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Commentary

THE 'end of year' statement issued by the Radio Communication and Electronic Engineering Association confirms that 1955 was a year of prosperity for the industry. The products of the members of the RCEEA, which includes communication equipment of all kinds, radio and television transmitters and ancillary equipment, radar, instruments and industrial electronic equipment, is estimated to account for approximately one third of the total output of the radio industry as a whole which is now believed to exceed £220m a year. These figures can be estimated only, of course, since, apart from other considerations, there is a large number of firms which do not belong to any of the manufacturers associations, while there are others which, although manufacturing considerable quantities of electronic equipment, are not, strictly, members of the industry. The figures err, no doubt, on the side of conservatism.

These figures include defence equipment which at the present time represents a considerable volume of production. For example, the Air Estimates for 1955/6 include £23m for electronic equipment and, while the Navy and Army spend somewhat less, the electronic equipment in a modern aircraft carrier is estimated to be worth around £1m.

The value of exports of electronic equipment during 1955 was some £13m which is the highest so far achieved. This does not include equipment which forms a part of larger goods as, for instance, equipment which was fitted into ships or aircraft, etc.

Television transmitting equipment has continued to provide a considerable volume of trade and one of the most notable achievements of the year was the designing and manufacturing of the equipment for the first television station in the ITA's commercial network in a time of less than seven months. The BBC has also purchased a considerable amount of new equipment including 17 cameras and two pairs of transmitters for its proposed alternative programme service. In addition, the overseas market for television and microwave link equipment is still thriving and when television starts in Australia most of the equipment at her first four stations will be British, the value of this equipment being approximately £½m. The market for 'entertainment' equipment has been further boosted by the rising popularity of f.m. broadcasting, both in this country and abroad; the BBC have now ordered no less than 46 v.h.f., f.m. transmitters.

In the communications field the highlight of the year was the installation, at the G.P.O. station at Rugby, of 28 new high power transmitters which makes this one of the largest and most up to date transmitting stations in the world. Activity in this field also produced many well worthwhile export orders, both for civil and military equipment; an order for a single type of packset from the Indian Government being worth nearly a quarter of a million pounds.

Some of the largest export orders received during the past year were for aeronautical and marine radar equipment, a sphere in which Britain is pre-eminent. The U.S.A Navy alone has ordered, during the past two years, some $30m of radar equipment from two British firms. It is estimated that more than 5 000 ships are now fitted with British made radar and last year one company alone received orders for radar equipping 1 221 vessels from 43 different countries.

Industrial electronics is, so to speak, the 'baby' of the radio industry, and a very hearty one it is proving to be, for industrial electronic equipment is now becoming an increasingly important part of the output of the industry. The initial prejudice against electronic equipment is fast being dispelled and electronic measuring and control apparatus is now an accepted part of many industries—from steel rolling mills to cigarette making factories.

The picture as presented by the RCEEA statement is indeed a pleasing and encouraging one and it is a pity that it should have to end with the now monotonous cry for more trained labour. It is calculated that the whole radio industry needs approximately 1 000 recruits a year in the professional engineering class and some 10 times this number of other technicians. The radio industry is, of course, well advanced in providing its own training schemes and young persons entering the industry can certainly be assured of a secure and profitable future, but the competition for manpower with other industries is keen.

A Silicon Junction Diode Modulator

For use in Junction Transistor Direct-Current Amplifiers

By N. F. Moody*

Although d.c. amplification is an essential function in many circuits, the designer using transistors must use d.c./a.c. modulation to compete with the high impedance and extreme sensitivity yielded by vacuum tubes.

Both mechanical modulators and the usual semi-conductor bridges show serious shortcomings; the former is restrictive in bandwidth and is bulky; the latter yields inadequate zero stability and sensitivity.

By exploitation of certain features unique to the silicon junction diode these limitations may be overcome and, indeed, figures surpassing vacuum tube performance are often obtainable.

This article commences with a study of silicon diode properties, which are evaluated in relation to a selected modulator circuit, whose design is then synthesized to optimize performance. A sub-miniature unit results which weighs ½ ounce and whose zero stability is in the order of 10^{-10} watts.

A wide range of input impedance is available so that, on the one hand, a unit of 300,000 ohms input impedance is sensitive to currents of 10^{-8} amperes or even less; while on the other, modulators of a few thousand ohms are suitable for a few millivolts of input signal. The performance is maintained, without serious degradation, to at least 80 degrees centigrade.

A key feature of the circuit is its freedom from critical conditions. No balance controls are needed. No selection of rectifiers is required. Most designs tolerate tenfold variations of reference level.

While the major study is devoted to modulators whose reference frequency lies in the audio range, it is shown that the performance is maintained in the carrier and lower radio frequency ranges.

MODERN vacuum tube circuits gain important flexibility by their free use of low-level d.c. amplification. Time delay circuits such as the Miller integrator, the adding and other operational functions of an analogue computer, or the input stage of a d.c. operated servo, are typical of the d.c. stages required. Such circuits demand an active device of high input impedance capable of responding to voltages of 10^{-1}V or less at currents as low as 10^{-7}A.

The designer using transistors finds them unsuitable for such purposes, for the input impedance is low, and all accessible terminals bear currents which swamp the figure quoted. The currents, furthermore, are often variable and always temperature dependent.

A solution must be sought by use of a modulator which can convert the d.c. signal into a.c. and allow impedance matching prior to amplification in the transistors. While mechanical modulators may be used, they impose severe bandwidth restrictions, are often large compared to the associated transistor amplifier, and are unsuitable for operating environments which impose severe shock and vibration.

Electronic modulators allow adequate frequency response, are small, and are resistant to shock. Unfortunately, units based on thermionic valves, or the earlier semi-conductor devices, have failed to yield adequate sensitivity and freedom from spurious signals—even when components were matched by selection.

The most stable modulator element has always been the simplest and hence least variable: the diode. The most stable circuit, in turn, relies on a balance between elements, such as exists in the ring bridge of Fig. 1(a). This circuit will be the basis of the discussions which follow.

