eastern Region are shortly to undertake experiments in the use of radio-telephony in their collection and delivery services. In order to test the possibilities, a contract has been placed with Pye Telecommunications Ltd of Cambridge for radio-telephone equipment which is to be installed on a number of goods delivery vehicles at Newcastle, and also on a number of parcels delivery vehicles at Leeds.

The Scientific Instrument Manufacturers' Association is holding its Sixth Annual Convention at Eastbourne from 24-27 October.

The British Scientific Instrument Research Association announce that the interest shown in their forthcoming series of open days, originally planned for the week 7-11 October, has been so great that the period has now been extended to include Monday, 14 October. Admission is by invitation only and enquiries should be addressed to the Director, SIRA, Southill, Elmstead Woods, Chislehurst, Kent.

Borough Polytechnic, Borough Road, London, S.E.1, announces a course of twenty lectures on Transistors and Allied Devices, to be given on Tuesday afternoons and evenings beginning 8 October. The charge for the course is £2 10s. and further information may be obtained from the Secretary.

A Course of Six Lectures on The Technique of Technical Writing will be given by Mr. G. Parr at Borough Polytechnic, commencing on Friday, 11 October. The course fee is 10s.

Twickenham Technical College has arranged a course of twenty-two special lectures on Pulse Circuit Design commencing on Thursday, 10 October. The fee for the course is £2 and further information may be obtained from the Principal, Twickenham Technical College, Egerton Road, Twickenham, Middlesex.

The English Electric Co Ltd have announced that five DEUCE computers will be in Government service by next spring. The latest order for a machine has been placed with the English Electric Co by the Mechanical Engineering Research Laboratories, East Kilbride, which is part of the Department of Scientific and Industrial Research. The D.S.I.R. already have a DEUCE working at the National Physical Laboratory, Teddington, and there are two installed at the Royal Aircraft Establishment, Farnborough, and one at the Atomic Weapons Establishment, Aldermaston. The D.S.I.R. is the third organization to purchase a second DEUCE since, in addition to the twin installation at Farnborough, there are also two machines at work for the Bristol Aeroplane Co. Six DEUCE computers are engaged in work connected with the aircraft industry, with two more to be delivered this year. The most recent installations of DEUCE have been made at the London offices of the British Petroleum Co and at Short Bros and Harland, Belfast.

Automatic Telephone & Electric Co Ltd have received an order for two 3-channel carrier telephone systems which are being provided for training purposes under the Colombo Technical Aid Plan, sponsored by the Commonwealth Relations Office. The equipment, which is being supplied to the Pakistan Posts and Telegraphs Department through the Crown Agents in London, will bear special plaques recording that they have been provided under the Colombo Plan. Several systems of this type have been in use for a number of years in Pakistan.

British Sarozal Ltd have supplied a 5kW broadcast transmitter with a complete studio installation to Paia, Cape Verde Islands. The station will be operated by the Portuguese Government sponsored Radio Clube of Cape Verde. British Sarozal Ltd have developed this transmitter to give an overall frequency response of ± 2dB between 30c/s and 15kc/s with distortion below 4 per cent. These characteristics apply to the complete system including studio equipment. The output valves for the transmitter have been developed by Siemens Edison Swan in conjunction with British Sarozal Ltd.

The British Tabulating Machine Co Ltd have received an order from the Durban City Council for a Hollerith Type 1201 electronic computer. The computer will be used for producing light and telephone accounts, at a rate of 1 000 an hour, and will prepare municipal employees' paysheets and leave records.

A subsidiary plant of Texas Instruments Inc is opening this month at Dallas Road, Bedford. The British plant and company will be known as Texas Instruments and will soon be in full production in the manufacture and marketing of Silicon semi-conductor products in Great Britain. Mr. Dudley Saward has been appointed Managing Director of Texas Instruments Ltd.

Fielden Electronics Ltd have opened a new north eastern sales and service office at Stockton-on-Tees.

Mr. H. E. Cornish, O.B.E., M.C., former Postmaster General of Sarawak, has been appointed General Manager of British Telecommunications Research Ltd, the research establishment jointly sponsored by Automatic Telephone and Electric Co Ltd and British Insulated Callender's Cables Ltd, at Taplow, Bucks.

