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Commentary

WHEN, in 1939, the television service of the BBC closed down at the outbreak of war, there were about 20 000 television receivers in operation in this country, confined mainly to the London area.

Today, the figure for the country as a whole has risen to over eight million and it is still growing.

This is one of a number of interesting items contained in a broadsheet* recently published by Political and Economic Planning and entitled "Television in Britain". Without comment on the results it has obtained, the document sets out some of the facts behind the phenomenal growth of television, together with the characteristics and viewing habits of the television public, the development of commercial television, and the public attitude towards it.

One of the difficulties which is immediately apparent on reading the broadsheet is the absence of any reliable yardstick by which the growth of television can be measured. The number of television receivers in a country is one indication, the number of television transmitters is another. The countries with the large densely populated areas tend to have the larger number of receivers, but they require, of course, a larger number of transmitters. Countries of small area and populations, on the other hand, are able to provide a greater coverage—almost complete in some cases—with relatively few transmitters. How, then, does one compare the growth of television in the smaller countries, for example, Monaco, with countries as vast and populous as the United States and the U.S.S.R.?

The broadsheet attempts to overcome this difficulty by measuring the growth of television by the somewhat arbitrary ratio of "number of persons per television receiver". It lists the countries of the world in which there is a regular television service and arrives at this ratio by dividing the population by the number of television receivers in the country.

But, whatever standards are chosen, it is only to be expected that the greatest growth would be in the United States. With an estimated forty million receivers—more than *twice* the total number in the rest of the world—and a daily transmission period of 19 hours—the United States heads the world with a ratio of four persons per receiver. Hawaii comes second with a ratio of five.

With nearly nine million television receivers, Britain now ranks third with Canada with a ratio of six but, even so, most other European countries fall far below these levels.

(For those who are interested in what goes on at the other end of the scale, it may be worth while bringing to their notice that at the very bottom is S. Korea with only one television receiver for every 8 400 of its population. In Europe, this privilege is reserved for Spain with one for about 5 000.)

If one were tempted to put forward a claim for Great Britain, it could be argued that the greatest rate of growth has occurred in this country for, with eighteen television transmitters in operation, the BBC has now a coverage of 98 per cent of the population. It is certainly true that no other country in the world with a comparable population has achieved anything like this coverage.

In fairness to the Independent Television Authority, who did not arrive on the scene until 1955, it should, of course, be pointed out that with six television transmitters in operation they have a coverage of 75 per cent at the present time, but they aim to have eleven stations in 1959-60 giving a coverage of 90 per cent and hope later to extend this to 95 per cent, by which time, according to their Annual Report for 1956-57, "the way will have been cleared for the development of a second I.T.A. television service".

It was, perhaps, outside the terms of reference to review the trends in receiver design that have taken place in the post-war years, but an aspect investigated by the authors of the broadsheet—and one which has had a considerable influence on receiver design—is the changing size of the television screen.

In 1948 four-fifths of the sets in this country had 9in or 10in tubes and the remainder was made up almost entirely of 12in tubes. Three years later, the 12in tube had gained in popularity and accounted for about four-fifths of the total sales for that year. This was about the time that tube shape was changing from circular to rectangular and the 14in rectangular tube was beginning to make its appearance so that, in 1954, more than half the sets sold had 14in rectangular tubes and about a third had the still larger 17in tube.

In 1957 the 17in was the most popular and accounted for two-thirds of the sales, while the 14in accounted for about one quarter of the total. The larger 21in tube, perhaps on account of its size, and cost, does not appear to have caught the public's fancy, so that the average size is still smaller than in the United States where the preference is for the 19in tube, or larger.

On the thorny problem of colour television, note has obviously been taken of the American experience and quoting an American report that "after at least four years of reasonably heavy production and marketing the number of colour TV sets . . . still amounts to less than 1 per cent of the number of black and white sets", the broadsheet concludes its survey of colour by remarking that "there seems little prospect of colour television being started on a regular basis within the next year or two"—a view which, unless there is more going on behind the scenes than has come to our notice, seems to be slightly optimistic.

* *Planning*, Vol. 24, No. 420 (24 March, 1958): Television in Britain. (London: Political and Economic Planning, 1958) 3s. 6d.

Liquid Cooling of Electronic Equipment

By E. N. Shaw*, A.M.I.E.E.

The potentialities of liquid cooling are explored to determine the effectiveness of heat-exchangers built into electronic equipment as integral parts of an assembly.

It is found that some 95 per cent of the heat in a pressurized container can be extracted without undue elaboration, leaving so small a residue that the usual thermal hazards in compact assemblies no longer exist. Temperature-sensitive components can be segregated in an environment at—or very close to—room ambient, an obvious advantage in "mixed" units containing both valves and transistors.

As the container skin is not called upon to play any part in the process of heat extraction, outer surfaces may be lagged to make the unit completely independent of unfavourable environments, with the internal conditions controlled and stabilized by adjustment of coolant inlet temperature and rate of flow.

EXPERIENCE gained over the past decade or so has shown that the miniaturization of electronic equipment is not just a simple exercise in mechanical ingenuity, but a process that is complicated by thermal hazards. Packing larger numbers of hotter components into smaller enclosures can raise internal temperatures to intolerable levels, resulting in the loss of stability, reduction in life-expectancy and the probability of early catastrophic failure of the more temperature-sensitive components.

Of the two methods of combating this threat to reliability, the most popular to date has been for equipment designers to transfer the responsibility to component manufacturers by demanding that their products should maintain their stability and reliability at ever-increasing temperatures. The wisdom of such demands is arguable since, with the present disparity in component operating temperatures, any one heat-dissipating component that is made to function at a higher temperature becomes a greater menace to those that cannot be up-graded.

The alternative is for equipment designers to recognize the fact that disparities exist and to accept responsibility for cooling their assemblies to a level at which reliable operation may be expected from conventional components, with temperature-sensitive items segregated from heat-dissipating elements and sequential cooling employed to ensure that all components in an assembly operate well within their thermal categories. It has been shown[1] that intelligent use of heat control in instrument cases and rack-mounting equipments can produce very substantial temperature reductions within electronic equipments cooled by natural means. The techniques developed during these investigations are now applied to the more exacting case of pressurized equipments in which compact assemblies cannot operate without forced cooling.

A customary approach to this problem regards equipments as 'black boxes' with the total dissipation determining the skin temperature of the container, which acts as a barrier between hot interior and cold exterior. Internal temperatures can be reduced by external cooling of the container, but only to the same extent as the reduction in container temperature, with a high gradient remaining in the interior as the motive force required for the transfer of the entire dissipation through the container barrier.

Liquid cooling has the initial advantage of being adaptable to sequential cooling within a container, and is thus able to augment any spatial segregation incorporated in the assembly by being applied first to temperature-sensitive components to counteract any heat transfer from hotter areas. Progressing to intimate contact with the heat source, the coolant absorbs heat and withdraws it bodily, obviating the need for forcing it through container surfaces. With heat extracted by direct cooling, the container becomes thermally neutral and can be lagged to prevent heat ingress in unfavourable environments.

Scope of Experiments

With the intention of verifying the above contentions and exploring the potentialities of liquid cooling, a series of some ten assemblies were built and tested. It soon became apparent that direct liquid cooling had so high an extraction efficiency that caution was thrown to the winds and a policy of cramming as many watts as possible into the available space was adopted—not without a guilty feeling that such a practice was most reprehensible—to present more pungent contrasts with conventional cooling methods.

The Experimental Assembly

The size and shape of the assembly were determined by the dimensions of a container that happened to be available; a restriction willingly accepted because it reproduced a condition that must be familiar to most designers who have been required to pack enormous quantities of power into incredibly small and unsuitable spaces. In this instance the container was 7in in diameter and 10½in long, with a surface area (including the front panel) of 308in^2 and a volume of 808in^3. To make the assembly representative it had to incorporate both temperature-sensitive components and heat-dissipating valves in close juxtaposition, so the first step was the prudent one of segregating the two by setting aside one quarter of the total volume for components, leaving some 606in^3 in which to deploy the heat source, consisting of 22 flying-lead valves operated as cathode-followers at their maximum rated dissipation.

Preliminary work had proved that a modicum of mechanical ingenuity would permit the cramming of up to 250W total dissipation into the available space, although accumulated experience predicted that such a high thermal loading would, quite literally, melt the assembly in a matter of minutes in the absence of a highly efficient cooling system. As it was desired to study the assembly for long periods under widely different conditions, the power supply was made very flexible, it being possible to reduce anode dissipation to zero, leaving a total of 70W heater dissipation, which could in turn be reduced by lowering heater voltage. The procedure under new and unexplored conditions was to commence at low dissipation levels and to increase the input gradually until temperature measurements indicated a close approach to melting point.

Mechanical features of the final arrangement are shown in a cross-section, Fig. 1, 'exploded' to emphasize the

* *Royal Radar Establishment, Malvern*

segregation that was an essential feature of this design. Heavy lines indicate structural members extending the entire length of the assembly from front panel to rear support. Symmetrically disposed between these members were two identical small sub-assemblies each containing six valves on a chassis 1½in wide and 8in long. Each chassis was in thermal contact with both upper and lower structural members, with the valves held in split brass tubes soldered to supporting plates that were thermally insulated from the chassis—a further example of thermal segregation. The liquid cooling system consisted of some ¼in copper

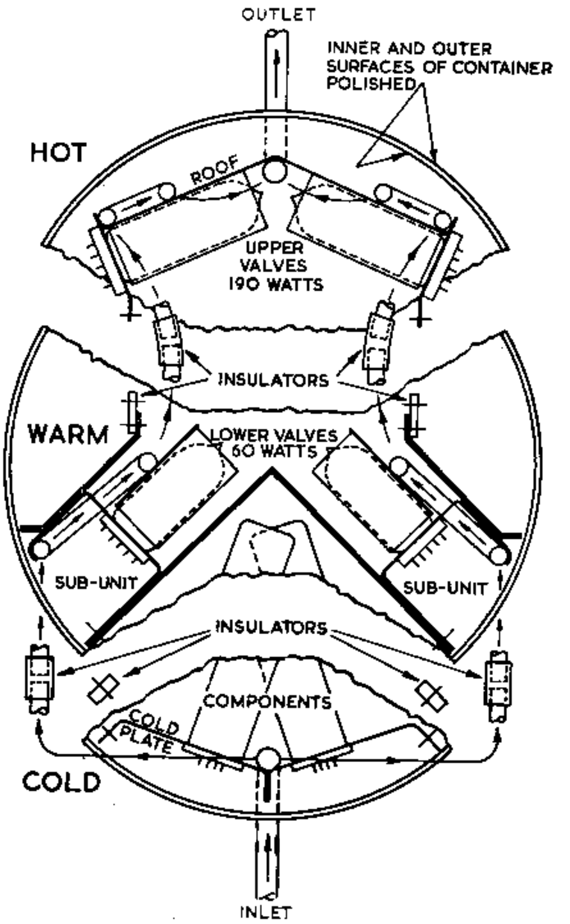

Fig. 1. Sectional view of experimental assembly, ' exploded ' to stress thermal segregation between hot, warm and cold portions

tube bent into hairpin shape, with the lower limb in good thermal contact with the chassis and both structural members, and the upper limb soldered to the rear of the valve supporting plate. Heat transfer between valves and chassis was prevented by the flow of coolant, from lower limb to upper limb, absorbing heat from surfaces in intimate contact with the valves and conveying it to the larger upper assembly. Maximum dissipation from the two small units was 60W total.

The remaining 190W were located in the larger upper assembly comprising ten power pentodes (five each side) held in split brass tubes soldered to the underside of the roof-shaped upper section, which was supported by, but insulated from, the upper structural members. A hairpin of copper tube was soldered to each side of the outer roof surface, with the upper limbs joining the larger diameter tube running under the roof crest to form a common outlet passing through the front panel, from which it was thermally insulated.

A larger diameter tube was used as a common inlet, entering the assembly at its lowest point through the front panel, from which it was thermally insulated. This inlet tube was in good thermal contact with the component deck, or ' cold plate ' which was supported by, but thermally insulated from, the inverted V-shaped lower structural member that formed a radiation shield between the cold component compartment and the warmer section above it.

Having divided the assembly into hot, warm and cold zones, it was considered that unbroken lengths of high-conductivity copper tube would impair the carefully built up segregation by encouraging heat flow from the hotter to the colder areas, so the three sections were insulated from one another by short lengths of low-conductivity plastic tubing which had no effect upon the intended upward transfer of heat by the coolant, but prevented unwanted downward thermal ' feedback ' in the absence of coolant, or at very low rates of flow.

Space occupied by the cooling system amounted to less than 8in^3, or under 1 per cent of the container volume.

Temperature Measurements

Twenty double-ended (differential) thermo-couples were used for measuring the temperature rise above room ambient, which was held at 20°C. All quoted figures were recorded in a vertical plane through the centre (hottest) section of the final assembly, although many measurements were made at other points during the various stages of development as a means of assessing the correctness of deductions made and put into practice.

Valve temperatures were measured by flattening the hot junction and pinching it between the glass envelope and the split brass tube holding it, while structural members had the hot junctions soldered directly to them. Measuring component temperatures presented a problem, which was solved by using six sealed relays as ' components '. These had the advantage of a fairly large surface area to absorb any heat transfer by radiation into the component compartment; they could be firmly mounted on the cold plate, and their copper windings connected in series formed a highly accurate resistance thermometer for measuring small changes in the (average) temperature rise of the ' components ' disposed along the entire length of the segregated compartment.

As the cylindrical container would have shown appreciable temperature variations along its surface, a circumferential winding of copper wire at the mid-length was used as a resistance thermometer to give the average temperature rise around the circumference in the hottest plane.

Ordinary mercury-in-glass thermometers were used to measure inlet and outlet temperatures of the coolant, although it was found that the readings on the outlet thermometer could not be relied upon for reasons referred to later.

Measurements were made under three test conditions, with the assembly (a) in free air, (b) in a polished container, and (c) in its container with water flowing through the cooling tubes.

Experimental Results

The point at which an assembly is liable to catastrophic failure being of vital interest to equipment designers, it is instructive to determine the thermal loading at which the assembly would have melted, as predicted by experience, by examining the results obtained under the three conditions stipulated above and recorded in Fig. 2 (a), (b) and (c) for dissipations up to the maximum of 250W. The thermal resistance figures on these curves are explained later.

With valve-holders attached to their supporting surfaces by solder melting just above 180°C, it was thought unwise to allow temperatures in this region to exceed 170° (150° rise above 20° ambient). Accepting this limitation, it is seen that in free air (a) the assembly was safe (as shown by full lines) up to 160W loading, with the certainty of catastrophic failure at slightly greater dissipation. The point of disintegration was reached at only 80W (b) with the assembly in an airtight container as a result of changing from direct convection cooling to indirect cooling of the container surfaces.

Plotted on a log-log scale, the relationship between the thermal loading and temperature rise is linear, permitting the easy extrapolation of temperatures that *might* have been recorded at the full 250W dissipation in the absence of catastrophic failure, as shown by the dotted lines in (a) and (b). In free air the upper valves would have reached their maximum temperature rating of 250°C and the lower valves would have attained a temperature exceeding 150°C which is marginal for some valve types. The components, on the other hand, would have risen only 20° as a result of segregation from the heat source. Cutting off direct convection cooling by placing the assembly in a container (b) would make full-load conditions completely intolerable, with upper valves reaching 460°, lower valves exceeding 300° and components sweltering at nearly 200° in a container with 146°C skin temperature. Thus, while the melting point of solder may be considered an artificial restriction to be easily overcome, it is indicative of temperature levels above which serious difficulties are likely to be encountered.

In contrast, a glance at the results under condition (c) shows that a flow of 25lb/h of water through an integral direct cooling system was more than adequate for the attainment of very temperate conditions indeed, without the slightest risk of disaster, even at the previously unattainable maximum dissipation of 250W. Temperature rise in the component compartment was only 5°, compared with a rise of 20° in free air, making this a very suitable location for temperature-sensitive items such as transistors. The safety factor was also remarkable, it being calculable that at this rate of flow it would need 1 060W dissipation to make the water boil at the outlet, and no less than 7·1kW to boil the assembly completely dry.

To give some idea of the scale of water flow involved, it might be mentioned that in circles associated with domestic plumbing a leaky tap dripping for one hour is estimated to waste 0·5 gallons of water at 5lb/h mass flow.

If this demonstrated potentiality of liquid cooling is to be fully exploited in any given application, it is important to obtain a clear understanding of the factors governing its effectiveness. This can be done by comparing it with forced convection cooling of the container, and then examining the means by which it was made to yield a cooling efficiency of 95 per cent in an experimental assembly designed to take every advantage offered by liquid cooling.

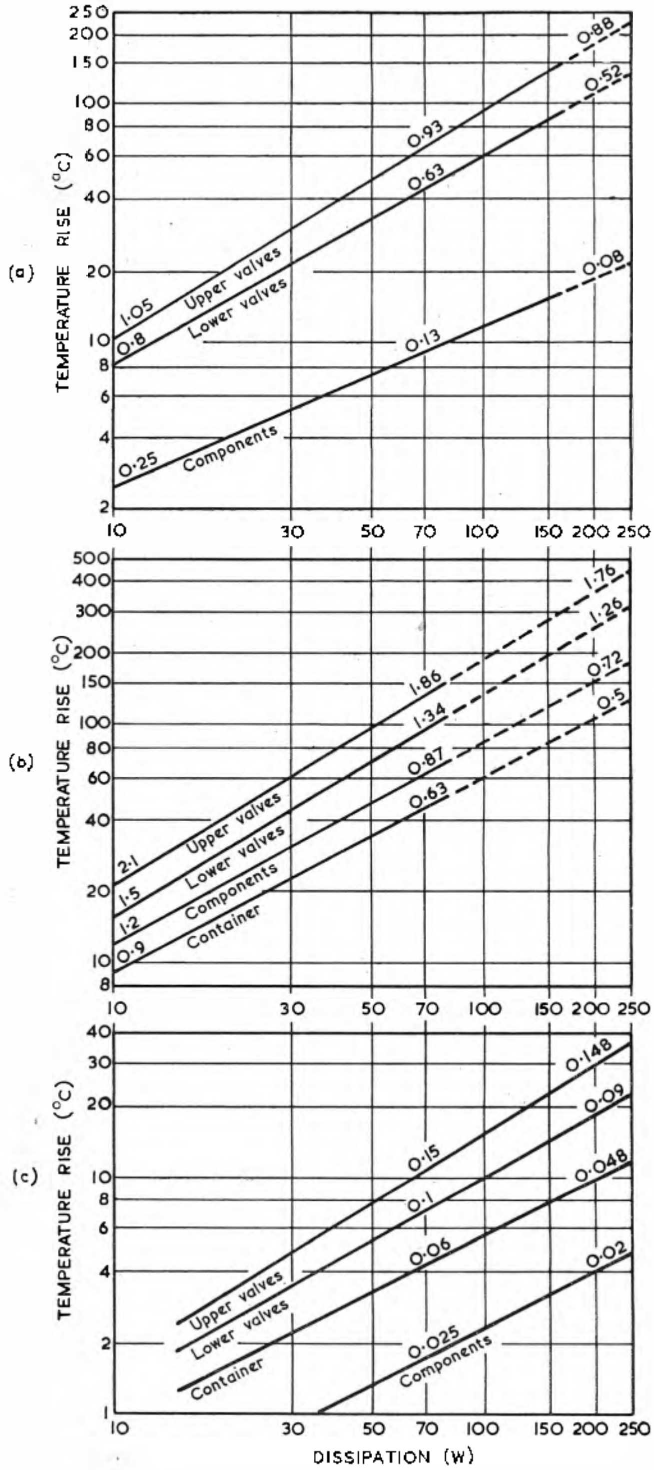

Fig. 2. Temperature rise/dissipation curves under three test conditions, with inset figures indicating the thermal resistance to ambient in °C/W

(*a*) *Assembly without container. Cooled by natural convection*
(*b*) *Assembly in container. Cooled by natural convection*
(*c*) *Assembly in container. Liquid cooled by 25lb/h of water at 20°C*

Thermal Resistance

Both studies are made simpler by adopting the concept of 'thermal resistance', R_t, expressed in °C/W, and defined here as the temperature difference between two points on the assembly, or between any one point and ambient, divided by the total thermal loading responsible for that difference.

Thus, in Fig. 2(a) the upper valves experienced a 65° rise at a loading of 70W, giving $R_t = 0{\cdot}93$°C/W. Values of R_t are marked along the curve, and it is seen that they vary from 1·05 at 10W to 0·88 at 250W—a reminder that an increase in heat flux is not accompanied by a proportional rise in temperature, although by using log-log scales one can show the relationship by means of conveniently straight lines. The 9° rise in component temperature at 70W gives a low figure of 0·13°C/W thermal resistance to ambient, and a high $\left(\frac{65-9}{70} = 0{\cdot}8\right)$ thermal resistance between components and upper valves. Had these components been mounted on the same chassis as the valves their temperature would not have been widely different, so the practice of thermal segregation may be likened to the introduction of a thermal attenuator with a high series arm and a low shunt arm.

With the assembly in its container, Fig. 2(b), values of R_t for the upper valves had doubled (i.e. the melting point was reached at half the previous loading), varying from 2·1 to 1·76°C/W, while R_t between container and ambient ranged from 0·9 at 10W to 0·5 at 250W. By calling the latter *external* thermal resistance, $R_{t.ext}$, and subtracting it from the R_t between the upper valves and ambient at common loads, a new set of figures is obtained for *internal* thermal resistance, $R_{t.int}$, ranging from 1·2 to 1·26°C/W. If so small a variation over a 25:1 change in load is disregarded, it can be stated that for all practical purposes the total thermal resistance between a point in the interior and the ambient is the sum of a constant *internal* resistance and a variable *external* resistance so that, for a thermal load W, a point in the interior will attain a temperature rise

$$\Delta t°\text{C} = W \cdot (R_{t.int} + R_{t.ext}) \qquad (1)$$

The significance of this expression becomes apparent at a point in the design procedure when it is realized that some reduction in internal temperature rise is imperative if the equipment is to survive. Should the required reduction be less than the temperature rise of the container, $W \cdot R_{t.ext}$, then there is some hope for conventional methods, such as increasing the container emissivity and resorting to forced convection cooling of the container, although it will be shown that methods hallowed by tradition can themselves introduce hazards that render futile the additional weight, expense and complexity devoted to producing such 'improvements'.

If the required reduction in temperature be greater than can be accomplished by reducing $R_{t.ext}$ to zero (or near zero, since there must be *some* temperature difference between the surface and ambient for heat transfer to occur), there are three courses available to the equipment designer. He can de-rate the assembly by removing heat-dissipating elements until tolerable thermal loading is attained; he can endeavour to reduce $R_{t.int}$, which is greater than $R_{t.ext}$ and much less tractable, or he can turn to liquid cooling for his salvation.

Examples of all these methods are given to demonstrate the magnitude of the various factors involved in the choice of a cooling system for an experimental assembly grossly overloaded by conventional standards.

Indirect Cooling by Natural Convection

The temperature rise above ambient at which a heated body attains thermal equilibrium in still air is usually related to heat flux by the expression

$$\text{W/in}^2 = h \cdot (\Delta t)^{1.25} \qquad (2)$$

where h is a coefficient depending upon the shape, aspect and surface emissivity of the body.

Once these factors have been established for a specific body, as in the experiment recorded in Fig. 2(b), its rise in temperature (Δt) becomes a measure of power dissipated; the expression for the container in this particular case is

$$\text{W} = 0{\cdot}59\ (\Delta t)^{1.25} \qquad (3)$$

or

$$\Delta t = 1{\cdot}5\ (\text{W})^{0.8} \qquad (4)$$

These relationships being a little cumbersome, it is easier to work in terms of thermal resistance, °C/W, so very simply derived from the container calibration curve and marked along it. The tendency is for thermal resistance to increase at lower levels of dissipation, as can be seen by substituting $\Delta t = 1°$ in equation (3) to give an equivalent loading of 0·59W, at which $R_{t.ext}$ is very high, being 1/0·59 = 1·7°C/W.

Assuming that the container is in truly still air at such very low dissipation, heat transfer can occur by conduction only, with the high thermal resistance accounted for by the low conductivity of air. Greater dissipation will increase surface temperature, heating the air in contact with it and decreasing its density until it is thrust upwards by colder and more dense air replacing it. This bodily transfer of heat from container surfaces is responsible for progressive reductions in $R_{t.ext}$, which continues to diminish as the increasing dissipation gives greater vigour to the natural, or self-generated, convection currents.

It was established that this natural process was not very effective, the assembly being in danger of melting at loads in excess of 80W, although continuous operation at 70W was proved to be safe, with a recorded temperature rise of 130° for the upper valves and 44° for the container, the $R_{t.ext}$ being 0·63°C/W.

With an assembly capable of dissipating 250W having to be de-rated to 32 per cent of this maximum in order to have a chance of survival, or to 28 per cent for continuous operation free from anxiety, the conventional reaction is to try to reduce container temperature by forced convection.

Indirect Cooling by Forced Convection

Before considering forced convection it is just as well to dispose of a fallacy that seems to be very prevalent. Forced convection does not 'blow off more watts', but merely extends the process described for natural convection by increasing the vigour with which cold air displaces warm air at the container surfaces, allowing the dissipation—which is unchanged—to take place at a lower equilibrium temperature. The ultimate aim is to reduce the container temperature to that of the cooling air, making $R_{t.ext}$ zero. Although this is impossible, it is worth a few moments consideration as an unattainable goal.

If air is forced across container surfaces in the previous example with such vigour that the 44° temperature rise falls to zero (due to elimination of $R_{t.ext}$) the upper valves would rise by the product of 70W and a constant $R_{t.int}$ of 1·23°C/W = 86°, or just 44° less than the 130° recorded with natural convection. The internal temperature drop would thus be the same as that produced externally.

A check was made in an improvised experiment by blowing air at the container without any attempt at attaining uniform air flow over the whole surface. At 70W dissipation $R_{t.ext}$ fell to 0·086°C/W, with a container rise of only 6° instead of 44°, a reduction of 38°—while internal temperatures fell by 34°. It was considered that this was sufficiently close to prove the point without fitting an outer skin and blowing air through the annular space, as is often done to ensure a uniform flow over the whole surface area, which is sometimes increased by providing fins in an effort to reduce $R_{t.ext}$.

Returning to the hypothetical case of a container cooled to an ambient of 20°, the maximum power that could be dissipated by the assembly would still be restricted by the 150° rise considered to be just safe for the upper valves. Melting point would be reached at 150/1·23 = 122W, or 49 per cent of the maximum dissipation.

Finally, at the full dissipation of 250W there would be a difference of 250 × 1·23 = 308° between valves and container skin. If the temperature rise of the upper valves were to be restricted to the completely safe figure of 130° above 20° ambient, the container would have to be cooled to

178° below ambient, or −158°C. A 20° rise in container temperature caused by decreased air flow or rise in air temperature would threaten the assembly with complete disintegration.

Disadvantages of Indirect Cooling

From the above examples it may be concluded that indirect cooling is strictly marginal. Placing the assembly in its container increased thermal hazards by reducing the permissible dissipation from 64 to 32 per cent of the available maximum, while forced convection with unlimited quantities of air and a container assumed to be at ambient failed to restore the conditions in free air, returning an optimistic 49 per cent. This is in sharp contrast with the very moderate temperatures achieved by liquid cooling in Fig. 2(c) at the full loading of 250W—which could have been increased by several times, if mechanical ingenuity had been equal to the task, without reaching dangerous temperatures.

Lest it be thought that a change from air to liquid cooling without any modification to a conventional assembly would be sufficient to achieve this desirable result, it must be pointed out that indirect cooling by forced convection, or by pumping liquid through tubes wound round a container, can only produce the same result—a container at (nearly) the coolant temperature, with the cooling medium having no effect at all on internal temperatures, which are determined by $R_{t.int}$.

Irrespective of the medium used, indirect cooling has many disadvantages. If a reduction of $R_{t.ext}$ is attempted by an increase in container emissivity, such an 'improvement' can be effective only while the container can 'see' the ultimate heat sink, free air. Placing it in a hot nest of containers at the same temperature will cancel any advantage and may, if adjacent surfaces are hotter, reverse the direction of heat flow and raise internal temperatures to levels higher than those current before the 'improvement'. Conditions such as these occur frequently in aircraft installations and will be increasingly difficult to combat when aircraft exceed the speed of sound for long periods and attain skin temperatures above 100°C due to kinetic heating. An equipment able to 'see' the hotter aircraft skin will become a heat sink vainly endeavouring to cool the aircraft—an undertaking that could prove fatal to the enclosed components.

Even at sub-sonic speeds difficulties arise in maintaining an adequate mass flow of air at the higher altitudes because air blowers are essentially constant-volume devices unable to supply the weight of rarefied air needed to satisfy thermal equations at sea level, although decrease in air temperature with altitude affords some compensation, but at sea level in tropical climates they function at full efficiency, belching out masses of 'cooling' air at the local ambient, which might be as high as 50°C.

Faced by such vagaries, the designer has to accept the fact that components with temperature coefficients in the interior will respond to changes in container temperature, and that he will have to de-rate his assembly to a thermal level such that catastrophic failure cannot be caused by *any* temperature rise of the container *plus* the constant internal rise, $W.R_{t.int}$.

Internal Thermal Resistance

The magnitude of the internal thermal resistance, $R_{t.int}$, is determined by the assembly geometry—a term that covers the component disposition and the means of heat transfer between individual components, and between the assembly and container.

Changes in the assembly geometry will permit variation over quite wide limits of $R_{t.int}$ between any point on the assembly and the container, and of the thermal resistance between the individual components, as exemplified by segregation of the temperature-sensitive components from hot valves by breaking the conductive path between them, by shielding components from radiation transfer and locating them at the lowest point in an internal convection pattern. Should it be desired to decrease the resistance to heat transfer from the valves to container, the opposite procedure would apply, with every effort being made to provide short conductive paths, to increase radiation transfer by raising the emissivity of the heat source and the absorptivity of interior container surfaces, and to encourage convection transfer from the heat source to upper surfaces of the container. All these expedients would have a beneficial effect in reducing thermal hazards by decreasing $R_{t.int}$, and thus the maximum interior temperature rise.

However, with indirect cooling of the assembly subject to so many imponderables, variations in container temperature would still react upon components with temperature coefficients, with no improvement in the thermal stability of the equipment.

It will be noticed that, in addition to very careful thermal segregation, all expedients used in the experimental assembly pictured in Fig. 1 were designed to *increase* $R_{t.int}$ by keeping inner and outer container surfaces polished to reduce their absorptivity and emissivity, by insulating heat sources from structural members, and by shaping the upper assembly in such a way as to impede heat transfer by convection from the assembly to upper surfaces of the container.

Direct Cooling With a Thermally Neutral Container

The apparently retrograde step of deliberately increasing an internal resistance that has been shown to bear the major responsibility for thermal hazards met in indirectly cooled assemblies is justified by the much greater effectiveness of liquid cooling applied directly to the heat source. Without being asked to play a significant part in the process of heat extraction, a thermally neutral container can have external surfaces polished like a calorimeter to make it relatively immune to environmental conditions. Such a container could be stacked among others having similar properties without appreciable heat transfer between them, or may be lagged to prevent heat ingress under exceptionally severe conditions, such as exposure to a heated aircraft skin.

Polishing the container interior maintains a high $R_{t.int}$ between it and the assembly, so that the container temperature would have to be raised quite considerably before the inward heat flow could have any marked effect upon a direct cooling system with a thermal resistance as low as the 0·148°C/W for the upper valves and 0·02°C/W for the components recorded in Fig. 2(c) for the full 250W load. Under these conditions an electric fire two inches from the container surface raised the skin temperature by 23·5°, while the valves and components rose by 3·5° and 2·5° respectively. This is an alternative form of the thermal attenuator principle mentioned earlier, with the high series arm formed by $R_{t.int}$ and the shunt arm maintained low by intimate contact with a coolant.

Because the value of $R_{t.int}$ is a constant set by the system geometry, the extent to which a thermally neutral container can protect an assembly from environmental conditions will depend upon the design of the assembly and the efficiency with which heat is withdrawn by the coolant.

Water as the Cooling Liquid

Although not necessarily ideal under all conditions, the use of water as the coolant in experimental work has certain advantages. It is readily available, stable, non-toxic, non-inflammable, and has a most suitable combination of physical properties, such as viscosity, density and high specific heat (by definition, unity). These properties make it valuable as a reference liquid in a cooling system designed to establish empirical data for use as a basis for judging the suitability of other coolants.

The cooling capacity of water is given by

$$W = 0{\cdot}527\,M\ .\ \Delta t \quad \ldots\ldots\ldots\ldots\ldots \quad (5)$$

where W is cooling capacity (W)

M is mass flow (lb/h)

Δt is temperature rise (°C)

Because thermal conditions in the assembly must depend upon the coolant inlet and outlet temperatures, the Δt term is of the greatest interest and equation (5) can be re-written as

$$\Delta t = 1{\cdot}89\,(W/M) \quad \ldots\ldots\ldots\ldots\ldots \quad (6)$$

This relationship is plotted in Fig. 3 for a wide range of cooling capacities. While directly applicable to water, the relationship for any other coolant may be obtained by taking into account the specific heat of that coolant. For example, a coolant of specific heat 0·5 would have half the indicated cooling capacity, or an indicated cooling capacity would need twice the product of Δt and M.

Cooling Efficiency

Knowing the power supplied to the assembly and the rate of mass flow, the expected temperature rise, Δt, of the coolant can be found from equation (6), or from Fig. 3. Compared with this theoretical maximum, the temperature rise obtained is a measure of the efficiency of the heat extraction process.

Simple as this might appear, it was found most difficult to put into practice, for the following reasons. At a maximum flow of 25 lb/h equation (6) gives the expected temperature rise as 18·9° at 250W input, or 13·2W/°C, demanding measurement to an accuracy of better than 0·1° if the thermal content of the coolant is to be estimated to the nearest watt. At reduced flow the required accuracy is not so great, but as temperatures increase a gradient appears at the outlet and a true measurement of outlet temperature is obtainable at one point only in the stream. This combination of varying accuracy and elastic temperature gradient made the measurement of efficiency at different rates of flow somewhat erratic.

Fortunately, it was possible to overcome this difficulty by using container temperature rise as a measure of residual wattage, which could either be calculated from equation (3), or read off the container calibration curve in Fig. 2(b). Subtracting residual wattage that had to be *dissipated* by the container from total input wattage to the assembly gave the wattage *extracted* by the coolant, and hence the extraction efficiency. Measuring efficiency in terms of residual wattage had the added advantage of drawing attention to the benefits conferred by a thermally neutral container, it being obvious that the smaller the part played by the container, the higher the extraction efficiency.

Control of Heat Transfer

Whereas an indirectly cooled system requires all the heat to be transferred from the assembly to the container before being dissipated from external surfaces, an assembly with an integral cooling system has most of the heat extracted by the mass flow of the coolant, with the container called upon to dissipate only the residual wattage. The effectiveness of this process can, and should, be controlled by the designer, whose purpose it must be to so construct the assembly that residual wattage is minimized by encouraging heat flow into the coolant, thus attaining a high extraction efficiency and a reasonably close approach to the concept of a thermally neutral container.

Presented with a choice of alternative heat transfer paths —from assembly to coolant, or to container—the object should be to manipulate heat flow by reducing the thermal resistance between heat source and coolant and increasing $R_{t.int}$, the thermal resistance between heat source and container, the two lines of attack being complementary. For optimum results it is necessary to control all three forms of heat transfer, by conduction, convection and radiation.

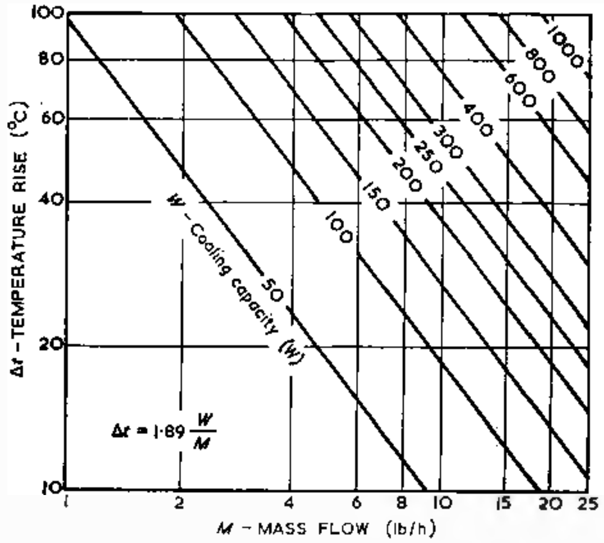

Fig. 3. Cooling capacity of water

Having exercised such control in the experimental assembly, it forms an excellent example of how to achieve 95 per cent efficiency at a mass flow of 25lb/h of water.

Dealing first with conduction, both inlet and outlet tubes were insulated from the front panel to prevent a surface with high conductivity 'short-circuiting' the coolant temperature rise, which determines its thermal content. The conductive path formed by copper tubing in the interior was broken for the same reason, and to improve thermal segregation between hot valves and temperature-sensitive components. Because the structural members, front panel, and container were all in good thermal contact, the cold plate and hot upper assembly were both insulated from the structural members, to which only the sub-unit chassis (with the valves insulated) were attached. With the conductive path to the container broken, steps were taken to encourage heat flow into the coolant by providing the most intimate contact between the split brass tubes holding the valves and the copper tubes containing the coolant. It was found that indifferent contact could result in gradients as high as 70°, which were reduced to the 13° recorded in the final assembly by soldering the valveholders to copper supporting plates, to the outer faces of which the coolant tubes were also soldered.