The unique properties of silicon diodes set new standards of bridge performance, so that the target figures quoted are readily exceeded without resort to selection and over a wide temperature range. The promised advantages are achieved only by correct operating conditions, which differ so radically from orthodox practice that Fig. 1(a) should be regarded as a new circuit. In order that its design may be synthesized the circuit must first be studied to determine the parameters which control sensitivity and stability. An investigation of diode characteristics in the light of these parameters then allows optimum conditions to be determined.

Simple Bridge Theories With Perfect Rectifiers

Consider the circuit of Fig. 1(a) in which R_s is the signal source impedance, and R_L' the value of the load resistance R_L as reflected to one-half of T_2 primary. Transformer T_2 evidently acts as a matching device between loads; and transformer T_1 as a means of injecting the reference signal. The whole structure exhibits symmetry, and hence balance with respect to input, reference, and output. In practice it is convenient to use an unbalanced input signal and terminal v will be grounded.

The introductory part of the discussion may be based on certain simplifying assumptions. In particular, it is helpful to idealize the silicon diodes, whose characteristic plotted to a linear voltage/current scale is shown in Fig. 1(b) curve (i). To the degree of resolution shown in the figure, the characteristics differ strikingly from earlier semi-conductor rectifiers: for at forward voltages below an approximate threshold E_B of some 0·55V, and under reverse excitation, no conduction is apparent. The rapid onset of current above the threshold suggests that a preliminary picture of diode action may be based on the simple broken line characteristic of curve (ii) in the figure; which represents a perfect voltage sensitive switch of zero forward

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impedance above voltage E_B and of complete open-circuit below this threshold. The ring of rectifiers may thus be represented by the biased diodes of Fig. 1(c) and the bridge operation will be introduced on this basis. A key feature of this concept is that the four paths WX, XY, YZ and ZW are open-circuit at all times when the excitation voltages are less than the value of E_B. Unlike conventional modulators, the paths from input signal (via T_1) to output (via T_2) exist only when the appropriate driving voltages are applied.

In the initial work the thresholds E_B will be assumed equal; the diodes free from capacitance and carrier storage effects; and the transformers lossless. It will be convenient to base the preliminary analysis on a reference signal of 50/50 square wave shape, whose injected power is constant and is far greater than that of the signal to be modulated. Then the rectifiers are set to their alternate states solely at the control of the reference signal. Consideration will be given later to the modifications required for other wave shapes and power ratios.

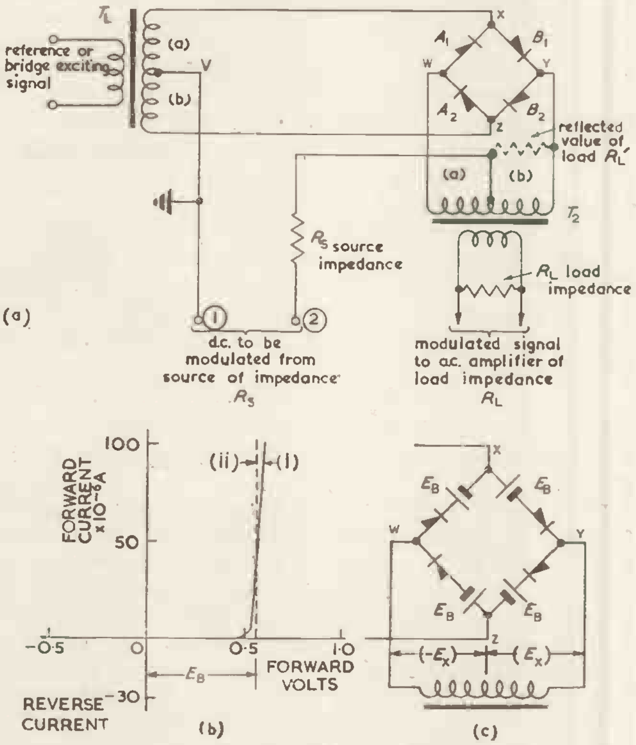

Fig. 1 (a) Ring bridge modulator.
(b) Forward and reverse current curve at a silicon diode (i), together with idealized curve (ii)
(c) Introduction of the thresholds E_B.

If a square wave reference generates a voltage of at least $\pm 2E_B$ across XZ, the paths XYZ and XWZ will alternately become conductive and the bridge then acts as a simple commutator, free from all errors.

The mechanism is as follows: A current injected into terminal (2) may accept only one of the paths (a) or (b) in the transformer T_2, for dependent on the reference phase only one of the paths XYZ or XWZ is conductive. Whichever path is open at that instant, the signal current will pass to the terminals X and Z as two equal components. These components, in returning to the terminal (1), will pass through the windings (a)(b) of transformer T_1 in inductive opposition. Thus the signal has been commutated with respect to the transformer T_2, but not with respect to transformer T_1, and no induction occurs in the latter.

Bridge Limiting Due to the Threshold E_B

Since the diodes are assumed to be perfect, and to have equal threshold voltages E_B, the centre point of a conducting rectifier pair will be at the same potential as V (assumed zero). Suppose, during one half cycle of reference, that Y is brought to earth potential. Then a signal current I_{dc} will set up an instantaneous voltage $E_x = I_{dc}R_L'$ across winding (b) of T_2 (see Fig. 1(c)) thus causing point W to be depressed $2E_x$ volts. In order that conduction shall be restricted to one pair of rectifiers the voltage appearing across rectifier A_2 must not exceed E_B, that is $(2E_x - E_B) \leqslant E_B$. For modulators handling the low powers discussed in this article the relationship is normally satisfied: if it is not, a form of saturation sets in and for this region the simple theory outlined above is not met.

Rectifier Imperfections and the Resultant Displacement of Null Balance

Practical rectifiers exhibit defects which prevent perfect commutation. As a result, spurious voltages and currents are injected into the bridge, and these disturbances cannot be distinguished from the signal to be modulated. It follows that the linear relationship between input and output is displaced and the null no longer occurs at zero input signal. The analysis which follows will calculate the input signal required to restore the null to the origin.