Notes from ———

NORTH AMERICA

Thermo-Compression Bonding

Research by O. L. Anderson, H. Christensen and P. Andreatch (at Bell Laboratories) has shown that a combination of heat and pressure can be employed to provide a firm bond between various soft metals and clean, single crystal semiconductor surfaces.

Called thermo-compression bonding, this new technique provides a bond that is stronger than the lead itself. Temperatures and pressures required are not high enough to affect the electrical properties of the semiconductor material.

One method of forming a suitable bond is to employ a heated element such as a wedge, a flat or a point, to press the metal against the heated semiconductor with a pressure sufficient to cause a slight deformation of the lead. Adhesion occurs within a matter of seconds.

Another useful connexion may be made by butting the balled (or headed) end of a wire against the heated semiconductor by means of a capillary tube.

Thermo-compression bonding has a number of advantages over other methods of attaching leads to semiconductors. The bond is stronger; the technique is more readily adaptable to mass production; no chemical flux or other chemical contaminant is involved in the process; and leads may be attached to much smaller areas, an invaluable aid in fabricating high-frequency transistors.

Adhesion takes place in seconds with pressures of a few thousand pounds per square inch and temperatures well below the melting point of the alloy of the metal and semiconductor. A gold-germanium bond appears to be the easiest to make, but gold, silver, aluminium and a number of alloys can be readily bonded to either germanium or silicon.

Intensive investigations are underway to determine the mechanism of this bonding process, and to further study its practical applications. The technique is already being employed on a laboratory basis in the fabrication of transistors.

Microwave Ferrites

New ferrite materials having properties particularly well adapted for a broad range of microwave applications have been developed by L. G. Van Uitert of Bell Telephone Laboratories. The desirable properties include controllable saturation magnetization, low dielectric loss, and a high degree of reproducibility.

The new materials are essentially magnesium, manganese, aluminium ferrites or nickel manganese ferrite with a small amount of copper replacing some of the magnesium or nickel. The addition of the proper quantities of copper and manganese to the basic ferrite is advantageous from several points of view. By increasing the reactivity of the mixture, copper decreases the necessary firing temperature by at least 100°C. Under comparable conditions this results in lower porosity and improved uniformity in the fired material. The manganese addition decreases electrical conductivity and thus the dielectric losses in these low porosity materials.

Microwave ferrites with low saturation magnetizations are obtained by the modification of magnesium ferrite. The saturation magnetization of this ferrite can be decreased in a controlled way by substituting aluminium for a part of the iron. While materials compounded in this fashion are basically satisfactory, their refractory nature makes it difficult to reproduce the magnetic properties required for many microwave applications. The added copper minimizes this difficulty, and also increases slightly the Curie temperature for comparable saturation magnetization.

Microwave measurements on these new ferrites indicate that in general they are comparable to or better than similar materials containing no copper. Applications include microwave gyrators, isolators, and similar devices.

Fusion Research

It was recently announced by Dr. Suits, vice-president and director of research of General Electric, that the G. E. is conducting a substantial research programme in the study of atomic fusion and the production of power the fundamental process of fusion as used in the hydrogen bomb. Dr. Suits forecasts, however, that it will need some twenty years of research effort before pilot-plant production of fusion power could begin.

Galvanometer Amplifier

A new and highly sensitive galvanometer amplifier, incorporating three cut-off filters, has recently gone into production at Nevada Air Products Company.

Features include a frequency range of 2c/s to 10kc/s, an input impedance of more than 100Ω and a gain of 85dB, together with an overall physical size of only 8in × 4in × 5in high.

The unit will handle large input signals of up to 15V without overloading the stages preceding the filter networks. Output voltage across 55Ω is 1·5V at 2c/s, with any of the filter networks in circuit. Noise and ripple is less than 0·5 per cent of rated output and, if still higher output voltage is required, the unit may be switched to provide characteristics of 4V across 55Ω between frequency limits of 5c/s to 10kc/s, ± 0·5dB.