Convection transfer by air currents takes place from all heat sources in an upward direction, with the rising stream frustrated by the upper surface of the container, to which it gives up its heat as it divides into two symmetrical streams down the container sides. This contact with the container upper surface was prevented by the roof-like shape of the upper assembly, made to act as a 'convection catcher' by conveying intercepted heat into the cooling tubes attached to the roof. An early assembly with a detachable roof proved the principle by an increase of some 8W of residual heat without a 'convection catcher'

Radiation transfer was minimized by reducing both the absorptivity of the container interior and the emissivity of the assembly surfaces, with great care taken to ensure that heat sources could not 'see' the container interior. Ten vitreous resistors used as cathode loads for the ten upper valves were located inside the roof to prevent radiation (and convection) transfer to the container, which amounted to an additional 10W with these resistors mounted above, and parallel to, the upper structural members. Heating of the component compartment by radiation from the lower valves was prevented by the screening action of the inverted V lower structural member, removal of which nearly doubled component temperature rise.

By expedients such as these an initial extraction efficiency of 75 per cent was worked up through a series of experimental assemblies to a peak of 95 per cent or, in terms of residual wattage, the initial 62·5W were reduced to 12·5W. In general, the objective was intimate contact between coolant and heat source, with escaping heat intercepted by a surface, preferably cooled, to prevent it reaching the container.

It is interesting to note that the 1·23°C/W achieved in the final assembly compares with an $R_{t.int}$ of about 0·8 in more conventional layouts, so that while a conventional assembly might have just escaped catastrophic failure at 250W load without a container, under the conditions of Fig. 2(b) it would have melted at about 110W. Another interesting point is that with a coolant inlet temperature the same as ambient, as was the case in the recorded experiments, it is impossible to withdraw all the heat in the container because in order to absorb power the coolant temperature must rise—in this case by 18·9° at maximum flow—with the mean temperature rise of the assembly providing the motive force for some heat transfer across $R_{t.int}$. A condition of balance, without heat transfer to the container, could be established with a coolant temperature below ambient and a mass flow rate such that Δt just brought the mean internal temperature up to ambient, allowing the container to remain truly neutral.

From this it follows that extraction efficiency depends upon coolant inlet temperature as well as its rise in temperature, which is determined by its specific heat and rate of mass flow.

Effect of Varying Mass Flow

Changes in extraction efficiency and internal temperatures with varying rates of mass flow are shown in Fig. 4. As mass flow decreases the coolant temperature rise increases in accordance with equation (6) and the higher mean temperature of the assembly permits a larger proportion of the 250W to escape across the constant $R_{t.int}$ to the container, with a consequent drop in extraction efficiency.

The wisdom of insulating the coolant inlet tube from the front panel is shown by the very small change in cold plate and component temperatures, which are seen to be lower than those of the container and sub-unit chassis, representative of the structure temperature. Thanks to this insulation, and to the effective thermal segregation, it can be stated that component temperatures depended almost entirely upon coolant inlet temperature, being only slightly effected by changes in the rate of mass flow.

On the other hand, while heat source temperatures will rise and fall linearly with coolant inlet temperature, they react to changing flow in a manner that must satisfy equation (6), rising more and more steeply with decreasing flow until the coolant boils at the hottest portion of the assembly, which remains at a constant temperature (in this case 100°C). It might be thought that invoking the latent heat of the coolant would increase extraction efficiency, as in an evaporative cooling system, but the only effect is to establish a maximum temperature rise in the region where boiling takes place, with further reductions in flow

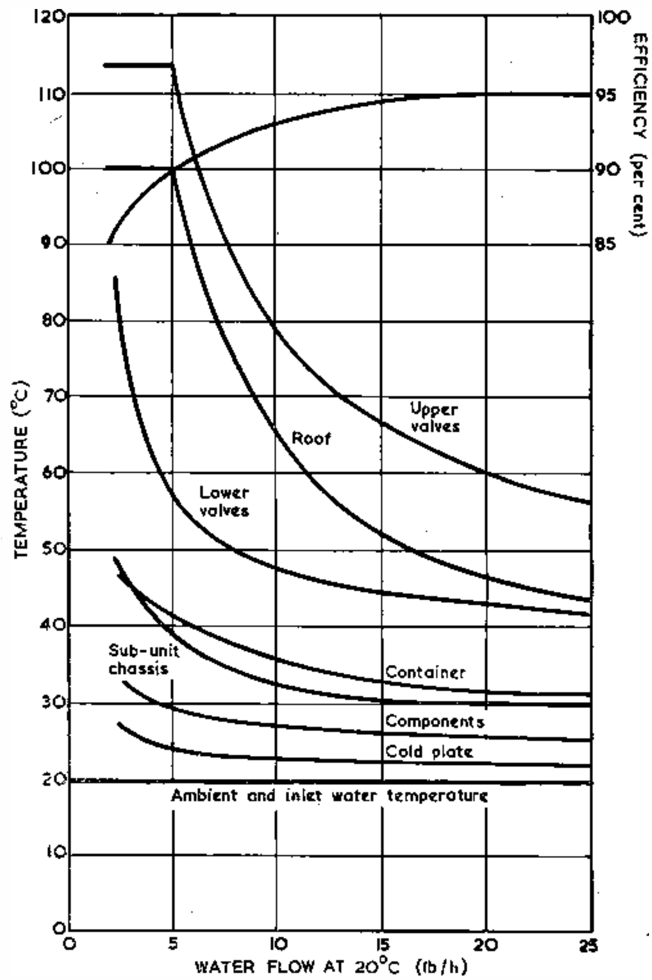

Fig. 4. Temperature/mass flow curves for assembly at maximum loading of 250W

increasing the area of this region and the mean temperature of the assembly, causing extraction efficiency to diminish.

The latent heat of vaporization being 284W-h/lb for water, the experimental assembly would have boiled dry below a flow of 1lb/h, experiencing a catastrophic failure that was carefully guarded against during the course of experiments.

Practical Applications

Having shown that internal temperatures depend upon coolant inlet temperature and rate of mass flow, it becomes possible to establish constant operating conditions within an assembly by controlling both these quantities, with a thermally neutral container protecting the interior from environmental changes.

Mobile apparatus can usually carry only a limited supply of coolant, thus restricting its operational time. For example, a tank containing 25lb of water at 20°C would maintain the terminal conditions in Fig. 4 for an hour if allowed to run to waste through the experimental assembly at 25lb/h or, by halving this rate of flow, the conditions appropriate to 12·5 lb/h could be maintained for two hours, and so on.

If coolant waste is not permissible, a closed-circuit pump-driven circulatory system could be used, with the inlet taken from the bottom of a 2·5 gallon storage tank with a large vertical dimension, and the outlet pumped to the top. The low conductivity of water would ensure a constant inlet temperature for an hour at 25lb/h, followed by a rise of $18{\cdot}9 \times 0{\cdot}95 = 17{\cdot}9$°C. At half this flow conditions appropriate to 12·5lb/h could be maintained for two hours, followed by a rise in inlet temperature of $37{\cdot}8 \times 0{\cdot}935 = 35$°C. By using a mixture of about 5lb of water with added anti-freeze and 20lb of ice, the latent heat of fusion of ice (42·2W-h/lb) would increase the coolant storage capacity by 844W-h, maintain inlet temperature at nearly 0°C, and improve extraction efficiency.

When a storage system proves inadequate for equipment that has to operate continuously, or even intermittently, for long periods, the logical step is to use a smaller quantity of the coolant purely as a transfer agent between the assembly in its non-dissipating container and a heat-exchanger rejecting the extracted heat into the atmosphere; the ultimate 'heat sink' of infinite capacity. In an aircraft, for instance, the heat-exchanger could be located in the fuel feed to the engines, with the transferred heat adding calorific value to the fuel immediately before being dissipated by combustion in the engines, thus obviating any storage problems.

The return line from the heat-exchanger would form a common inlet to all the electronic equipment on board, with an inlet temperature dependent upon the rate of fuel flow and storage temperature in the fuel tanks. Since both these are variable, the inlet line could be refrigerated to a desired temperature maintained by thermostatic control. All the equipment would thus have a common inlet temperature, but the rate of flow through each could be adjusted to meet individual requirements by providing assemblies with their own small pumps controlled by the temperature rise at some convenient point within the container. As heat extraction depends upon rate of mass flow, and not on the total quantity of coolant, the latter can be very small in circulatory systems rejecting the extracted heat through a heat-exchanger.

Choice of Coolant

One of the disadvantages of using water as a heat transfer agent in circulatory systems is that a rise in temperature causes the air and carbon dioxide dissolved in the water to separate out, at the rate of about 0·8in^3/lb, forming compressible pockets that might inhibit circulation. Boiling in any portion of the assembly is an extreme case and should be avoided at all costs.

Water is by no means an impossible transfer agent for use in closed-circuit circulatory systems if the necessary care is taken to avoid freezing, boiling, and the formation of air pockets, although a substance with a wider temperature range will give greater latitude to the designer. It should have a high specific heat and density; be non-toxic and non-corrosive, and should not liberate dissolved gases.

The use of air as a transfer agent is made difficult by its low (0·24) specific heat and its very low density. To obtain results equal to those given by a flow of 25lb/h of water the equivalent mass flow of air would be $25/0{\cdot}24 =$ 104lb/h which, at a density of 0·0764lb/ft^3, is 1 360ft^3/h. For such a volume of air to pass through the small-bore tubes in the experimental assembly, with a cross-sectional area of only ·00035ft^2, it would have to be forced through at a supersonic velocity of 3 900 000ft/h, or 738mile/h. The nearest approach made experimentally was the attachment of a compressed air line at a pressure of 60lb/in^2, which had a cooling effect equivalent to about 15lb/h of water, but the noise and vibration introduced by high-velocity air flow in the tubes could make its use impossible with microphonic equipment. In contrast, a flow of 25lb/h of water was maintained with a pressure drop across the assembly of only 0·6in of water, or 0·02lb/in^2

The choice of coolant must therefore take into account the product of specific heat and density which, in the extreme cases above, is $0{\cdot}24 \times 0{\cdot}0764$ for air and $1 \times 62{\cdot}4$ for water; a ratio of 1 : 3 400. Since the cooling capacity is determined by mass flow, a low-density coolant will require to be pumped through the cooling tubes at greater velocity, with the pumping power increasing as the square of velocity, unless the resistance to flow is reduced by providing ducts of larger cross-sectional area at the expense of component space within an assembly.

Heat Transfer Materials

Materials are currently available in which an integral duct pattern can be formed by roll-welding together two thicknesses of aluminium. This material was developed for refrigeration work and should prove useful for liquid cooling of electronic equipments.

The copper tubing in the experimental assembly simulated a material of this type, with the added advantage of a greater ease of manipulation in the early stages of evaluating the potentialities of liquid cooling.

Conclusions

Liquid cooling of miniaturized equipment is shown to be so effective that perfectly ordinary components may be used in very compact pressurized assemblies with less thermal hazard than encountered by the same assemblies in free air. More adaptable and less marginal than the usual indirect cooling methods, which tend to ignore disparities in component temperature categories, liquid cooling encourages the designer to introduce spatial segregation in the knowledge that it can be augmented by sequential cooling of the zones so created.

Weight and volume penalties of integral heat-exchangers can be very small for high-density coolants, and heat extraction efficiency can be made very high by logical expedients shown to be under the control of the assembly designer. The more these powers of control are applied, the closer the approach to a thermally neutral container protecting the interior from changes in environmental conditions.

Thermostatic temperature control of coolant inlet supplies to an entire equipment, with automatic flow rate adjustment for individual assemblies, will improve the stability of the equipment by establishing constant low-temperature conditions in the interior unaffected by changes in external temperatures or pressures.

Liquid cooling is particularly applicable to high-power transistors requiring a low thermal resistance to a cooling source at maximum dissipation.

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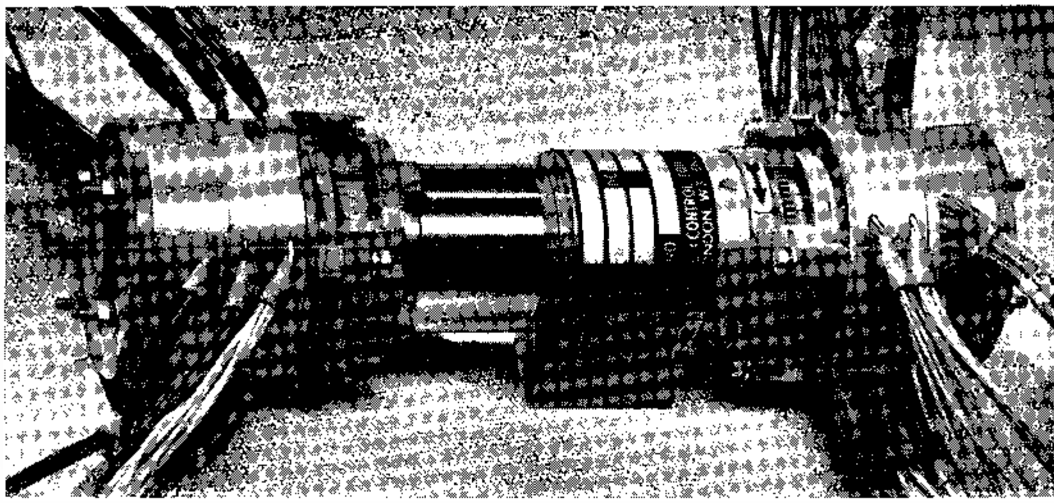
A High Speed Rotary Switch and some Applications

By M. Lowenberg*, A.M.I.E.E.



The general requirements of high speed switches are discussed with particular reference to their applications in information sampling systems.

Details of a new high speed rotary switch are given and typical applications are described.

THE most obvious way of connecting a circuit temporarily to another is by means of a switch, and if say, the circuit is to be connected in turn to several others, then the natural development is the rotary stud switch. Manual operation is preferably replaced by automatic selection, and here again the simplest step is to motorize the rotating pick-off arm.

The uses of rotating or sampling switches are numerous and some typical applications are given. References for further reading are given at the end of the article.

For sampling and metering applications it is important that the 'on-off' ratio is as high as possible, since the 'on' time determines the information period. The sampling speed is directly controlled by the maximum frequency of the sampled phenomena, since in order to reconstruct a continuous sine wave it is theoretically necessary to obtain a minimum of just over two samples per cycle. With discontinuous phenomena and complex waveforms, however, ten samples are desirable for the reproduction of higher harmonics.

It is in this respect that electronic switches will score heavily[1,2,3], since the gating of information presents little difficulty at say 500kc/s. To counter this advantage, the complexity of a purely electronic type is great when compared with its mechanical counterpart, thus in some applications where space is important this consideration would decide against its use. Many electronic switches have been described in the literature and those which use electronic valves, Dekatrons, rotating beam tubes and photoelectric devices have regretfully this disadvantage of high bulk and possible limitations to a specific type of input impedance.

A recent device[4], which has been proved very effective impinges a rotating jet of mercury on to stationary pins, and another which is electromagnetic in principle, utilizes a permanent magnet rotor to sequentially close a ring of hermetically sealed contacts.

Mechanical Switches

The simplest unit, which has the merit of minimum size and least complexity is the purely mechanical switch. This type which has now gone through several stages of evolution originated as the rotary stud switch. The defects of this design are in the province of the metallurgist who had firstly to find contacts which have good wearing qualities at high surface speeds, and secondly to provide the rubbing surfaces with as good a finish as possible. It has therefore undergone fluctuations of favour with each new development of materials and disuse as the demands for higher switching speeds were increased. One such switch[5], used for airborne telemetry, permitted rotations of 5 rev/sec, but a new development which will be described permits rotational speeds of up to 200 rev/min.

Noise

One requirement of any system of information transmission is to keep the signal-to-noise ratio as high as possible[6,7,8,9]. The noise produced by mechanical devices is largely dependent on the voltage and current passed and also bears a relation to the sliding surface speed as shown in Fig. 1. Fluctuating contact resistance due to oxides on the rubbing surfaces must be avoided by the choice of noble metals, and abrasive dust must at all cost be prevented from falling on to the surface.

Generated thermal e.m.f. caused by local heating between wiper and surface and partial or complete discontinuities

* *Vactric (Control Equipment) Ltd.*

The above illustration shows a complex Vactric high speed rotary switch unit causing changes in circuit resistance, can be reduced by putting a number of wiper blades in parallel.

LIFE

Rotational speed is probably the most important factor in determining the life of mechanical switches[10], with the destruction of rubbing surfaces by arcing following a close second. Therefore speeds compared with electronic devices are relatively low, and switching capacity is usually limited to a few volts and milliamperes.

Many variations of the basic design have been devised in order to reduce wear compatible with the life and noise level required, it appearing important to reduce the diameter of the unit to as small as is feasible so that surface speeds can be low. Immersion in oil does help, but adds the problem of providing an oil tight seal on the motor shaft.

Since irregularities in the stationary surface will create brush bounce, several methods have been tried which have as their basis contacts embedded in an insulating material[8].

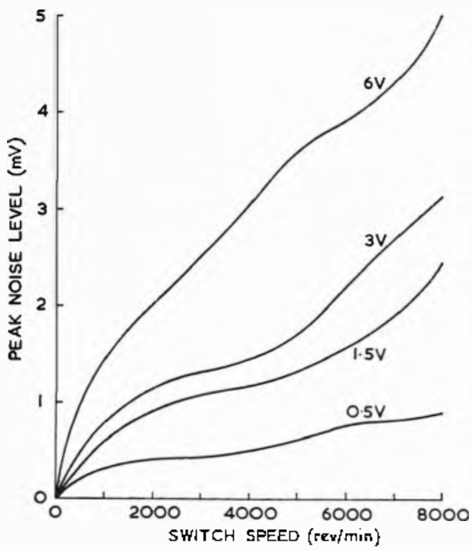


Fig. 1. Noise produced by mechanical devices

Insulation and contact are machined to a fine surface finish, the reliability of the switch then depending on the adhesion between the two, distortion due to temperature and the stability of the materials. For this type of unit, typical figures are a life of 30 hours at 6 000 rev/min and 1mV noise level with frequent cleaning.

A new type of mechanical switch will now be described which has all the advantages of its type plus a considerably extended life, giving low noise levels at high rotational speeds.

A New Type of High Speed Rotary Switch

The switch is based on a commutator made with printed circuit techniques. As distinct from the normal type of conductor which is left on the surface of the insulating card, the switch conductor is pressed into the insulation until the surface is flush. Since copper is poor as a sliding electrical contact, the metallic surface of the switch is plated with about ·00006in of rhodium before flush bonding.

Alloys of gold, silver and platinum are used for brushes which are mounted on an insulating block forming the rotor. The brushes form the conducting path from outer segment to inner collector ring, and according to the present design there can be 24 or 48 segments equi-spaced on a track of Melamine or resin-bonded glass fibre. This arrangement makes a very compact unit, a track and wiper being contained in a disk 1½in diameter and ½in thick.

A complex unit is shown on page 524. This has three high speed switches and three low speed switches driven via a precision gearbox, and is 6in long.

In applications where r.f. is to be switched, intersegment capacitance is important. In this design the capacitance is between segments ·0015in thick lying in the same plane and is negligible compared with the capacitance of the lead-out wires.

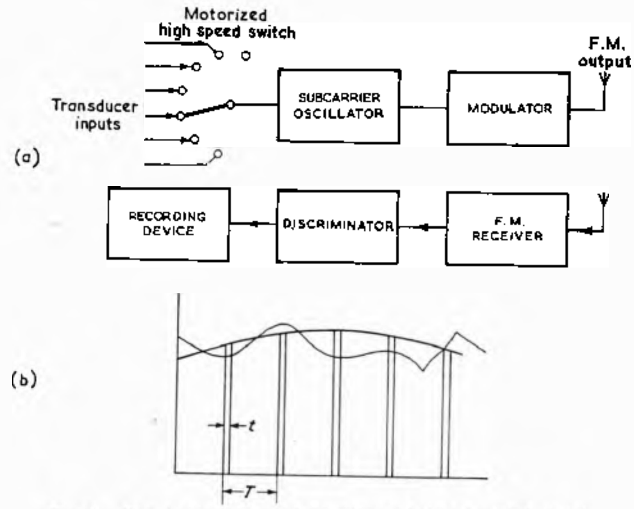


Fig. 2. The transmission and reception of time sampled signals

(a) Basic block diagram

(b) The sampled waveform. Time for one revolution of commutator is Tsec sampling time per channel is tsec

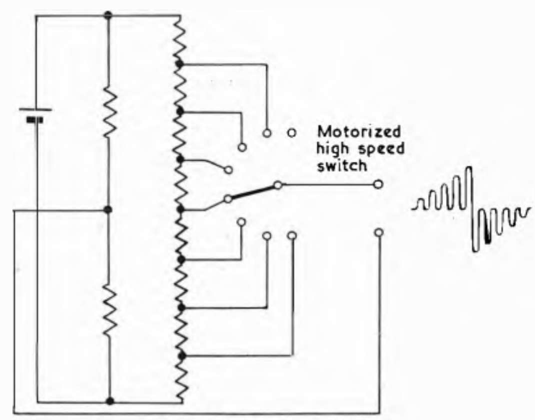


Fig. 3. Waveform shaping

Connexion with the segments is made by flexible leads which can be screened if necessary, and electrical contact with the segment is made by a small rivet into which the lead end is soldered. A circular clamp ring secures the wires away from the soldered joint, thus minimizing breakages when vibration is present.

Applications

TELEMETRY[6,7,11,12,13]

The requirement for greater economy in the use of radio telephony channels has resulted in a wide choice of time-division multiplexing systems. Whichever system is used, i.e., pulse amplitude, pulse coded, pulse length, pulse position etc., the sampling switch is an essential component. Fig. 2 gives the basic block diagram for the transmission and reception of time sampled signals, and shows how a low switch speed will not sample a rapidly changing input and the necessity of as wide a sampling time as the other switched channels will allow.

Waveform Shaping

A convenient and versatile method of producing recurrent waveforms is by sampling voltages along a potentiometer chain. The potential tappings are arranged according to the required law and are connected as in Fig. 3.

Thermocouple Sampling

Fig. 4 shows a simple block diagram of a sampled thermocouple metering system. As mentioned previously, the

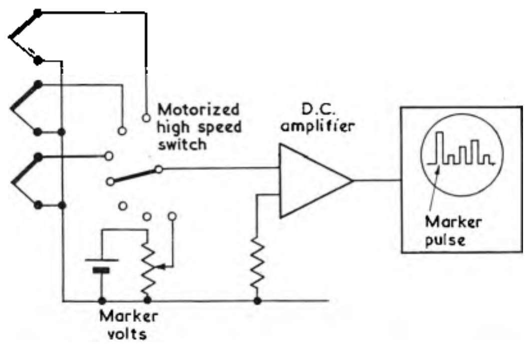


Fig. 4. High speed sampling of thermocouple voltages

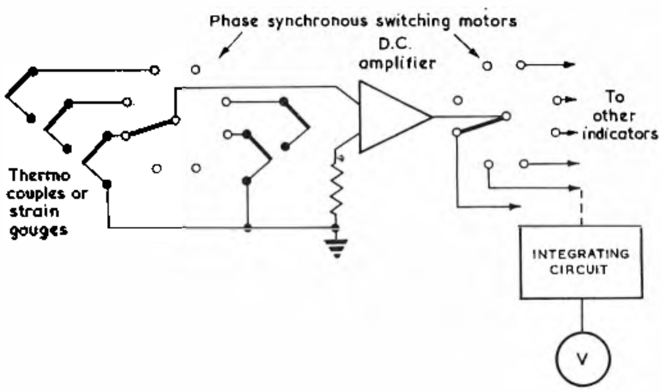


Fig. 5. Synchronous monitoring

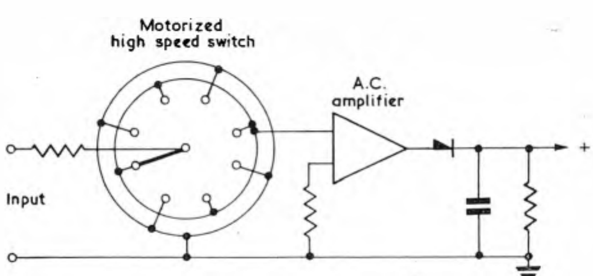


Fig. 6. Simple input chopping circuit

sampling speed is related to the maximum frequency of the signal, so that in this application thermocouple time constants are a determining factor.

A synchronous monitor is another variation of a remote telemetering system, and requires two synchronized rotary switches as shown in Fig. 5. The indicator should have a fast rise-time and a slow decay time in order to respond rapidly to the applied signal and to maintain its reading during the 'off' period.

D.C. Amplifiers and Stabilization[14,15]

In most amplifier applications, the widest frequency range possible is a very desirable feature, but in analogue computers amplification of signals of zero frequency

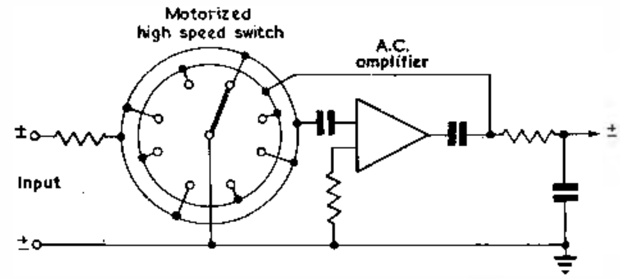


Fig. 7. Phase conscious synchronous chopping circuit

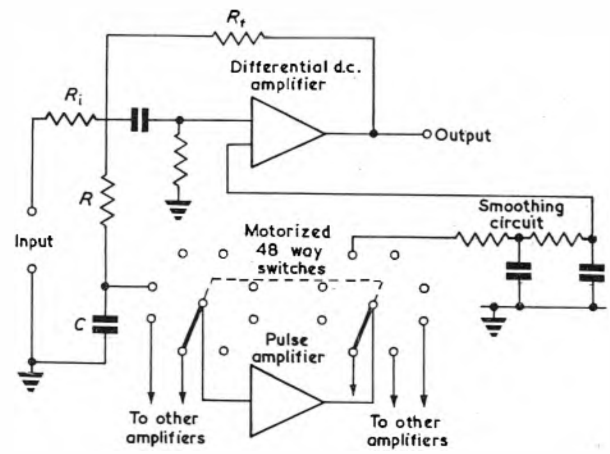


Fig. 8. Stabilization of d.c. amplifier

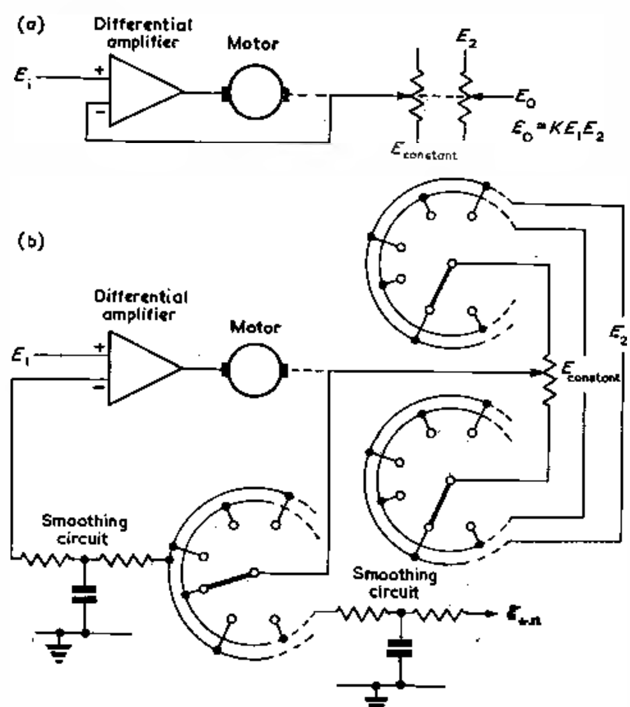


Fig. 9. (a) Conventional servo multiplier; (b) Servo multiplier using 3 high speed rotary switches

becomes imperative. The simple directly coupled amplifier suffers from drift, that is with the input shorted, the output is not zero. In order to overcome this defect, the incoming d.c. can be modulated or chopped into a square wave which is then amplified in an a.c. coupled amplifier and reconverted to d.c. by rectification. This method is shown in Fig. 6. and an alternative which provides synchronous rectification is shown in Fig. 7. The disadvantage of these methods is the inability to amplify signals much more than 1/10 of the modulation frequency and in the figure shown, this would be about 240c/s with a 48-way switch running at 6 000 rev/min.

The remedy lies in a solution shown in Fig. 8. The input circuit directs the input voltage to the differential amplifier if it has a frequency 10c/s to 10kc/s, and to the pulse amplifier if it has a frequency of 0 to 10c/s. Since the amplifiers are connected in cascade, the low frequency gain is the product of both gains, whereas the high frequency gain is only that due to the differential amplifier. In this way, the freedom from drift of a chopper amplifier is combined with the high frequency performance of a direct-coupled differential amplifier giving a frequency response of zero to 10kc/s.

ABRIDGED SPECIFICATION

No. of switches ..	Up to 12 each end.
No. of ways ..	24 or 48.
No. of poles ..	Single pole per switch, make-break or break-make.
On-off ratio ..	24-way—10 : 1, 48-way—5 : 1.
Scanning speed ..	Up to 12 000 rev/min, set as required $\pm 10\%$.
Timing accuracy ..	$\pm$ 10 minutes of arc.
Ganging accuracy	$\pm 1°$.
Contact resistance	$0{\cdot}2\,\Omega$ (static), $1{\cdot}52\,\Omega$ (running).
Thermal contact potential ..	$0{\cdot}9 \times 10^{-6}$V (ambient). $1{\cdot}7 \times 10^{-6}$V (85°C).
Intersegment resistance	$5\,000 \times 10^{6}\,\Omega$.
Collector to segment resistance	$5\,000 \times 10^{6}\,\Omega$.
Segment to collector and segment-segment capacitance	(*a*) with 4in co-axial lead 2×10^{-12}F. (*b*) with 12in p.t.f.e. leads $1{\cdot}5 \times 10^{-12}$F.
Noise	100μV r.m.s. maximum for full-life of switch at 1·5mV signal level when feeding into $1\text{k}\Omega$.
Life	Over 200 hours at 6 000 rev/min. The life of the switch is considerably extended at lower speeds and when greater noise levels can be tolerated.
Power consumption	Dependent on motor, but for design as illustrated, about 4W.
Wiring	To customer's requirements internally flyleads externally.—
Voltage	6V to 24V for d.c. motors. 6V to 115V for 400c/s and 50c/s a.c. motors.
Mounting	As required.

Servo Multiplier

The basic servo multiplier, Fig. 9, using a follow-up and multiplying potentiometer can be improved by use of a rotary switch.

The two potentiometers which would have to be matched are now replaced by one which is shared on a time basis. Low-pass networks which have a long time-constant compared with the switching speed are added in order to smooth the input signal to the amplifier and to filter the output signal.

Acknowledgments

Thanks are due to the Directors of Vactric (Control Equipment) Ltd., for permission to publish this article.

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A New Automatic Test Equipment

The electronics division of The General Electric Co. Ltd. has recently developed an automatic test equipment for the high speed measurement of some 300 inter-unit electrical signals and conditions in complex assemblies of electronic apparatus. While at present in use with radar installations this equipment is equally applicable to the testing of any type of complicated network where a large number of static or dynamic signals need checking regularly.

Automatic test equipment suitable for military use was first developed by G.E.C. to accompany a large complex airborne radar system. The operational requirement specified that the Service test equipment for this radar should be of the 'Go/No Go' type, permitting:—

(1) The accurate measurement of the complete performance of the radar to determine whether it is within acceptable tolerances in all important respects.

(2) The unambiguous location of faulty major units (or assemblies) so that these may be immediately replaced.

(3) Operation by relatively unskilled personnel.

To meet the first and second of the above requirements, a large number of different conventional test sets would have been required. In addition, several specially designed test sets would have been needed to combine adequate accuracy with the necessary robustness for airfield and aircraft carrier use.

The successful use of all these test sets would have necessitated the employment of skilled technicians, and even so, a considerable time would have been needed to test each radar.

It was therefore decided to employ an automatic test equipment, programmed to measure sequentially some 300 inter-unit electrical signals and supplies in the radar under controlled conditions. By using uniselector techniques it is possible to do this within a few minutes, and individual pre-set rejection tolerances are incorporated. Each measurement is individually timed. For example, d.c. voltages are measured at the rate of four per second to an accuracy of 0·2 per cent with minimum rejection limits of ± 1 per cent. The test set stops when a fault is encountered. Normal homing and reset facilities are provided.

Some of the tests are of a static nature, but many are dynamic, involving the injection of test signals and measurement of the results. In certain cases secondary tests are automatically made only where primary tests indicate a fault condition.

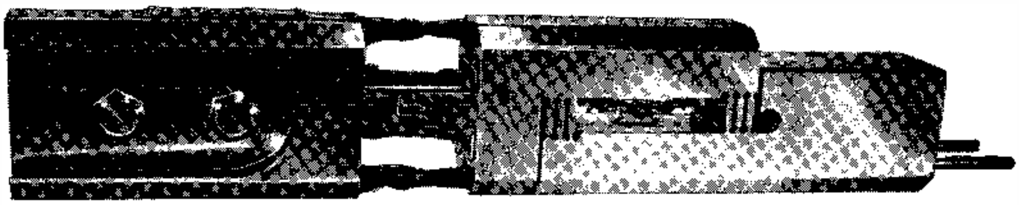
A fault pattern display is at present incorporated, but print-out facilities could be added.

Automatic testing of a limited nature has been used for production purposes for many years, but it has generally been confined to a few simple types of measurement of relatively low accuracy such as resistance, voltage, capacitance.

The present equipment may be adapted for the application of rapid high-accuracy measurements to any complex equipment where a large number of signals (static or dynamic) need testing regularly. These signals may be of a complex nature, and adverse environmental conditions may prevail. The test equipment has also been designed as a transfer standard of good reliability and high stability (e.g. 200 hours without re-alignment).

Precision Strain Gauge Techniques

By A. Tiffany*, B.Sc., A.M.I.E.E.,
and J. Wood*



New methods of data handling have necessitated a re-appraisal of fundamental measurements. Here the strain gauge techniques evolved for use with automatic digital equipment are reviewed. While far from an exact science the procedures yield precisions matching the performance of present day recording instruments. Imminent developments such as the fast digital voltmeter will demand further advances.

RESISTANCE strain gauges have long been accepted as a simple and reliable method for the measurement of static or dynamic stresses in structural work where, generally speaking, large numbers of observations are required but moderate accuracies suffice: measurements to 1 or 2 per cent of full scale are readily achieved by familiar techniques[1] and the principal problem is the expeditious handling of the large quantities of data and their subsequent analysis.

Outside somewhat specialist fields of application such as transducer design, load cells and wind tunnel balances it is not generally recognized that the resistance wire strain gauge can be made to yield precisions approaching 0·1 per cent of full scale when applied to essentially structural problems. It should be understood from the outset that this performance is achieved only with systems that can be calibrated before use as metric elements to the required accuracy, e.g. by dead weight loading: the so called 'gauge factor' or absolute strain sensitivity is not relied upon for calibration. It is not, however, necessary to continue calibration loading to full working value if sufficient attention is given to the design of the associated measuring instruments. Moreover, the overall calibration which, of course, includes the effective gauge factor together with the secondary factors of gauge misplacements, adhesion, stress concentrations and the like may be relied upon for several months so that precise comparisons are possible even when absolute calibration is impracticable.

In what follows the experience over the last three years with one particular type of Bakelite bonded gauge of American origin (the SR4 series of the Baldwin Lima Hamilton Corporation) is outlined. The authors are not aware of any other currently available commercial gauge capable of similar precisions in these applications, but it is only fair to add that no other gauge has been subjected to such thorough scrutiny as yet. A very great deal of development has been done both by manufacturers and users over several years[2,3] aimed at the production of gauges of less intractable properties than the familiar kinds. It is hoped that this short account of the encouraging results obtained by one user may encourage exploitation of the enhanced precisions in other fields of application.

Factors Affecting Precision

When accuracies of the order of 0·1 per cent are to be attempted two separate sets of problems arise; those associated with the gauge bridge itself which are largely manifest on slight reflection and whose solutions are largely matters of technique, skill and patience—in short the art of 'gauging-up'. Closely interwoven are the requirements placed upon the measuring equipment with which observations are to be made and which can be satisfied if adequate attention is paid to seemingly unimportant approximations normally accepted in strain gauge measurements, i.e. to the science of bridge measurements.

Briefly stated the main problems of the gauge bridge are

(a) Selection of gauge sets of reasonable resistive matching in the unstrained condition: and coupled with it the ageing of gauges prior to selection.

(b) Selection of gauge sets from the first selections having adequately matched thermal behaviour with varying ambient temperature: the compensation of thermal mismatch and the permanence of such compensations.

(c) The mechanical design of the metric elements themselves to give reasonable thermal 'symmetry' so avoiding thermal distortions, while yielding suitably large stress signals.

The principal problems of the measuring system are

(a) The achievement of adequate scale linearity—particularly where automatic instrumentation is concerned.

(b) Effects of gauge resistance unbalance in the unstrained condition upon the calibration of the instrument.

* *Aircraft Research Association Ltd.*

The above illustration shows a six component strain gauge balance

(c) Difficulties inherent in the determination of the calibration of a measuring channel to the required accuracy and adjustment of it to predetermined values.