Forward Voltage Drop

When a rectifier conducts a voltage drop appears across it. The thresholds E_B can be regarded as such a drop and, if unequal, can introduce an error term in the bridge. Writing the forward voltage drops of the four rectifiers as E_{A_1}, E_{A_2}, E_{B_1}, E_{B_2}, it follows from the symmetry of the structure that if the voltage drop of each member of pair B, or of pair A, is equal, no error results. In the absence of such balance there is an error signal E_E between the points WY of peak magnitude:

$$E_E = 0{\cdot}25\,[\,|(E_{B_1} - E_{B_2})| \pm |(E_{A_1} - E_{A_2})|\,] \quad \ldots\ldots \quad (1)$$

where the alternate sign shows the change of magnitude in E_E which results from reversal of positions of the rectifiers in one limb.

Reverse Current

When the reference drives a rectifier to the non-conduction region there will remain a back resistance and thus a reverse current flows through it. Once more, symmetry indicates that equal net reverse currents in a pair A, or B, will cause a nulling of the error. Inequality will lead to an error current I_E between the points WY, whose peak magnitude is:

$$I_E = 0{\cdot}5\,[\,|(I_{B_1} - I_{B_2})| \pm |(I_{A_1} - I_{A_2})|\,] \quad \ldots\ldots \quad (2)$$

and where the alternate sign has the same significance as in equation (1).

Total Error Power

The apparent error, expressed as power required at the input to restore null output, is readily calculated for square wave excitation on an assumption which is met in practice: that the rectifier forward and reverse impedances are respectively low and high compared to the net signal path impedance $R_s + R_L'$. For optimum efficiency, i.e. 50 per cent power transfer, R_s and R_L' will be equal, and this condition is the one taken. The error power W_E is then approximated by summing the two components appropriately and is:

$$W_E \sim 0{\cdot}5\,E_E^2/R_s \pm 0{\cdot}75\,I_E^2 R_s \quad \ldots\ldots\ldots \quad (3)$$

Conclusion to Preliminary Discussion

The elementary treatment of this section has outlined modulator operation based upon simplifying assumptions, both for the circuit and the rectifiers. In the following section silicon diodes are studied in detail and comparison is also made with other types of rectifier. This work will show that the simplifying assumptions are not valid for all rectifiers, nor for any rectifier under all operating conditions. It will be seen, however, that the silicon diode when properly used results in a bridge which is adequately treated by the simple theory, and that the null errors are of the form postulated above.

Characteristics of Silicon Rectifiers and their Influence on Modulator Design

In this section a study of rectifier characteristics will be made, directed to a choice of operating conditions which renders the null balance insensitive both to the rectifiers and to the reference waveshape and amplitude. As an introduction to this study, it will be useful to expand the concepts on sources of null balance error.

The error due to differences in conduction threshold has been discussed. It is now necessary to consider the nature of such thresholds in greater detail. The forward voltage drop in a conducting rectifier may conveniently be treated as having two components. One of these in an internal source of e.m.f. and the other a forward resistance.

For a reference of constant amplitude and square waveshape, such as has hitherto been assumed, the distinction can be ignored; for balance of a rectifier pair merely requires coincidence of their characteristics at a single value of current. For a reference whose amplitude is variable, or which is a continuous function of time (in the manner of a sinusoid), the rectifier forward characteristics of a pair must be coincident over the whole range explored by the reference. This is satisfied if the internal e.m.f.'s and incremental forward resistances are equal.

A more detailed consideration of rectifier reverse characteristics suggests that, once again, two components should be included. These are a saturation (or non voltage dependent) reverse current and a reverse resistance.

By reasoning similar to that applied to the forward characteristic it may be seen that reverse curves coincident at a single point suffice for balance with constant square wave reference, whereas a variable reference demands a complete matching of curves. Such matching exists if the reverse saturation currents and incremental reverse resistances are equal.

According to the foregoing discussion the rectifier properties are completely characterized (for low frequency modulators) by four parameters,

(1) Incremental forward resistance
(2) Incremental reverse resistance
(3) Internal forward e.m.f.
(4) Saturation reverse current

and the study which follows will be based on this classification.

Evaluation of the Parameters

Incremental Resistances

For forward and reverse voltages, V, of a few tenths of a volt, the current I of a silicon diode follows a law which is sufficiently approximated by

$$I = I_s\ [\exp\ (V/CKT) - 1] \quad \text{.............} (4)$$

where KT is the energy equivalence of temperature; C is a constant for the particular rectifier, having a value around 1·5; and I_s is the theoretical reverse saturation current.

Since at $V=0$, $I=0$, it follows trivially that the rectifier produces no e.m.f. in a non-excited state. It may be noted that this property is true for all semi-conductor rectifiers, but that it does not hold for the thermionic diode.

The incremental resistances may be obtained from equation (4) by differentiation as follows:

$$dI/dV = I_s/CKT\ \exp(V/CKT)$$

but from equation (4):

$$I_s\ \exp(V/CKT) = I + I_s$$

Hence:

$$dI/dV = (I + I_s)/CKT$$

and

$$R = dV/dI = CKT/(I + I_s) \quad \text{..........} (5)$$

The reverse resistance may be calculated for the typical case $I_s = 10^{-9}$A; $C = 1{\cdot}5$; $KT = 40^{-1}$V. Then at zero voltage the incremental resistance $R = 4 \times 10^7\Omega$ and R approaches infinity when $I + I_s \rightarrow 0$, at a few tenths of a volt reverse potential. Evidently for all practical purposes the reverse incremental resistance may be treated as infinite, so that one source of bridge error is completely eliminated.

In the forward direction a conducting rectifier will be effective only when its impedance is low compared to the sum of source and load impedances. In a high impedance modulator the sum of these loads may well be 100kΩ and the rectifier impedance falls to this figure at $I \sim 4 \times 10^{-7}$A, when the forward voltage is some 0·4V*. Thus interest centres on the forward characteristic only at currents in the order of 10^{-6}A and above, where equation (5) simplifies to:

$$R_F = C/40I_F \quad \text{..............} (5(a))$$

The subscripts have been added to signify forward resistance and current; I_F is in microamperes and R_F in megohms.