A high impedance monitor output is also provided with 5V r.m.s. into a 100Ω resistive load. Total harmonic distortion is less than 1 per cent, and total current drain is less than 35mA. An external voltage regulated power supply is required.

The controlled frequency cut-off characteristics are as follows:

1kc/s filter position...2c/s to 1kc/s $\pm\frac{1}{2}$dB
5kc/s filter position...2c/s to 5kc/s $\pm\frac{1}{2}$dB
10kc/s filter position...2c/s to 10kc/s $\pm\frac{1}{2}$dB

Proceedings of RETMA Symposium

The proceedings of the second RETMA symposium on applied reliability have now been published and are available from Engineering Publishers, G.P.O. Box 1151, New York, 1, price $5.

This Proceedings contains the full versions, with illustrations, of 14 papers delivered at the Second RETMA symposium on Applied Reliability held in Syracuse, New York on June 10 and 11, 1957. The basic theme of the Symposium was the achievement of reliability in electronic equipment by emphasis on mature design. For this purpose mature design may be taken to mean that a design has "matured" when it has reached the limit of its development allowed by current knowledge, materials, and techniques.

The four technical sessions were entitled: Selection and Use of Components in Mature Design; Principles and Techniques of Mature Mechanical Design; Measurement and Proof of Mature Design; and Case Histories of Mature Equipment and Systems.

Information and Sample Service

Madison Trading Co. Inc. has announced the formation of a new Division to be known as 'Samples Services Division' which will have its offices at 274 Madison Avenue, New York City.

The function of this Division will be to aid foreign manufacturers and technicians in procuring technical information about and samples of American products.

It was announced recently that the Company would welcome the opportunity to represent foreign manufacturers and through its resources would be able to procure samples of new American products to forward abroad.

Full details of this service can be obtained by writing for Bulletin FP 937 which the Company would be prepared to forward to interested parties.

Ultra Wide Band Amplifier

Spencer-Kennedy Laboratories have recently introduced a new amplifier having a stable gain of 20dB over the bandwidth of 600c/s to 320Mc/s. The amplifier utilizes a chain, or travelling wave, circuit. The nominal gain of the amplifier is 24dB, it has linear phase shift and a rise-time of less than 0·002μsec.

Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

North Western Section

Date: 3 October. Time: 6.30 p.m.
Held at: Reynolds Hall, College of Technology, Sackville Street, Manchester, 1.
Lecture: High Quality Sound Equipment.
By: K. Davin and F. C. Gibson.

Scottish Section

Date: 10 October. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow.
Lecture: Reception Problems of Scottish Television.
By: W. Boyd.

West Midlands Section

Date: 9 October. Time: 7.15 p.m.
Held at: Wolverhampton and Staffordshire Technical College, Wulfruna Street, Wolverhampton.
Lecture: Transistors, Circuits and Applications.
By: M. D. Cooper.

South Midlands Section

Date: 4 October Time: 7 p.m.
Held at: Gloucestershire Technical College, Cheltenham.
Lecture: Radio Engineering.
By: C. T. Lamping.

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings, unless otherwise stated, will be held at the Institution, commencing at 5.30 p.m.

Ordinary Meeting

Date: 3 October.
Inaugural Address as President.
By: T. E. Goldup.

Measurement and Control Section

Date: 8 October.
Chairman's Address.
By: H. S. Petch.

Date: 22 October.
Lecture: The Measurement of High Voltages with Indicating or Recording Instruments.
By: G. W. Bowdler.

Informal Meeting

Date: 14 October.
Discussion on The Social Consequences of Automation.
Opened by the President.

Radio and Telecommunication Section

Date: 16 October.
Chairman's Address: Some Radio Aids for High-Speed Aircraft.
By: J. S. McPetrie.

Date: 28 October.
Informal Meeting: Domestic High-Fidelity Reproduction.
Talk by: J. Moir.

East Midland Centre

Date: 8 October. Time: 6.30 p.m.
Held at: Loughborough College.
Chairman's Address and Annual General Meeting.
By: J. D. Pierce.