The following sections deal with these problems in turn and indicate the solutions successfully adopted and operated over the last two years at a major wind tunnel facility.

A Typical Application—the Wind Tunnel Balance

As an illustration of the structural and thermal considerations relevant to the design of strain gauged metric elements the wind tunnel balance will repay some study. 'Balances' of this sort are commonly used to determine the resolved forces and moments acting at the centre of gravity of a wind tunnel model restrained thereby, but the theory has been found adequate to study the stressing of such elements and their synthesis.

(a) Beam in contraflexure, Fig. 2(a), is used to measure lift force and side force by means of gauges placed at roots of the central built in beam. The section moduli are proportioned to accommodate the relative full size forces sxpected —Table 1: lift force is always much the greatest loading in normal models so that the central beam is relatively deep. The strain gauges are wired into a four-active-arm additive bridge as indicated in the figure.

(b) Fig. 2(b) shows how a pair of beams of smaller section is employed to measure moments. Two such pairs in the vertical (pitching moment) and horizontal (yawing moment) planes are employed. It will be seen that the uniform stresses produced in these members by pure

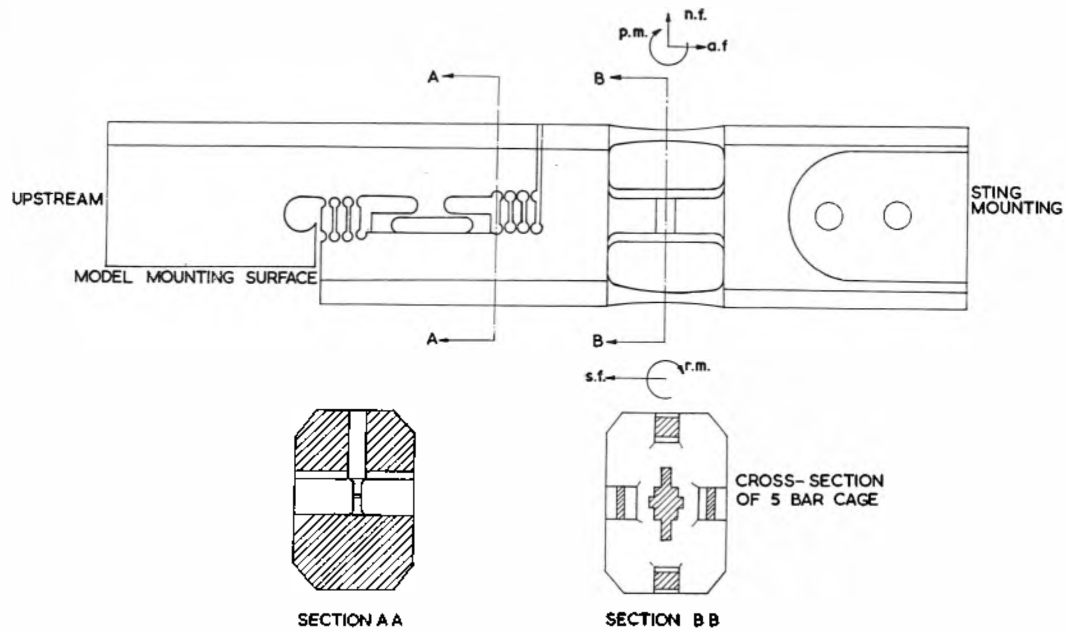


Fig. 1. A six component internal balance

approach is clearly suited to various kinds of 'weighbeam' techniques, e.g. crane gantry or hook load determinations, wagon tare measurements and hopper charge and mix control.

A six component internal balance in common use for mounting within a model is illustrated in Fig. 1. The design of such a balance is aimed at producing reactions in the various members that separately restrain motion in the six degrees of freedom available to the model considered as a rigid body, so producing distinctive stressings proportional to the individual force and moment loadings at specified 'gauge stations'. Modern aerodynamic shapes with slender wings and waisted fuselages present difficulties in obtaining sufficient compression of the outside dimensions of the balance while retaining adequate separation of the individual stress responses. Moreover if highly linear and precise responses are required the normal forms of engineering joints become impermissible, e.g. bolting and/ or dowelling. As far as possible balances are machined from the solid. In one particularly difficult instance a joint in shear was only made to exhibit zero hysteresis by seam welding the outer edges.

The type of balance shown, generically known as a cage balance, may be resolved into three basic mechanical systems which considered with the appropriate gauges constitute three types of metric element, though by no means the only satisfactory types. Engineers bending moments must be altered in the presence of vertical and lateral loads; the relatively large modulus of the central beam reduces the effect of such interfering loads and the careful positioning of the gauges at the midspan of the moment beams, i.e. at the points of contraflexure of the beams if deflected by forces in the co-ordinate directions should ideally eliminate all response to such forces.

(c) Axial force (drag) is measured by means of a tension-compression strip of the sort illustrated in Fig. 2(c). The upper and lower parts of the device are connected by vertical webs very rigid in all directions except the longitudinal. The strips themselves are placed in the line of action of the net axial force to minimize responses to other loadings.

Fig. 2(d) indicates the method of detecting rolling moment (torsional) loads utilizing the roots of the pitching-moment bars. The relative strength of the central beam and the

TABLE 1
Typical Loading of a Model in a Large Transonic Wind Tunnel

COMPONENT	LOAD (lb or lb-in)	STRESS SIGNAL (lb/in^2)
Normal force	2 440 lb	18 200
Side force	610 ,,	5 270
Axial force	300 ,,	10 800
Pitching moment	10 000 lb-in	12 460
Rolling moment	2 940 ,,	14 400
Yawing moment	4 000 ,,	9 330

sense of wiring of the strain gauges make the configuration insensitive to both yawing and lateral loadings. Table 2 indicates some of the interaction factors obtained with such a design.

Gauging-Up

Of the three phases in the production of the gauged-up metric element, viz preselection of the gauge sets, the adhesion to the gauge station and the subsequent treatment of the complete bridge the first is by far the most far reaching in its effects upon the final performance.

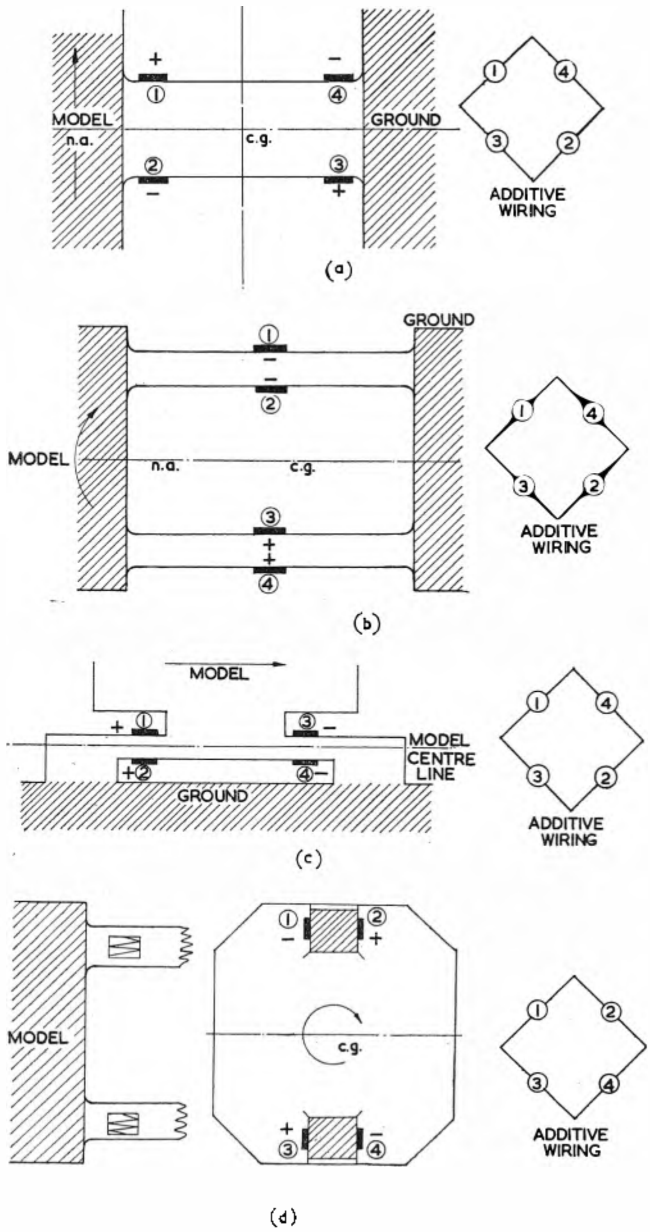


Fig. 2. The basic mechanical systems used in the 5 bar cage
(a) *Normal force and side force*
(b) *Pitching moment and yawning moment*
(c) *Axial force*
(d) *Rolling moment*

TABLE 2
Interference Factor Matrix (Percentage Response)

	N.F.	P.M.	A.F.	R.M.	S.F.	Y.M.
N.F.	100	0·3	0·65	−0·48	0·4	−0·16
P.M.	0·67	100	12·7	0·34	-0·15	0·78
A.F.	0·0	−0·22	100	−0·1	0·0	0·216
R.M.	−1·74	0·15	− 0·98	100	−0·1	- 1·65
S.F.	−2·63	1·58	2·23	1·25	100	−0·73
Y.M.	0·54	0·05	1·03	1·26	0·0	100

Selection of the Gauge Sets

Ageing

The selection process proper is preceded by an ageing process which is effective in stabilizing both absolute resistance and temperature coefficient of resistance of the gauges as received. The gauges are taken through about six heating and cooling cycles from room temperature to 150°C while laid loosely in an open tray in a laboratory oven. Resistance changes during the process may be of the order of 0·2Ω and the temperature coefficient of resistance decreases during successive cycles approaching a steady value. The gauges usually employed are designated types AB7 and AB11 by the makers; both comprise a ·001in diameter gauge wire formed into a grid and bonded in a Bakelite matrix, Fig. 3 and Table 3. The

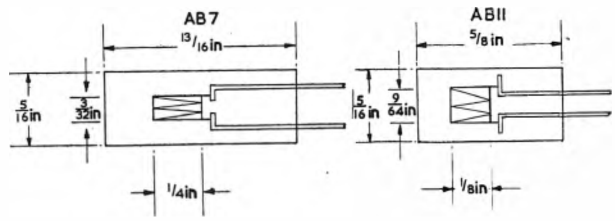


Fig. 3. The AB7 and AB11 strain gauges

temperature coefficient of resistance is very low—about one-fifth of that for copper but is very sensitive to the temperature and work history of the gauge wire and may even vary in sign over the working range of temperature (30°C rise from room ambient). Hence the temperature coefficient is never determined exactly but matched under similar conditions against those of other gauges as is discussed below.

Resistance Matching

The attainment of reasonable temperature compensation makes at least approximate matching of the absolute resistances of the gauges constituting a single bridge a prerequisite. Moreover, as will be shown in the sections concerned with instrument techniques, undue disparity of the resistance of gauges within a bridge can lead to serious errors of calibration. Hence a first selection of gauges into groups having ohmic values within 0·1Ω of one another is made. In practice the aged gauges are often sorted into 'families' whose resistance values are within 0·01Ω and put into stock so grouped: it is found that the individual resistances of such a group drift randomly within a band 0·1Ω wide with time but after several months the families may be relied upon to yield sets matched to 0·1Ω without further checking.

Temperature Coefficient Matching

The total thermal effect upon resistance of a gauge must be of the form

$$\Delta R/R = [A + K(C_s - C_g)]\,T$$

TABLE 3

GAUGE TYPE	RESISTANCE			GAUGE FACTOR	
	NOMINAL (Ohms)	SPREAD (Ohms)	TOLERANCE (Ohms)	NOMINAL	TOLERANCE
AB11	120	± 1·5	± 0·5	1·9	± 5%
AB7	120	± 1·5	± 0·5	2·0	± 5%

Note:—The Resistance Spread defines the limits between which the nominal resistance may lie, as supplied by the manufacturers. The Resistance Tolerance refers to the limits inside which will lie the value of one particular nominal resistance as quoted on an individual packet.

where $\Delta R/R$ is the fractional resistance change of the gauge for a temperature rise T

A is the thermal resistance coefficient of the gauge wire.

C_s, C_g are the coefficients of linear expansion of the gauged surface and of the gauge respectively.

K is the gauge factor, i.e. the fractional resistance strain per unit fractional mechanical strain.

Thus $K = (dR/R)/(dL/L)$ and is usually about 2 for metallic resistance wires.

Hence the effective coefficient 'a' for an individual gauge may be defined from $\Delta R/R = aT$ but will be seen to vary considerably from gauge to gauge since K and C_g differ from batch to batch in production, and A as has been implied above is almost indeterminate.

The combination of four such gauges into a bridge yields a temperature error which may be written[4] as

$$\Delta R/R = a_1T_1 - a_2T_2 + a_3T_3 - a_4T_4$$

(the assumption is made that the individual gauge resistances are very nearly equal). It is apparent that if the temperature change is merely a uniform rise of ambient temperature one can write the expression as $da.T$, where da is a measure of the disparity of effective thermal coefficients. However, if temperature gradients exist then the expression is more realistically expressed by $a.dT$.

The former situation can be dealt with fairly effectively by means of short lengths of small gauge copper or similar wire inserted in the bridge in such a sense as to offset the observed drifts. Typically 1in of 38 s.w.g. copper wire has been found necessary to compensate an apparent stress of 60lb/in² in steel. Early attempts at such compensation were made by this method alone using gauges resistively matched before adhesion to within 0·1Ω; in such cases the temperature errors remaining to be compensated by wire insertion were usually between 200 and 600lb/in² apparent stress in steel specimens. This method, however, becomes cumbersome when more than 200lb/in² has to be offset, since it is difficult to dispose the added wire in suitably intimate contact with the gauges being compensated. More important, however, is the fact that such compensations are not permanent and it is found that after three to six months the temperature errors are as large or larger than before treatment. It was noted that those bridges that had originally required very small correction were less liable to alter in this way.

The presence of temperature gradients in the metric system is a more severe problem and the mechanical design should preferably be aimed at producing thermal symmetry and small thermal distortion. Some relief may be obtained if gauges having equal and opposite thermal coefficients are available, giving symmetrical design.

Both the desire to alleviate the temperature gradient condition and the impermanence of the simple wire matching techniques stimulated efforts to obtain at least comparative measures of the effective thermal coefficient of individual gauges.

Temperature Coefficient Measurement

Attempts at absolute measurement were soon found to be abortive: nor is this surprising when it is considered that the accuracy of measurement needed was such as to allow changes of 50μΩ to be reliably detected in a gauge of 120Ω nominal resistance.

The method finally used is based on the seemingly crude thesis that gauges exhibiting similar behaviour under some arbitrary set of test conditions are likely to show some coincidence of properties under some unrelated set of conditions experienced in common. Nonetheless the results so far achieved indicate that the technique amply repays the modest testing work involved. Some of these results are shown in Table 4.

The testing consists of inserting gauges one at a time as the unknown arm of a Wheatstone bridge circuit, the detector of which is some sensitive, rapidly responding device; a d.c. chart recorder of 1mV full scale deflexion is normally used. The bridge is brought to balance with the test gauge immersed, freely suspended, in a bath of paraffin at room temperature. The gauge is then transferred to a similar bath thermostatically controlled at 20 or 30°C above room temperature and the ensuing detector deflexion observed after a few seconds of settling time. This settles rapidly to a steady reading but if observation is continued for some minutes the reading will be found to drift slowly and often continues to do so for an hour or more. These long term drifts are ignored and the response of the first few seconds only is used as a basis for comparison. It is found that this 'transient thermal coefficient' can be repeated a large number of times within close limits and over many days for any individual gauge and that pairings made on this basis continue valid in the face of repeated trials.

For gauges of the two types tested the rejection rate on the grounds of unusually large transient coefficient or inability to find a pairing gauge is quite small, largely due to batch similarities, and the rejects are more than absorbed by calls for mock-up gauge bridges in normal development work. On the average no more than twelve gauges, previously grouped to 0·1Ω in resistance have to be tested to obtain two matched pairs—which suffice for the formation of a four gauge bridge in the absence of thermal gradients. However the isolation of two pairs of equal and opposite signed transient coefficients appears to require test populations of upwards of 50 gauges and may not be feasible at all unless large stocks are available. Overall reject rate is obviously an inverse function of the available gauge stock. The author's laboratory has a turnover of 200 to 300 gauges per annum with total stocks of about 500 at any time.

As a rough measure of success it is found that bridges constructed from gauges selected by the above methods are found to require no additional treatment (after adhesion and wiring up) in over 50 per cent of cases, the remainder having residual errors below 200lb/in² apparent stress over a temperature range of 15 to 45°C.

Adhesion

The authors have found the manufacturers routine

TABLE 4

Four Sample Bridges Performance After Selection by Transient Thermal Coefficient

BRIDGE	OHMIC CHANGE OVER 30°C RISE	ABSOLUTE RESISTANCE COLD	INITIAL DRIFT
A ..	−0·09 −0·091 −0·086 −0·087	121·249 121·243 121·798 121·227	less than 30lb/in²
B ..	−0·072 −0·072 −0·070 −0·070	120·284 120·184 120·057 120·070	less than 30lb/in²
C ..	−0·084 −0·084 −0·090 −0·090	120·368 120·393 121·401 121·332	80lb/in²
D ..	−0·083 −0·083 −0·0745 −0·074	120·975 120·979 120·908 120·942	55lb/in²

methods of adhesion adequate for all normal working conditions. It may be noted that reliable indications have been obtained during tests to destruction involving stresses upwards of 100 000lb/in², the gauges in some instances remaining adhered intimately to surfaces violently distorted by plastic yield.

The phenolic resins normally employed involve baking temperatures of up to 150°C and clamping pressures upwards of 75lb/in². Where these are not practicable and for repair work an alternative Epoxy resin adhesive successfully employed is Shell Adhesive 6('Epikote') which allows intimate adhesion with very light clamping (e.g. a crocodile clip) and baking temperatures below 100°C. These adhesives loose their shear strength above 80°C and are therefore unsuited for higher temperature operation.

WIRING-UP

When d.c. measuring instruments are to be employed with the gauges it becomes particularly important that the wiring should be such as to minimize the risk of thermo-electric e.m.f.'s arising from the interconnexion of the copper supply and signal leads with the constantan tails of the gauges. Accordingly it is good practice to cut back the tails and make all inter-gauge connexions in copper; this ensures that each copper-constantan junction is in proximity to an opposing junction. The small physical sizes of the gauges used make appreciable gradients of temperature between the tails of a gauge unlikely to persist.

Measuring Instruments

Obviously the precision of strain gauge measurements can be no better than that of the instruments used for observation or recording and it has been found in practice that the achievement of instrument accuracies of the order of 0·1 per cent of full scale for inputs of the order 2 to 20mV is in itself a matter requiring considerable care and refinement. While the bases of the principal methods employed—viz a.c. Wheatstone bridge and d.c. potentiometric procedures are commonplace in instrument practice some of the less obvious points constitute major pitfalls if precisions of 1 part in 1 000 are to be maintained.

There is little inherent difficulty in setting up and calibrating a simple manually operated channel to the required accuracy when direct-loading calibrations of the test specimens are carried out, though several such channels may be needed to give complete calibration of the test piece, each having individual calibration constants. Operationally however it is essential that measuring channels be rapidly interchangeable and therefore able to be calibrated to the accuracy needed independently of the test piece, e.g., by some purely electrical means. It is also a great convenience if all channels can be set to a common calibration sensitivity.

It is increasingly the practice to utilize self-balancing analogue recorders or digital read-out methods[5,6] where multi-channel observation is required. This imposes a further requirement on the measuring channel of linearity to the same degree or better than, the expected accuracy. In particular, when analogue instruments are involved (e.g. chart recorders) $\pm 0{\cdot}1$ per cent linearity represents about the practicable limit of performance for contemporary instruments.

In what follows some of the points found to require attention in an installation in daily use for the past two years are reviewed, the elements of strain gauge measurement being taken as read.

Effects Associated with the Balancing Controls

Fig. 4 illustrates in essentials the method of zero-balancing a strain gauge bridge, i.e. of backing-off the initial unbalance signal arising from the differences in unstrained resistance of the constituent gauges (see Table 4). The equivalent circuit, derived via the delta-star transformation, shows how this balancing is achieved; viz. by effectively shunting one pair of gauges (R_1 and R_2) until their resistance ratio is the same as that of the other pair (R_3 and R_4). Thus the final position of the shunt slider and the fraction n of the total shunt resistance R_p tapped off is dictated by the relative resistance of the gauges R_1 and R_2 and of the other pair: i.e., n is a function of the initial resistance mismatch.

Now a combination of a strain gauge and a fixed shunt will exhibit less fractional change of resistance than the

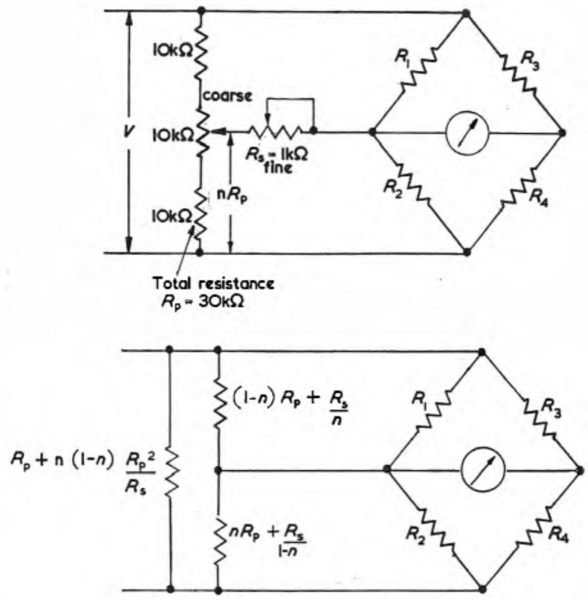


Fig. 4. Zero balancing controls for strain gauge bridge

gauge itself when the latter is subjected to a strain, according to

$$\delta R/R \simeq \Delta G/G \cdot \frac{S}{S+G} \text{ (see Appendix 1)}$$

where Δ, δ are the actual and apparent resistance strains

G = unstrained resistance of the gauge

S = fixed shunt resistor

R = combined resistance = $SG/(S+G)$

or more briefly

$$\delta = k\Delta$$

where k may be termed the desensitizing factor = $S/(S+G)$.

In general if a gauge of resistance pG is shunted down to G ohms the requisite shunt is

$$S = \frac{pG}{p-1}$$

and the desensitizing factor resulting is

$$k = \frac{p - p/(p-1)}{p/(p-1)} = 1/p$$

Hence a resistance strain in the shunted gauge of Δ would yield an apparent strain of $\delta = \Delta/p$.

From the foregoing it may be concluded that:—

(a) The balancing controls should provide only a limited range of accommodation of initial resistive mismatch so as to discriminate against grossly unbalanced gauge sets. Appendix 1 shows that a change of calibration of about half the percentage mismatch may arise.

(b) The balancing controls of channels intended to be interchanged must be identical and any balancing components utilized during a calibration or mock-up test must be the same as those employed in the regular measuring instruments.

(c) Balancing controls should not be unduly low in resistance otherwise valuable sensitivity may be lost. Further, if the shunt component is low several combinations of the balancing controls may be found to yield balance since only a ratio needs to be set, and these can give significant alterations of sensitivity (10 per cent changes have been experienced from an ill-considered choice). Practical values have been found to be those indicated in Fig. 4; the fine control may be omitted entirely if the shunt slider is of the multiturn helical variety.

Zero Shift

Clearly the method of balancing described above might be used to obtain up-scale shift of zero or zero-suppression to accommodate large ranges of readings. However, since this involves alteration of the state of differential shunting it will lead to changes in overall sensitivity. For instance if in Fig. 4 the gauges R_3 and R_4 are assumed to be of equal resistance then it would be required to shunt R_1 and R_2 not to equality but to differ by a factor ρ, slightly different from unity so generating a shift voltage $\rho-1/(\rho+1)\ .\ V/2$ across the detector terminals—taking the detector as having a relatively high impedance. The desensitizing factor arising from this procedure is simply

$$k = \rho$$

If the gauges R_1 and R_2 are initially unequal by a factor p then a desensitizing factor

$$k = \rho/p$$

is appropriate.

A general expression is shown in Appendix 1, but the results are summarized in Table 5. These effects are more serious where the detector impedance cannot be omitted from the calculation and this has been confirmed experimentally: as much as 2 per cent change of sensitivity has been observed to occur after using the balancing controls to obtain an up-scale shift of zero equivalent to 5mV. It may be further observed that the backing off of tare loadings during direct loading calibrations, e.g., the weight of scale pans, can lead to similar errors if the tare is an appreciable fraction of the full range load.

Effects Arising from the Measuring Circuits

Fig. 5 shows the elements of one method of measuring the unbalance of a bridge due to strain after the initial mismatch has been removed by the balancing controls. Essentially the same as the balancing network it is a null method, but whereas the balancing circuits may be required to deal with mismatch unbalances of up to 0·5 per cent the measuring circuit may well be used for full scale unbalances as small as 0·02 per cent of gauge resistance and discrimination to about 1000th of this is needed.

It is clear that the position of the slider of R_p is again a measure of the unbalance. The resistor R_s serves to control the sensitivity, since an increase in R_s will necessitate additional excursion of the slider of R_p to effect a given differential shunting. (Alternatively R_s is sometimes considered as determining the source impedance of a 'balancing current' applied to the corner of the bridge). Thus the relationship of the factor n to the strain in the gauges is determined *inter alia* by R_s.

TABLE 5

Relative Sensitivities or Desensitizing Factors

GAUGE MISMATCH FACTOR p			UNBALANCE FACTOR ρ	EQUIVALENT VOLTAGE SHIFT
0·995	1·000	1·005		
1·000	0·9975	0·9950	0·995	−1·25mV/V applied
1·0025	1·000	0·9975	1·000	No shift
1·005	1·0025	1·000	1·005	+1·25mV/V applied

The position of the unstrained null setting would clearly be at the mid-position of R_p, i.e. at $n = 0{\cdot}5$, if a perfectly symmetrical set of gauges were employed. In this position also variations of R_s would have no effect, since there would be no potential difference across it. With a practical set of gauges it is necessary to adjust the balancing controls until this latter condition exists, testing for it by varying R_s, and the resulting value of n will be set uniquely by the initial mismatch of the gauges.

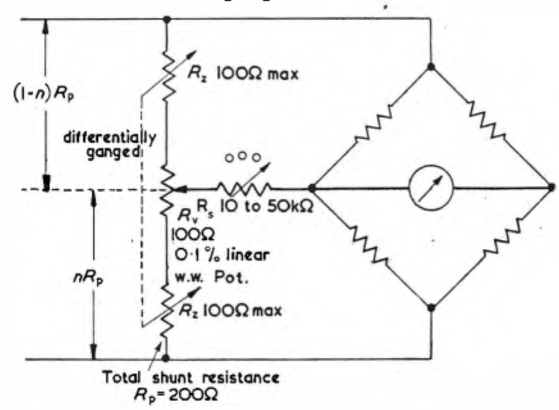


Fig. 5. Measuring system for automatic strain gauge bridge

If this initial position is not convenient means must be found for moving the electrical centre of the system relative to the mechanical centre of R_p. This is done in Fig. 5 by means of the resistors R_z which are differentially ganged so as to keep the total value of R_p constant. However, imperfections in the tracking of the R_z's in practice give rise to small variations in calibration—up to 1 per cent—so that zero adjustment must be followed by recalibration.

Linearity

Appendix 2 derives an expression relating the value of n to the resistance strain x of the gauges in terms of two circuit parameters

$$a = R_s/R \qquad b = R_p/R$$

where R is the gauge resistance, as

$$2bxn^2 + (1 - 2bx)n - (\tfrac{1}{2} + 2ax) = 0 \quad \ldots\ldots \quad (1)$$

or

$$x = \frac{\frac{1}{2}(n - \frac{1}{2})}{a + bn + bn^2} \quad \ldots\ldots\ldots\ldots \quad (2)$$

If x is near zero then n is about 0·5 and the slope in this region is obtained by differentiating equation (2) and inserting these values as

$$dx/dn = \frac{1}{2(a + \frac{1}{4}b)} \quad \ldots\ldots\ldots\ldots \quad (3)$$

If n changes from 0·5 by an amount dn equation (2) yields

$$dx = \frac{dn}{2[a + \frac{1}{4}b - b(dn)^2]}$$

and the percentage error from the linear law of equation (3) is

$$e = -\frac{(dn)^2\ .\ b}{a + \frac{1}{4}b}\ .\ 100 \text{ per cent}$$

The requirements for accuracy are thus manifest: the reduction of b is limited by the lowest feasible value that can be produced in linear potentiometers. For 0·1 per cent linear card wound potentiometers the lower limit is about 100Ω for single turn devices so that the value of R_{μ} will be about 200Ω at least since the resistors R_z require to have the same ohmic value as the slidewire R_T. No improvement can be achieved by limiting dn since the value of a is dependent upon this choice—see equation (3)—and cancels any advantage gained exactly, since a is generally much greater than b.

Appendix 2 also shows that the error can be expressed directly in terms of the slidewire resisitance to gauge resistance ratio for a given range of measurement, thus:

$$e = 2 \,.\, R_T/R \,.\, x \,.\, 100 \text{ per cent}$$

It may be noted that the use of comparatively low resistance gauges aggravates the problem of obtaining adequate linearity. The values shown in Fig. 5 have given adequate performance for strains up to 800 microstrain with 120Ω gauges.

Calibration

The means of obtaining a reliable calibration of the measuring instruments require some care especially when d.c. excitation of the gauge bridge is employed. The use of decade resistance boxes in test bridges is hazardous unless great attention is paid to the contacts—e.g., contact resistance variations of 100μΩ are important and thermo-electric e.m.f.'s of 2μV, corresponding to 1/20°C difference with copper-constantan junctions, are detectable.

The simple expedient of shunting one arm of the bridge with a high fixed resistor often utilized in simple measurements is no longer sufficiently accurate due to the slight but important differences in absolute resistance of the gauges constituting a bridge. The method is, however, valid as a comparative check of the stability of a calibration made by other means.

Two main methods of calibration appear to be usable, particularly where d.c. gauge excitation is used:

(a) The use of a dummy bridge of precise and stable resistors, one or more arms of which can be shunted at will by the operation of thermal-free key switches or similar contactors. Such a device will give an output in the form of a resistance strain but the voltage output will clearly be no more stable than the voltage source used to supply the bridge.

(b) By calibration against a manual potentiometer of good quality, however, because of the difficulty of constructing stable and repeatable low voltage supplies few precision strain gauge measurements are made in terms of absolute millivolts.

Gauge Supplies

D.C. potentiometric methods can be used for strain gauge work if means are available for compensating the variations in the gauge supply voltage. A simple technique is to arrange for the potentiometric instruments to be standardized not against a standard cell but against some suitable fraction of the gauge supply voltage. The calibration is then sensitive only to variations of supply voltage in the inter-standardizing period—say half an hour, or to any variation in the voltage division ratio of the network used to step down from the supply voltage, usually about 6V, to the standardizing level, around 1V. The latter is readily met by good quality (0·02 per cent stability) resistors while low voltage supplies of good short-term stability may be derived from modern mercury cells or silver-zinc accumulators: a car battery of moderate capacity (60Ah) can give satisfactory results if used in the middle of the gravity range. Zener diodes and other forms of low voltage reference offer the possibility of mains driven low voltage supplies of adequate short term stability. It may be noted that, given means of standardizing against the gauge supply, the absolute value of voltage is immaterial within fairly wide limits. Although it is outside the scope of the present article it is relevant to note that a developed form of a commercial chart recorder has been adapted to give an overall accuracy of 0·1 per cent as a strain gauge recorder when operated with car batteries at anything between 5 and 7V. Some progress has also been made towards a 6V supply at upwards of 50mA having an absolute stability of 1 part in 1 000 or better.

Conclusions

The present survey of the problems involved in precise strain gauge measurements shows that existing techniques are sufficient to make observations to 1 part in 1 000—but only just. By attention to detail in the preparation of the test specimens, in the design of the measuring instruments and of the means of calibration these precisions can be maintained for considerable periods of time. Any significant advance on the present accuracies is only possible if as well as improved strain gauges, instruments of a startling specification become available.

Acknowledgments

The authors are grateful to the Chief Executive of the Aircraft Research Association, Mr. R. Hills, for permission to publish this work. The careful and patient measurements undertaken by members of the instrumentation department, Messrs. K. Milsom, W. R. Wignall and G. B. Sims made possible the successes achieved.

Appendix 1

Desensitizing of Strain Gauges by Differential Shunting

If an active strain gauge of ohmic value G is shunted by a fixed resistor of S ohms the resistance of the combination is:

$$R = \frac{S \,.\, G}{S + G}$$

If now G is strained to become $G(1 + \Delta)$ where Δ is very small order 10^{-3} to 10^{-4}

$$R(1 + \delta) = \frac{SG(1 + \Delta)}{S + G(1 + \Delta)}$$

and

$$\delta R/R = [1 + (G/S)(1 + \Delta)]^{-1}$$

which reduces on expansion to

$$\delta = \Delta \,.\, \frac{S}{S + G} = k\Delta$$

where k is the desensitizing factor.

Bridge Strain Sensitivities for Gauges of Unequal Gauge Factor

If in Fig. 6(a) the two active gauges exhibit additive but unequal strains δ_j and δ_k then the unbalance voltage will be given by

$$dV/V = \frac{120(1 + \delta_k)}{120(1 + \delta_k) + 120(1 - \delta_j)} - \frac{1}{2}$$

$$= \frac{\frac{1}{2}(\delta_k + \delta_j)}{2 + (\delta_k - \delta_j)}$$

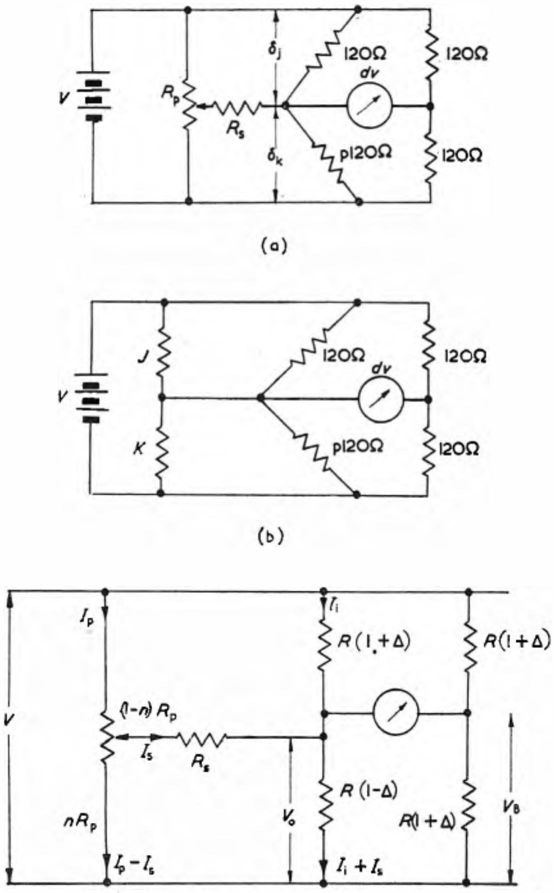
and if δ_k and δ_j are both $\ll 2$ then dv/V approaches a mean strain sensitivity

$$dV/V = \tfrac{1}{4}(\delta_k + \delta_j)$$

If the inequalities of resistance strain arise from an ohmic mismatch factor p then $\delta_k = \delta_j/p$ and

$$dV/V = \tfrac{1}{4}\,\delta_j\,(1/p + 1)$$

and the bridge sensitivity differs from that of a perfectly matched bridge by a factor $\frac{1}{2}(1/p + 1)$
e.g. if two gauges are of resistance 120 and 120·6Ω then $p = 1{\cdot}005$ and $dV/V = 0{\cdot}9975\,\delta_j/2$
i.e. Percentage calibration change is about half the percentage of mismatch.



(c)

Fig. 6. Analyses of strain gauge bridge
(a) Effect of unequal strains
(b) Effect of deliberate unbalance
(c) Relationship of n to strain

EFFECT OF DELIBERATE MISMATCH

The balancing controls may be replaced by the equivalent delta network shown in Fig. 6(b). If now the shunted gauges are set not to equality but so that

$$\frac{K \cdot p120}{K + p120} = \rho \cdot \frac{J \cdot 120}{J + 120}$$

where ρ is a factor slightly greater or less than unity, a shift voltage $V_s = V/2\,(\rho - 1)/(\rho + 1)$ is produced. Now $J/(J+120)$ and $K/(K+p120)$ are the desensitizing factors acting on the respective gauges so that

$$p\delta_k = \rho\delta_j$$

and $dV/V = \dfrac{\delta_j/2\,(\rho/p + 1)}{2 + \delta_j\,(p - 1)}$

An exact analysis of the circuit yields the general expression

$$dV/V = \Delta \,.\, V/2 \,.\, \frac{J}{120 + J} \cdot \frac{\rho(p + 1)}{p(\rho + 1)} + \frac{\rho - 1}{\rho + 1} \quad V/2$$

The first term of this gives the strain response of the bridge the second the constant upscale shift. Thus even with a perfectly symmetrical arrangement ($p = 1$) the strain sensitivity is seen to alter with the unbalance factor.