From equation (5(a)) it is seen that rectifier incremental forward impedance is inversely proportional to forward current, and so by control of reference current injected into the system the rectifier impedances can be similarly controlled. It follows, furthermore, that since the rectifiers in the A or B limbs are in series, and are thus traversed by identical currents (signal source zero), the incremental resistances will be perfectly balanced for all values of reference current, so long as the constant C is identical for the pair.

The study performed in this article has been carried out using the Texas Instrument 601 silicon diode, and typical forward characteristics, together with values of the constant C, are shown in Fig. 2. A few units, some 5 per cent of those tested, showed a grossly inferior forward or reverse characteristic and these were rejected.

For currents above 10^{-4}A, equation (4) and its derivatives do not hold, because of the onset of spreading resistance†. It is essential that the excitation current shall not exceed this figure, for thereafter the R_F values become highly variable and the bridge balance suffers gross degeneration. As Fig. 2 shows, the bridge excitation current is extremely voltage sensitive, for conduction starts at 0·35V (10^{-7}A) and the linear region ends at 0·6V. For this reason the reference signal should never be generated by a low impedance voltage source. Instead, transformer T_1 should have its primary excited from a constant current generator. The

This forward voltage may be regarded as the threshold E_B of Fig. 1(c), and thus has a value of approximately 0·4V. The use of conducting threshold is abandoned in this section when considering bridge balance, for it is now replaced by the two components of incremental forward resistance and an internal e.m.f. However, the concept of a threshold is still of sufficient accuracy for the determination of bridge limiting described previously, and it is retained for this sole purpose.

† *Since this article was written a number of new silicon diodes have appeared on the market. Reverse currents as low as 10^{-10}A. are obtainable, and the sensitivities quoted are accordingly raised by one order—particularly does this apply at high temperatures.*

excitation may then vary as much as 50:1 for square wave reference, and some 15:1 for sinusoidal reference, with little deterioration of low level performance**.

Internal Forward E.M.F.

The spread curves of Fig. 2 allow the curves of individual diodes to be displaced with respect to each other on the voltage axis, apart from any differences due to variability in the constant C. While units may be matched in pairs, by measurement of forward voltage at two currents, this procedure would be necessary only to obtain the ultimate in performance. No balancing or selection is envisaged for the modulators herein described. Such unbalance errors of voltage will be seen to be but a few millivolts and represent the first term of equation (3). A typical magnitude is $E_E = 5$mV, and when the values of R_s and R_L' are each $10^5\Omega$ this results in an apparent input d.c. error which, expressed as a power, is about 10^{-10}W.

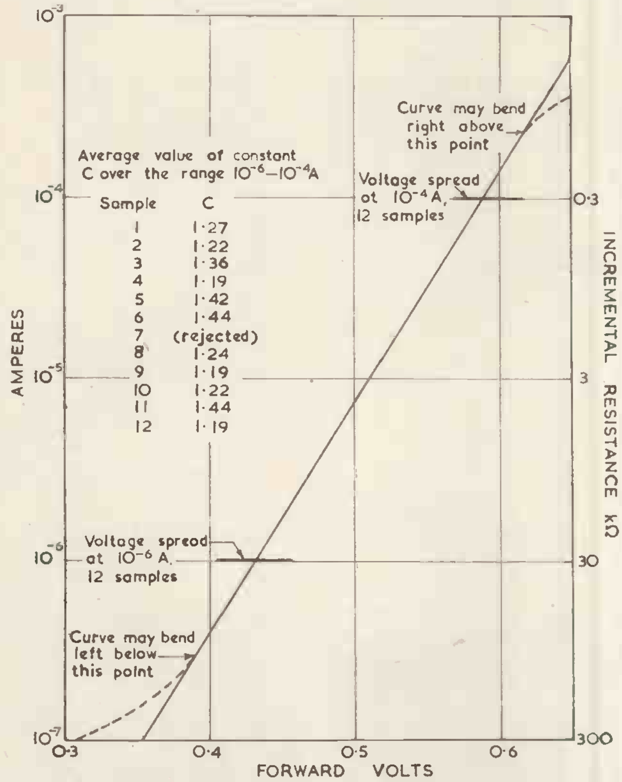

Fig. 2. Typical forward characteristics of texas 601 silicon diode together with spread voltage and slope variations.

Saturation Reverse Current

Typical reverse currents at room temperature, Fig. 3, show that this component does not exceed $1{\cdot}5 \times 10^{-8}$A under operating conditions. Assuming an unbalance I_E of 5×10^{-9}A the error power given by the last term of equation (3) is 2×10^{-12}W: a negligible addition to that due to the voltage error.

Summary of Silicon Diode Properties and Comparison with Other Rectifiers

From the foregoing discussion it has been seen that bridge balance depends on the rectifying elements obeying equation (4), with matched values of the constant C and negligibly small values for the constant I_s. In addition, it has been found necessary to modify the forward voltage V in equation (4) to include a small error term, and this correction should be small and constant in rectifiers chosen for modulators.

The silicon unit fulfils these conditions almost ideally, yielding forward and reverse resistances which may be treated as completely matched. There remain only the two error sources of saturation reverse current (typical value 10^{-8}A) and the residual correction to V, (typical value 5mV). The influence of these residuals is often small enough to neglect, but it may be calculated when required by means of the error formulæ (1), (2) and (3). Examples of such calculation already given show that zero errors are typically less than 10^{-10}W.