Cambridge Radio and Telecommunication Group

Date: 8 October. Time: 6.30 p.m.
Held at: The Cambridgeshire Technical College.
Chairman's Address.
By: N. C. Rolfe.

Date: 29 October. Time 8 p.m.
Held at: The Cavendish Laboratory, Free School Lane, Cambridge.
Informal Evening: The Importance of Research in Hearing and Seeing to the Future of Telecommunication Engineering.
By: E. C. Cherry.

East Anglian Sub-Centre

Date: 7 October. Time: 7.30 p.m.
Held at: Assembly House, Norwich.
Chairman's Address.
By: G. E. Middleton.

Date: 22 October. Time: 6.30 p.m.
Held at: The Technical College, Cambridge.
Lecture: Engineering Education in the Soviet Union.
By: G. S. C. Lucas.

North-Eastern Centre

Date: 14 October Time: 6.15 p.m.
Held at: The Neville Hall, Newcastle upon Tyne.
Chairman's Address.
By: T. W. Wilcox.

Date: 28 October. Time: 6.15 p.m.
Held at: The Neville Hall, Newcastle upon Tyne.
Lecture: 138kV Submarine Power Cable Interconnection between the Mainland of British Columbia and Vancouver Island.
By: T. Ingledow, R. M. Fairfield, E. L. Davey, K. S. Brazier and J. N. Gibson.

North-Eastern Radio and Measurements Group

Date: 21 October. Time: 6.15 p.m.
Held at: King's College, Newcastle upon Tyne.
Chairman's Address.
By: W. Gray.

North Midland Centre

Date: 1 October. Time: 6 30 p.m.
Held at: The Central Electricity Authority, Yorkshire Division, 1 Whitehall Road, Leeds.
Chairman's Address.
By: A. J. Coveney.

Date: 10 October. Time: 6.30 p.m.
Held at: Yorkshire Electricity Board, Ferensway, Hull.
Lecture: A New Meter for the kVA Demand Charge.
By: P. Baxter.

Sheffield Sub-Centre

Date: 16 October. Time: 6.30 p.m.
Held at: The Grand Hotel, Sheffield.
Chairman's Address.
By: O. I. Butler.

Date: 30 October. Time: 7 p.m.
Held at: The Technical College, Scunthorpe.
Lecture: The Construction of the 275kV Transmission Lines.
By: J. D. Pierce.

North-Western Centre

Date: 1 October. Time: 6.30 p.m.
Held at: The Engineers' Club, 17 Albert Square, Manchester.
Chairman's Address.
By: F. R. Perry.

North-Western Radio and Telecommunication Group

Date: 9 October. Time: 6.45 p.m.
Held at: The Engineer's Club, 17 Albert Square, Manchester.
Chairman's Address.
By: K. J. Butler.

North Lancashire Sub-Centre

Date: 9 October. Time: 7.15 p.m.
Held at: North Western Electricity Board Demonstration Theatre, 19 Friargate, Preston.
Chairman's Address.
By: H. G. Cope.

North Scotland Sub-Centre

Date: 4 October. Time: 7.30 p.m
Held at: Robert Gordon's Technical College, Aberdeen.
Centre Chairman's Address.
By: E O. Taylor.

Date: 17 October. Time: 7 p.m.
Held at: Electrical Engineering Department, Queen's College, Dundee.
Centre Chairman's Address.
By: E. O. Taylor.

South-East Scotland Sub-Centre.

Date: 1 October. Time: 7 p.m.
Held at: The Carlton Hotel, North Bridge, Edinburgh.
Centre Chairman's Address.
By: E. O. Taylor.

Date: 15 October. Time: 7 p.m.
Held at: The Carlton Hotel, North Bridge, Edinburgh.
Sub-Centre Chairman's Address.
By: J. Mendelson.

South-West Scotland Sub-Centre

Date: 2 October. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow, C.2.
Centre Chairman's Address.
By: E. O. Taylor.

Date: 15 October. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow, C.2.
Sub-Centre Chairman's Address.
By: A. J. Small.