APPENDIX 2

Referring to the arrangement of Fig. 6(c) the balance condition is simply

$$V_o = V_b.$$

Applying Kirchoff's laws

$$I_p(1 - n)R_p + I_sR_s - I_1(R + \Delta R) = 0 \;\ldots\ldots\; (4)$$

$$I_1(R + \Delta R) + (I_1 + I_s)(R - \Delta R) = V \;\ldots\ldots\; (5)$$

$$I_p(1 - n)R_p + (I_p - I_s)nR_p = V \;\ldots\ldots\ldots\; (6)$$

Solving for I_s yields

$$I_s = V \cdot \frac{(R + \Delta R)/2R - (1 - n)}{R_s + \frac{1}{2}R - (\Delta R)^2/2R + n(1 - n)R_p}$$

and from equation (5)

$$I_1 = (V/2R) - I_s \cdot \frac{(R - \Delta R)}{2A}$$

Now

$$(I_1 - I_s)(R - \Delta R) = V_o \text{ and } V_b = \frac{(R - \Delta R) \,.\, V}{2R}$$

so that at balance

$$\frac{2x}{1 - x^2}[a + \tfrac{1}{2} - (x^2/2) + n(1 - n)b] = (x/2) + \tfrac{1}{2} - 2n$$

where $x = (\Delta R/R)$ $\quad a = (R_s/R)$ $\quad b = (R_p/R)$

In a practical arrangement a is of the order 200 and b of the order 1, while x is of the order of 10^{-4} so that one may write without error,

$$2x = \frac{n - \frac{1}{2}}{a - b(1 - n)n}$$

LINEARITY

Differentiating

$$2x[a + (bdn/dx) - 2bn\,(dn/dx)] + 2(a + bn - bn^2) = dn/dx$$

If $x = 0$ then $n = 0{\cdot}5$ assuming all gauges of equal resistance and at this condition

$$dx/dn = \frac{1}{2(a + \frac{1}{4}b)}$$

EFFECT OF CHOICE OF SLIDEWIRE

The equations of the text shown linearity error to be

$$e = \frac{100 \,.\, (dn)^2}{a/b + \frac{1}{4}} \text{ per cent}$$

But

$$x/dn = \frac{1}{2(a + \frac{1}{4}b)}$$

where x is the full scale strain to be measured.

Full scale $dn = R_v/R_p = R_v/Rb$, where R_v is the variable part of R_p. Inserting these in the error-expression

$$e = 100 \,.\, R_v/R \,.\, 2x \text{ per cent}$$

where R is the unstrained gauge resistance.

With the values of Fig. 5 the expected error from linearity for a maximum span of 0 to 800 microstrain with 120Ω gauges would be

$$e = 100 \,.\, 100/120 \,.\, 16 \,.\, 10^{-4} = 0{\cdot}133 \text{ per cent}$$

(taking the gauge factor as 2·0Ω/Ω/unit strain)

The peak errors would be plus and minus half this amount if the calibration were set up at the extremes of the range and at the mid-span, since the characteristic curve is S shaped about the mean slope at zero strain.

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Accuracy Control in a File Processor

By J. C. Hammerton*, M.A.

In this article the accuracy controls incorporated into a particular logical machine called a file processor are described. The methods of handling errors detected by the accuracy controls are also described. The purpose of the machine and the fundamentals of its design are outlined.

THE problem of accuracy control in large electronic systems has been discussed in a previous article[1]. In the present article the philosophy already developed is illustrated by application to a particular logical machine. The functions and the construction of the machine are described in general terms in order to provide the necessary background information

The machine, which is of the type known as a file processor, is considered as part of a system in which the working records of a small insurance company (300 000 policies) are stored as magnetized locations on reels of magnetic tape.

The records are consolidated into one master file of policies ordered by policy number. By processing the file according to a regular cycle of operation the company maintains up-to-date information about policy status, premium billing, premium accounting, dividend history, loan history, etc.

The principal functions which have to be performed by the system are as follows:

(1) To order new transaction information, e.g. policy changes, into an ascending sequence by policy number thereby producing a 'transaction' tape.

(2) To locate the data on the policies being serviced which are recorded in the master policy file.

(3) To amend the existing data in accordance with the new transaction information.

(4) To extract from the master file information on policies which require some action, e.g. premium billing.

(5) To perform calculations on 'active' policies, e.g. dividend calculations.

These functions can be fulfilled in several ways, all of which can be considered as compromises between two extremes. At one extreme all the functions are incorporated in a single machine and they are performed in a predetermined sequence every time an 'active policy record is located in the master policy file (Fig. 1(a)). At the other extreme the functions can be distributed between several specialized machines.

One method of distributing the functions is to use a special purpose machine to find the active information in the master file and to 'extract' it (functions 2 and 4). Consequently, only active information is presented to the computer in the system. With this method it is, of course, necessary to re-introduce the up-dated information to the master file by 'merging' it with the previous information.

A machine of this type is called a file processor. It can be employed economically when a cycle of operation involves only a small fraction of all the policies listed in the master file. The method of operation is depicted in Fig. 1(b).

There are three input tapes and two output tapes.

The blocks of data in the policy file are matched against those in the transaction tape. When an equality is detected the block of data from the policy file is transcribed to an 'extract' tape. The extract tape and the transaction tape are used by the computer to up-date the master records. At the same time the up-dated information from the previous cycle of the operation is merged with the old policy file to produce an up-dated policy file.

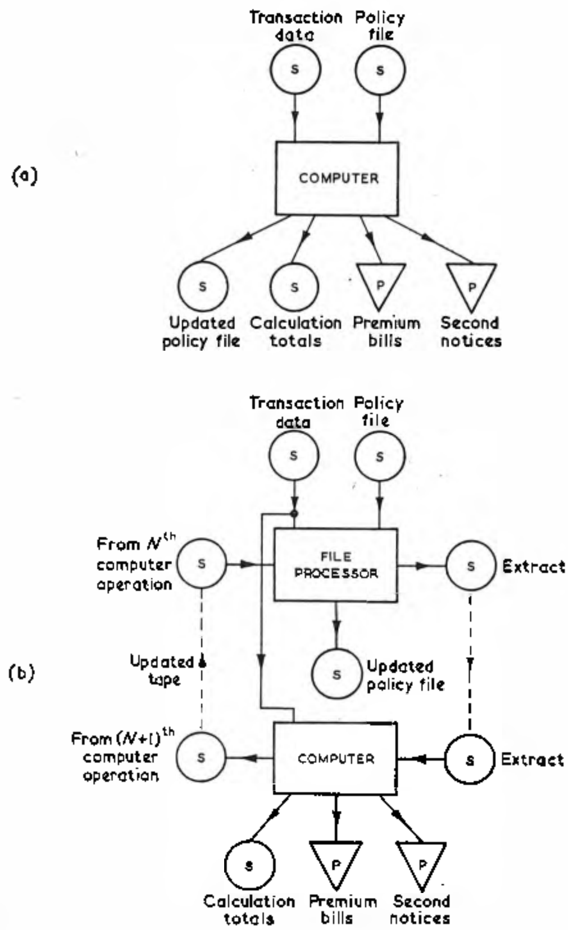

Fig. 1(a). All purpose computer
(b). Computer and file processor

Description of the File Processor

The only information required by the file processor is the policy number of each block of data. This is positioned at the beginning of a block so that the file processor reads the policy number as soon as it starts inspecting the block.

The rules governing the operation can be readily determined, but will not be enumerated here. The principle of operation is, however, important. The machine does not have a static memory large enough to store a block of data. Therefore, a block of data is held on an input tape in a position to be read until the rules indicate that it is to be transferred to an output tape. The operation proceeds by selective re-transcription of blocks of data from input tapes to output tapes. Thus, taking the 'extract' function as an

* Radio Corporation of America.

example, the method of operation is as follows:

(a) To find and store the policy number of the first block of data on the transaction tape, L_1.

(b) To run the master file inspecting the policy number of each block of data and comparing it with L_1; when a number F_1 is found equal to L_1, the block of data of which F_1 is a part is transcribed to the "extract' tape.

(c) To find and record the policy number of the second block of data on the transaction tape and so on.

A pre-requisite is that the blocks of data on all the input tapes are arranged in ascending order by policy number.

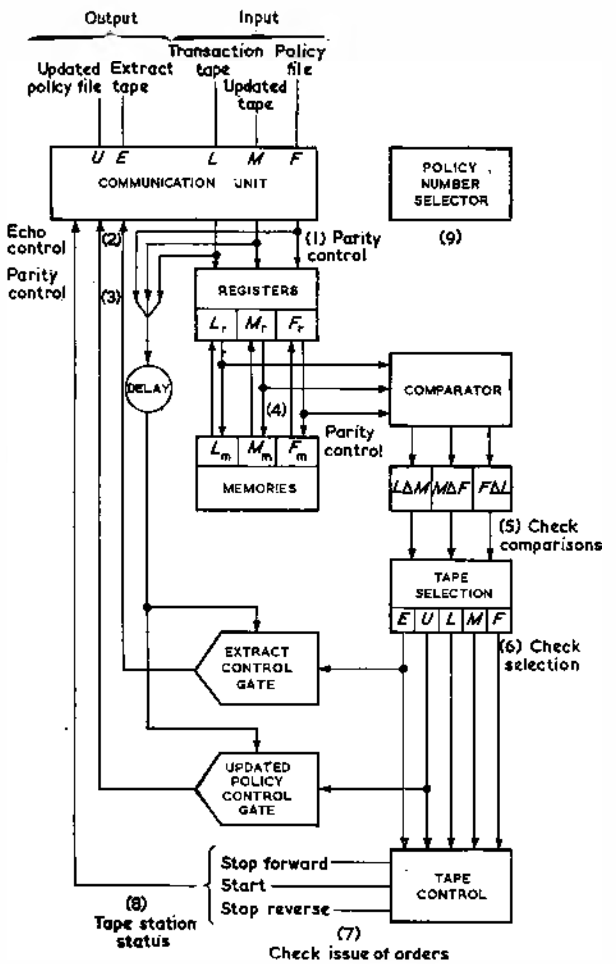

Fig. 2. Schematic of file processor

The equipment necessary to mechanize the extracting and merging functions is shown in Fig. 2.

It consists of the following parts:

(1) A *communications unit* which connects the machine to the required tape stations, three input and two output.

(2) *Registers*, L_r, M_r and F_r to staticize the information received from the input tapes.

(3) *Memories* L_m, M_m and F_m to store the policy numbers of the three blocks of data which were last inspected (one from each tape).

(4) A *comparator* which compares the three stored policy numbers and ascertains their relative magnitudes.

(5) A *tape selection* function which interprets the output of the comparator in terms of the operating rules and decides which tapes to start and which to stop.

(5) A *tape control* function which issues the necessary orders to the tape stations at the appropriate times.

(7) A *delay* so that after reading the beginning of a block of data, all the above functions can be performed before an attempt is made to transcribe the block to output tape.

The delay can, for example, be a drum on to which the block of data is transcribed. By the time that the first character appears at the reading head the policy number has been selected and compared. Consequently at this time it has been decided either to transcribe the block of data to the extract tape or to the up-dated policy file, or to both, or to neither. This selective transcription is accomplished by conditioning the two control gates appropriately (Fig. 2).

Accuracy Checks

Having defined the purpose of the file processor and outlined its mechanization, the problem of accuracy control in the machine can be examined.

Communications

In accordance with the philosophy already developed[1] the first checks to establish are those which secure the communications between machines.

At any one time the file processor is connected to five tape stations, three input and two output. However, during the course of an operation each tape trunk may be connected to a number of tape stations. For example, the *F* trunk which handles the policy file is connected in turn to each of the tape stations which house the file. (It is assumed that the system is such as to eliminate hand-carrying of tapes either wholly or in part).

In a system of this type it is economical to reduce the checking equipment in the tape stations to a minimum. Hence the input trunks to the file processor are monitored by parity checkers (1) and the output trunks by echo checkers (2). (The numbers in parentheses refer to Fig. 2).

The echo-checking takes the form of registers which are 'set' by the outgoing unit bits and reset by versions of these bits returned to the user machine from the tape station[1].

The commands issued to the five tape stations are monitored in a similar manner. However, since the status of the stations is frequently of interest in connexion with the logic of the machine, a more embracing check can be employed. The status signals (usually d.c. levels) received from the tape station are 'moving or stopped and 'forward or reverse'. The station adopts these conditions in response to three commands from the file processor, namely, 'stop forward', 'start', and 'stop reverse'.

With the status signals available a check can be established which covers the transmission of the commands to the tape stations and the receipt from them of their status (8). The error inspection pulse (Fig. 3) inspects the two control gates to determine whether the status of the tape station is compatible with the orders issued to it.

Information Channels

The information channels in the file processor are relatively simple. The minimum accuracy control is provided by a parity checker which monitors the output information (3).

The selected parts of the information, i.e., the policy numbers, are transferred from the registers L_r, M_r, F_r, to the memories L_m, M_m, F_m, and circulated as required from memory to register and back again. The importance of the policy numbers in controlling the operation and the complexity of the circuits which handle them make it advisable to use parity checkers to monitor the circulation paths (4).

Logical Functions

The three comparator circuits each have three possible outputs.

For example,

L is greater than M

L is equal to M

L is less than M.

Since these three possibilities are mutually exclusive more than one of them at a time indicates a malfunction. Furthermore, one of them must be true.

Every time a new policy number is read into the machine the outputs of the comparators are checked to see that one and only one of the three possible outputs has been registered (5).

Checking the tape selection function cannot conveniently be as rigorous. The nature of the operation is such that either, neither or both output tapes may be in motion at any particular time. This makes it difficult to devise a comprehensive test to cover selection of the output tapes. With regard to the input tapes, however, it can be stated that,

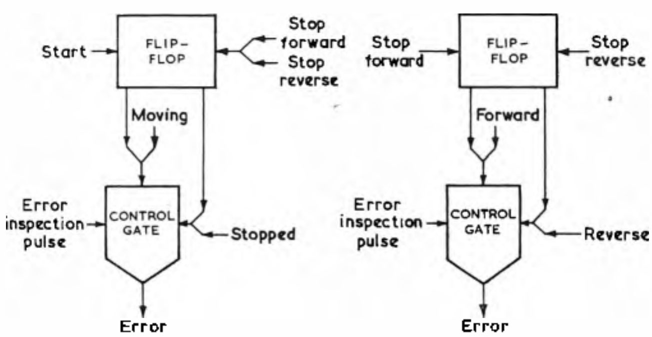

Fig. 3. Tape station status check

(a) If the machine is operating, it must be supplied with information.

(b) The machine cannot handle information from more than one source at a time.

Consequently, a check is used to establish that one and only one source of input information is selected (6).

In general the selection of a tape is only one of the conditions which must be fulfilled before a tape is run. Hence it is advisable to check that the selection of a tape is followed by ordering the appropriate tape station to start (7). Furthermore, in the case of tapes which have been running but are not reselected, it is important to check that they are ordered to stop. For example, a particular block of data may have been transcribed to the up-dated policy tape and to the extract tape. If the following block is only required in the up-dated policy file, the extract tape is not reselected. Failure to order it to stop will, however, enable transcription to be effected.

Assessment of Control Checks

A reappraisal of the accuracy control checks which have been enumerated so far reveals an important omission.

Each block of data contains the information pertinent to a policy. The operation outlined is concerned with the routing of these blocks to output tapes according to their policy numbers. To do this the policy number is selected from each block and used as a factor in determining which of the tapes to run. A selection is, therefore, made from every block of data. If no selection is made during the $(n + 1)^{th}$ block, it is transcribed in the same manner as the n^{th} block since no further orders are issued to the tape stations.

To prevent this a check is made at an appropriate time after the start of a block of data to see that the policy number has been selected from it (9).

Supplementary Control Checks

The nine control checks described cover the logical functions and the information handling abilities of the file processor. At this point an estimate of the extra equipment involved can be made. It is then possible to determine whether supplementary checks can be included in the design.

The supplementary checks are directed towards one or more of the following objectives:

(1) Better identification of the cause of a shut-down.

(2) Checking parts of the machine which because of the nature of the equipment used or the small amount of equipment, are considered highly reliable.

(3) More rigorous checking of complex parts of the machine.

For example, a parity checking circuit positioned after the delay in the information channels enables maintenance personnel to distinguish errors introduced in the delay from those introduced in the control gates.

The second objective is exemplified by the three memory circuits. These, if constructed from magnetic cores, are considered highly reliable. A parity checker in each circulation loop is normally quite adequate to guard against pick-up from 'noisy' cores or loss of unit bits from marginal cores. However, it is possible for a whole character to be missing due, for example, to failure to address the locations in which its bits are stored.

As a final example, some additional check of the complex comparator circuits can be included. The check described above ensures that the output of each of the three comparators is unique. A further check can be made to see that the three outputs are compatible in an algebraic sense, e.g. if L is less than M and if M is not greater than F, then L cannot be either equal to or greater than F.

Error Correction and Indication

The primary purpose of the accuracy control checks is to establish that an error of some sort has occurred in the machine. It is then possible to initiate a correction routine at the appropriate time. The objective of correction routines is to erase the basic group of characters involved in the malfunction and to repeat that part of the operation which was being processed[1]. For the purpose of illustrating what is involved in the design of the machine, two cases are considered. The first is an error which occurs in the policy number of a block of data. The second is an error which occurs in transcribing a block of data from an input tape to an output tape.

Organization of Data

For convenience a review of the way in which the data are organized in a system of this kind is included. The data are grouped into 'blocks' each containing the information pertinent to a particular policy number. For example, a block contains the policy number, the issue date, the plan, age at issue, name of policyholder and so on. These details are recorded as a number of characters, each of which consists of a coded combination of unit and zero bits. For the present purpose it is supposed that the bits forming a character occur in parallel but that the characters occur in series.

The blocks of data are separated by gaps of blank tape, the minimum length of which is specified to ensure that the tape can be stopped and restarted between blocks without loss of information.

Timing of Functions

The delay (Fig. 2) in the information channels and the time taken to traverse the minimum gaps between blocks are such that the policy number of the $(n+1)^{th}$ block occurs at the input registers while the back end of the n^{th} block is still in the delay prior to transcription. This means in effect that all the logical functions relating to the $(n+1)^{th}$ block are completed before the n^{th} block has been fully transcribed.

Fig. 4 provides a graphical representation in which time is the abscissa. At a time after the start of the $(n+1)^{th}$ block at which it is known that the n^{th} block has been transcribed, an 'error assessment pulse' (e.a.p.) is generated. This pulse inspects the system to see if an error has occurred and, if so, whether the error is in the transcription of the n^{th} block or in the logical functions connected with the policy number of the $(n+1)^{th}$ block.

The basic error cycle associated with the error assessment pulse of the $(n+1)^{th}$ block only covers these two parts of the operation. A malfunction, for example, in the process-

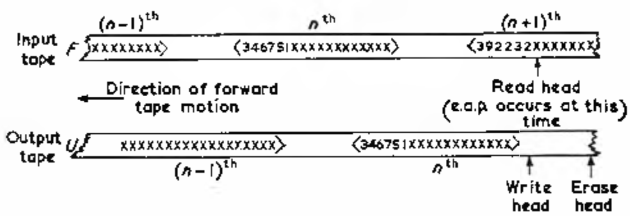

Fig. 4. Timing of read/write function

ing of the policy number of the n^{th} block would have been detected by the error assessment pulse of the n^{th} block.

Error in the Logical Functions

If on assessment the error proves to be associated with the policy number of the $(n+1)^{th}$ policy the following actions are necessary:

(1) The moving output tapes are stopped but remain 'forward' (it is inferred that the n^{th} block was transcribed without error).

(2) The input tape (F in Fig. 4) is stopped and then reversed until the reading head is in the gap preceding the $(n+1)^{th}$ block.

(3) The three input tapes are run again in the predetermined sequence in order to rediscover the policy numbers of the next blocks of data on the three input tapes; these numbers are stored in the three memories (Fig. 2).

At this point the machine has sufficient information available to it to continue the operation.

In re-running the F tape, of course, the error may be re-detected if it is a real error rather than a transient one. If the error is redetected, the error correction routine is re-initiated.

Error in Transcription

If the system, on inspection, is found to have a transcription error, e.g. a parity error in the output information, this must have occurred in the n^{th} block of data. The error is assessed at the e.a.p. (Fig. 4) as before. (The correction of this type of error takes precedence over an error discovered in the $(n+1)^{th}$ policy number).

The error correction routine in this case is more complex. The objective of it is to repeat the transcription of the n^{th} block of data. To do this the reading head of the input tape station (F) is positioned in the gap preceding the n^{th} block. At the same time the transcribed n^{th} block is erased from the output tape.

These functions are accomplished by the following actions:

(1) Run the input tape in reverse by issuing the commands 'stop reverse' and 'start'.

(2) Stop the input tape when the reading head is in the gap between the $(n-1)^{th}$ and the n^{th} blocks of data and return it to its 'forward' status.

(3) Put the output tape in reverse.

(4) Turn off the 'erase' function.

(5) Adjust the status of the output tape station so that the information on the tape (U of Fig. 4) can be read.

(6) Start the tape in reverse.

(7) When the 'write/read' head detects the gap between the transcribed $(n-1)^{th}$ and n^{th} blocks, continue running the tape for a sufficient time to ensure that the 'erase' head has entered the gap and stop the tape: at this time the 'write/read' head is positioned in the $(n-1)^{th}$ block of data.

(8) Allow the status of the output tape station to revert to its writing condition.

(9) Switch on the erase function.

(10) Advance the tape a distance which is sufficient to ensure that the 'write/read' head clears the $(n-1)^{th}$ block; part of the transcribed n^{th} block is erased in the process.

(11) Stop the tape.

(12) Generate an initiation pulse and re-run the three input tapes in the predetermined sequence in order to re-discover the policy numbers of the next blocks of data on the tapes.

At this point the machine has sufficient data available to it to continue the operation.

If the n^{th} block of data is transcribed to both output tapes, the actions relating to the output station must be performed for tape E as well as tape U

Shut-down

A correction routine is repeated either until the error is not re-detected or until, after a number of unsuccessful correction attempts, the machine is shut down. The number of attempts necessary to infer that a detected error is a real one rather than a transient one is a matter for discussion. If the error is due, for example, to loose oxide on the reading head of the input tape station, the oxide may be dislodged by moving the tape back and forth over the head. The important point is that once the machine has been shut down some time will elapse before it can be brought back into operation. As compared with this time, error correction routines are cheap. Hence it is better to err on the side of too many attempts rather than too few.

Error Indication

If the transcription or the criterion error is of transient origin, the operation is continued after the appropriate error correction routine has been made. If, however, the error arises from some permanent cause the operation is discontinued.

The information about the type of error becomes important at this time in order to assist in determining how the operation can best be continued.

In general, the error can be attributed to either a malfunction in the processing machine or a malfunction in a tape station or an error in the input data.

In the first two cases the operation can be continued either by repairing the defective machine or by substituting a machine of the same type for it. In both cases it is necessary to re-initiate the operation at some point in the middle of it.

If the error is attributable to the input data, an 'error-override' option can be exercised (see next section).

The most economical manner of indicating errors uses a register of the requisite number of bits. A four-bit register is adequate for the different errors enumerated above. If the extra equipment is available, the errors can usefully be grouped according to the three possible origins discussed above.

By-passing an Error on Tape

The machine is shut-down at the time when both input and output tapes have been reversed and have been stopped forward. The reading head of the input tape station is positioned in front of the block of data containing the error. The output tape has had the transcribed error block (if any) partially or wholly erased.

To continue the operation a function is included in the processing machine whereby the block of data containing the error can be processed regardless of the error. Subsequently the machine is returned to normal before continuing the operation. Some record of the shut-down is made to enable the error to be located and evaluated at a later time.

The ability to get past an error is of great importance in obtaining the most work from a machine. It is, of course, an abnormal procedure. If desired, it can be made fully automatic.

Conclusion

This article has endeavoured to demonstrate how the general principles previously established for accuracy control in a data-handling system can be applied to a file processing machine.

Acknowledgments

The author is indebted to the Radio Corporation of America for permission to publish this article and to his associates for their helpful criticism.

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A Television Translator

The BBC began transmitting programmes on July 14 1958 from a new type of low-power television transmitter, known as as a 'translator' which is undergoing extended service trials at Folkestone. This town is typical of small populated areas which are within or adjacent to the service areas of the main BBC stations, but are prevented by surrounding hills from obaining satisfactory reception

A translator converts the sound and vision transmission frequencies from one channel to another without demodulation to audio and video frequencies, which occurs when a normal receiver and transmitter relay installation is employed. This simplification increases the reliability of the equipment which can therefore be arranged for automatic operation without attendant staff. Because the equipment is small it can conveniently be housed in weather-proof and insect-proof cabinets, thus dispensing with the need for a station building. The BBC utilizes cabinets of existing design and special precautions have been taken to provide sufficient cooling. The power supply is included in the steel cabinets, so that the translator can work direct from the mains.

The translator must be on high ground where good reception is possible from an existing station in the BBC network and from where its transmissions can be radiated over line-of-sight paths to the area to be served. In this way only a very low-power output is required, which therefore does not add to the already serious co-channel interference problem.

The Folkestone translator is installed at Creteway Down where the receiving aerial is in line-of-sight from the Dover BBC television station and the transmitting aerial has a commanding position overlooking Folkestone. The receiving aerial system consists of a double 3-element array and the transmitting aerial has four tiers of single, folded dipoles.

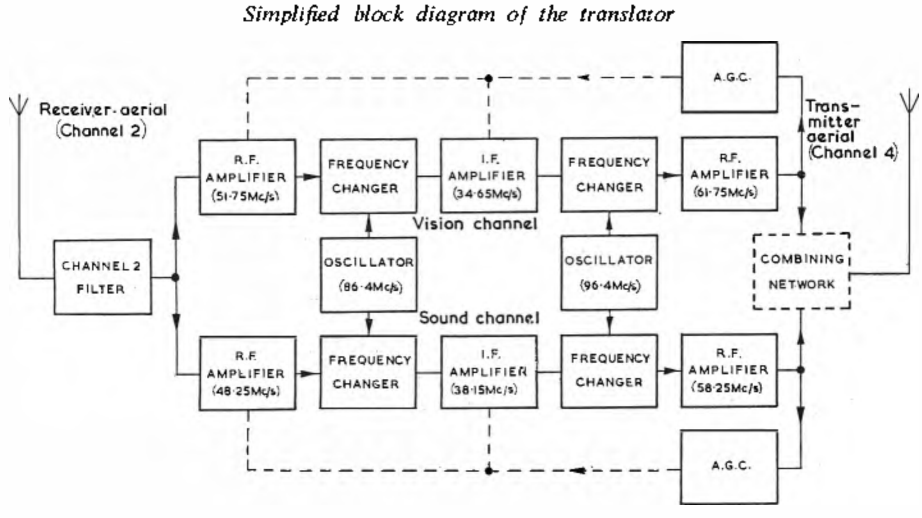
If the sound and vision signals of a television system are amplitude modulated, difficulties are introduced if they share a common amplifier in the translator because of the greater possibility of intermodulation. These difficulties are eased in systems such as those used on the Continent in which the sound signal is frequency modulated. It is hoped that these and other difficulties have now been overcome with the system developed by the BBC. The simplified schematic diagram for this equipment shows that separate channels have been provided for the amplification of the sound and vision signals using common frequency-changing oscillators. The separate channel arrangement reduces the risk of intermodulation and enables separate automatic gain control to be employed to combat the effects of differential fading between the sound and vision signals, which occur if a translator were dependent upon reception from a really remote BBC station. The automatic gain control voltages are also used to initiate an automatic changeover to reserve translator equipment should the normal unit become faulty.

The double frequency-changing process facilitates the rejection of spurious signals and provides additional protection against "in band" feedback. The first frequency-changing process resembles that in a normal television receiver, producing vision and sound intermediate frequencies of 34·65Mc/s and 38·15Mc/s respectively, and the second frequency-changing stage produces vision and sound frequencies in the required channel.

The Folkestone translator peak white vision power output is 1·5 watts and in conjunction with the type of transmitting aerial used, gives an effective radiated power of 7 watts in the direction of maximum radiation It is hoped that the installation of similar low-power translators which have been developed by the BBC's Designs Department will enable at least some of the small populated areas in which satisfactory reception is difficult to be given an improved service.

The BBC intends also to develop translators to extend the BBC'S v.h.f. Sound Broadcasting Service to small areas which are prevented by intervening high ground from obtaining direct reception from the main stations.

Simplified block diagram of the translator



An Amplitude/Frequency Response Display Using a Ratio Method

(Part I)

By H. L. Mansford*, B.Sc. (Eng.), A.C.G.I., A.M.I.E.E.,
and K. M. I. Khan†, B.Sc. (Eng.), A.M.I.E.E.

A wobulator for amplitude testing often gives errors because of unwanted signal amplitude variations during the sweep cycle. Part 1 of this article outlines a new ratio method for balancing-out the errors. Part 2 describes equipment for presenting a family of curves generated from independent variable potentials by using a stroboscopic sampling technique. A modified television receiver displays the waveform with its reference cursors.

Signal Ratio

If a sweep signal frequency generator or wobulator be constructed to cover a useful video frequency range of say 50kc/s to 10Mc/s or more, the amplitude variations of this test signal are found to exceed the errors tolerated in the equipment under test. These may be minimized by feeding the signal to a wideband amplifier fitted with automatic gain control; but there are residual errors, and difficulties may arise from the absence of the signal during the frequency sweep return time.

Alternatively, source errors may be eliminated by using the input and output signal amplitudes of the equipment under test to generate a ratio signal. The technique is made most effective by converting the potential values to a logarithmic scale, as this allows the ratio to be expressed as the difference between two potentials representing logarithmic values.

If also the display is to be that of the logarithm of the signal amplitude ratio, as a function of frequency, then:—

$$y = a \log E_o/E_i$$

where y is a potential controlling the height of the c.r.t. spot on the screen.

E_i is proportional to the input signal amplitude.

E_o is proportional to the output signal amplitude.

a is a controlled variable factor.

The expression may be rewritten:—

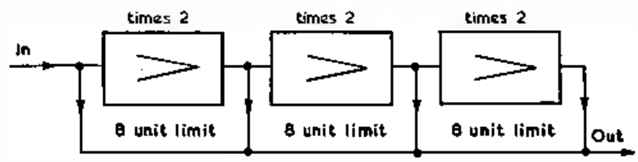
$$y = a (\log E_o - \log E_i)$$

which indicates that the signal source may be split into two channels; the one having a value representative of the signal E_i at the input to the unit under test, and the other to be a function of the output signal E_o. Then two identical linear-to-log transfer circuits may be used for the conversion of these two signals, followed by a difference circuit from which is taken a suitable display potential; or the appropriate logarithmic signal may be generated with reversed polarity so that a simple resistance adding circuit may be used, as in the method to be described.

Early circuits[1] for a conversion to a logarithmic scale took the form of selected valves chosen to give the required ratio of anode current to control grid potential. Such a method has obvious disadvantages.

Another method[2] makes use of a multiplication of the signal amplitude, by cascading a number of similar stages of voltage amplification, each designed to have a fixed upper limit for the value of output amplitude. Incremental signals may then be tapped from successive stages and an addition made in a combining circuit.

A block schematic circuit diagram for signal multiplication and addition is outlined in Fig. 1. Alternatively some advantages are offered by a similar system which relies on signal division and addition; and in this case the division may be provided by passive networks and diode switching[3].



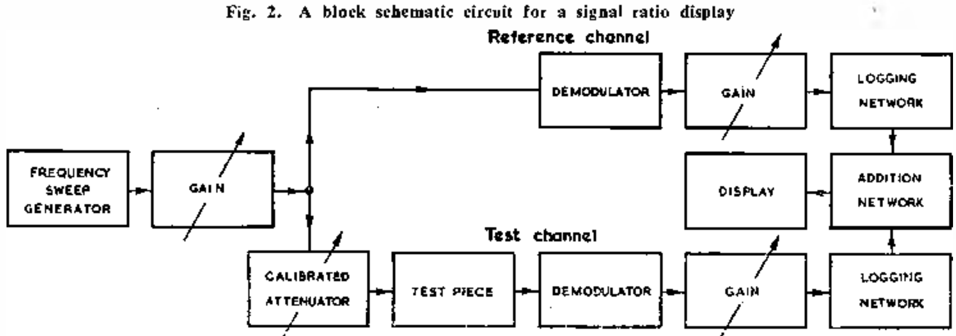
IN	dB	OUT	Δ
1 unit	0	1+2+4+8=15 units	0
2 unit	6	2+4+8+8=22 units	7
4 unit	12	4+8+8+8=28 units	13
8 unit	18	8+8+8+8=32 units	17

Fig. 1. A block schematic circuit giving a logarithmic signal relationship by muliplication and addition

A potential divider having attenuation which varies automatically with the amplitude of the input signal may be constructed with diode switches to bring into circuit additional parallel paths to produce greater attenuation by reducing the shunt resistance. The law connecting the input and output potentials is controlled both by the resistance values chosen for these shunt paths and by the bias potentials applied to the diodes. This is the method of this article and detailed design data will be given.

Another attractive method of conversion employs a potential from a resistance-capacitance exponential charging curve, together with a flip-flop valve circuit to balance the input potential against that of the exponential.

Fig. 2. A block schematic circuit for a signal ratio display



* Bell Punch Co. Ltd.
† E.M.I. College of Electronics
both formerly with E.M.I. Electronics Ltd.

The system[4] is an information sampling (stroboscopic) technique.

For the division and addition system, now to be described, design parameters are taken as:—

- Maximum display range 0 ± 20dB.
- Maximum sensitivity to show 0·05dB changes.
- Maximum source amplitude variations 6dB.
- Maximum display error due to this 0·1dB.

Fig. 2 shows a block schematic circuit of the ratio forming apparatus, which starts from a wobulator supplying a nominal 0·1V (r.m.s.) signal level. This is too little since the test equipment may give further attenuations at some frequencies, and the demodulators require sufficient signal for linear operation. Accordingly, an adjustable gain amplifier follows the wobulator, before the signal division into the two channels in order that errors in its frequency response shall not add to those of the unit under test.

A calibrated 75Ω attenuator is included in the test channel to adjust and balance the signal levels at the two demodulators, and as each is followed by an amplifier having continuously variable gain control, this enables a complete balance to be made of the signals fed to the logging networks. The circuit details of the demodulators, amplifier, feed valves, logging networks and the adding network, are given in Fig. 3.

From the adding network a signal proportional to $y = a \log E_0/E_1$ is fed to the display unit which is the subject of Part 2 of this article.

The signal at the demodulator input is a sine wave having a frequency varying with time, and with some unwanted amplitude variations during the sweep. In the test channel, additional amplitude variation is added by the attenuator and by the equipment under test. The test signal is also suppressed for a time to allow the frequency to sweep back to its first value.

Following the demodulator, the signal output of the

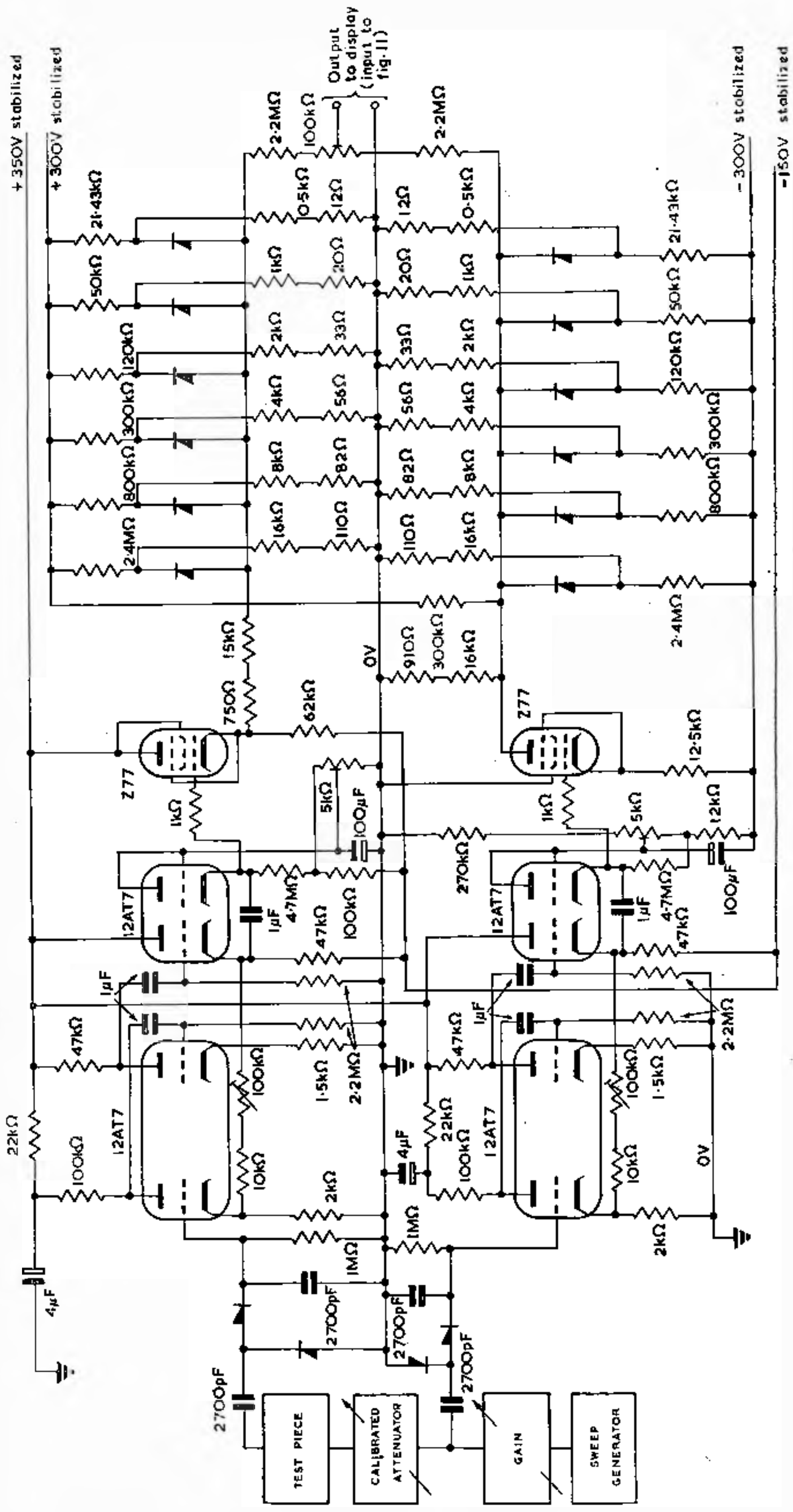


Fig. 3. Details of the Fig. 2 circuit

reference channel takes the form shown in Fig. 4. It then has a 50c/s major component based on the sweep repetition frequency. Other components extend up to say 5kc/s, which is a limit dependent upon the rate of change of the signal amplitude.