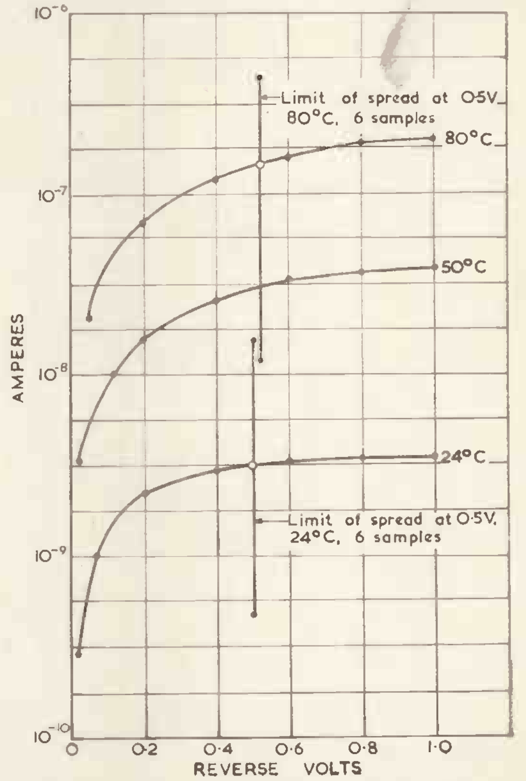

Fig. 3. Typical characteristics of texas 601 silicon diode showing temperature dependance.

It is interesting to compare this performance with that given by other types of rectifier. The vacuum diode yields an excellent semi-logarithmic relationship between forward voltage and current and, furthermore, may be designed to show reverse currents even smaller than the silicon diode. The performance is marred by contact and thermal emission velocity potentials which may introduce changes of a tenth of a volt or more in V during life. In transistor circuits the need for heater power, with the attendant problems of heater-cathode insulation and hum pick-up, makes the device an undesirable ancillary.

The older multi-crystalline rectifiers, such as the copper oxide and selenium units, do not follow any accurate theoretical law. Thus considerable resistive unbalance errors occur and reverse saturatio currents are usually many orders greater than those shown by silicon junctions.

The germanium point contact or junction diode can, in certain instances, meet equation (4) with an accuracy equal to the silicon diode. These units are unsuitable for modulating small currents, however, owing to the magnitude of the term I_s. A simple substitution of germanium

** *Other diodes extend the logarithmic characteristic to 10^{-3} A. or more, and thus the upper limit is also extended. However, the overload point may depend on the reference current and this, and the problems of sinusoidal excitation will be treated later.*

for silicon in a given structure may well raise I_s by three orders.

Modulators fed from Signal Source of Infinite Impedance

In circuits such as Miller integrators the source impedance may be several megohms, and a match between source and load prove difficult. The input terminals of the modulator then look towards an impedance, much larger than that presented by the modulator, and thus the d.c. signal appears as a constant current source. In this event the first term of equation (3) vanishes, and the concept of error power must be replaced by the concept of an error current. This error current is simply the I_E of equation (2) and may be considered to be injected at the bridge terminals (1) and (2) (Fig. 1) in parallel with the signal. Thus such modulators will operate, at room temperature, down to currents in the order 10^{-8} to 10^{-9}A.

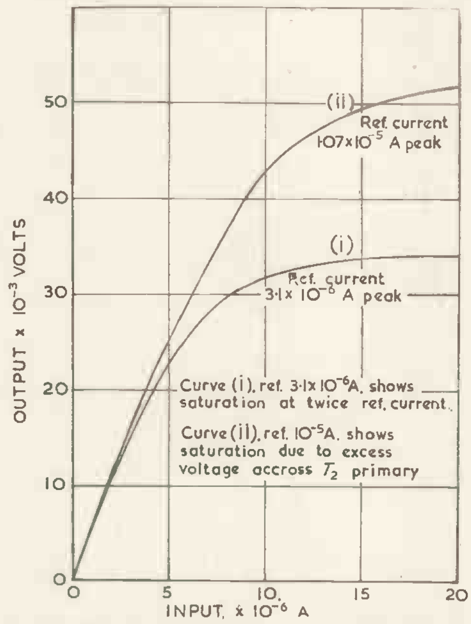

Fig. 4. Input/output curve of circuit shown in Fig. 5.

Curve taken at 25°C. Modulators operating at this scale of currents show negligible disturbance at 80°C. Curve plotted in quandrant 3 is identical to that shown.

The Influence of Reference Waveshape and Magnitude

The preceding analysis has been based on the fact that the signal current is small compared to the reference current. In fact inspection of the circuit shows that the modulator must saturate when the signal rises to twice the value of the peak reference current.* Thus for square wave excitation the bridge cannot handle signal currents in excess of a few hundred microamperes (based on the Texas 601).

In practice sinusoidal reference will probably be employed and this leads only to slight modification of operation. Efficient transfer of energy occurs only during those portions of the reference cycle when the diode impedance has a value equal to or less than the net load; the power transfer therefore occurs over an angle less than 360° per cycle. In order that the conducting angle shall approach the maximum, the peak reference current should be considerably greater than the signal current. A factor of 5 to 10 is suitable.

For 10^{-6}A signal current, 10^{-5}A excitation current, and $R_s = R_L' = 10^5\Omega$, it may be deduced from Fig. 2 that the conducting angle will be about 358°. The conversion efficiency approximates the ratio of actual conducting angle to its maximum value, and is found to be substantially constant for signal currents up to 0·3 of the excitation current. As the signal is increased beyond this point the efficiency gradually falls, and the input/output response becomes non-linear. Typical conversion curves are shown in Fig. 4 and apply to the modulator circuit of Fig. 5 (to be described). At a reference current of $3{\cdot}1 \times 10^{-6}$A peak, curve (i) exhibits saturation at a d.c. input equal to twice reference current, i.e. at some 6×10^{-6}A. When the reference is raised to 10^{-5}A curve (ii), saturation sets in by the voltage limiting mechanism at some 0·8V across the primary of T_2. The limiting feature of the bridge is valuable, and may be used to prevent saturation and blocking in amplifiers subsequent to the modulator. Since the peak reference current of modulators of this class should not exceed 1 to 3 × 10^{-4}A, the sensibly linear region of modulation will be limited to some ± 10^{-4}A for sine wave excitation.