South Midland Centre

Date: 7 October. Time: 6 p.m.
Held at: The Grand Hotel, Birmingham.
Chairman's Address, Annual General Meeting and Conversazione. (Admission by ticket)
By: L. I. Tolley.

South Midland Radio and Measurements Group

Date: 28 October. Time: 6 p.m.
Held at: James Watt Memorial Institute, Great Charles Street, Birmingham.
Informal Evening on: Electronics and Automation—the Use of Nucleonic Devices.
By: Denis Taylor.

Southern Centre

Date: 2 October. Time: 6.30 p.m.
Held at: The C.E.A. Portsmouth.
Chairman's Address.
By: L. G. A. Sims.

Date: 16 October. Time: 6.30 p.m.
Held at: The Brighton Technical College.
Lecture: Electrical Floor Warming.
By: J. W. Moule and W. M. Stevenson.

Date: 25 October. Time: 6.30 p.m.
Held at: The South Dorset Technical College, Weymouth.
Lecture: The Transatlantic Telephone Cable.
By: R. J. Halsey.

Western Centre

Date: 14 October. Time: 6 p.m.
Held at: The South Wales Institute of Engineers, Cardiff.
Chairman's Address.
By: J. F. Wright.

Western Supply Group

Date: 21 October. Time: 6 p.m.
Held at: The South Wales Institute of Engineers, Cardiff.
Chairman's Address.
By: C. W. Priest.

South-Western Sub-Centre

Date: 17 October. Time: 3 p.m.
Held at: The Electric Hall, Union Street, Torquay
Chairman's Address.
By: C. H. Foulkes.

West Wales (Swansea) Sub-Centre

Date: 10 October. Time: 6 p.m.
Held at: Conference Rooms, South Wales Electricity Board Showrooms, The Kingsway, Swansea.
Chairman's Address.
By: I. G. Evans.

THE TELEVISION SOCIETY

Date: 4 October Time: 7 p.m.
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.
Lecture: Recent Investigations into the Operation of Image Orthicon Camera Tubes.
By: R. Theile.
Date: 17 October. (Time and place as above).
Discussion: Servicing Modern Television Receivers.
Date: 31 October. (Time and place as above).
Lecture: Performance of Television Receiver Turret Tuners.
By: K. H. Smith.

PUBLICATIONS RECEIVED

PROCEEDINGS OF THE RETMA SYMPOSIUM ON APPLIED RELIABILITY include all the papers delivered at this symposium in Los Angeles in December 1956, sponsored by the Engineering Department of the Radio-Electronics-Television Manufacturers' Association. In addition to the papers, there is a supplement entitled "A General Guide for Technical Reporting of Electronic Systems Reliability Measurements". Engineering Publishers, GPO Box 1151, New York 1, U.S.A. Price $5·00.

TRANSISTOR ENGINEERING REFERENCE BOOK by H. E. Marrows. The author of this book has brought together information on the commercial aspects of the industry. The contents includes five sections under the following headings; General Survey of Transistors; Reference Data on Commercial Transistors; Reference Data on Commercial Transistor Components and Test Sets; Reference Data on Commercial Applications of Transistors; and a Manufacturers' Directory. John F. Rider Publisher, Inc., New York. Chapman & Hall Ltd. 37 Essex Street, London, W.C.2. Price 80s.

LE CALCUL DES TUYAUTERIES A HAUTE TEMPERATURE by A. Gage is a book which will be of considerable value to engineers working on tubing systems for carrying steam, gas or high temperature liquids. The author offers a method of resolving the problems of metal fatigue by the use of electronic calculating machines. The second half of the book deals with the practical applications and contains many relative tables and diagrams. Dunod Editeur, 92 rue Bonaparte, Paris vie. Price 2 900F.

ATOMIC POWER FOR PEACE GLOSSARY OF COMMON ATOMIC TERMS AND WHAT THEY MEAN has recently been published by Atomic Power Review, 414/416 Elizabeth House, 18 Pritchard Street, Johannesburg, South Africa. It is written not only for engineers in the field of nuclear engineering, but also for all non-scientific readers interested in the use of atomic energy for peaceful purposes.