The output of the demodulator in the test channel differs from Fig. 4 by the gain changes introduced by the test piece. This difference forms the basis of the amplitude/frequency graph to be displayed.

Fig. 5 illustrates the nucleus of the design for a logarithmic law network, relating E_o and y'. That is $y' = \log_2 E_o$. (The log base two has been chosen for the sake of ease in the practical construction. No account has been taken of the crystal diode resistances in the simple design theory. The results justify the approximation thus made).

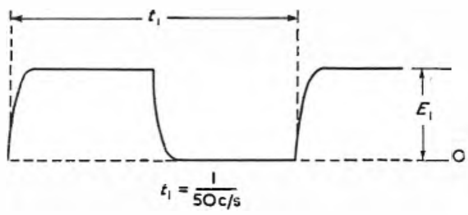


Fig. 4. The signal waveform at the demodulator output, showing its suppression during the return time

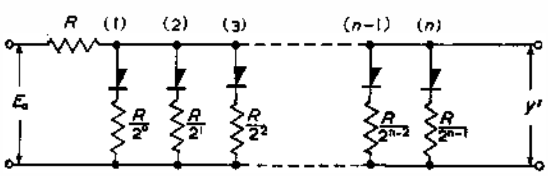


Fig. 5. Circuit nucleus for the relationship $y' = \log_2 E_0$

Fig. 6 introduces the practical requirements of crystal bias potentials for the control of the order of switching as the E_o signal potential rises. It shows the resistive elements $r_1, r_2 \ldots\ldots r_n$, which have been included in order to compensate for the reduction of resistance caused by the new shunt paths $R_1, R_2 \ldots\ldots R_n$. The bias voltages indicated on this circuit against each diode cathode are maintained in the diode 'off' condition by current flowing from the positive supply E, and these ensure that a rise in the signal level applied at E_o shall give a switching sequence in accordance with the values given in Table 1.

Returning again to Fig. 6, it is found that as the value of the input signal E_o increases from zero:—

(1) *No diode conducting.*

$$y'/E_o = 1 = 1/2^0$$

(2) 1st *diode conducting*

Shunt resistance $= R = R/2^0$

$$\therefore y'/E_o = R/(R + R) = 1/2 = 1/2^1$$

(3) *Two diodes conducting*

$$\text{Shunt resistance} = \frac{R.R/2}{R + R/2} = R/3 = \frac{R}{(2^2 - 1)}$$

$$y'/E_o = \frac{R/(2^2 - 1)}{R + [R/(2^2 - 1)]} = \frac{1}{2^2 - 1 + 1} = 1/2^2$$

(4) *Three diodes conducting*

$$\text{Shunt resistance} = \frac{R/3 \,.\, R/4}{R/3 + R/4} = R/7 = \frac{R}{(2^3 - 1)}$$

$$y'/E_o = \frac{R/7}{R + R/7} = 1/8 = 1/2^3$$

(5) *n diodes conducting*

$$\text{Shunt resistance} = \frac{R/(2^{n-1} - 1) \,.\, R/(2^{n-1})}{R/(2^{n-1} - 1) + R/(2^{n-1})}$$

$$= R/(2^{n-1} + 2^{n-1} - 1) = R/(2^n - 1)$$

$$y'/E_o = \frac{R/(2^n - 1)}{R + [R/(2^n - 1)]} = \frac{1}{2^n - 1 + 1} = 1/2^n$$

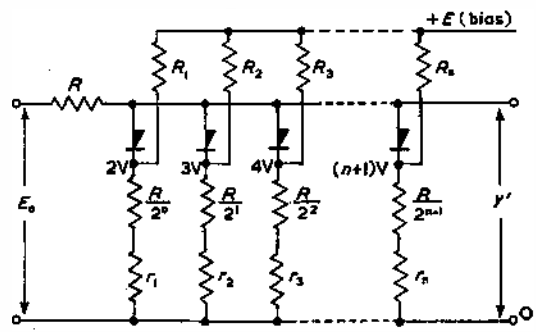


Fig. 6. The Fig. 5 circuit in a practical form

Values of R_n and r_n are determined in terms of R, E, and n. (See Appendix.) Design values may now be found.

It is evident from the values of Table 1 that six sections should be used to cover the design range 1 to 100V for E_o.

R is given a value of 16kΩ, and E is given a value of 300V, in order to provide R_n with values readily obtainable from a normal stock.

$$\text{From } R_n = \frac{ER}{2^{n-1}\,(n+1)} \text{ and } r_n = \frac{R\,(n+1)}{2^{n-1}\,(E - n - 1)}$$

$R_1 = 2{\cdot}4$MΩ	$r_1 = 107$Ω
$R_2 = 800$kΩ	$r_2 = 81$Ω
$R_3 = 300$kΩ	$r_3 = 54$Ω
$R_4 = 120$kΩ	$r_4 = 34$Ω
$R_5 = 50$kΩ	$r_5 = 20$Ω
$R_6 = 21{\cdot}4$kΩ	$r_6 = 12$Ω

Two identical demodulators derive the Fig. 4 type waveform from their respective channels, and each is followed by an a.c. coupled amplifier having variable negative feedback with a gain control for a maximum of times one hundred. A 10V peak-to-peak signal at the output is considered as 0dB level, so that the range at the output becomes 1 to 100V peak-to-peak (0 ± 20dB).

It is essential that the two signals be 'd.c. restored', before each is fed to its logging network. This means that each signal then has an absolute level above a zero baseline. The zero level occurring between each sweep —the return time period which is blanked by the sweep generator—gives a zero reference level for d.c. restoration by means of diodes, which conduct for this period and restore the charge lost by the output coupling capacitor during the previous sweep period.

One network receives only the source amplitude changes, which by definition must not exceed 6dB. The other must be capable of handling a 40dB range. However, both channels are designed for the full range in order that the

TABLE 1
Circuit potential analysis for Fig. 6

E_o RANGE	SWITCHING	y'	SLOPE OF INCREASE
0— 2	no diode 'on'	0—2	1
0— 4	1 diode 'on'	2—3	$\frac{1}{2}$
4— 8	2 diodes 'on'	3—4	$\frac{1}{4}$
8—16	3 diodes 'on'	3—5	$\frac{1}{8}$
2^n—2^{n+1}	n diodes 'on'	$(n+1)$—$(n+2)$	$1/2^n$

working signal level may be varied to suit the limitations imposed by the unit being tested.

To enable the summing network to produce a difference signal, it has been shown that the two logging network outputs must supply opposite polarities. The feed to that giving the positive signal is handled satisfactorily by a cathode-follower supplying a potential to the 16kΩ input of the network; 16kΩ being the requisite design source resistance for correct operation of the logging circuit.

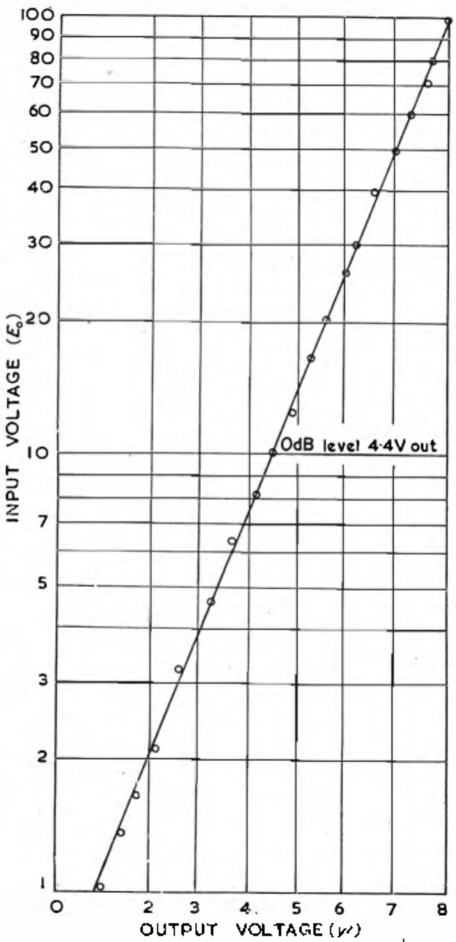


Fig. 7. Plot of experimental data, using the Fig. 3 circuit values

But the feed to the negative output network cannot be handled satisfactorily in this way because the current feeding the circuit is reversed and a load of this type would leave a cathode follower-valve to pass progressively less current as that in the load increased. A current feed from the relatively high anode impedance of a pentode forms a better circuit, which also gives the required phase reversal. A large amount of cathode-grid (negative) feedback is employed to raise the effective anode impedance, to restrict variations due to valve changes, and to allow a signal input amplitude equal to that of the other side of the pair, at the cathode-follower grid.

Fig. 7 is a plot of input and output voltage levels on each of the logging networks, showing that the 10V (input 0dB) level develops a 4·4V output 0dB level. The addition network (Fig. 3) balances the test channel +4·4V against the reference channel −4·4V for a true zero output.

Fig. 7 also shows a 0·9V output for an input 20dB below zero level. This corresponds to an unbalance voltage of $\frac{4{\cdot}4-0{\cdot}9}{2} = 1{\cdot}75$V or 87·5mV/dB at the output of the addition network.

The experimental values relating the input and output signal levels of the log network, which are shown on the curve of Fig. 7, make it evident that although the relationship is clearly logarithmic, the slope of the curve is correct for a base of 1·93 and not 2. In view of the fact that the resistances of the crystal diode switches were omitted from the design, this is quite good agreement.

Crystals were picked and sorted for use in the circuit position most suited to the respective characteristics. Those with a low forward resistance were allocated to the lower resistance shunts, and those with high reverse resistance to the higher resistance circuits.

Fig. 8 is a curve plotted from experimental values of the addition network output voltage when the signal level to both reference and test channels was varied over the 40dB range.

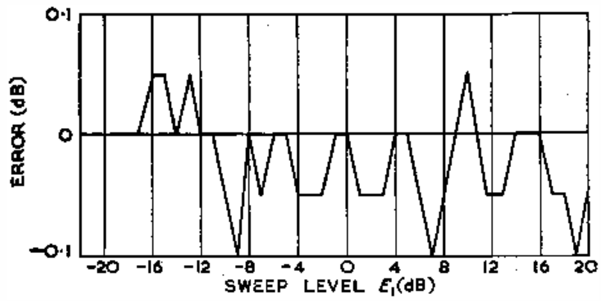


Fig. 8. Plot of experimental data, using the logarithmic signal ratio network

APPENDIX

Let $R/2^{n-1} + r_n = \rho_n$

Two conditions must be satisfied in the choice of values for R_n and ρ_n.

(1) *Bias voltage*

$$(n+1) = E \times \rho_n/(R_n + \rho_n)$$

(2) *Parallel resistances*

$$R/2^{n-1} = \frac{R_n\rho_n}{R_n + \rho_n} \quad \text{by definition}$$

$$\therefore \quad \frac{R_n\rho_n}{R_n+\rho_n} \times \frac{R_n+\rho_n}{\rho_n} = R/2^{n-1} \times \frac{E}{n+1} \text{ from equation (1)}$$

$$\text{Giving } R_n = \frac{ER}{2^{n-1}(n+1)} \qquad \ldots\ldots\ldots\ldots\ldots \text{(A)}$$

Also by inverting equation (2):

$$2^{n-1}/R = 1/R_n + 1/\rho_n$$

$$\therefore 1/\rho_n = 2^{n-1}/R - \frac{2^{n-1}(n+1)}{ER} \quad \text{From equation (A)}$$

$$= [2^{n-1}/ER]\ [E - (n+1)]$$

$$\therefore r_n = \frac{ER}{2^{n-1}[E-(n+1)]} - R/2^{n-1}$$

$$= R/2^{n-1}\left[\frac{E-(E-n-1)}{E-n-1}\right]$$

$$= \frac{R(n+1)}{2^{n-1}(E-n-1)} \qquad \ldots\ldots\ldots\ldots\ldots \text{(B)}$$

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(To be continued)

An Electronic Transformer

By T. G. Clark*, A.M.Brit.I.R.E.

A method of transmitting signals along a cable is described. No electronic device is required at the sending end and negligible loading is imposed upon the source. Voltage transformation independent of the impedance transformation is possible. The cable may be matched, or restriction of bandwidth may be accepted when operating in the unmatched condition.

THE originally posed problem that resulted in the development of the electronic transformer was that of conveying a function of 100V amplitude and of several milliseconds duration from a master generator to a number of separate displays, being part of a complex system. A number of possibilities were reviewed and rejected either on the grounds of inadequate frequency response or of excessive loading upon the source. Finally, a property of the anode-follower, or see-saw, circuit was utilized. This property may be simply stated, the impedance at the pivot of the see-saw is low and nominally equal to Z_2/A, where Z_2 is one arm of the anode feed-back network and A is the

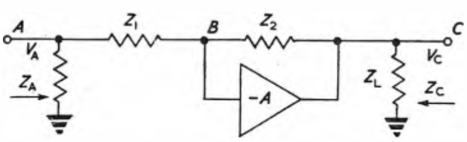


Fig. 1. The anode-follower

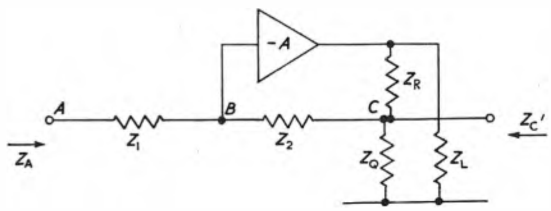


Fig. 2. The d.c. coupled anode-follower

circuit gain in the absence of feedback. Fig. 1 shows the configuration considered and other parameters of the circuit.

Provided that the loop gain of this circuit is sufficiently high, the operation is defined by the following parameters:

$A_o = -(V_c/V_A) = -(Z_2/Z_1)$, the system gain.

$\beta = \dfrac{Z_1}{Z_1+Z_2}$, the feedback factor.

$Z_B = (Z_2/A)$, the impedance at point B.

$Z_c = (Z_L/A\beta)$, the output impedance.

Point B is the pivot of the see-saw and is at virtual earth, i.e., provided that the amplifier is operated in the linear regime then this point is substantially at earth potential. Typically, for $A = 100$ and $Z_2 = Z_1$ this point will move by 1V for an input signal of 100V. The output impedance at point C is low, of the order of 200Ω for a simple configuration.

Where a shift of d.c. level is required, as in the case of a d.c. coupled amplifier, the configuration of Fig. 2 may be adopted. In this case the output impedance is given by:

$$Z_c' = \frac{Z_R Z_L}{A\beta}\left\{\frac{1 + Z_L/Z_R}{Z_L}\right\}$$
$$= Z_R/A\beta\left\{1 + Z_L/Z_R\right\}$$

* *Decca Radar Ltd.*

Thus a shift of d.c. level may be achieved while retaining a substantially low output impedance.

Reactive frequency compensation may be applied to Z_1 and Z_2 and to Z_R and Z_Q.

Consideration of the impedance at point B indicates that a capacitance may be connected from this point to ground provided that the value of the capacitive reactance is greater than the impedance at point B, i.e.

$$C < (A/\omega Z_2)$$

A screened cable offers such a capacitance, provided that it is electrically short, and Fig. 1 may now be re-drawn, as shown at Fig. 3, to illustrate the final configuration. Point B is now connected to point D by means of a screened cable.

Provided that the bandwidth of the basic feedback

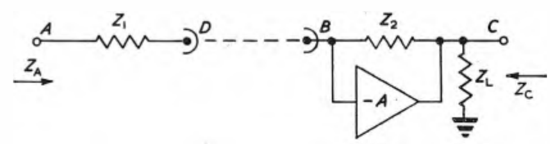


Fig. 3. The electronic transformer

amplifier is not a limitation, then the system bandwidth is given by:

$$B_s = B_c = \frac{A}{2\pi\, klZ_2}$$

where B_s = system bandwidth

B_c = bandwidth due to cable capacitance

l = cable length

k = cable capacitance per unit length

The bandwidth of the basic feedback amplifier is given by:

$$B_o = \frac{A\beta}{2\pi\, R_L C_L}$$

where B_o = amplifier bandwidth

R_L = parallel resistive component of Z_L

C_L = parallel capacitive component of Z_L.

The system and individual bandwidths are related by the expression

$$B_s = \frac{B_o B_c}{\sqrt{(B_o^2 + B_c^2)}} = \frac{B_o}{\sqrt{[1 + (B_o/B_c)^2]}}$$

When it is imperative to match the cable then Z_B may be made equal to Z_c, the cable characteristic impedance. However, the terminating impedance, Z_B, would be a function of amplifier gain. The reflection coefficient of a cable is comparatively tolerant of a small mismatch, but it is better practice to arrange that $Z_B \ll Z_o$. A padding impedance may then be inserted to 'swamp' the variation of Z_B and properly terminate the cable.

The circuit used during the early investigation is shown at Fig. 4. For experimental purposes the gain was varied by means of VR_1 and VR_2. Without suitable precautions variation of Z_2 would affect the amplifier gain, since Z_2 is basically in parallel with Z_L. A cathode-follower, V_2, effectively disconnects this shunting effect permitting the value of Z_2 to be much lower than Z_L.

In order that the amplifier characteristics would be unaffected by the measurement, gain was measured with the feedback loop closed. There are several methods, of which two were utilized, viz:

(1) With zero length of cable between B and D and with a signal applied to point A, a resistance from point B to ground may be adjusted until the output is halved. The value of this resistance is then equal to Z_2/A and, knowing Z_2, A may be calculated.

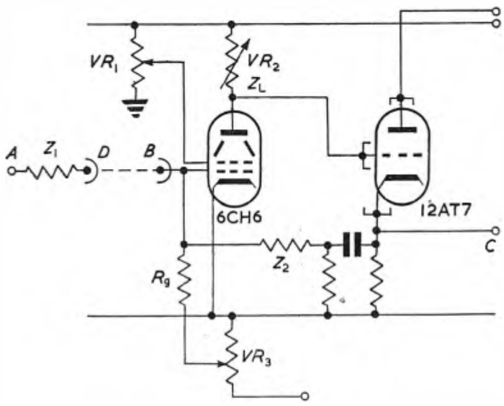


Fig. 4. The experimental circuit

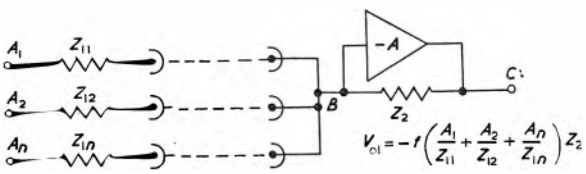


Fig. 5. The electronic transformer as a mixer

Care should be taken that the d.c. conditions remain unaltered during the measurement; for example, a capacitance having a low reactance at the signal frequency may be connected in series with the measuring resistance.

(2) With the initial conditions as for the previous case the input frequency is varied to ascertain the amplifier bandwidth.

A close tolerance capacitor is chosen experimentally to provide a system bandwidth appreciably less than the amplifier bandwidth. The frequency is ascertained at which the response is -3dB relative to the low frequency exposure. Then,

$$A = \omega C Z_2$$

In practice configurations rather more sophisticated than that of Fig. 4 have been used for a variety of applications. The initial requirement of transmitting high level functions has found wide application, especially in arrangements permitting the mixing or splitting of signals. These two applications are shown in Fig. 5 and 6 from which it is seen that 'weighting' may be introduced by adjustment of the appropriate feedback impedance. In the case of Fig. 6, however, the several outputs must be equal and it will be seen that the value of the Z_1 impedance is a function of the number of 'splits'. A further adaption of Fig. 5, shown at Fig. 7, allows signals to be injected at numerous points along the cable. Fig. 8 offers an example of time coding. In this case the cable may be lumped line or high impedance distributed delay line and, as discussed earlier, the line may be matched. Function weighting may also be introduced. Other applications include the provision of oscilloscope monitor points within an equipment, the construction of oscilloscope probes, the transmission of narrow band signals, e.g. magslip information, over

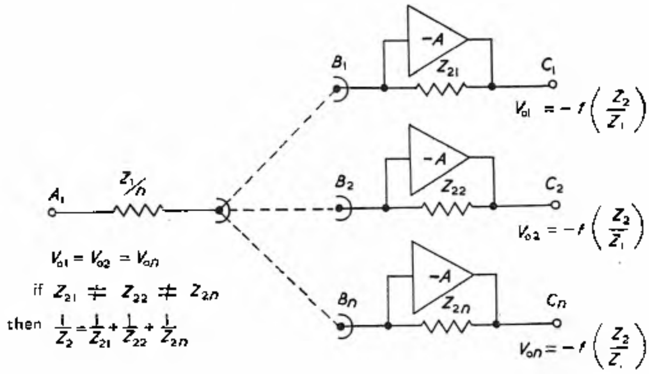


Fig. 6. The electronic transformer as a splitter

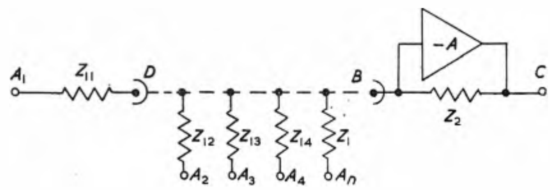


Fig. 7. Another electronic transformer as a mixer

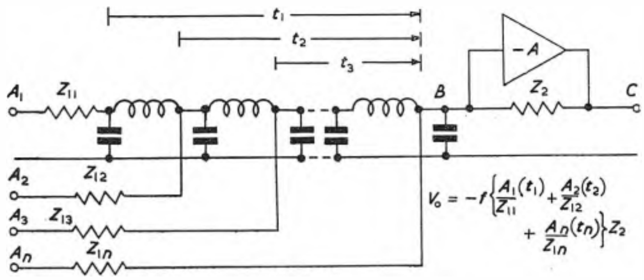


Fig. 8. A time coder

proportionately longer distances. A sawtooth of ± 100V amplitude, of 250μsec duration, has been transmitted over distances of up to 300yd without impairing accuracy. Experimentally, a mile of cable has been used in conjunction with a high gain amplifier.

The system is subject to interference pick-up on the outer screen of the cable, this being induced on to the inner conductor. The signal gain and the gain to interference assumed to be injected at point B are related by the expression

$$G_s/G_i = \beta\, Z_2/Z_1$$

where G_s = system signal gain.

G_i = interference gain.

However, the effect of interference may be minimized by ensuring that the pivot impedance is kept low at all frequencies of interest. The pivot impedance is given by $Z_2/(A + 1)$ and A is a complex quantity approximating to a parallel CR circuit. If Z_2 is given the same characteristic the pivot impedance is substantially constant until the loop gain has fallen to a level where the approximations no longer apply. In severe cases of interference the

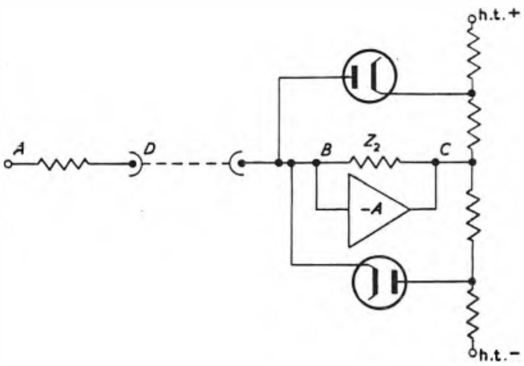


Fig. 9. Protection circuit

inner and outer conductors of the cable may be connected to a difference amplifier, e.g. a long-tailed pair. Since the interference signal is equal on both conductors, only the signal is present in the output.

Throughout the discussion it will have been appreciated that the system operates provided that the amplifier is operated in the linear regime. An overload signal saturating the amplifier will cause the cable to charge and, hence, disable the amplifier for a long time depending upon the cable capacitance and the unmodified Z_2 impedance. A method of overcoming this situation is illustrated at Fig. 9. The diodes conduct when the signal exceeds a set positive or negative level, thus the impedance at the pivot is then a function of Z_2 shunted by a low impedance. This system permits an effective signal of ± 800V to be limited to ± 40V output with a total variation, after limit, of less than 0·1V. Alternatively, the diodes may be replaced by cathode-followers utilizing the low output impedance to modify Z_2 under limit conditions.

Gain Bandwidth Factor

The gain bandwidth factor of an amplifier is given by:—

$$GB = g_m/2\pi C$$

However, the series feedback amplifier that has been discussed has a gain bandwidth factor as follows:—

$$GB = g_m/2\pi C \times \beta Z_2/Z_1$$
$$= g_m/2\pi C \times 1/(1+1/A_o)$$

That is, naturally, true whether used as an electronic transformer or otherwise.

Results

Table 1 shows the initial experimental results. In this case $Z_1 = Z_2 = 10\text{k}\Omega$. The cable used was Telcon T3171 having $Z_o = 70\Omega$ and a capacitance of 22pF/ft.

The calculated bandwidths were calculated on the basis of the cable capacitance only, discrepancies in the table are due to normal experimental error and the finite amplifier bandwidth.

Fig. 10 shows the actual system bandwidth plotted against cable length for a variety of conditions. Referring to Fig. 1 the conditions were as follows:—

Curve 1
$Z_1 = 15\text{k}\Omega$, $Z_2 = 15\text{k}\Omega$ in parallel with 3·3pF.
$A = 58$, $A_o = 1$.

Curve 2
$Z_1 = 3{\cdot}9\text{k}\Omega$, $Z_2 = 15\text{k}\Omega$ in parallel with 3·3pF.
$A = 55$, $A_o = 3{\cdot}9$.

Curve 3
$Z_1 = 56\text{k}\Omega$, $Z_2 = 15\text{k}\Omega$ in parallel with 3·3pF.
$A = 57$, $A_o = 0{\cdot}27$.

Curve 4
$Z_1 = 15\text{k}\Omega$, $Z_2 = 56\text{k}\Omega$
$A = 55$, $A_o = 3{\cdot}7$.

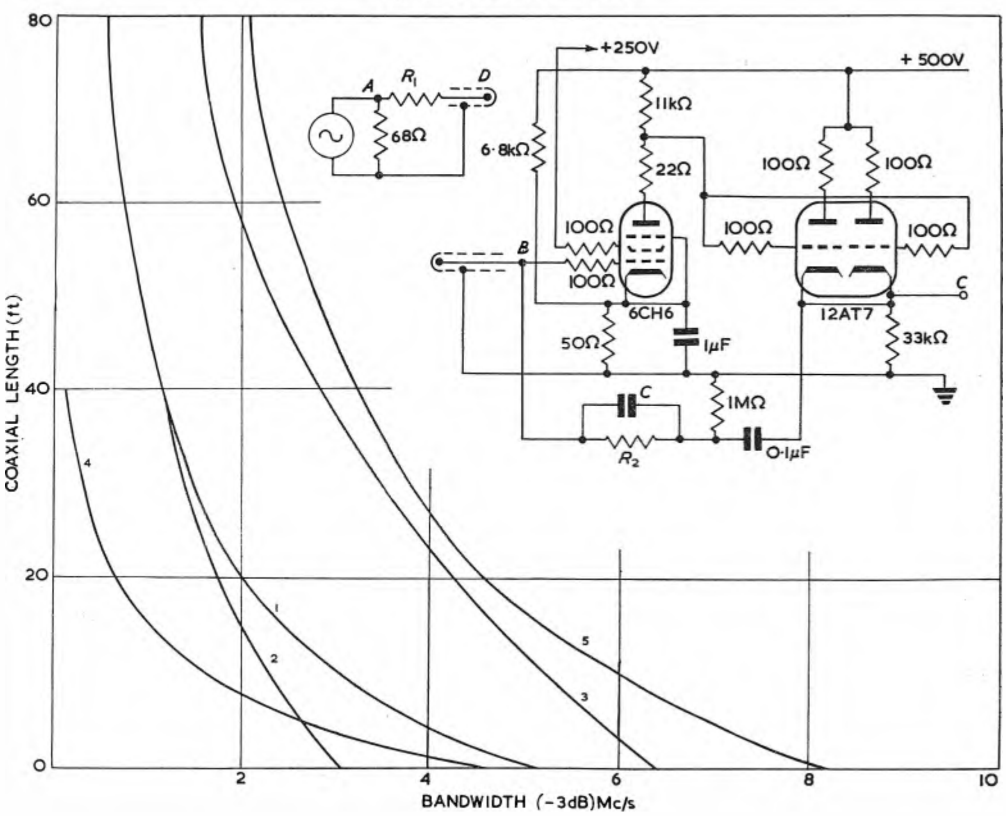
Curve 5
$Z_1 = 15\text{k}\Omega$, $Z_2 = 3{\cdot}9\text{k}\Omega$ is parallel with 12pF.
$A = 55$, $A_o = 0{\cdot}26$.

These curves are offered as early experimental results,

TABLE 1

CALCULATED LENGTH (Feet)	BANDWIDTH (Mc/s) $A=25, Z_L=22$ kΩ.		BANDWIDTH (Mc/s) $A=70, Z_L=100$ kΩ.	
	CALCULATED	ACTUAL	CALCULATED	ACTUAL
60	0·3	0·2	0·84	0·8
40	0·45	0·4	1·3	1·2
20	0·9	0·8	2·5	1·9
10	1·8	1·6	5·0	2·8
0	—	3·2	—	4·4

Fig. 10. Experimental curves



giving an indication of the performance that could be obtained with a simple configuration and, hence, encouraging further investigation. The response of the amplifier was adjusted to be substantially flat with zero cable length. However, it was found that the frequency compensation should, ideally, be adjusted in conjunction with the required cable length.

Summary of Formula

System gain $A_0 = V_c/V_A = \left\{[A\beta/(1+A\beta)]Z_2/Z_1\right\} \sim Z_2/Z_1$

Feedback factor $\beta = \dfrac{Z_1}{Z_1 + Z_2}$

The impedance at the pivot

$$Z_B = \frac{Z_2}{A + 1} \sim Z_2/A$$

The output impedance of Fig. 1

$$Z_c = Z_L/A\beta$$

The output impedance of Fig. 2

$$Z_c' = Z_B/A\beta \left\{1 + Z_L/Z_B\right\}$$

The output impedance of Fig. 4

$$Z_c'' = \frac{1}{g_m A\beta}$$

The bandwidth due to the cable

$$B_c = \frac{A}{2\pi\ klZ_2}$$

The bandwidth due to the basic feedback amplifier

$$B_o = \frac{A\beta}{2\pi\ R_L C_L}$$

The system bandwidth in terms of B_c and B_o

$$B_s = \frac{B_o}{\sqrt{[1 + (B_o/B_c)^2]}}$$

Relative gain of an applied signal to an interfering signal considered to be applied to point B.

$$G_B/G_1 = \beta\ Z_2/Z_1$$

Gain/bandwidth factor

$$GB = \frac{g_m\beta}{2\pi C_L}\ Z_2/Z_1 = \frac{g_m}{2\pi\ C_L} \times \frac{1}{1 + 1/A_0}$$

Summary of Properties

(1) Voltage transformation independent of impedance transformation, both being controllable.

(2) Negligible loading upon the source

(3) No electronic apparatus at the source.

(4) System bandwidth inversely proportional to cable length.

(5) The cable may be matched if desired.

(6) Provided that the system is operated linearly the response is substantially equal for negative and positive going signals.

(7) Susceptible to noise pick-up, but this may be obviated.

No new principles are involved in the operation of the system, for this reason full derivation of fundamental equations has been omitted. It is felt that the applications are novel and may provide a more flexible device than the ubiquitous cathode-follower.

Acknowledgments

Acknowledgments are due to Mr. M. H. Easy, Director of Development, Decca Radar Ltd, for permission to publish this article. Also to Mr. R. Mathews and Mr. D. Bollard of the Decca Radar Display and Data Laboratory, for helpful discussion and additional information.

Post Office Electronic Letter Sorter

The first of the twenty Electronic Letter Sorting Machines, ordered by the Post Office for extended trials throughout the United Kingdom, has now been delivered by The Thrissell Engineering Company, Bristol.

Using the G.P.O. Experimental Machine as a basis and working in close co-operation with Post Office Engineers, a team of Thrissell designers has evolved a machine which has a number of improvements over the prototype. Electronic Instruments Limited of Richmond have worked in close collaboration with The Thrissell Engineering Company, and have been responsible for the design and manufacture of the electronic equipment used in these machines.

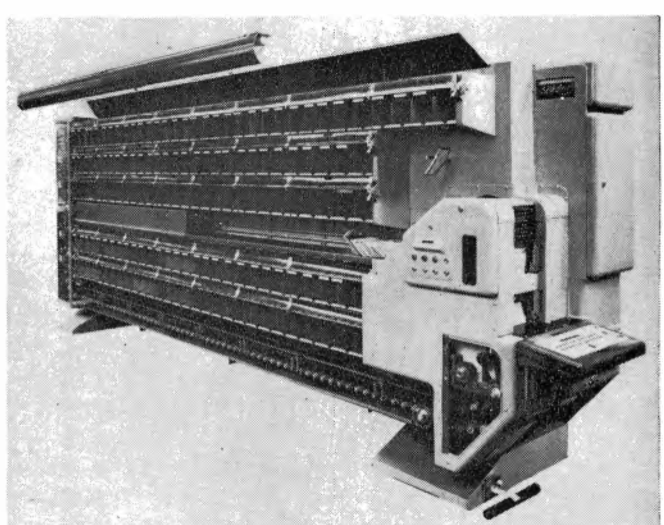
Comfortably seated at a desk, an operator will be able to sort letters on these machines very much faster than by hand, with the added advantage that letters can be directed into 144 separate boxes as against the usual 48 associated with manual sorting.

The speed with which the letters can be sorted is entirely under the control of the operator, and is governed by his speed of reading. A skilled operator can deal with 70 letters a minute. Although the machine has a cycle time rhythm, the operator can key at random just whenever he chooses, and the machine automatically compensates and adjusts the timing to synchronize as necessary.

Letters are automatically fed to a viewing window. The operator simply notes the destination and then presses one of 12 keys with his left hand and one of a similar group with his right hand. The letter is then automatically moved into a waiting compartment and the next letter comes into the reading position. While the previous letter is in the waiting compartment, the destination is remembered electronically, and if the operator considers that he has misread the address, he can cancel his previous keying by pressing a cancel key.

The previous letter is despatched into the machine conveyor system just as this next letter is read and coded. The electronic memory now transfers its information to banks of mechanical memories which operate diverters, and thus direct the letter first to one of five levels and then to its appropriate box. The machine is 17ft long and weighs just under 3 tons. The electronic equipment contains 270 long-life valves and over 1 000 components.

The electronic letter sorting machine



Efficiency-Diode Scanning Circuits

(Part 2)

By K. G. Beauchamp*, C.G.I.A., A.M.Brit.I.R.E.

Derivation of E.H.T.

A USEFUL by-product of the scanning process is the derivation of e.h.t. for the cathode-ray tube. This is made possible by the rapid transference of energy, previously stored in the magnetic field of the scanning coils during the scan period, into the circuit capacitance. For a loss-less circuit these energies may be equated viz:

$$\tfrac{1}{2} \,.\, L \,.\, i_{1(max)}{}^2 = \tfrac{1}{2} \,.\, C \,.\, E_{max}{}^2$$

giving the potential developed across the coils during the retrace period as:

$$E_{max} = i_{1(max)} \sqrt{(L/C)}$$

In practice a finite Q-factor is experienced, and taking this into account the expression:

$$E_{max} = e^{-0{\cdot}83/Q} \,.\, i_{1(max)} \sqrt{(L/C)} \quad \ldots\ldots\ldots \quad (8)$$

may be derived.

This potential is increased by transformer action to develop a large peak potential at the pentode anode. By winding an additional coil over the transformer windings this pulse potential can be further increased in value and may, after rectification, be used as e.h.t. for the cathode-ray tube.

The figure for C in equation (8) must now include the winding self-capacitance and the value of this will limit the e.h.t. capable of being developed. To reduce this capacitance to a minimum a narrow wavewound coil is used for the e.h.t. overwind. It is preferable to arrange that only a single coil is used in order to develop a constant potential gradient across the entire transformer winding.

E.H.T. Source Impedance

It can be shown that in an ideal loss-less system the e.h.t. source impedance is:

$$Z \simeq (T_s/C) \quad \ldots\ldots\ldots\ldots\ldots \quad (9)$$

where T_s = period of one cycle.

C = total circuit capacitance reflected into the pentode anode winding.