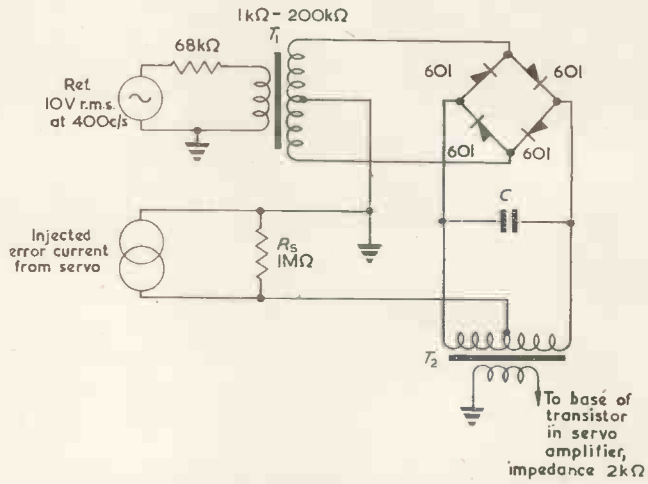

Fig. 5. Modulator for servo amplifier.

Transformer turns ratio, 1 : 14, primary to whole secondary.

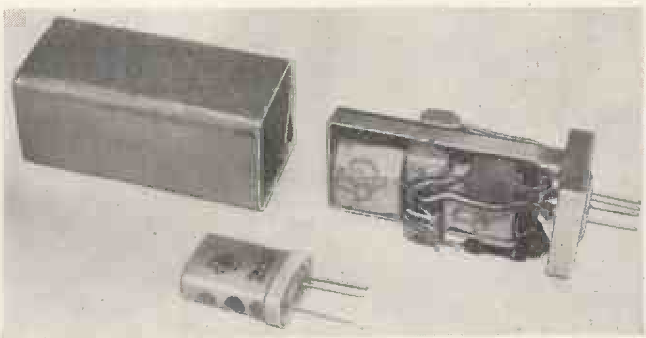

Fig. 6. The complete unit photographed against a transistor

An Application Example

It was required to make a subminiature modulator as the input stage of a d.c. error actuated servo amplifier. The complete unit, photographed against a transistor in Fig. 6, weighed ½oz and set severe limitations on transformer design. The circuit is shown in Fig. 5. Since the servo motor is a 400c/s unit, this frequency was used as a reference, enabling the modulator to be matched directly into the tuned transistor amplifier which drives the motor. The

* *As has been shown previously, saturation will occur below this level if the signal voltage appearing across T_2 primary reaches $2E_b$ (=0·8 V.).*

specification gave the source impedance as 1MΩ and physical size limits the impedance presented by the modulator to 70kΩ. Thus the modulator is essentially current operated and full torque is given at 4×10^{-7}A. The circuit may be regarded as a prototype of modulators fed from a constant current source.

The transformers are by Texas Instruments, type 201D, especially centre tapped for this purpose. In this design the transistor amplifier has 2kΩ input impedance which is reflected as $R_L' = 100$kΩ. Harmonics are minimized, and phase shift due to magnetizing current eliminated, by tuning the output transformer primary to 400c/s by means of $C(= \cdot 0017\mu F)$.* The dynamic impedance of the transformer then shunts R_L' with a 150kΩ loss resistance, so that some 67 per cent of the d.c. signal power appearing at the bridge terminals is converted to a.c. at the transistor base. The excitation current of the bridge, 10^{-5}A, is obtained from 10V r.m.s. via a series resistor of 68kΩ, which injects some 10^{-4}A into the primary of T_1. The secondary source impedance is sufficiently high to approximate the required

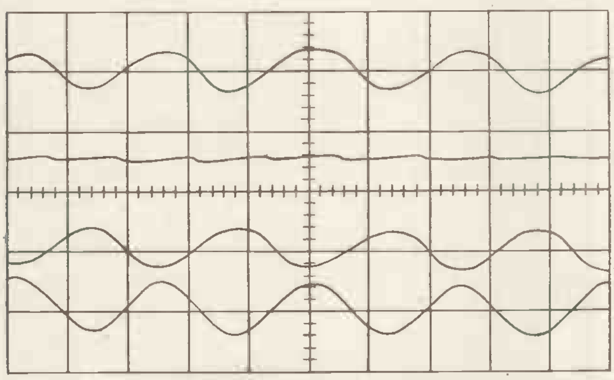

Fig. 7. Waveforms of the modulator.

constant current source: it is substantially ωL, about 200kΩ at 400c/s†. The output waveform of the modulator is shown in Fig. 7 where curves (A) and (C) represent the response to input currents of $\pm 10^{-6}$A, and (B) the null waveform at zero injected current. Fig. 7 (D) is the reference voltage as applied to the 68kΩ series resistor. The input/output responses shown in Fig. 4 have already received discussion in the preceding section.

Modulators for Use in the Carrier and Lower r.f. Ranges

It is reasonable to suppose that modulators based on the silicon diode can be used at reference frequencies up to several hundreds of kilocycles per second. Evidently diode shunt capacitance and carrier storage time must receive consideration in this region.

Although no serious study has been made on such high frequency applications, an experimental modulator was constructed to operate at 100kc/s. This unit, which differed from Fig. 5 only by the addition of a simple capacitive balancing circuit and substitution of suitable transformers, showed little deterioration from low frequency performance.

It may be useful to remark that modulators of this class are not restricted to the modulation of d.c. signals. Like the ring modulator, the device may be employed as a suppressed carrier modulator.

High Temperature Performance

Theory predicts that silicon diode modulators should continue to give excellent performance at high temperatures, and the author's measurements taken up to 90°C confirm this. Some degradation is to be expected, and this is predominantly due to increase in reverse saturation current with temperature.