There are two conflicting factors present here. In order to obtain a high value of e.h.t. it is necessary to reduce the circuit capacitances to a small value and, on the other hand, good regulation demands a large value for C. The losses present in a practical circuit tend to mitigate the effect of C on regulation and, as shown in Fig. 19, the effect is less than would be expected from equation (9). The influence of C on e.h.t. potential is also shown and is seen to follow the relationship $E_{(max)} \propto (1/\sqrt{C})$ stated previously.

A practical expression for the initial source impedance can be derived from consideration of the reactive power circulating in the transformer winding. Let this be P watts, then source impedance, $Z = (V_s{}^2/P)$, where V_s is the open-circuit potential developed across the secondary.

Now P will include the reactive power developed in the system, and using the notation of Table 1, the source impedance may be taken as:

$$Z = \frac{V_s{}^2}{P_i - P_s - P_{a1} - P_{a2} - P_{g2} + P_b} \quad \ldots\ldots \quad (10)$$

* *Associated Electrical Industries Ltd., formerly with General Electric Co. Ltd.*

By way of example the design worked out later produces an e.h.t. of 13kV, while the total reactive power is 29W giving Z as 5·8MΩ, which compares favourably with the measured figure of 6·9MΩ.

To some extent the effect of poor regulation on picture size may be mitigated if the frame sawtooth generator is supplied from the boosted h.t. potential. The frame oscillator will require a current of only a few milliamperes and is thus quite a small load on this source of supply. As the tube draws more beam current the reaction on the line scanning circuit is such that the boost voltage is reduced. This in turn reduces the scan amplitudes of both line and frame sawtooth currents and compensates for the increase in picture size which would otherwise have taken place. Thus the effect is to allow a rather poorer regulation to be tolerated, and finds extensive use in commercial television receivers.

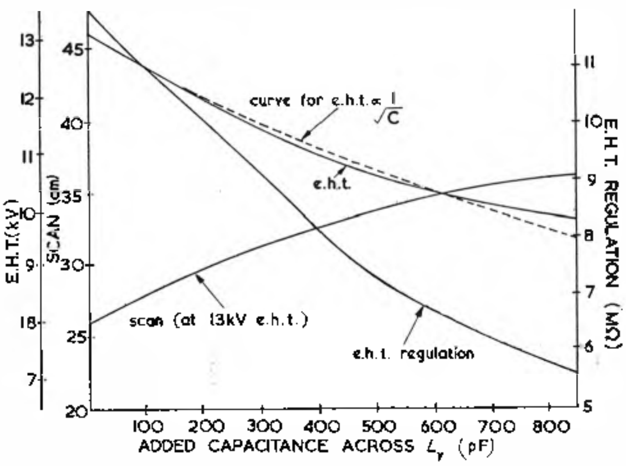


Fig. 19. Effect of circuit capacitance

Maximum Value of E.H.T.

It has been mentioned previously that a limiting factor lies in the value of self-capacitance for the transformer windings. If a sinusoidal retrace waveform is assumed and occurs in one half cycle then the instantaneous value of scanning coil current during this period is:

$$i_y = (I_y/2) \,.\, \sin \omega t$$

where $\omega = 2\pi$ + retrace frequency.

TABLE 1

Measurement and Calculation of Power Distribution in a Typical Series Resonant-return Scanning Circuit

		WATTS
Calculated scanning power		6·10
Pentode anode dissipation	P_{a1}	5·81
Pentode screen-grid dissipation ..	P_{g2}	2·80
Diode anode dissipation	P_{a2}	1·35
Dissipation in g_2 series resistor ..	P_s	0·64
E.H.T. power generated		0·52
E.H.T. diode rectifier filament		0·64
Transformer core and copper losses		2·20
	P_i	20·06
Measured input power		19·40
Recovered power due to resonant-return action	P_b	19·20

From the simplified circuit shown in Fig. 20 the e.h.t. developed is:

$$E = NL_y \cdot (di_y/dt) = \frac{NL_yI_y\omega}{2}$$

now:

$$\omega = 1/\sqrt{(LC)} \text{ and } L = N^2 \cdot \frac{L_sL_y}{L_s + L_y}$$

where C is the total circuit capacitance reflected into the primary winding.

Therefore:

$$E = (L_yI_y/2) \cdot \frac{1}{\sqrt{\left[C\,\dfrac{L_s + L_y}{L_sL_y}\right]}}$$

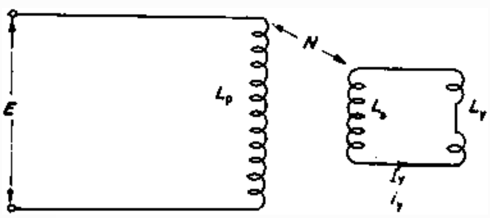


Fig. 20. Derivation of maximum e.h.t.

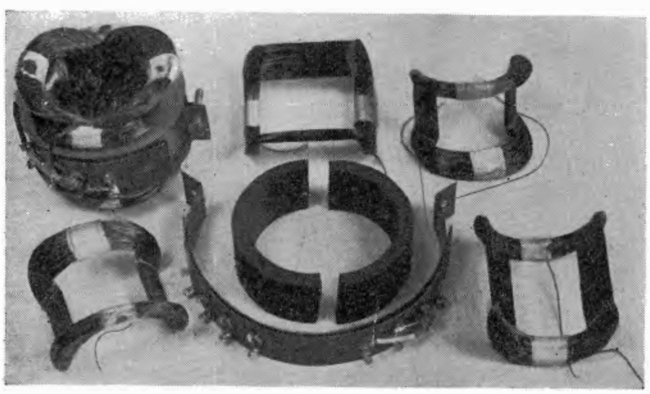


Fig. 21. Scanning coil construction

i.e.:

$$E = \tfrac{1}{2}I_y \sqrt{\left[\frac{L_y(L_s + L_y)}{CL_s}\right]} \quad \ldots\ldots\ldots\ldots (11)$$

This is the maximum e.h.t. available irrespective of turns ratio, the addition of more turns to the overwind beyond that required to produce this figure will result in a reduced value of e.h.t. due to increase in winding capacitance.

Practical Design Procedure

A suitable commencing point for the design is to decide the inductance of the line scanning coils. In order to reduce transformer resonance and increase circuit efficiency, the inductance of the scanning coils should be high, and ideally large enough to enable direct connexion to be made across the diode tap 1-4 shown in Fig. 3. This requirement is, however, limited by the maximum peak potential permissible across the coils. With saddle type coils as shown in Fig. 21 a working figure of about 2kV represents a safe maximum, and leads to a figure of 10mH for the scanning coil inductance. While designs for higher peak potentials are possible, especially if castillated yoke construction is used, consideration of ease in mass-production leads to the acceptance of this figure as a fair compromise.

SCANNING COIL DESIGN

The terms of reference in the design of television scanning coils can be summarized as follows:

(1) Good focus uniformity—a small spot of uniform size over the entire picture area.

(2) Correct raster shape—a rectangle free from geometric distortion.

(3) High sensitivity and low losses, especially for the line coils, where a high Q-factor is essential.

(4) Ability to maintain these standards in large quantity production.

The first two factors are, to some extent, incompatible and a compromise solution must be reached. This usually takes the form of allowing a small amount of pin-cushion distortion to be present in order to achieve good focus uniformity. With the large tubes designed for 90° deflexion this distortion is often corrected by means of the field produced by four permanent magnets disposed around the cathode-ray tube neck or bulb.

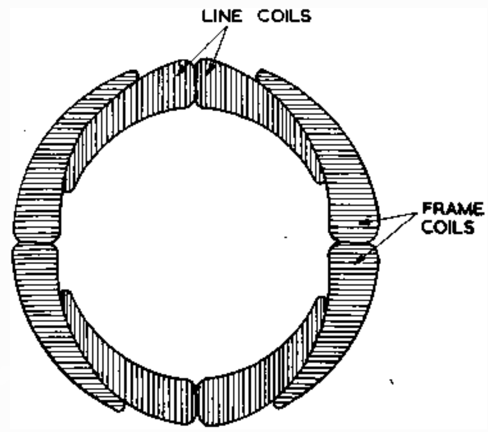


Fig. 22. Scanning coils having 'cosine-squared' wire distribution

In order to provide a uniform magnetic field within the scanning coil yoke the distribution of wires forming a pair of coils should be sinusoidal in shape but with the flat-faced wide-angle tubes now in common use, a design of this nature will give excessive pin-cushion distortion and a better compromise is obtained if the law of wire distribution is amended to:

$$t = k \cdot \sin^2\theta \text{ for the line coils}$$
$$t' = k \cdot \cos^2\theta \text{ for the frame coils} \quad \ldots\ldots (12)$$

This arrangement also facilitates assembly in that the total winding depth results in a true annulus.

Several practical means are available to achieve this:

(1) Graded winding[9]—use of several different gauges of wire during the winding of each separate coil.

(2) Insertion of spacers of different thickness in the coil during winding.

(3) Tapered winding to the shape shown in Fig. 22.

With the aid of specially designed coil winding machinery it is possible to 'form-wind' coils directly into this tapered shape, including the bent-up end sections normal to the tube axis. This operation is made considerably easier by the use of enamelled copper wire covered with a thermo-setting coating ·001 to ·002in thick. This plastic coating softens at about 120°C, and by passing current through the coils while still in the forming jig, the wires may be bonded together by fusion of the coating. When cool the coil is removed from the jig and will have the solid form impressed upon it during its formation. A coil yoke formed in this may be seen in Fig. 21, together with its constituent parts. The yoke core material used is ferrite to enable a high sensitivity to be realized.

The sensitivity of the completed coil can be expressed as:

$$S = \frac{d \, . \, \sqrt{E}}{I_y} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (13)$$

where d = deflexion (cm)

E = e.h.t. in (V)

I_y = scanning current (A).

The coil sensitivity is also shown in graphical form in Fig. 23 and shows the deflexion obtained when a known d.c. is passed through the windings. From these curves a value of $I_y = 0{\cdot}85$A is obtained at the desired e.h.t. (13kV).

Auto-transformer Connexion

Continuing with the design the h.t. voltage e_s can be assumed (say 190V) and the circuit details decided. It has been shown[19] that the use of auto-transformer connexion permits not only considerable reduction in size, but

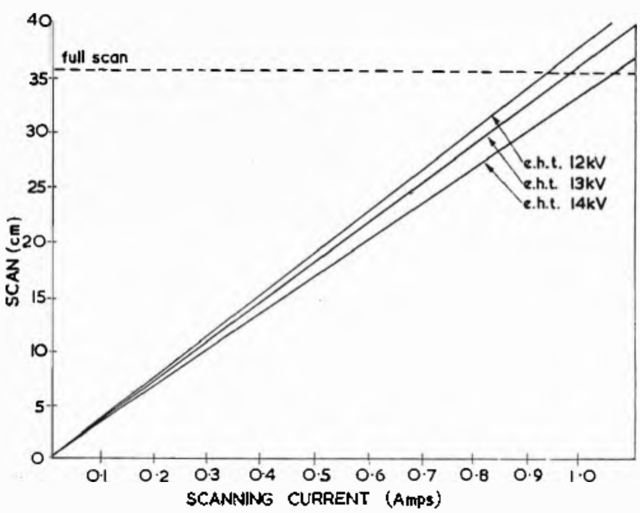


Fig. 23. Scanning coil sensitivity. (S = 1450)

increases overall circuit efficiency and reduces undesired transformer resonances. Its use, however, involves the operation of the boosting diode with a very high pulse potential between heater and cathode of about 4 or 5kV. It is possible to use a low capacitance heater transformer, suitably insulated, for this purpose; but two other methods find wider application.

The first employs a bifilar winding having in each winding a number of turns equal to the diode winding N_{1-4} and wound upon the other limb of the transformer core, (Fig. 24(a)). The heater current is passed through this winding so that the cathode and heater are subject to the same amplitude of pulse potential developed in windings 1-4 and 7-8 respectively, (Fig. 24(b)). Care must be taken to avoid leakage inductance resonances in these added windings and to prevent this a capacitor C is added between heater and cathode. Its value is kept fairly low ($\simeq 0{\cdot}02\mu$F) to avoid injection of 50c/s supply frequency into the scanning circuit.

A second method is to use a specially constructed valve (e.g. Osram U329) having a peak heater-to-cathode rating of up to 7kV. The cathode construction takes the form of a double concentric tube, the outer tube being coated with emissive oxide material. The heater wires raise the adjacent tube temperature and allow the outer tube to be heated by radiation from this inner tube.

The Drive Circuit

The controlling waveform for V_3 can be derived in a number of ways. Nearly all of these incorporate a simple integrator circuit, the capacitor of which being either rapidly charged through a low impedance and allowed to discharge slowly through a resistor R; or alternatively is charged slowly through R and discharged rapidly. In either case the slow change of potential across C occurs at a rate governed by the CR product and is exponential in form.

For the second case:

$$v = E \, . \, (1 - e^{-t/CR}) \quad \ldots\ldots\ldots\ldots \quad (14)$$

where v = the instantaneous potential across C.

t = the time from which C commenced to lose its acquired charge.

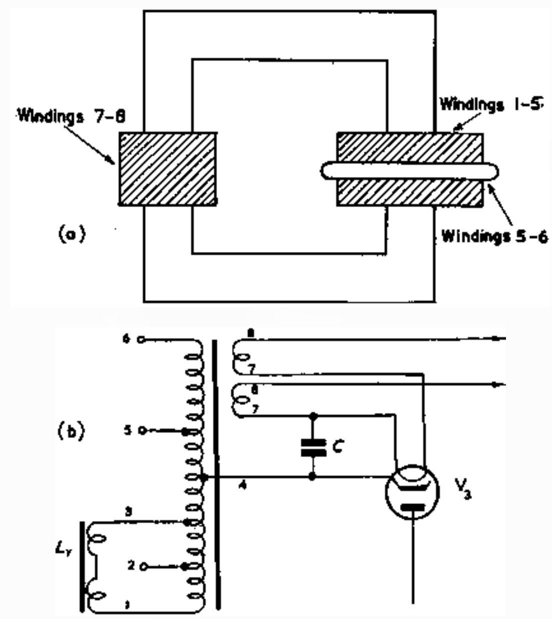


Fig. 24. Auto-transformer connexions using bifilar heater winding for V_3

If the rise in potential is required to be substantially linear only a small amount of the possible rise in potential is permitted before C is discharged once more.

One simple arrangement for controlling the discharge of C is to shunt the capacitor by a valve whose impedance is reduced to a low value when a positive pulse is applied to its grid. The feedback circuit shown in Fig. 2 is of this type but suffers from a number of practical difficulties leading to the use of relaxation oscillators to control the discharge of C.

These may take the form of a multivibrator or blocking oscillator circuits. The multivibrator requires a very small synchronization pulse and can provide a very stable oscillator suitable for d.c. frequency control, and for this reason is often found associated with 'flywheel' synchronization circuits.

The blocking oscillator is perhaps the most widely used sawtooth generator and a typical circuit is shown in Fig. 25. Here the cathode, grid and screened grid of V_1 function as a triode oscillator, with the transformer connexions phased to give positive feedback from g_2 to g_1. Upon switching on, g_1 begins to draw current and capacitor C_1 is rapidly charged negatively and the valve is cut-off. This negative charge then leaks slowly away through R_1 at a rate dependent on C_1R_1, to the potential to which R_1 is returned When the potential across C_1 reaches the grid base potential the valve conducts and the cycle repeats.

The oscillator transformer core losses usually provide sufficient damping across the transformer circuit to prevent large swings of grid potential after C_1 has reached its maximum negative charge. It has been shown elsewhere[11] that this is important if a rapid retrace is to be obtained.

Synchronization is provided by the application of a positive pulse to the grid via the transformer windings, slightly before the valve is rendered conductive by the rising potential across C_1. In the circuit given in Fig. 26 the sawtooth waveform across C_1 is not made use of directly, but a second integrating circuit CR is included in the V_1 anode circuit and in this way inductive elements are eliminated from the discharge path and a more rapid retrace obtained.

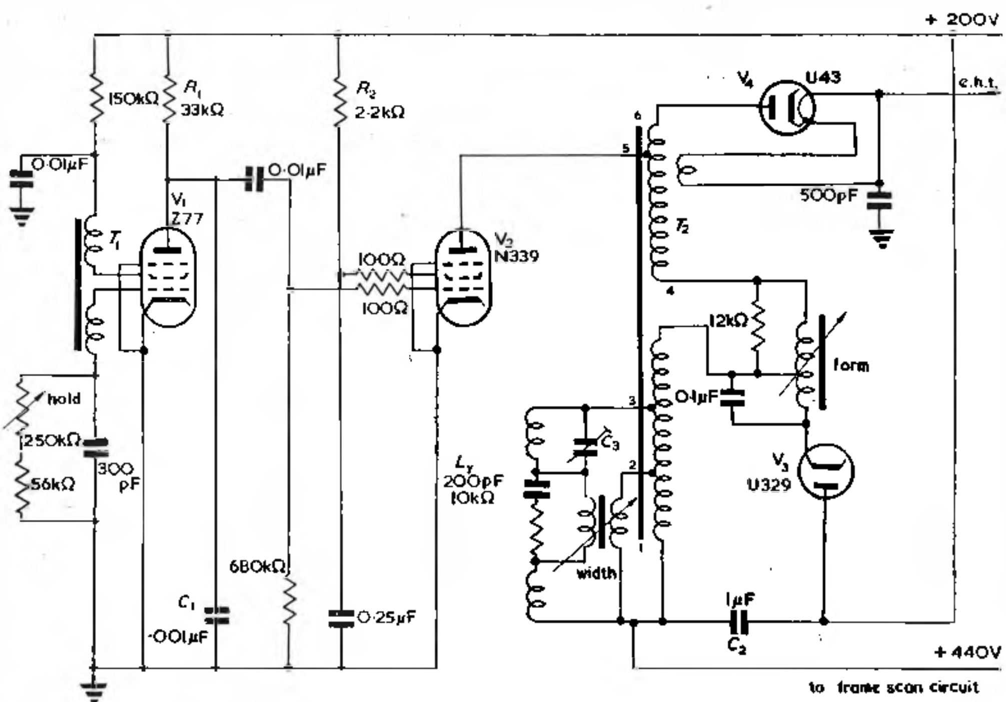


Fig. 26. Complete scanning circuit

The Complete Scanning Circuit

The complete circuit is drawn in Fig. 26 incorporating the features described in the preceding sections. A differential width control is included and a controlled-diode linearity circuit used. The major circuit element yet to be discussed is the transformer T_2 and it is proposed to consider this now in relation to the theory previously described.

Transformer efficiency has been made the subject of an analysis by Friend[13] based on the equivalent circuit for the output transformer shown in Fig. 27(a). From this an expression may be derived for the deflexion factor F_d which may be defined as the ratio of scanning coil ampere-turns required by a practical transformer to that necessary with an ideal transformer. This has been shown to be:

$$F_d = \left\{L_s/L_y\,(1/k^2 - 1) + (2/k^2 - 1) + \frac{1}{k^2 \, . \, L_s/L_y}\right\}^{-1} \quad \ldots\ldots (15)$$

where L_s is the secondary winding corresponding to L_{1-3}. This expression may be differentiated to give a figure for maximum F_d,

$$F_{d(max)} = \frac{k}{1 + L_y/L_s} \quad \ldots\ldots\ldots\ldots (16)$$

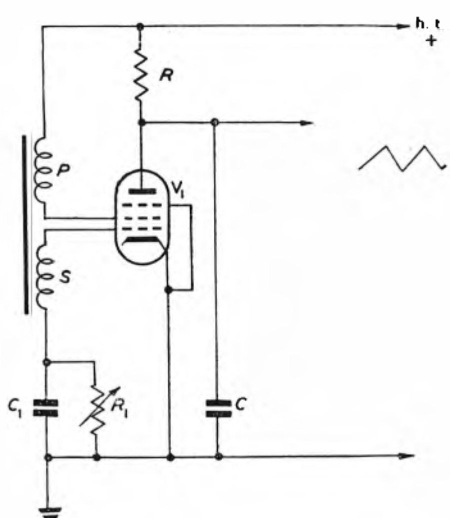


Fig. 25. Blocking oscillator sawtooth generator

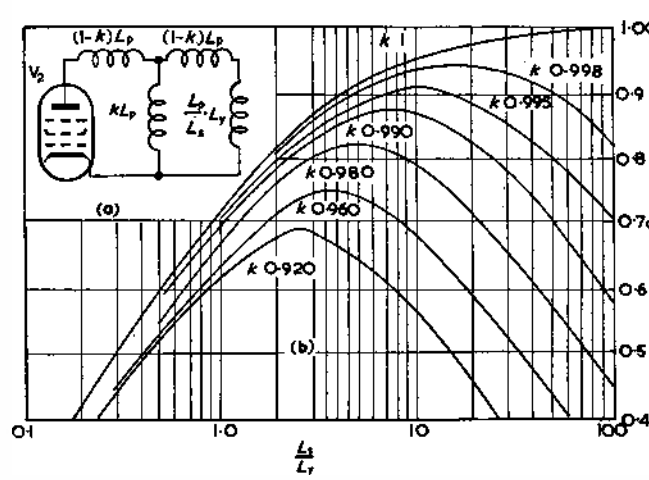


Fig. 27. Efficiency of line scanning transformer

The efficiency of the transformer is therefore very dependent on the coupling factor k and a series of curves has been compiled from this expression in order to illustrate this point and are shown in Fig. 27(b). One may use these to find the optimum value for L_y/L_s, given a figure for k (obtained from a test transformer winding). Due to leakage inductance resonance it is preferable to use a somewhat lower value than that found from expression (16) and a figure of 0·7 $L_{s(max)}/L_y$ is often used.

With $k = 0{\cdot}998$ (a typical value using ferrite material for the transformer core)

$$L_s/L_y = L_{1-3}/L_y = 12$$

So that for $L_y = 10\text{mH}$,

$$L_{1-3} = 84\text{mH}$$

From Fig. 27, $F_{d(max)} = 0{\cdot}92$ showing a loss of 8 per cent scanning power. To this loss can be added that due to the inclusion of linearity and width coils, say 20 per cent, giving a total scanning current requirement of:

$$I_y \times 1{\cdot}28 = 1{\cdot}09\text{A}$$

Now the voltage developed across L_y during the scan period is given as $V_y = L_y \, . \, dI_y/dt$ volts, and as this is constant during the scan, one may write:

$$V_y = L_y I_y / T_s \text{ volts} \quad \ldots\ldots\ldots\ldots (17)$$

where T_s = Total scanning period in seconds.

The volt-amperes required by the scanning circuit is therefore: VA = $1{\cdot}09V_y$, and substituting the figures given above $V_y = 110\text{V}$, so that VA = 120VA.

To find the peak current flowing in winding 1-5, the potential developed across this will require to be known. This will be the difference between the boosted potential, $V_b = e_s + e_b$, and a minimum anode potential V_a for V_2. This latter will vary a little with different valves, but in any case will be small compared with the boosted potential. A limit can be assigned to V_b, say 450V, and V_a taken as about 60V, giving:

$$I_{p\text{-}p} = \frac{VA}{V_b - V_a} = 285\text{mA}$$

V_2 will be required to supply just over one half of this, say 55 per cent, so that the peak anode current for this valve will be approximately:

$$i_p = 0{\cdot}55 I_{p\text{-}p} = 157\text{mA}$$

A mode of operation is chosen such that the diode is just cut-off when the pentode anode current reaches this figure. This gives the root of the diode curve immediately below this value of i_p, so that:

$$\frac{\text{Potential across 1-5}}{\text{Potential across 1-4}} = \frac{450 - 60}{450 - 190} = N_{15}/N_{14} = 1{\cdot}5$$

This gives the transformer ratio required between V_3 and V_2. The boosting diode tap must be such that V_3 is at or near cut-off when V_2 is conducting, and can be found by equating V_3 cathode and anode potentials during this period. i.e.:

$$N_{14}/N_{13} = \frac{V_y}{V_b - e_s} = 2{\cdot}25$$

The transformer may now be wound (less e.h.t. winding), but before doing so it is necessary to determine the working flux density to ensure that the core material is not saturated by the large currents flowing through the transformer windings.

The choice of core material for resonant-return transformers will be governed by the permissible losses, and although nickel-iron and dust-cores have been used, the very small eddy-current loss and high permeability of ferrite materials renders their use very attractive indeed. Ferrite materials are of ceramic-type composition and characterized by their very high specific resistance (of the order 10^2 to $10^6\Omega/\text{cm}^3$) and consequently low eddy-current losses. Of the two main types available commercially in this country, the Mn . Zn ferrite (Mullard Ferroxcube 'A') has been found most suitable for scanning transformer service. As the material tends to saturate at a comparatively low flux density a gap may be necessary to increase the reluctance of the magnetic path.

Using Ferroxcube A2, the saturation flux density is $B = 350\text{mWB/m}^2$ and the relative permeability $\mu_r = 400$. The total transformer flux is given as:

$$BA = \frac{\text{m.m.f.}}{\text{total reluctance}} \text{ Wb}$$

so that with a small air-gap,

$$B = \frac{F}{l/(\mu_r/\mu_0) + l'/\mu_0} \,.\, \text{Wb/m}^2 \quad \ldots\ldots\ldots (18)$$

where l = length of magnetic path (m)

l' = thickness of air-gap (m)

μ_r = incremental permeability

μ_0 = magnetic space constant = $4\pi \times 10^{-7}$

F = m.m.f. = $N_{15} \times i_{p(r.m.s.)}$

The core shape shown in Fig. 28 has a magnetic path length of 15·6cm and the substitution of values, stated pre-

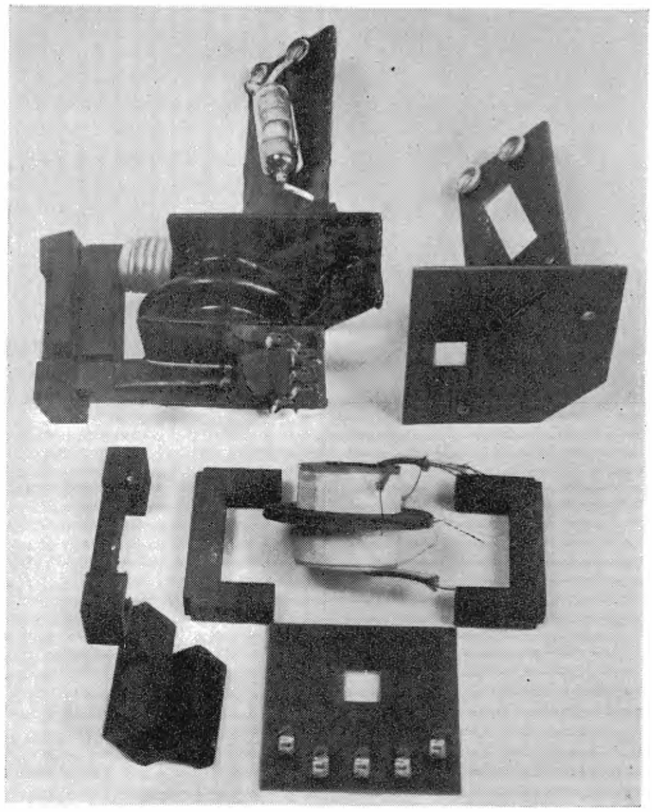


Fig. 28. Scanning transformer construction

viously, in equation (18) gives l = ·008in, thus two paper gaps, each of ·004in are required, placed in each limb of the transformer core.

The transformer is layer wound and provision included for adjusting the diode and pentode taps on either side of those calculated. The circuit of Fig. 26 is set up and preliminary adjustments made; a separate source of e.h.t. being used for the tube. R_2 is set to ensure that g_2 dissipation is within the valve rating, and the drive adjusted by either C_1 or R_1, so that the picture does not quite 'fold' about the screen centre. This indicates correct amplitude of drive waveform, approximately twice the valve grid-base potential. Adjustment of transformer tappings can be carried out to secure optimum performance, and the peak potential at V_3 anode measured.

The e.h.t. overwind step-up ratio can then be determined and this winding added to the transformer. To avoid breakdown during initial testing a film of wax is spun over the e.h.t. overwind prior to assembly. Resonance effects within the e.h.t. winding can effectively alter the expected value of e.h.t. dependent on the phase relationships of the pulse voltage components. This is especially true of the Tourshou circuit and is responsible for the lower maximum value of e.h.t. obtainable with this circuit.

Finally another transformer is wound, less intermediate tappings, and performance checked after wax impregnation. The addition of a coat of wax over the transformer will increase the self-capacitance and reduce the derived e.h.t. by some 3 to 5 per cent and due allowance should be made for this.

The heater supply for the high-voltage diode rectifier may also be taken from the scanning transformer. A few turns of Polythene insulated wire around one of the core limbs as shown in Fig. 28 will provide sufficient flux linkage for the 3 to 4W of heater power required.

Transformer Impregnation

To avoid corona with its possibility of causing surface tracking and eventual breakdown it is essential to take great precautions in the finish and impregnation of the completed transformer design. Porous hygroscopic materials, such as press-pahn, should be avoided in transformer assembly and the lead-out terminals spun into high-grade Bakelite or similar material (a ceramic moulding has been used by one manufacturer).

Two prevalent causes of premature failure are moisture and small air-pockets trapped within the windings. Small air-pockets are a particularly potent cause of breakdown as the sudden change of dielectric in the pocket region can give rise to high potentials and intermittent arcing across the air-space. The value of a constant potential gradient has been referred to earlier and the use of vacuum impregnation techniques will provide an additional safeguard. A procedure found very effective by the author is to dry the transformer thoroughly for several hours in a temperature of 220 to 240°F and then vacuum impregnate at this temperature in Ozokerite wax. Immediately upon removal from the vacuum tank the transformer is given a dip in bitumen wax at a temperature of 300 to 350°F. This provides a protective coating against damage due to careless handling and also serves to anchor rigidly the lead-out wires.

Magneto-striction Effect

One practical disadvantage of using ferrite core material is the very pronounced 10kc/s whistle produced with the British 405 line system. This is due to magneto-striction vibrations at the butt joints of the U-shaped cores, and can be overcome by bonding together the core so that the transformer vibrates as one homogenous mass. The transformer may then be rubber-mounted and little audible whistle transferred outside the receiver cabinet. A very suitable method of core bonding for ceramic-type materials is to make use of the synthetic-resin techniques of which 'Araldite' is probably the best known example. This is an epoxy-ethane resin and polymerizes without giving off any volatile by-products, an important factor as ferrite is not a porous material and any by-products would be trapped within the joint and prevent a good bond from being obtained.

Scan Distortion

A fundamental aim in scanning circuit design is to produce a linear trace free from variations in scan velocity. One type of distortion has already been discussed, i.e., the expansion at the commencement of the scan due to curvature of the diode characteristic. Several other causes of distortion are possible and may initially be present in a preliminary design.

One of these arises out of the necessity to scan linearly a tube having a large radius of curvature for the screen face. This difficulty has been discussed elsewhere[13], wherein it was shown that to produce a linear scan under these conditions the scanning current waveform should assume the 'S' shape shown in Fig. 29.

To produce this a parabolic correction waveform should be added or subtracted from the linear rise of current through the coils The correction waveform may be derived from the potential developed across a capacitor C inserted in series with the scanning coils, the amplitude of the correcting waveform being given as:

$$v = (1/C) \int_0^{T_s/2} . \, i \, . \, dt \quad \ldots\ldots\ldots \quad (19)$$

where $i = kt$ = rate of change of sawtooth current through coils.

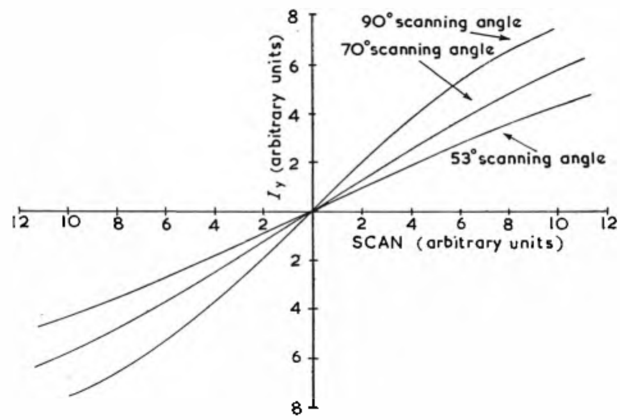


Fig. 29. Scanning current requirements for flat-faced cathode-ray tubes

Adjustment for this correction is obtained by choice of C, a value of between $0{\cdot}1\mu$F and $1{\cdot}0\mu$F being usually required.

Alternatively the ripple potential across the boost storage capacitor C_1 (Fig. 3) may be utilized and added to the supply for V_2. This provides an economical method and has been adopted in Fig. 26 by suitable choice of C_2.

Transformer Resonances ('Ringing')

A second form of distortion is due to unwanted resonances in the transformer and associated components. A simplified circuit which may be used to illustrate this is given in Fig. 30, and shows the leakage inductances and stray capacitances of the scanning circuit.

Such a system is capable of oscillating at several frequencies simultaneously. The main magnetic flux around the circuit will resonate at a frequency of about 50kc/s during the retrace period. Superimposed upon this will be smaller oscillations of a much higher frequency, which are due to leakage reactance resonances. These may modulate the constant potential across L_y during the scan period and hence affect the scanning current through it. The effect of this is displayed on the screen as a series of vertical striations of varying light intensity and diminishing effect towards the right-hand side of the picture. In particular the e.h.t. overwind has a large number of turns, and usually is constructed as a narrow wave-wound coil. Consequently its leakage inductance is large, giving rise to a large storage of energy in L_5 during the retrace period. This energy will cause oscillations to take place around the circuit C_5, L_5, L_3, V_5 and an alternating potential is developed across L_3 which will have the effect of modulating the current in

L_Y during the first part of the scan. Also when V_2 is cut-off at the beginning of the retrace period the stored energy in L_4 will flow into C_4 and cause oscillation between these two components. Oscillations between L_3 and C_3 are possible in a similar manner when V_3 is cut off. The leakage inductance L_2 is not important as it forms part of the scanning coil resonant circuit and oscillates at the fundamental resonant frequency.

Fortunately by using an auto-transformer connexion and ferrite core material coupling coefficients of $k = 0{\cdot}998$ can be obtained and L_3 and L_4 are very small. It is not possible to obtain such a high figure with the e.h.t. overwind and for $k = 0{\cdot}88$ (a typical figure), $L_5 = 100$mH and will be responsible for almost all the ringing experienced with a completed transformer design.

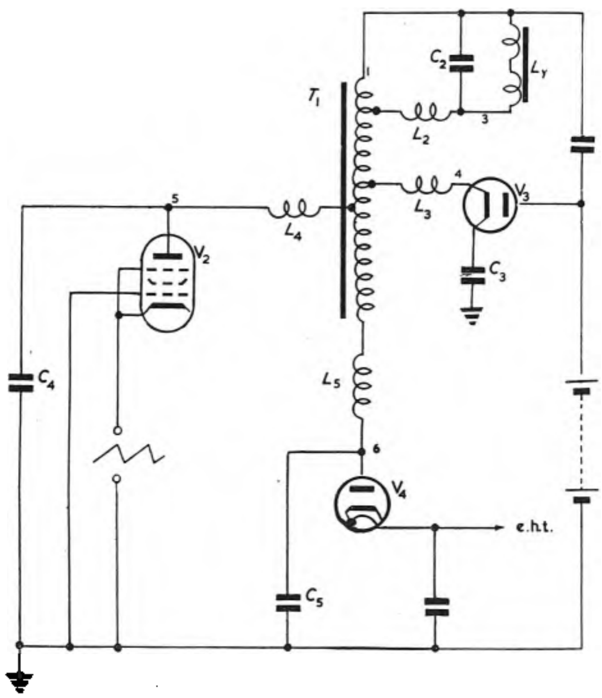


Fig. 30. Simplified scanning circuit showing leakage inductances and stray capacitances

One method of reducing this ringing is to increase the impedance around the path C_5, L_5, L_3, V_3 by winding the e.h.t. overwind with resistance (Eureka) wire. The Q-factor can in this way be reduced by a factor of 40 to 100 times which is sufficient to reduce the ringing to negligible proportions.

Although this method is effective it has two attendant disadvantages. A certain amount of power is dissipated within the e.h.t. overwind reducing the overall transformer Q-factor and affecting efficiency, and secondly the dissipation of this power in the form of heat within the winding will increase the coil temperature and may soften or even melt the protective wax covering. Another solution is to insert a strip of nickel-iron alloy such as Mumetal or Radiometal between windings 1-5 and 5-6 in a somewhat similar manner to the electrostatic copper screen inserted between primary and secondary of some mains transformers. This screen should not form a complete turn otherwise the shorted turn so formed will absorb a large amount of the circulating energy.

As the added Mumetal strip will be situated within a strong alternating field it will localize the flux near the strip and energy will be required to enable the flux in the material to go through a cycle of magnetization. This power loss becomes quite large as the frequency is raised, so that losses at the ringing frequencies, 200 to 600kc/s, will be large compared with those at the fundamental resonant frequencies. As hysteresis losses can be represented by a series resistance, the Q-factor of the e.h.t. overwind can be considered to be decreased at the higher frequencies. Thus only the energy stored in the leakage inductance is dissipated as heat losses, and as this is relatively small, little loss of overall transformer efficiency is experienced.

Other Circuit Resonances

Other forms of ringing are possible and one prevalent source is due to resonances within the scanning coils. This form of ringing also occurs at the left-hand side of the screen and is caused by assymetry of the scanning coils. This may be removed by tuning the half of the scanning coils having the least stray capacitance with a mica-compression trimmer of some 10 to 100pF capacitance[14] suitably insulated to withstand the high pulse potentials present across the coils during the retrace period. (C_3 in Fig. 26.)

Mention has already been made of the resonances due to the series section of the width coil. Similarly resonance of a saturated reactor coil is also possible, and appropriate precautions need to be taken. It is worth mentioning at this point that although the added inductive elements may be rendered non-resonant by the addition of a damping circuit, the complete circuit can behave as a capacitance at certain frequencies. This capacitance can resonate with the transformer leakage reactance and some ringing may occur. The solution lies in placing the coil, together with its associated *CR* damping circuit between the two scanning coils as shown in Fig. 26. The leakage inductance is now no longer directly associated with the circuit and this source of ringing obviated.

An unusual form of ringing is possible with the Tourshou circuit briefly described previously. It is a condition of normal operation with this circuit that V_3 must remain conductive during the entire scan period (Fig. 4). Should V_3 cease to conduct at any point in the cycle then, as the scanning coils are not magnetically coupled to the inductance L_a in the anode circuit, there is nothing to prevent resonances of either L_Y or L_a from varying the steady potential across L_Y. This condition is most likely to occur towards the end of the scanning period when the energy stored in L_a is greatest. Consequently ringing due to this cause will be observed at the right-hand side of the screen and continue with diminishing intensity towards the picture centre.

Performance Measurements

Apart from routine measurements of component values, valve operating conditions, etc., certain measurements are peculiar to scanning circuit design and to complete this article a few are given below.

Calculation of anode dissipation for power valves handling non-sinusoidal waveforms, such as the line output valve, is difficult, although an approximate method suitable preliminary calculation has been stated elsewhere[15]. A more accurate method is to use the heat radiated from the anode as a measure of the power dissipation. A thermo-pile may be used to produce an e.m.f. proportional to the anode dissipation under normal working conditions. This reading may be noted, and the valve then operated with d.c. potentials applied direct to its electrodes. These are adjusted until the same reading is obtained as before; providing the screen grid dissipation has been kept constant then the anode dissipation can simply be evaluated.

Unfortunately the method does not allow continuous monitoring of anode dissipation to be made while circuit adjustments are carried out, also due to thermal inertia each measurement takes some time to make. A direct-reading electronic watt-meter has been described by Garrett and Cole[16] in which these limitations are avoided. The instrument is rather complex and consists of a 10Mc/s carrier oscillator which is modulated first by the current and later by the voltage waveform of the output valve to be measured. The instantaneous product is recovered and may either be integrated to give an average power figure or passed through a peak detector for peak power measurement.

To measure the heater power for the e.h.t. diode rectifier a thermal meter may be used which will give an r.m.s. reading independent of waveform. Sufficiently accurate results can sometimes be obtained by comparing the brightness of the diode filament with another diode fed from a measurable d.c. source.

The final criterion of a scanning circuit is its linearity, and in order to check this, some form of pattern generator is required, or alternatively the BBC test card used. In either case measurements are taken of the distance between pairs of vertical bars along the screen face. The linearity can be expressed either graphically or as a percentage variation about the mean bar spacing. A single figure can be given of the greatest departure from exact linearity as a percentage of mean bar spacing.
e.g.

$$\text{Non-linearity} = \frac{200\,(a - b)}{a + b} \text{ per cent} \quad \ldots\ldots\ldots \quad (20)$$

where a = width of widest bar spacing

b = width of narrowest bar spacing.

For greatest accuracy care must be taken to include the error arising due to the curved face of the tube[17] which can be comparable with the largest percentage of non-linearity tolerable.

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New Radar Speed Check

Marconi's Wireless Telegraph Co. Ltd announce the production of the first radar speed check equipment to be designed and manufactured in Great Britain. The device is known as 'PETA' (Portable Electronic Traffic Analyser).

'PETA' removes one of the major objections to the use of radar as a means of checking the speed of road vehicles in that it can discriminate between individual vehicles even under conditions of dense traffic. It will provide separate readings on vehicles which are only 8ft apart and will automatically 'lock on' to any individual reading if so desired. The speed range of the equipment is from 2 mile/h to 80 mile/h, but this can be extended if required. The accuracy of speed assessment is within 1 mile/h.

'PETA' employs the Doppler effect and this in conjunction with the extremely narrow horizontal beamwidth of the radiated signals (between 3° and 4°), provides simplicity of design, portability and high accuracy of measurement, giving the user a device which calls for no technical skill in its operation. An extensive use has been made of transistors and printed circuits in order to achieve compactness and light weight.

The complete equipment, which weighs only 21lb, consists of a transmitter/receiver unit with its associated aerial, and a meter unit.

In practice, the aerial unit forms a baseplate for the transmitter/receiver unit. The equipment is placed in the open boot of a car with the aerial facing the oncoming traffic, or taken outside as convenient. The natural 'squint angle' of the aerial is such that when the latter is physically at right angles to the road the transmitted beam is directed at the required angle of 20° to the line of traffic.

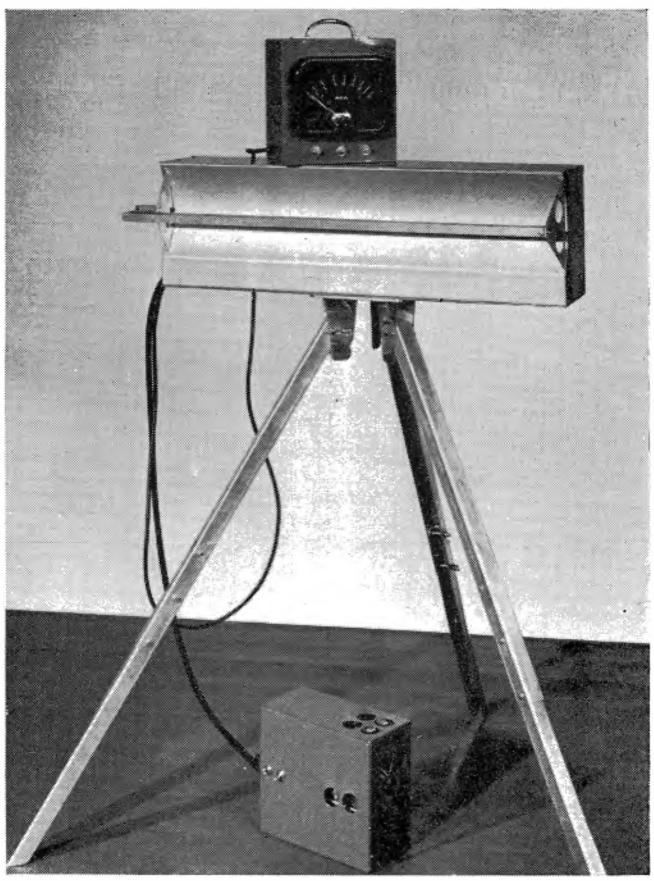
The transmitted beam is reflected from the vehicle concerned and is picked up by the receiver, which measures the difference in frequency between the transmitted signal and the received one. This change is occasioned by the movement of the vehicle; the frequency difference is proportional to the speed at which the latter is travelling. A voltage signal, proportional to the change, is transmitted over a convenient length of cable to a meter which is directly calibrated in mile/h or km/h and thus provides instantaneous information upon the speed of the vehicle.

The potential uses of 'PETA' are by no means confined to the operation of speed checks. By the use of a 35mm camera attachment and an analyser unit which is in the course of development, a record can be provided of the speed and frequency of traffic at any particular road point.

The standard version of the equipment incorporates a single channel system; if discrimination between near and far lane traffic is required a double channel system is available for fitting as an optional extra.

The equipment operates in the X band and gives an f.m./c.w. output of 4 to 10mW from a klystron feeding into a slotted waveguide aerial with flare.

The speed check equipment



Measurement of Small Time-Intervals in an Electronic Torquemeter

By H. Rakshit*, D.Sc., F.Inst.P., F.N.I., and S. C. Mukherjee†

The measurement of torque or twist of a rotating shaft transmitting mechanical power, in terms of time-interval and revolutions per minute is discussed. When the speed of revolution is high, the time of rotation through this twist angle becomes very small, of the order of a fraction of a microsecond. A technique for measuring such small time-intervals from a remote distance by the method of coincidence of two pulses with electrical delay circuit is suggested. Operation of a number of delay circuits has been specially studied and the Sanatron type has been found most convenient. With the help of this circuit and by using a new technique of coincidence, it has been possible to measure time-intervals with an accuracy of 0·1μsec and hence torque at very high speeds. A suitable device for aural-cum-visual detection of coincidence has also been developed for operation of the torquemeter in noisy surroundings.

It is well known that a shaft transmitting mechanical power is subjected to a torque or twist whose magnitude is a measure of transmitted power. Such power is often measured by means of a suspension and a weighing apparatus, but this method is not suitable for systems with high speed revolution or for operation from a remote distance. Stroboscopic methods are also unsuitable for remote operation. The twist in the rotating shaft can also be measured by using suitable strain gauges and this can conveniently be operated from a remote distance but the speed limit of the shaft is restricted. Thus an early torquemeter due to Ford[1], based on change of inductance between two similar transformers fixed at a known distance apart on the shaft, which was designed to measure torque on the propeller shaft of ships having limited speed, is not suitable for high speeds.

A new type of electronic torquemeter has recently been described by Rakshit and Mukherjee[2] in which the torque or twist of a rotating shaft, transmitting mechanical power can be measured with a high degree of accuracy from a remote distance by a method based on the principle of delayed coincidence of pulses. Development of such an instrument was necessitated by the absence of any suitable apparatus for measuring the power transmitted by a shaft rotating at very high speed (say 15 000rev/min).

In the new system two sharp pulses of voltage are proproduced in two magnetic pick-up coils due to the change in flux caused by two balanced blades of magnetic material fixed at a known separation on the rotating shaft. Initially, when the shaft is running free without any load, the positions of the pick-up units are so adjusted that the two pulses occur at the same instant. When the shaft transmits power, it will undergo torsional deflexion and as a result the pulses will be separated by a small interval of time, which is dependent upon the angle of twist and the rotation speed of the shaft. It has been suggested[3] that the speed of the shaft can be measured with great accuracy by the method of coincidence of two sets of recurrent pulses. Thus the main problem in the electronic torquemeter is to measure the very small time interval between the two pulses. This interval T is given by

$$T = \Theta/6n \text{ second} \qquad (1)$$

where Θ = amount of twist angle in mechanical degrees between the two blade positions,

and n = speed of the shaft (rev/min).

Thus when n = 15 000rev/min and $\Theta = 0{\cdot}1°$ which is the permissible full load twist angle per foot of mild steel shaft, T becomes approximately 1·1μsec.

Theory of Operation of Torquemeter

The term time-interval in connexion with this work can be defined as the time that elapses between the occurrences of identifiable portions of two waveforms generated separately in two pick-up coils. The nature of the generated waveform as seen on the screen of a cathode-ray oscillograph at a shaft speed of 1 500rev/min is shown in Fig. 1, in which it is evident that the portion of the curve through the point A is most well defined.

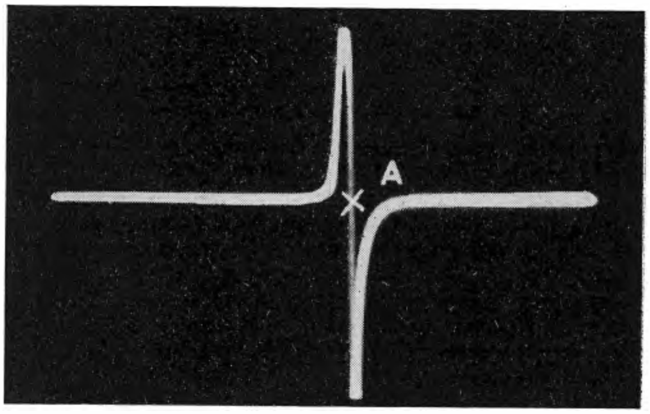


Fig. 1. Generated voltage in pick-up

The negative going pulse from point A of the waveform has therefore been taken as the identifiable portion of the pulse for measurement of time-interval. Since the two sets of pulses are generated by the same source having a constant repetition frequency, the first pulse occurring in advance on the time scale is ahead of the second pulse by the same amount every time. As a result, if the first pulse is delayed with respect to the second by the appropriate amount then these pulses will always coincide. A technique of detecting coincident pulses by aural and visual methods has been specially developed for the purpose and will be described later. In the operation of the torquemeter unit it will be noted that the delay of the small time interval should be continuously variable. Electrical delay circuits have special advantage in this respect and the delay can be smoothly varied over a wide range

* *Indian Institute of Technology, Kharagpur.*
† *Bengal Engineering College, Howrah.*

as required. The minimum delay cannot however be reduced to zero. This difficulty can be overcome if two delay circuits are incorporated in the two channels corresponding to the two pick-up coils, so that the output pulses from both of the sources initially undergo the same amount of delay. Alternatively, an initial fixed delay of not less than the minimum delay of the circuit may be introduced in one of the channels and the variable delay of the other channel adjusted for initial coincidence. This initial fixed delay can of course be introduced by fixing the two blades with a pre-determined angular separation dependent upon the operating shaft speed according to equation (1). Since the angular separation is to be altered as the speed is changed, this arrangement makes the torquemeter less flexible. Further, since $\delta T = -\Theta.\delta n/6n^2$, it follows that if only one delay circuit is to be used, along with initial twist Θ_0, there will be a jitter proportional to Θ_0 if the speed varies. Under load conditions the jitter due to variation in speed will increase since the overall Θ then increases. To keep such jitter to a minimum it is therefore desirable to keep Θ_0 to a minimum, obviously showing that it is preferable to use two delay circuits with the requisite initial delays and thus work with zero initial twist. The block diagram of the arrangement is given in Fig. 2.

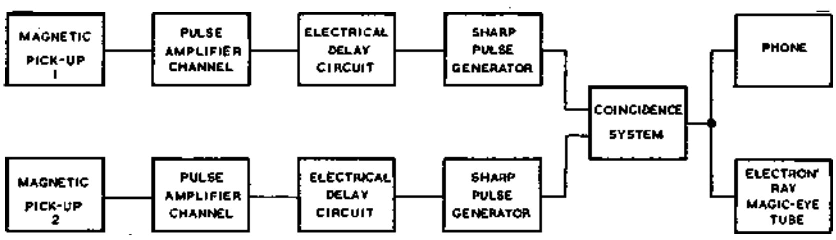


Fig. 2. Block diagram of the torquemeter

The pulse from each pick-up is fed to a pulse amplifier channel whose output triggers an electrical delay circuit. The delayed output pulse triggers a thyratron type pulse generator and two such negative outputs obtained from the two corresponding pick-up units are applied to the coincidence system, which gives a large output only when the two pulses coincide.

The different units in each channel are associated with a small inherent delay due to their operation. For satisfactory measurement of a small time-interval, it is essential that the inherent delays of these units should be as small as possible and highly stable, as otherwise the jittering effect between the two delayed pulses will not give accurate results.

Electrical Delay Circuits

The basic operating principle of all electrical delay circuits is similar to that of the relaxation oscillator, which when properly triggered by a pulse will generate a single sweep. The sweep speed is determined by an *RC* differentiator in the circuit and the time during which the sweep waveform runs can be controlled by d.c. voltage with a precision potentiometer. From the terminating point of the sweep an output pulse delayed with reference to the input triggering pulse is obtained. Taking into consideration the nature of the sweep waveform, electrical delay circuits can be classified in two groups:

(1) Exponential sweep waveform, i.e. triggered multivibrator or univibrator.

(2) Miller sweep generator, which includes the Phantastron, Sanatron and Sanaphant circuits.

Owing to the exponential nature of the timing waveform of the multivibrator circuit, the delay cannot be controlled linearly. The Miller sweep generator or Miller integrator produces a waveform which varies linearly with time. In an idealized Miller integrator, which is essentially a feedback amplifier, the output voltage is given by

$$E_{out} = A\,E_{dc}\left[1 - \exp\left(-\frac{t}{(1+A)RC}\right)\right]$$

where E_{dc} is the d.c. voltage applied to the series grid resistor, A is the stage gain and RC is the time-constant of the timing circuit. If $t \ll (1 + A)\,RC$,

$$E_{out} = E_{dc}.t/RC, \text{ approximately for large values of } A.$$

The use of a linear rather than an exponential timing waveform leads to a very important advantage, namely that the delay can be made a linear function of an input control voltage through a precision potentiometer with a linear scale.

The Phantastron, Sanatron and Sanaphant are the three main delay circuits, the sweep waveform of which can be varied linearly by controlling a d.c. voltage. The total variation in delay required in an electronic torquemeter for covering the entire range of operation is of the order of 20μsec. The Sanatron and Sanaphant although more complex than the Phantastron, can generate fast waveforms with a delay as short as 1μsec. On the other hand there is a limitation to the Phantastron's ability to generate a short delay, as by mere reduction of the time-constant, *CR*, the sweep speed cannot be increased owing to reduced screen dissipation[4]. Another disadvantage is that the end of the sweep, or Miller run-down, in the Phantastron is governed by the 'bottoming' of the anode potential, which may vary occasionally. As a result the total delay is not constant and hence the timing accuracy not very good in the Phantastron. In the Sanatron and Sanaphant, on the other hand, the sweep is ended at a fixed d.c. potential of the anode by the conduction of a diode. It has been found experimentally that the timing accuracy of a single Sanatron generating 20μsec delay is very high and the short-time variation does not exceed about 0·01μsec. The performance and complexity of Sanaphant is intermediate between the Sanatron and Phantastron. In comparison to the Sanatron the Sanaphant shows larger variations in delay when the operating conditions change[4]. It has been shown in Fig. 2 that two electrical delay circuits are required for measuring the time interval. With such an arrangement, a time interval of 20μsec can be measured by using either two Phantastrons or two Sanatrons in the two channels, although the generated delay in a practical Phantastron circuit is comparatively larger than 20μsec.

Stability of Phantastron and Sanatron

Before finally selecting the Sanatron in view of its advantages as discussed above, the stability of the Sanatron and Phantastron was estimated experimentally. The arrangements used for the purpose are shown in Figs. 3 and 4. It will be seen from Fig. 3 that the negative reference pulse simultaneously triggers the two Phantastrons. The delayed output positive pulses from these are amplified by suitable pulse amplifiers and are finally used to excite two blocking oscillators. The delays in the two Phantastrons are so adjusted that the sharp output pulses from the two blocking oscillators differ by a very small interval of time. The pulse occurring earlier is used to trigger a fast time-base and the other pulse is viewed on this time-base. The repetition frequency of the negative reference pulse was 50p/s. The arrangement for the Sanatron is very similar (Fig. 4). In order to trigger the Sanatron the reference pulse was derived from a low impedance cathode-follower source. For such an operation it is also essential that the negative pulse should have sharp leading edge because the triggering of Sanatron is done through a series diode.

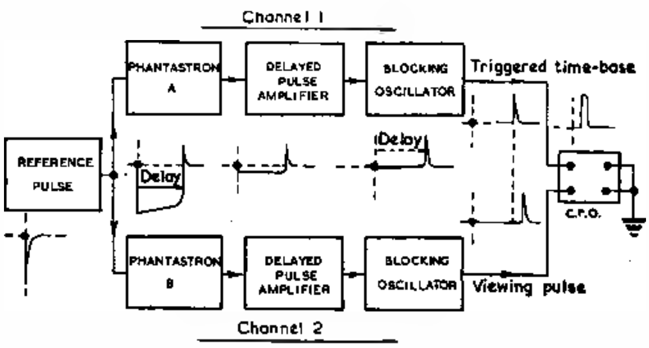


Fig. 3. Block diagram for phantastron testing circuit

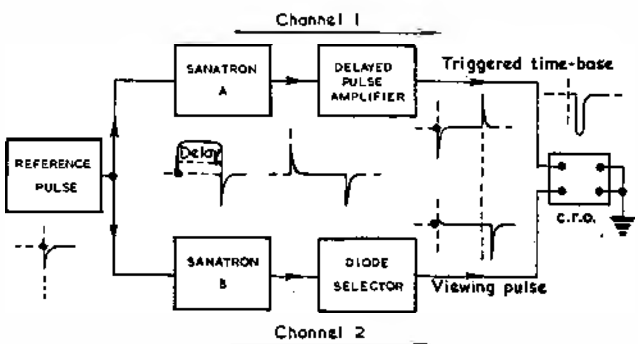


Fig. 4. Block diagram for Sanatron testing circuit

If the timing stability of the two channels is high, the viewing pulse from the second channel will remain stationary on the triggered time-base of the cathode-ray oscilloscope. On the other hand the position of the viewing pulse will shift back and forth in case of instability or jittering of the timing circuit.

The observed timing stabilities of the two types of circuit are given in Table 1. It is seen that the percentage jitter is smallest in the Phantastron with 2 500μsec maximum delay but in the Sanatron it is the same for different maximum delays. In the present work the absolute value of jitter is more important, since a very small time-interval, of the order of 1·0μsec will have to be measured with precision. The Sanatron designed for small maximum delay would therefore be more useful. The leading and trailing edges of the output waveform in the Sanatron being sharp, the delayed output is obtained by a process of differentiation and rectification. The delayed negative output is further phase inverted by means of a pulse amplifier in order to trigger the sharp pulse generator in the next stage.

TABLE 1.
Experimental Observation of Phantastron and Sanatron.

CIRCUIT	MAX. POSSIBLE DELAY (μSEC)	TIME-CONSTANT R(MΩ)	TIME-CONSTANT C	JITTER (μSEC)	PERCENTAGE JITTER
Phantastron	2500	1	0·01μF	0·05	0·002
Phantastron	150	1	500pF	0·02	0·013
Sanatron	2500	1	0·01μF	0·25	0·01
Sanatron	40	0·22	200pF	0·004	0·01

Sharp Pulse Generator

The accuracy of measurement of a small time-interval between two pulses is dependent upon the sharpness and also the duration of the pulse. The delayed output from a Sanatron being unsuitable for precision measurement, the operation of a number of triggered pulse generators was studied. It has been observed that by a proper choice of circuit components a triggered thyratron can generate very sharp pulses with rise-times and width estimated to be less than 0·01 and not more than 0·04μsec respectively. The stability of operation of a thyratron has been found to be very high in comparison to blocking oscillator and other sharp pulse generators. The inherent delay between the input triggering pulse and the sharp output pulse of thyratrons has been observed to be different for different valves, and this delay changes whenever the repetitive frequency of the X triggering pulse or the time-constant RC of the pulse forming circuit is changed. By using type 2050 or 2051 valves, which possess steep anode voltage-grid bias conduction characteristics, this difficulty has been avoided. In order to prevent interaction due to firing between the thyratron and the timing circuit with the Sanatron, a cathode-follower has been used between the stages.

Coincidence Technique

In a typical coincidence method, the two pulses are applied to the grids of two valves with a common anode load resistance. A large output is obtained only when the two pulses occur simultaneously. When the pulses are not in unison the amplitude of the output is negligibly small. The resolving time which is defined as half the maximum time separation for which the pulses will remain coincident is obviously dependent upon the pulse width. In the case of two identical rectangular pulses the resolving time becomes equal to the width of each pulse. In general, the resolving time and hence the accuracy of coincidence is a function of the pulse width. If the common anode load in the conventional Rossi circuit is not very small, with resulting decrease in output, the accuracy of coincidence is low for pulses of small length. In order to increase the accuracy of coincidence without undue reduction of output a new method of coincidence has been developed as shown in Fig. 5.

Valves V_1 and V_3 are two triggered thyratrons for generating very sharp pulse. The resistance r controlling the width of the output pulse from V_1 and V_2 is common, so that when the two pulses are in unison, a large negative pulse is obtained across r the shape of which does not distort due to the additive effect of the two individual sharp output pulses. When the pulses are not in unison, the output becomes less than half of the amplitude of the coincident pulses. Valve V_3, normally a triggered thyratron, has been used as an amplitude discriminator for the output across r and the negative grid bias is so adjusted that the valve V_3 fires only when the output across r exceeds a certain peak value. The amplitude of the input discriminating voltage can be adjusted to 80V (peak) and the accuracy of discrimination to 4V (peak). Using a headphone across the output r' the accuracy of coincidence has been found to be better than 0·02μsec.

Aural-Visual Detection

The time interval between two pulses from two sources can thus be measured with the delay circuit by the method

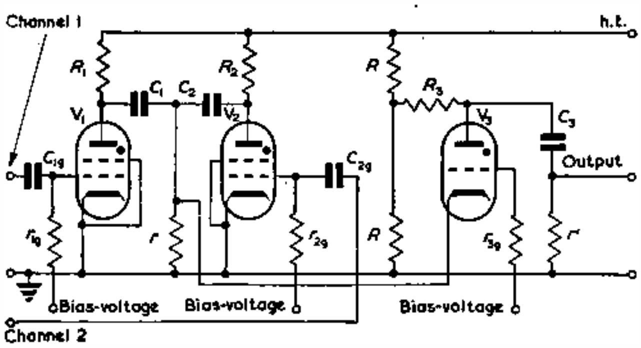


Fig. 5. Coincidence circuit in torquemeter

of coincidence and the coincidence can be detected by a pair of earphones. The repetition frequency of the output pulse across the earphone will be the same as that from the two sources, but the output pulse being sharp the response in the earphone is not loud. Moreover visual detection using a magic-eye type electron ray tube has to be provided in the torquemeter for proper operation in noisy surroundings. But the average change in illumination in the magic-eye is not prominent when the input pulse is sharp. The width of the output pulse across r' in Fig. 5 cannot be much increased by increasing C_3 without decreasing the pulse amplitude. A pulse broadening circuit using a triggered multivibrator has been incorporated which can generate negative pulses of 600μsec duration. This broad pulse can now be detected by either an aural or visual method. The accuracy of measurement of the entire unit, when the reference pulse is derived from one source having repetition frequency equal to 400p/s has been found to be 0·01μsec.

Conclusion

The above principle of time interval measurement has been utilized for measuring a very small angle, Θ, between two projected blades fixed 6in apart on a shaft rotating at approximately 1 500, 3 000, 6 000 and 9 000rev/min. Instead of a twist in the shaft produced by a load, the twist was simulated in the following way: Initially the delays of the two channels were adjusted to minimum values and made nearly equal. The shaft was then made to rotate at the desired speed and the pulses were made coincident by fine adjustment of the delay of one of the Sanatrons. One of the pick-up coils was then shifted by a measurable amount by a micrometer arrangement, specially fixed for the purpose. This produced the measurable angular shift Θ between the two pick-up coils. Coincidence was then restored by varying the delay of one of the Sanatrons.

The accuracy of the observed values of angular deviation Θ as obtained from the readings of the delay potentiometer has been found quite satisfactory. Details of observations on the performance of the torquemeter are being separately communicated.

Acknowledgment

The authors are grateful to the Council of Scientific and Industrial Research for financial help in carrying out the investigations and grant of a scholarship to one of them (S.C.M.). Thanks are also due to the Council for permission to publish the results.

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A Simple Transistor Amplifier for Energizing a Hall Multiplier

By D. J. Lloyd*, B.Sc.(Eng.)

A Hall-effect multiplier in which the Hall plate is indium antimonide presents a low, variable resistance to the source which supplies it. Since true multiplication occurs between current in the plate and flux density, the current source should have a high impedance if the multiplication is to be accurate. The amplifier described has this property, together with low distortion and economy of power requirements.

If a magnetic field is maintained through and perpendicular to the plane of a thin rectangular slice of conducting or semiconducting material, and a current is passed through the lamina parallel to one pair of sides, then in most materials a potential difference proportional to the product of current and magnetic flux density can be detected between opposite points on these sides. This is called the Hall effect, after its discoverer, and the slice of material a Hall plate.

For a given temperature rise, length and breadth of plate and flux density, the greatest power output is available from a Hall plate of indium antimonide[1]. A plate of this material has, however, two disadvantages which impose unusual limitations on the source which supplies its input terminals. The first disadvantage is its low resistivity, as a result of which it is difficult to make the plate input resistance more than about 1Ω. The second is its large magneto-resistive effect; since accurate multiplication occurs between input current and flux density, the input quantity should be a current rather than a voltage if errors in multiplication due to this cause are to be avoided. The order of current commonly used is 0·5A.

The special characteristics required of the amplifier feeding the indium antimonide Hall plate terminals are, therefore, that it shall be capable of supplying a low-resistance load with a current of 0·5A or more, and under these conditions behave as a constant-current source. A similar amplifier can be used for energizing the coil producing the magnetic field, but the requirements here are less stringent because the impedance of the winding is largely under the control of the designer.

Specific Requirements

In a particular application a Hall multiplier was required

* *Electrical Engineering Department, University of Bristol.*

to operate with input signal frequencies in the range 0 to 30c/s. Its input resistance was approximately 1Ω. The current available was that from a demodulator consisting of a pair of CV448 (OA71) diodes and this had to be raised to the order of 1A at the input of the multiplier by means of a direct-coupled amplifier. In this particular case a standing current (d.c.) of up to 0·75A could be tolerated.

The Amplifier

Thermionic valve circuits were first considered but these would have been inefficient, wasteful of power, and cumbersome. Low voltages and high currents precluded the use of these but suggested that a transistor amplifier—with a high power type such as an OC16 in the final stage—might be used.

The curves of collector current against collector voltage for common-emitter operation of the OC16 all have pronounced 'knees' at about V_c − 0·5V, and above this collector voltage the curves are almost constant-current lines, the output impedance of an OC16 amplifier in this region being comparatively high. For a given supply voltage the load line can be made to pass through the constant-

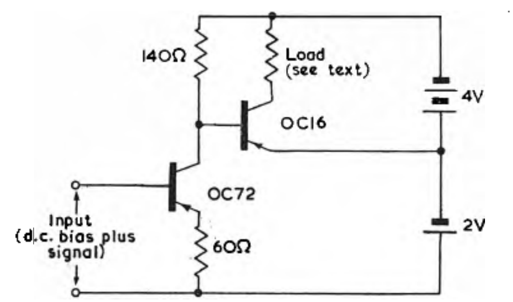

Fig. 1. Circuit diagram showing an OC72 directly-coupled to an OC16

current lines only if the load resistance is small, but this condition is automatically fulfilled when the load is an indium antimonide Hall plate.

A current gain of 1 000 or more was required and with this in mind an OC72 was directly-coupled to an OC16 in the circuit shown in Fig. 1. The values of the resistors were altered until a combination of good linearity and reasonable thermal stability was obtained, and the figures given in the circuit diagram are those which were eventually chosen for one particular OC72-OC16 combination. The measured curve of input current against output current for this combination is given in Fig. 2, and the optimum standing output current can be seen to be approximately 0·6A.

Distortion

As would be expected, the distortion produced in the amplifier depended on the source impedance of the input circuit. Since a near-linear relation was observed between input and output currents and the input voltage-current characteristic of the OC72 was not linear, it was assumed that least distortion would occur when the input was fed from a high-impedance source. This was found to be so.

A sine-wave input was superimposed on the d.c. base bias to the OC72. With a 40kΩ input resistance, a standing output current of 0·6A and a peak-to-peak output swing of 0·8A the second harmonic component was approximately 0·5 per cent of the fundamental, and the third and higher components were less than this. On decreasing the input resistance to 7·5kΩ and adjusting the d.c. base bias to the OC72 to make the other conditions the same as before, the 2nd harmonic component increased to 2·5 per cent of the fundamental output.

Frequency Response

The response of the amplifier was found to be adequate for the purpose; it was flat within ½dB up to 1kc/s. At higher frequencies the response fell, the following figures being obtained:

4·5kc/s	− 3dB
8·5kc/s	− 6dB
14·5kc/s	− 12dB

Output Impedance

This was not measured directly but it was obtained for low frequencies from curves of collector current against collector voltage for the OC16. For an output current of 0·6A the ratio $\frac{\text{output impedance}}{\text{load impedance}}$ was greater than 100 for all load resistances less than 2Ω. Thus even if the resistance of the nominal 1Ω Hall plate increased by 100 per cent due to temperature or magneto-resistive effects, the input current to it would change by less than 1 per cent when energized by the amplifier described.

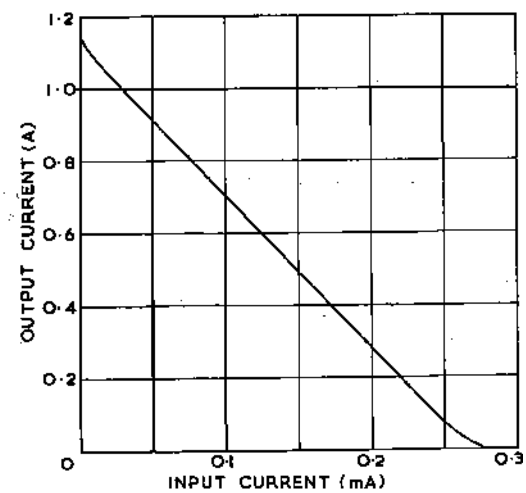

Fig. 2. Measured curve of input current against output current

Stability

For the particular purpose for which the amplifier was required the short-term stability, over times of the order of seconds or a few minutes, was much more important than the long-term stability.

After an initial warming-up period of about 20min the short-term drift was not measurable on standard instruments. The long-term drift was checked as a matter of interest and the change of d.c. output current was approximately 1 per cent in two hours. The stability of the circuit was also directly affected by that of the battery supply.

Conclusions

The amplifier is suitable where a low resistance load has to be fed from a constant current source at frequencies from zero upwards. This combination of requirements is somewhat unusual so it is not expected that the amplifier will find wide application; it does, however, adequately perform its original function of energizing the input terminals of a Hall multiplier.

Its advantages are simplicity and low distortion. In other applications the frequency response and long-term stability might prove inadequate. The standing d.c. output current might also be undesirable but this could presumably be backed off with an equal and opposite current from another high impedance source.

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LETTERS TO THE EDITOR

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

Oscilloscope Pre-amplifier

Dear Sir,—In the July issue Mr. Aitchison described a useful oscilloscope pre-amplifier using transistors. A similar circuit is shown in Fig. A, which may have features of general interest, particularly when many such units are required as part of complicated equipment.

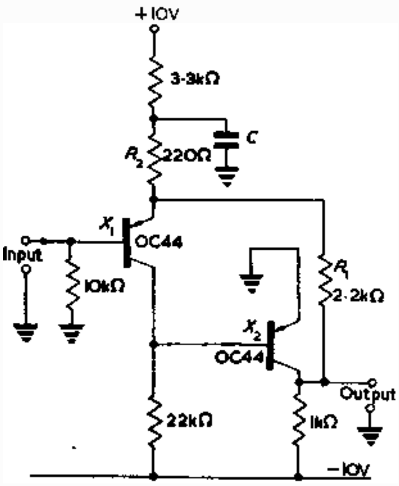


Fig. A. The transistor amplifier circuit

The resistors, R_1 and R_2, determine the voltage gain. The frequency response is determined by the value of C at low frequencies and the high frequency response can be controlled by a shunt capacitor connected across R_1. The gain of the circuit is $\times 10$ and the upper frequency at the 3dB point is 2Mc/s. Further improvement can be obtained by inductive compensation of the collector of X_1.

There are a number of advantages in using a d.c. connected circuit. The transistors operate under constant current conditions, which improve the overall stability of the circuit. The number of components is reduced and high value capacitors avoided, these capacitors may be large and if electrolytic, unreliable. The d.c. potential of the output terminal is defined and may be used as an overall check. This last point is regarded as an important feature of any electronic circuit, and it is of very great help in commissioning and maintaining complicated equipment.

The circuit has been tested up to 40°C and no change in gain, or in the d.c. conditions, have been observed. Three such units have been operated in cascade giving an upper frequency limit greater than 0·5Mc/s and an overall gain of 1 000. Low frequency transistors may be substituted without changing the component values.

Yours faithfully,

B. Collinge

Nuclear Physics Research Laboratory,
University of Liverpool.

Dear Sir,—The appearance of Mr. R. E. Aitchison's letter in the June issue prompts the forwarding of the similar circuit shown in Fig. B. The feedback pair illustrated offers improved performance by the omission of the coupling capacitor from the gain determining feedback loop and further by returning the bias resistor of X_1 to the emitter of X_2 rather than the collector of X_1. Although these alterations are simple, significant advantages accrue.

By omitting the coupling capacitor two useful consequences result:

(1) The loop frequency is held uniform rather than the rising characteristic which would commence at approximately 180c/s with the components shown by Mr. Aitchison. Uniform feedback is desired to combat the familiar l.f. noise of transistors.

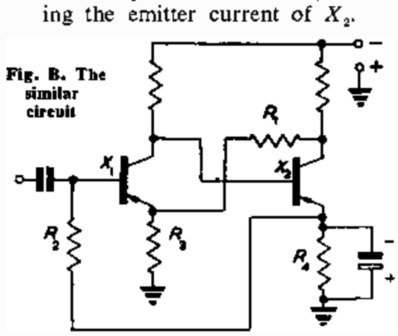
(2) D.C. feedback now embraces the full open gain of the amplifier to increase bias stability.

Returning the bias resistor R_2 to the emitter of X_2 achieves the following desirable characteristics:

(1) A.C. voltage feedback is removed from X_1 while retaining d.c. stabilization. This increases open loop gain as well as raising the input impedance.

(2) A degree of bias stabilization is added to X_2 since a new d.c. feedback loop has been formed containing the emitter current of X_2.

Fig. B. The similar circuit



As can be seen, all d.c. feedback loops aid stabilization. The self-current feedback due to R_3 and R_4 add stability to X_1 and X_2 respectively. The voltage feedback through R_1 and R_2 combat variations in either of both X_1 and X_2.