The reverse saturation current I_s should be temperature dependent according to a law:

$$I_s' = I_s \exp [a(t_2 - t_1)] \quad \ldots\ldots\ldots\ldots \quad (6)$$

where I_s is the original saturation current at room temperature t_1, I_s' is the new value at higher temperature t_2, and a is a constant. It is quite common for surface effects in the junction units to cause values of I_s to exceed the theoretical values calculated from equation (4) and in such cases

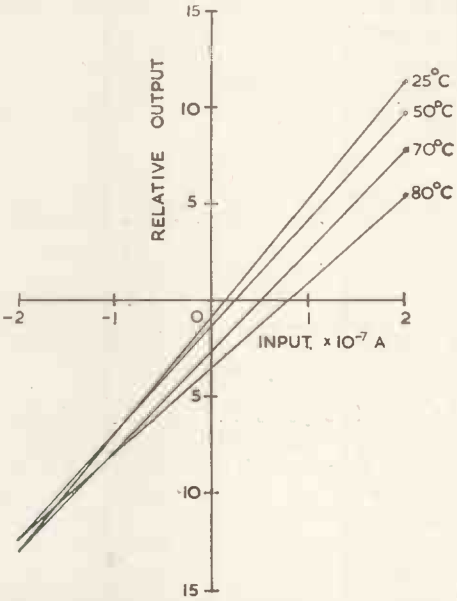

Fig. 8. Operation of the modulator at low currents.
Increased temperature causes a fall in conversion efficiency and a displacement of origin.

equation (6) is not well obeyed. Fig. 3, a plot of I_s' versus temperature, shows that reverse currents may rise to the range 5×10^{-8} to 5×10^{-7}A at 80°C. From such measurements it is deduced that values of $0{\cdot}045 < a < 0{\cdot}085$ give reasonable guidance on behaviour. The magnitudes of I_s' make it wise to avoid working at currents below 2×10^{-7}A at elevated temperatures unless a simple matching of I_s and a is permissible*.

In addition to the change in reverse current there will also be a shift in the forward characteristics at high temperature. For diodes having small surface effects the true value of a is given from equation (6) and by substitution of the right-hand member of equation (6) in equation (4) this shift can be calculated. Schaffner and Shea[1] have shown that the voltage for a given forward current is changed according to the law:

$$dV/dt = -(a/CKT)$$

and on the basis of $a = 0{\cdot}08$, $C = 1$, they find the voltage axis to be displaced by 2mV/°C. Tests by the author on the Texas units yield $2{\cdot}4 \pm 0{\cdot}2$mV/°C.

The calculation given above shows that the voltage/current curves at different temperatures remain sensibly parallel, and thus the forward resistance balance is unchanged. However, the displacement of the curves requires that a voltage term be introduced into the first member of equation (3). The peak magnitude should never exceed

* *To be selected according to primary inductance.*
† *The secondary inductance L is some 200H.*

* *The newer diodes should give improved performance. See earlier footnote.*

0·4mV/°C and it can always be reduced below this figure by reversing the positions of one pair of diodes. A typical unbalance from 25 to 80°C is found to yield a value of E_E = 5mV.

These predictions are confirmed by tests at 80°C on the modulator of Fig. 5. For the signal levels shown on Fig. 4 no disturbance is expected to be visible and indeed the curve appears unchanged. At the extremely low levels of Fig. 8, both a displacement of origin and change of conversion efficiency become apparent.

Conclusions

It has been shown that a silicon diode modulator, when operated under the correct conditions, can convert d.c. signals to an a.c. carrier over a range which extends from audio to at least 100kc/s.

A zero stability of 10^{-10}W or better is achieved at room temperatures, corresponding to currents in the order of 10^{-8}A for units fed from a constant current source, or voltages of a few millivolts for the zero source impedance counterpart.

Operation at 80°C and above is still acceptable, the performance being degraded by about two orders in terms of error input power, and one order each for the current and voltage fed forms.

Acknowledgment

In conclusion the author wishes to acknowledge his indebtedness to Mr. G. Thistle, of the Canadian Armament Research and Development Establishment, and to Mr. B. Jones of this laboratory, for their careful measurements on the diode characteristics; and to the Defence Research Board of Canada for their kind permission to publish. The work was undertaken in connexion with Project No. D48/55/66/01.

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A Low Noise Level H.T. Eliminator

By R. J. Luke*, B.Sc., A.M.I.E.E.

A simple stabilizer for h.t. supply for high-gain differential input amplifiers is described. The noise level is better than the equivalent of 1μV at the input when the bandwidth of the output is 0·5 to 75c/s.

FOR the past five years the Mark I Ediswan e.e.g. in many centres has consumed the power from two 120V h.t. batteries each month. Towards the end of their working life the regulation of the battery supply gradually deteriorates and it is this which usually necessitates replacement. Since the introduction by Mullard of the precision glow-discharge voltage regulator type 85A1 there was hope that a simple stabilizer using one of these tubes might be adequate to replace batteries. Preliminary experiments at the Burden Neurological Institute, Bristol by W. H. Shipton showed that the noise level using only the type 85A1 tube was too high. In 1951 a design similar to that shown in Fig. 1 was evolved and given preliminary tests. Since then a power pack and stabilizer which fitted into the existing space in the e.e.g. machine was constructed and was satisfactory in use for over a year, until the e.e.g. was replaced.

The eliminator (Fig. 1) utilizes a G.E.C. STV280/80 tube to provide an initially stabilized input to two type 85A1 tubes. This permits a 140V supply line to each of the 85A1 tubes so that approximately 55V are available across the series resistor. The output current fed to the preamplifiers remains almost constant at 7·5mA and the tube current is normally about 3·5mA. Under these conditions the noise level output of the amplifiers supplied by the stabilizer over a frequency band of 0·5 to 75c/s is lower than the equivalent of 1μV at the input (see Fig. 2).

The 85A1 tubes seem to be quite interchangeable and the six which were available were used in the circuit without any adjustment being required. This is not so, however, for the STV 280/80. Some of these when in circuit give rise to relaxation oscillations presumably due to a stepped characteristic such as that shown[1] for the type Cossor S130.