Component values would be identical for the two circuits if the same gain is desired. If compensation for supply voltage and temperature are the limiting criteria the circuit of Figure B is capable of higher gains with the same stability as Mr. Aitchison's circuit. If variation in transistor parameters is limiting, the results obtained with either circuit should be identical.

Yours faithfully,

T. C. Penn

Texas Instruments,
Dallas, Texas.

The author replies:

Dear Sir,—Dr. Collinge's amplifier is an interesting variation of the basic feedback pair design. Most of his comments on d.c. coupling of the amplifier[1] are met if the output capacitor is omitted; however in the original design it was included to avoid annoying d.c. shifts when changing from the amplifier to the straight through connexion. (Note that the sensitivity when the amplifier is used with a particular wide band oscilloscope is 0·5mV/cm.). The input capacitor (to protect the input transistor) is a Bipolar 1·25μF 50V tantalum type not a normal electrolytic as shown in error in the original circuit.

Most of the details of the original design are a result of the decision to keep the power consumption to a minimum, the input impedance high, and to use a single battery bias system for mechanical reasons (small size of battery, on-off switch, and battery mounting). The volume occupied by the other components is by comparison very small. Miniature electrolytic capacitors with a leakage of less than 0·5μA/μF have proved extremely reliable.

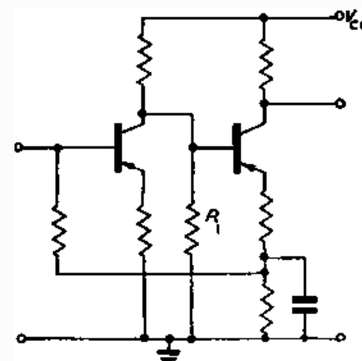


Fig: C. Direct coupled pair bias system, a.c. feedback not shown

A new one battery bias circuit, a modification of an earlier circuit[2], has been mentioned recently[3], the design being more flexible and the performance better than the normal two-stage design[4]. The arrangement is shown in the accompanying Fig. C; if a further resistor R_1 is included it introduces an extra degree of freedom in the design. This circuit is particularly suitable for pre-amplier feedback pair bias design, and the thermionic valve equivalent is a useful circuit for use in high gain direct coupled thermionic valve amplifiers.

The subsequent letter from Mr. T. C. Penn also discusses the circuit, and I endorse his comments on the advantages of this feedback pair design.

Yours faithfully,

R. E Aitchison

Department of Electrical Engineering,
The University of Sydney.

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Short News Items

The John Price Wetheril Medal of the Franklin Institute has been awarded to the British inventors of the cavity magnetron which provided the first high power source of centimetre radio waves for use in radar equipment during the war and since. The award is to Dr. H. A. H. Boot, now at the Services Electronics Research Laboratory, Baldock; Professor J. T. Randall, F.R.S., Wheatstone Professor of Physics in London University (King's College); and Professor J. Sayers, who holds the Chair of Electron Physics in Birmingham University. The award is "for their basic development which provided the first useful high power pulsed microwave magnetron and which established the fundamental principles upon which all later developments in this field were based." The cavity magnetron was developed in Birmingham University.

The newly introduced range of Murphy v.h.f. Mobile Radio Telephone equipment was recently subjected to exceptionally stringent tests when radio-telephone units were installed at the T.T. Races in the Isle of Man. With a mobile transmitter/receiver mounted on each travelling race marshal's motor cycle and a base station remotely controlled from the grandstand, race officials were able to maintain permanent communication with the marshals. The duties of a travelling marshal require that he shall patrol a section of the course during the actual progress of a race and often, speeds approaching 100 mile/h are maintained. Despite the exceedingly rough treatment that the equipment received both during trials and three days of actual racing, it continued to operate satisfactorily throughout. The value of radio-telephone equipment was again particularly apparent, when, after an unfortunate accident which brought down a telegraph pole in the path of oncoming riders, a travelling marshal was able to radio assistance to the spot and have the road cleared without further mishap.

R. H. Cole (Overseas) Ltd, the U.K. distributors for rectifiers made by the German Siemens and Halske organization, announce that their principals have now released a silicon half-wave rectifier measuring 7mm diameter × 17mm. This has an exceptionally high peak inverse voltage rating of 750V and is rated for 400mA d.c. load with a capacitive load of 500mA with a resistive load.

Cawkell Research & Electronics Ltd, formerly A. E. Cawkell, Electronic Engineers, have moved into larger premises, the new address being Scotts Road, Southall, Middlesex. The telephone number is Southall 3702/5881 as before.

Sir Lionel Harris, Engineer-in-Chief of the Post Office, recently laid the foundation stone of the new Electronic Switching Laboratories at Ericsson Telephones Ltd, Beeston. The new laboratories will rehouse and provide additional space for research and development of electronic telephone exchanges. Some three years ago the Post Office, Ericsson Telephones and other large telephone companies agreed to pool research into all electronic telephone exchange systems and, as a result of this joint development, manufacture will shortly begin of the first example of the new system which will be installed at Highgate Wood in North London. The new building has been designed by a team under the company architect, Mr. F. S. Eales, in collaboration with Dr. J. H. Mitchell, Research Director; Mr. J. R. Pollard, Head of Communications Research, and staff of the Electronic Switching Division under Mr. J. T. Stringer, who is in charge of the Ericsson share of the joint development programme. It is expected that the new building will be occupied early next year.

The Northern Polytechnic has been given permission by the Ministry of Education to hold a full evening course on the Principles and Practice of Colour Television. It is intended to cover the fundamentals of television transmission and reception in colour. The entry qualification required is at least City and Guilds Radio III or City and Guilds Final Certificate, and some knowledge of monochrome principles will be necessary. Polytechnic staff will be dealing with the bulk of the course but specialist lecturers will be engaged for items of particular interest. Provision has been made for practical work on colour receivers and, in the event of experimental transmissions taking place during the run of the course, full facilities exist for their display. Further details may be obtained from Mr. J. C. G. Gilbert, Head of Department of Telecommunications, the Northern Polytechnic, Holloway, London, N.7.

The Eleventh Annual Conference on Electrical Techniques in Medicine and Biology will be held in Minneapolis, Minnesota, from 19-21 November at the Nicollet Hotel. It is sponsored by the Institute of Radio Engineers, the American Institute of Electrical Engineers, and the Instrument Society of America through their Joint Executive Committee on Medicine and Biology. The theme for this conference will be Biology and Computers.

The Electronics Group of the Institute of Physics has arranged a meeting on "Solid State Memory and Switching Devices," for 26 and 27 September at University College, London. Further particulars may be obtained from the Institute, 47 Belgrave Square, London, S.W.1.

Dr. N. Levin has been appointed a Deputy Director of the Atomic Weapons Research Establishment.

Airmec Ltd have recently negotiated an agency agreement with Apparachhi Scientifici Federici of Milan for the exclusive rights to market their ultrasonic drilling and ultrasonic medical equipment in the United Kingdom.

Kelvin & Hughes (Industrial) Ltd have moved their administrative offices from 2 Caxton Street, Westminster, London, S.W.1 to their new building at Empire Way, Wembley, Middlesex. Telephone number Wembley 8888.

Fry's Metal Foundries Research Department have developed a special safety flux, Type S.31, to enable the nickel-chromium alloys to be readily tinned and soldered. The flux is in liquid form, it is non-corrosive to nickel-chromium and to copper and it has no detrimental action on paper or fabric insulation.

CIBA (A.R.L.) Ltd is the new name for Aero Research Ltd of Duxford, Cambridge. Aero Research was formed in 1934 to pioneer research into aircraft structures, including the development of new adhesives. The new name has associations with the past but emphasizes affiliation to the world-wide CIBA organization. The trade names of 'Aerolite,' 'Araldite,' 'Redux' and other proprietary products are unaffected.

Errata. In the article, "Triodes and Tetrodes for U.H.F./S.H.F. Operation" by Messrs. Tremlett and Williams, published in the May issue, the EC57 valve given in the table on page 339 is incorrectly stated to have a power gain of 4dB instead of 8dB and the power output should be 1·8W instead of 1·5W.

The caption to Fig. A in the letter from Mr. R. B. Mackenzie in the August issue (page 509), incorrectly reads, "Simpified drawing of Pacak's circuit." This is, in fact, the circuit dealt with in Mr. Mackenzie's article*, quoted as Reference 2 in Mr. Pacak's letter in the April issue (page 212), viz.

* MACKENZIE, R. B. A New Electronic d.c. Voltage Stabilizing Circuit. *Proc. Instn. Elect. Engrs 101*, Pt. 2, 59 (1954).

BOOK REVIEWS

Network Synthesis

By D. F. Tuttle, Jr. 1175 pp. 676 figs. Medium 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd., London. 1958. Price 188s.

THREE new books on the subject of passive network synthesis have recently arrived from the United States of America. The first to appear was Storer's *Passive Network Synthesis* which provided a concise survey of the field and the second was Guillemin's *Synthesis of Passive Networks* in which the theory of the subject was formerly developed. The latest arrival is the first volume of Tuttle's two-volume work, *Network Synthesis*.

This new addition to the literature of synthesis bears evidence of a tremendous amount of preparation. It deals only with the synthesis of one-ports according to prescribed behaviour in the sinusoidal steady-state; the second volume will deal with two-port networks. After an introductory review of the principles of modern network analysis and the notion of complex frequency, the book begins with a thoroughgoing exposition of Brune's theorem that if $W(s)$ represents the driving-point immittance of a passive one-port it is necessarily positive real; that is, it is real when s is real and confined to the right half plane when s is confined to the right half plane. Certain properties of Brune functions are then discussed in considerable detail and a test for positive reality is developed. Next to be considered are the special classes of Brune functions which are realizable as non-dissipative one-ports with two kinds of energy storage and dissipative structures with one kind of energy storage The methods of Foster and Cauer for realizing such functions are described and illustrated with examples. At this juncture the theory of Brune functions is developed further in preparation for the study of the general dissipative one-port containing two kinds of energy storage. The concepts of minimum reactance, minimum resistance, and their duals are introduced and the technique of reducing a given Brune function to a 'minimum immittance' function by reactance and resistance extraction is explained. In this connexion, the important work of Gewertz, Bayard, and Bode is presented at length. These discussions lead to Brune's demonstration that positive reality is not only a necessary condition but also a sufficient condition for a given function to represent a passive one-port. The theory of passive one-port synthesis is then extended to include the general three-element case and is thus completed. The treatment of realization concludes with such topics as the Bott, Duffin, and Pantell methods for avoiding mutual inductance.

The last three of the book's fifteen chapters are concerned with the approximation aspect of the synthesis problem and occupy a little less than half the book. The subject is covered with remarkable thoroughness. Among the techniques considered are: approximation by trial and error; the least-mean-square-error method; Taylor approximation; and Chebyshev function approximation. Methods based on the well-known analogy between complex-variable theory and two-dimensional potential theory receive exhaustive consideration.

It is possible here only to indicate the subjects treated and the sequence of presentation; it is not possible to give a true impression of the wealth of electric network knowledge represented in the book's eleven hundred pages. Numerous problems are provided at the end of each chapter and appendices contain computational information and an extensive list of references. It is a magnificent volume and will surely establish itself as the standard reference work in its field for some time to come.

S. R. Deards

Präzisionsmessungen von Kapazitäten, Induktivitäten und Zeitkonstanten (Precision Measurements of Capacitances, Inductances and Time Constants)

By Dr. Erich Blechschmidt. 166 pp., 74 figs. Small 8vo. Friedrich Vieweg & Sohn, Braunschweig. 1957. Price DM.11.50.

THIS booklet forms the second part of the second, revised edition of a monograph published in 1940. The first part, dealing with capacitances, dielectric losses and permittivities, appeared in 1956 and was reviewed in these columns in September 1957. The present little volume has the subtitle *Precision Measurements of Inductances, Coil Losses and Time Constants.* The purpose and the general treatment and arrangement of the subject matter are the same as in the first part and it may be used without knowledge of this first part. Therefore a few repetitions proved to be unavoidable.

A brief survey of the theory in the first chapter, in which the calculation of self and mutual inductivities is dealt with, is followed by a description of the various standards of inductivity in the second chapter. Measuring methods form the subject of the third chapter and they are subdivided into direct and indirect methods for determining the self and mutual inductivities respectively. Also the measurement of time constants is described in this chapter. The fourth chapter contains again a very extensive bibliography of about 20 pages with references till 1956. A name and subject matter index conclude this useful little book.

R. Neumann

Bibliography on Medical Electronics

91 pp. Crown quarto. Prepared by the Medical Electronics Centre of the Rockefeller Institute. Published by the Professional Group on Medical Electronics, Institution of Radio Engineers, New York. Paper cover. Price $2·50.

IN his introduction, Dr. V. K. Zworykin broadly defines medical electronics as "comprising any application of any of the branches of electronics, such as acoustics, communications, television techniques, spectrophotometry, or dielectric heating, to any problems of biological or medical research, therapy, public health and related fields." It is not surprising, therefore, that this bibliography is very far from being complete, a fact that Dr. Zworykin freely admits. The basic fault is that the fundamental definition is much wider than indicated by the title, and many references included could be more accurately listed under other headings such as Biophysics. Moreover, in a number of cases the subjects have apparently no direct connexion with medicine, one example being, 'A Simple High Temperature Microwave Spectrometer'.

This wide range of subjects may be contrasted with the limited area from which references have been drawn. Most of the papers were published in the United States, which is understandable since this bibliography originated as a card index at the Medical Electronics Centre, and was supplemented from the personal files of some prominent American research workers. The British contribution to Medical Electronics has not, however, been completely neglected, but many worthy names are noticeably absent from the Author Index. Foreign language references are almost non-existent. The compilers are keenly aware of this limitation, and one of the reasons for the recent International Conference on Medical Electronics was the formation of national groups capable of providing world-wide references for the projected annual supplements and future editions of this work.

The present bibliography is divided into three sections, the first 44 pages containing the full references listed in arbitary order and numbered consecutively up to 2 200. The subject index of 32 pages, which is cross-referenced and includes over 2 000 headings, indicates the appropriate numbers in the main section. A third index lists the senior authors in alphabetical order. It is stated that related subjects are grouped together within the main section, but this is not entirely so. Electrocardiography, for instance, is mainly covered by one group of 49 references (Nos. 301 to 349), and within this group the authors' names are alphabetically listed. However, a further 55 references to this subject are scattered about, apparently at random, either singly or in much smaller groups, but of course they may still be traced through the subject index.

Despite these criticisms, it must be emphasized that much previously scattered information is now rendered accessible for the first time through the pages of this bibliography. This alone makes it an important new publication, and the prospect of its being followed by supplements designed to increase its scope and balance make it a desirable addition to any library of Medical Electronics.

ANTHONY S. VELATE

An Electronic Organ for the Home Constructor

By Alan Douglas. 100 pp. 91 figs. Demy 8vo. Pitman & Sons Ltd. 1958. Price 15s.

THIS 'how to make it' book contains simple and straightforward instructions for building an inexpensive musical instrument specially designed for amateur use in the home. The circuits are a revised edition of the organ described in this journal for August and September, 1955, with a new solo generator and only two types of lamination for the main generator coils.

A high standard of tonal fidelity is ensured by adopting free-phase oscillators giving a multiple waveform, and the construction is such that only the ordinary tools available to the amateur are required.

Included in the book is a list of all materials and parts, with approved sources of supply. There are 91 constructional drawings and three photographs, as well as a list of valve base connexions, drilling and tapping tables for wood and metal screws, etc.

Electronic Semiconductors

By E. Spenke. 402 pp. 80 figs. Demy 8vo. McGraw-Hill Book Company Inc., New York, Toronto, London. 1958. Price 85s. 6d.

THERE has been a long need for a sound and well produced book on electronic semiconductors. The German reader and those familiar with German have for some time enjoyed the privilege of being able to use the German edition of the above book which was produced in 1955. It is therefore very welcome that an English edition has now been produced. This work deals fully and adequately with the background to the physics of transistors and electronic semiconductor devices.

The first part consists of a most excellent survey of the subject systematically arranged. The energy band concept, positive hole and lattice imperfection are introduced in a well and clearly understandable manner without too much confusion by mathematics. The part added by the translators on junction capacitance nicely completes the original author's attractive work. Here one might perhaps add that the author could have gone into greater details, especially as far as the theory of devices is concerned. However, many review and other articles on this aspect are now available and it may be excusable that the author thought he might lengthen the book intolerably if he proceeded too far in this direction.

The second and somewhat larger part of the book deals with the fundamentals of semiconductor physics. Here rigorous mathematical methods are used to describe in a very lucid way the basic theory connected with the energy band model, Brillouin zones, Fermi levels, etc. All problems connected with lifetime, surface states, contact potentials, etc., are lucidly described in a clear and adequate manner.

The book is well produced and clearly printed. The diagrams are even clearer than in the German edition which already were very good. It is almost certain that this book will be usefully added to the few good books available to the student in semiconductor physics, and the engineer and scientist doing research and development in transistors and electronic semiconductor devices.

K. HOSELITZ

Special Relativity for Physicists

By G. Stephenson and C. W. Kilmister. 108 pp. Demy 8vo. Longmans, Green & Co. Ltd. 1958. Price 18s.

THIS book is an up-to-date account of the special theory of relativity. It is suitable for students reading for an honours degree course in physics or mathematics and contains a large number of applications of interest to the research physicist and electrical engineer, including high-energy particle accelerators, classical meson theory, particle decay schemes, the Dirac wave equation, relativistic hydrodynamics, de Broglie waves, the relativistic Child-Langmuir law and many other topics. A detailed account is also given of the historical development of the subject and the relevant experimental work.

Conductance Curve Design Manual

By K. A. Pullen. 128 pp. 73 figs. Demy 4to. John F. Rider Publisher, Inc., New York. 1958. Price $4·25.

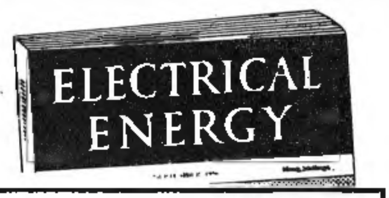
THIS manual explains the use of conductance curves in circuit designing and provides some 70 of the most representative thermionic valve conductance curves. It is divided into three main sections; a brief explanation of the special curves and their application in *RC* amplifier designs; a set of tables useful in making valve substitutions, and tables to simplify the selection of valves for given applications; and a special set of curves to facilitate valve circuit design.

An Introduction to the Theory and Practice of Semiconductors

By A. A. Shepherd. 206 pp. 93 figs. Demy 8vo. Constable & Co. Ltd. 1958. Price 18s. 6d.

THIS book has been written for advanced students and graduate workers in all fields of physics. The first chapters deal with the structure of solids, conduction in semiconductors, and the properties of single crystals. Rectification by semiconductors is dealt with next, transistors and light sensitive semiconductor devices. The technology of germanium and silicon device development is considered in some detail. The book concludes with a discussion of semi-conducting compounds.

The author is head of the semiconductor development laboratories of Ferranti Ltd.



ELECTRONIC EQUIPMENT

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

HIGH SPEED BATCH COUNTER

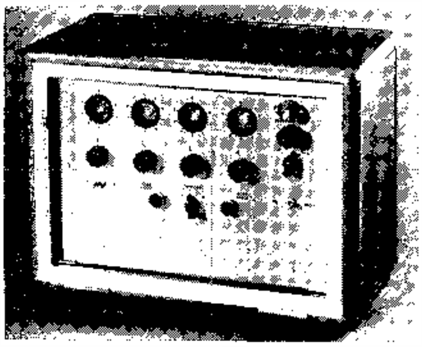
(Illustrated below)

Elcontrol Ltd, Wibury Way, Hitchin, Hertfordshire

THE Elcontrol high speed batch counter is of the fully electronic type making use of up to 4 Dekatron stages. The total count is displayed on the Dekatrons.

The complete standard unit embodies its own pulse-shaping circuits and will accept normal external pulses including those originating photo-electrically.

The unit can also be arranged to generate its own pulses and can, therefore, be used as a timer.



The standard unit can be used in a large number of ways including the following:—

As a batch counter

(1) Automatic-reset—Batch signal approx. 0·2sec.

(2) Reset on operation of external contacts.

(3) Reset on specified count of successive batch (works selected).

(4) Relay energized/de-energized alternate batches.

As a timer

100c/s signal provided within the unit. When used with above batching facilities the following timer functions are available:—

(1) Cyclic timer with 0·2sec fixed dwell.

(2) Process timer with external initiation.

(3) Cyclic timer with pre-fixed dwell period.

(4) Cyclic timer with equal times both 'on' and 'dwell'.

As a time measurer

(1) Time recorded between successive operation of external contacts. $\pm$ 10msec resolution. Mains frequency accuracy.

The maximum counting rate is 500 per second and the minimum recovery time between batches is 3msec.

The 4 decade units together with the photo amplifier circuit are assembled on individual chassis which can be separately unplugged from the complete rack. The power pack and trigger unit are independently assembled in a complete chassis housed on the main deck of the assembly. The complete rack embodying the electronic circuits can itself be readily withdrawn from the case.

The unit is housed in a sheet steel case assembled on a steel framework. A viewing window covers the Dekatrons and control knobs and can readily be opened to provide access to them. The window is itself mounted in the front panel which can be easily removed, thereby enabling the rack assembly to be withdrawn. The steel case is finished in grey stove enamel.

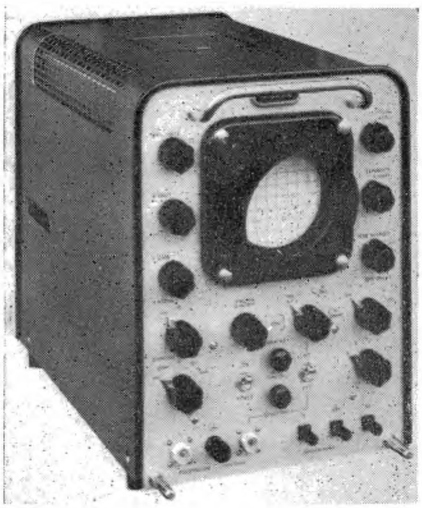
EE 1751 *for further details*

GENERAL PURPOSE OSCILLOSCOPE

(Illustrated below)

The Solartron Electronic Group Ltd, Solartron Works, Thames Ditton, Surrey

AN inexpensive and easily portable general-purpose oscilloscope, the CD 814 has many features normally found only in more specialized instruments. With a constant bandwidth (3dB) of 0·9c/s to 9Mc/s or to 12Mc/s (6dB), and a maximum Y amplifier sensitivity of better than 40mV/cm, this new Solarscope is equally useful for radio or television servicing or use in educational and technical training establishments. A brilliant, clear trace is assured at all frequencies by use of a type 4EP1 p.d.a. tube operating at 2·3kV. The rise-time of the Y amplifier is better than 40mμsec—without overshoot—and at 50c/s there is no visible droop on a square-wave display. Velocity calibration is provided by use of time-base marker pips, accurate to 5 per cent, in four ranges from 0·1msec separation; also Y input measurement by injection of 50c/s square-waves at amplitudes of 100mV, 1V 10V and 100V. The flexible time-base, which has facilities for internal or external synchronization or triggering has basic sweep of 0·5μsec/cm to 10msec/cm with symmetrical × 10 expansion. Television signal sync separation input facilities enable the time-base to be locked on a 'line' or 'frame' when presented with a video signal.



The CD 814 is provided with attachments for fitting standard oscilloscope cameras; has an illuminated graticule with variable intensity edge-lighting—and an adjustable Solartron viewpoise stand.

EE 1752 *for further details*

UNSPILLABLE BATTERIES

(Illustrated below)

Varley Dry Accumulators Ltd, By-Pass Road, Barking, Essex

THE Varley method of construction whereby the active material in the positive and negative plates is fully supported and maintained under compression by the separators provides a number

of advantages over conventional methods of construction. Included in these are: high power to weight ratio, increased efficiency at high rates of discharge, more resistance to vibration and corrosion and no free electrolyte to spill.

These batteries are available in a variety of types to suit most applications. Typical of the range are the types VPT 12-19/25 and the VPT 2-24/11/4 (illustrated). The former is a 12V 25Ah battery capable of discharges up to 1 000A for short periods. The latter is a 24V 4Ah battery intended for laboratory use. It is tapped every 2V and can be used to provide any voltage up to 24V or as individual 2V cells. The maximum discharge rate is 10A. Both batteries are supplied in polystyrene containers.

EE 1753 *for further details*

SOLDERING TOOL

(Illustrated above right)

Allied Distributing Corporation Ltd, 13-17 Rathbone Street, London, W.1

THE Allied Distributing Corporation announce that they are now marketing a new soldering tool known as the 'Victor V360'.

The Victor is based on the principle of heating the work direct instead of the conventional 'iron' method of transferring a small amount of heat from a preheated mass of metal.

Heating the work direct is achieved by gripping it with the twin arms of the tool. At the end of the arms, which act as leads for the electric current, are special carbon electrodes which convey the heat directly to the work.

To solder, the joint is held by the twin arms; the button pressed, the work immediately heats up and the solder runs. The button is then released thereby breaking the heating circuit, and the joint held for a few seconds for the solder to harden.

Adjustable stems and electrode holders, together with varying shaped electrodes, make possible all types of soldering with a single tool. The earth return lead permits the use of one arm only, if required.

The Victor will operate equally well with a.c. or d.c. supply and although 6V 5A is the normal power, it can operate on any voltage from 2 to 8 depending on the size of the job. For example, when soldering very fine work (30 s.w.g. and smaller) 2V supply would be sufficient for fast, efficient operation. The Victor will solder solid copper, brass or other solderable material from 1/64in to 3/16in diameter or equivalent surfaces.

For heavier work a model known as the 'Venom' is available which has a nominal rating of 6V 30A.

EE 1754 *for further details*

HIGH VOLTAGE FLASH TESTER

(Illustrated above right)

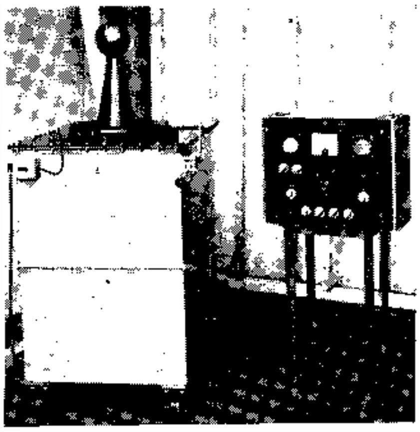
Airmec Ltd, High Wycombe, Buckinghamshire

THE 100kV flash tester type P.200 has been designed for a.c. and d.c. testing of insulating materials and component dielectrics at very high voltages. It not only provides a heavy current for breakdown but also incorporates a sensitive indication of ionization and leakage currents.

A continuously variable output voltage of from 0 to 140kV peak a.c. at 50c/s or 0 to 140kV d.c. is obtained from an e.h.t. unit which is housed in a steel, oil filled case mounted on wheels. A control unit, installed at any distance up to 36ft away and electrically isolated from the e.h.t. source, provides a remote control facility which can be operated with absolute safety.

The control unit contains a meter with an associated switch for the measurement of a.c. or d.c. output voltages and leakage current. The leakage current may be measured in two ranges 0 to 500μA and 0 to 5mA, the lower range being selected by a push-button switch.

A 2¾in cathode-ray tube is also incorporated on which either the leakage current or applied voltage waveforms may be presented, or an elliptical trace provided for the display of ionization currents. In addition, the presence of ionization currents is audibly indicated by a loudspeaker.



Breakdown or flashover is indicated by a glowing neon tube and a thermal overload switch in the control unit is operated if the test piece maintains full load for any appreciable time, but will re-set after a short period when normal operations can be resumed.

Since both c.r.t. and loudspeaker indication of ionization is provided, the voltage at which ionization starts is at once evident and tests may be discontinued before breakdown of the test piece occurs. Furthermore, by virtue of the loudspeaker facility it is frequently possible for an experienced operator to interpret specific faults in some insulating media.

The equipment operates from a 200 to 250V 50c/s a.c. power supply and has a total consumption of 1kVA on full load.

EE 1755 *for further details*

UNSPILLABLE LIGHTWEIGHT CELL

(Illustrated below)

Chloride Batteries Ltd, Clifton Junction, Manchester

A COMPLETELY unspillable Exide 2V 8Ah cell, type MFB9, has been produced by Chloride Batteries for portable light electrical duties. Filled and charged, this compact cell weighs only 1lb 5oz and is capable of providing high discharge currents, with high terminal voltage over a wide range of temperatures.

The container and lid of the cell are moulded in translucent high impact polystyrene, and the lid is a specially designed

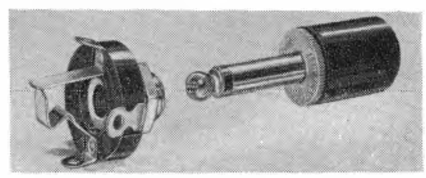
one-piece moulding cemented to the container by a technique which ensures a leak-proof joint.

A vent which provides an exit for gases, and which incorporates a trap for acid spray, is moulded integrally with the lid. The vent outlet is so designed that it is virtually impossible for it to become blocked either by dirt or some object resting on top of the cell.

A filler plug, with a broad slot for tightening with a coin, is located alongside the gas vent. This arrangement, keeping the acid trap and gas vent separate from the filler plug, enables a more effective type of trap to be used.

The cell terminals consist of metal inserts moulded in the lid into which are screwed small bolts with insulated heads. When these bolts are tightened the cable ends are clamped firmly against lead alloy strips. These strips are positioned in recesses in the cell lid and are connected by lead burning to the connecting pillars which rise from the group bar. Special rubber bushes are fitted round the pillars where they pass through the lid, and these bushes ensure a leak-proof joint at the cell lid.

EE 1756 *for further details*



MINIATURE JACK AND PLUG

(Illustrated above)

A. F. Bulgin & Co. Ltd, By-Pass Road, Barking, Essex

THE latest addition to the Bulgin range consists of a miniature jack (J.30) and jack plug (P.519). The plug has axial cable exit screw-on phenolic cover and plated members. The internal solder tags being silver plated for easy connexion. The jack has phenolic insulation, nickel silver contacts and tags with one pole live to fixing bush and a third contact (mated with tip contact when plug removed). It is rated at 1A 50V.

EE 1757 *for further details*

HIGH TEMPERATURE STRAIN GAUGES

Farnell Instruments Ltd, Wetherby Industrial Estates, York Road, Wetherby, Yorkshire

THE AVRO range of strain gauges will operate in temperatures up to 750°C.

The non-removable backings to the gauges are of two types. One is a resin-impregnated tissue, the other fibreglass cloth impregnated with a ceramic cement. The temperature range of operation has thus been extended to 750°C.

Three ranges are available:—

Up to 200°C—Epoxy resin base (Type 951)
250°C—Resin base (Type BF3)
750°C—Ceramic base (Type U527) (Type U529)

The Type BF3 gauge can be supplied for use on materials which could not be given the high temperature cure required by the ceramic cements.

These gauges are robust and flexible, even the ceramic gauges being capable of bending round small radii with care.

Measurement of strain at temperature requires an accurate knowledge of the variation of gauge factor with temperature, lead length, compensation device and gauge form. For consistent response, regularity of form and close tolerance resistance values are required. While a foil gauge has many advantages in shape and negligible transverse sensitivity, it is difficult to produce small gauges with reasonably high sensitivity.

The AVRO gauge, of 250Ω resistance, is manufactured on jigs ensuring control of form, and all the joints are spot welded. Gauges with differing lengths can be made to special order.

Resistance values other than 250Ω can be supplied, as can 45° or 60° rosettes (standard grid length 0·3in nominal resistance 100Ω).

Adequate bonding to the material under test is a prime essential of strain gauge work. The resin based gauges can be bonded on to almost any constructional material, while ceramic gauges are particularly suited to heat resistant steels, all stainless steels and Nimonic alloys. They can be applied to low alloy steels, cast iron, brass, titanium, copper and aluminium alloys, but the bond is not as effective.

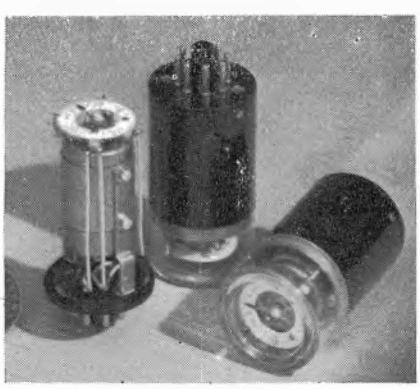
EE 1758 *for further details*

PRECISION MEASURING RELAYS

(Illustrated below)

Leland Instruments Ltd, 22-23 Millbank, Westminster, London, S.W.1

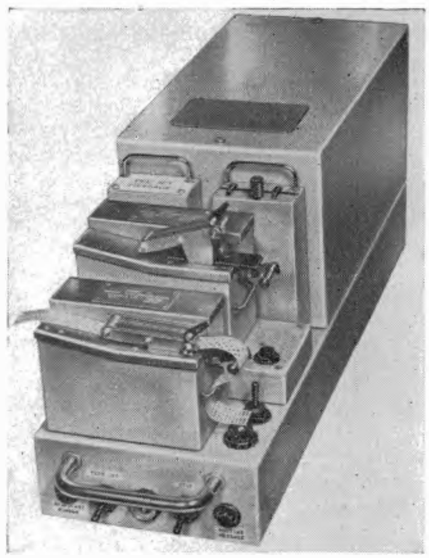
DESIGNED and developed by Brion Leroux & Cie of Paris, these relays are based upon their patented spring suspension moving coil unit. For the past seven years they have been used in increasing quantities by French industry and government departments, particularly by the French Post Office. They are now available in the United Kingdom and British Commonwealth and Empire from Leland Instruments Limited.



The relay consists of a galvanometer movement, complete with scale and pointer, to which has been added a moving contact. On each side of this are arranged two fully adjustable contacts which are instantly accessible beneath a transparent plastic dust cover, and are adjustable over the full range of the scale.

These relays are fully shielded and are available in 16 different models, with moving- coil resistances ranging from 2·5Ω to 7·5kΩ, and with full-scale deflexion current ratings ranging from 10mA down to 0·02mA. In the case of the most sensitive model the minimum change of control power necessary to make or break contact is of the order of 0·05μW (five-hundredths of a microwatt). The relay contacts have a power handling capacity of 200mW. Each of the 16 models is available either as a maximum-minimum indicator with scale covering 90° and calibrated 0 to 100 divisions, or in centre-zero form with scale calibrated 50-0-50 divisions.

EE 1759 *for further details*



NEW TAPE READING AND AUTO-NUMBERING APPARATUS

(Illustrated above)

Automatic Telephone & Electric Co. Ltd, Strowger House, 8 Arundel Street, London, W.C.2.

TELEGRAPH traffic by modern high-speed punched tape systems has led to the necessity for the numbering of messages. Ideally the numbering should be done automatically, and standard or 'routine' traffic should also be available without punching special tapes.

British Telecommunications Research Ltd have developed for Automatic Telephone and Electric Co Ltd, a self-contained piece of equipment with built-in facilities which are claimed to meet all the necessary operational requirements. This apparatus, known as the 'tape reading, auto-sending equipment' series TAA, comprises two main units. One section incorporates the tape reading heads and all the operator's functional controls; and the other, which can be remotely mounted on a standard 19in rack, is the associated electronic distributor.

The complete operator's terminal is housed in a unit measuring only 22in × 6½in × 9in and weighing 31lb. Two operating heads working on a unique step-by-step principal are provided. Both heads can be loaded simultaneously and are sequentially switched so that as soon as one tape has been read the other head comes into operation.

In addition to the two heads the terminal includes an automatic numbering device, a 100-character test message sender and two 40-character message senders. The first of the 40-character senders is used for the station call sign and the other provides for the automatic transmission of routine traffic. Both the call sign and the routine traffic messages are readily changeable to meet operating conditions. An automatic timing device permits the terminal to transmit test messages at predetermined times so as to test the circuit and associated equipment.

The electronic distributor, size 19in × 8¾in × 3½in, weight 23½lb, translates the information read by the tape reading heads, or message senders, into telegraph code.

The manufacturers claim many advantages for this very original equipment. It occupies less operator's space than conventional tape heads, provides complete flexibility under all operating conditions and permits maximum loading of several circuits by one operator.

EE 1760 *for further details*

'MOTOLOAD' RELAY

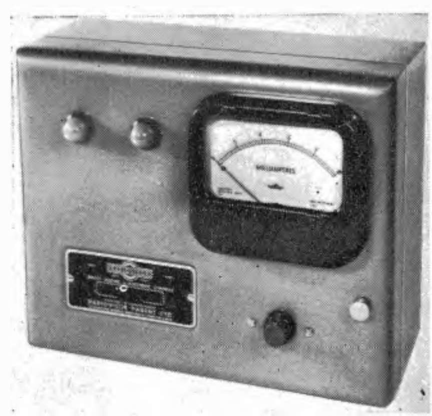
(Illustrated below)

Radiovisor Parent Ltd, Stanhope Works, High Path, London, S.W.19

THE Radiovisor electronic 'Motoload' relay unit is a versatile equipment applicable to any control function which may be initiated from a current change in the feed to an electric motor, caused by varying loads or other conditions.

The equipment is designed for use in conjunction with a current transformer, this latter having a standard output arranged to operate the electronic circuit of the control unit.

Alternative equipments are available with and without indicating instruments and auxiliary circuits are arranged to accept latching contacts for those applications where manual resetting is desirable.



The 'Motoload' is connected in series with the driving motor to be monitored and under steady load conditions the relay stays in the 'off' position. An increase in load causes an increase in motor current and the relay operates immediately to give warning of the new condition or to perform the desired switching operation.

EE 1761 *for further details*