Low operating 85A1 tube currents (i.e. 3·5mA) were chosen so that only four of the six channels could be used and the current would still remain within the recommended operating range. There appears to be no damage if no output is taken from the stabilizer unit even though the stabilized load is in excess of the makers recommended maximum. The time-delay switch was included to prevent overloading the stabilizer tubes when the equipment was switched on from cold. This switch permits the valve heaters to warm up sufficiently to allow anode current to flow immediately the stabilizer voltage is switched on. This was probably not essential, although necessary to keep to

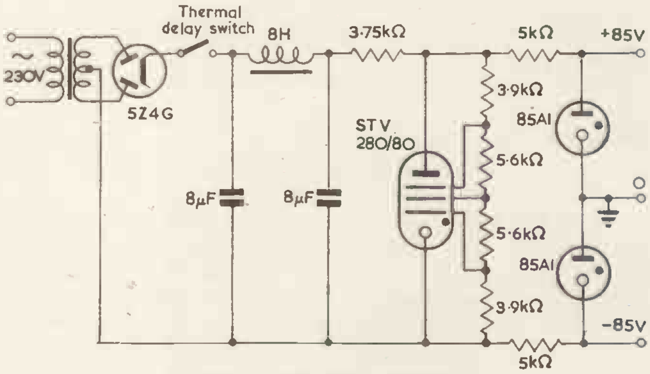

Fig. 1. The circuit described

Fig. 2. Recording of 5μV signal
Note that the noise level is approx 1μV

within the maker's specified maximum ignition voltage (125V) and current ratings.

Unselected carbon resistors were used and, rather surprisingly, these have not generated noise of sufficiently high voltage to be noticeable. It was found that no antimicrophonic mounting was necessary.

The power pack supplies 375V on load and is quite conventional—the only interesting point is the relatively small amount of smoothing required.

Acknowledgment

The author wishes to thank the Committee of Claremont Street Hospital for permission to publish this article.

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* *Formerly Claremont Street Hospital, Belfast.*

A Phase Controlled Oscillator

By P. G. Davis, M.Sc.

The article describes an oscillator so controlled that its frequency remains an exact multiple of mains frequency and constant in phase relation to it. The oscillator waveform is used for marking a scale of electrical degrees on a mains frequency waveform displayed oscillographically. The theory and design of the control are given together with a worked example.

WHEN it is required to record a recurrent waveform accurately a point by point measurement is frequently used. The instantaneous value of the wave is measured, by sampling circuit or oscillograph at equal intervals throughout the cycle and the curve plotted.

In the case of low power circuits the output waveform of which is to be measured, the driving or input wave may be obtained conveniently from a relaxation oscillator triggered by a source of higher frequency. The individual cycles of the high frequency then serve as time markers, subdividing one cycle of the circuit output waveform.

With apparatus working from the mains, however, this method is not immediately applicable.

The following summary compares various methods of subdividing one cycle of the mains waveform.

(1) A phase shifting transformer followed by a pulse forming circuit generating a single pulse which may be set anywhere in the cycle. This method is simple when a three-phase supply is available, but it is difficult to get comparable accuracy with change of mains frequency when the same device is used on a single-phase supply with phase shifting networks.

(2) A fixed frequency oscillator or ringing circuit triggered at one point in the mains cycle. This method will give equal time intervals, but for any mains frequency other than the nominal value it will not give equal fractions of the cycle. It is useful where measurements in terms of time are more important than measurements in electrical degrees.

(3) Generation of a harmonic by distorting the mains waveform and filtering. The filter corresponds with the ringing circuit in the previous case, so that the same remarks apply.

(4) An oscillator controlled to work at a multiple of mains frequency and with a given phase relation to it. A close approximation to equal fractions of a cycle may be obtained. Normal changes of mains frequency do not affect the system.

The instrument used in method (4) is termed the phase controlled oscillator (p.c.o.). The theory and design of such an instrument form the subject matter of this article.

Specification of the P.C.O.

(1) The p.c.o. must provide a recurrent waveform at an integral multiple N of mains frequency and constant in phase relationship with it.

(2) This p.c.o. waveform must contain clearly visible time marking points, as with a square, sawtooth or impulsive wave.

(3) The deviation of the phase of the marking points from the true $1/N$ cycle points must be less than, say $\pm 1°$ throughout the cycle.

(4) The change in phase of the marking points should be negligible for the allowed variation of mains frequency.

Principles of Operation

Fig. 1 gives the block diagram of the system. To supply the marker waveform a type of relaxation oscillator is chosen in which the frequency may be controlled by means of a voltage. This oscillator generates a frequency nominally at the required multiple N of the mains frequency f_i.

The output of this oscillator is frequency divided by a conventional circuit; e.g. multivibrator or blocking oscillator chain or by dekatrons or other aperiodic counter circuits. The final output of the counter chain which is a

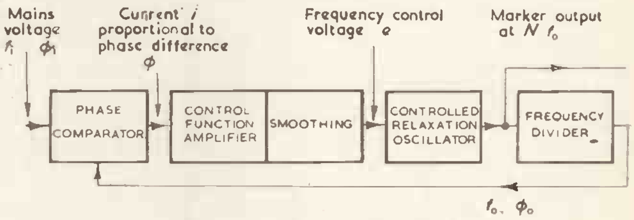


Fig. 1. Functional diagram of the system

square wave at $1/N$ of the frequency of the relaxation oscillator and therefore nominally at mains frequency is fed back to a phase comparator in which its phase is compared with that of the mains.

The phase difference is expressed as a current i and fed into the control function unit in which stabilizing and integral terms are generated. The reason for these terms is given later.

The output voltage of the control function unit is fed via smoothing to the relaxation oscillator, controlling its frequency in such a direction as to restore the phase difference between the frequency divider output and the mains to the required value.

The output current of the phase comparator will be pulsating at $2f_i$ (for a full wave circuit). This, flowing in the smoothing capacitor will cause a fluctuation in the frequency control voltage. Due to the presence of this ripple voltage the frequency of the relaxation oscillator will vary throughout one cycle of the mains, though the mean value will be accurately Nf_i. The marking points will consequently have a cyclic phase error about the $1/N$ points. Sufficient smoothing is therefore required to reduce this error to an acceptable value. The smoothing capacitor is incorporated in the control function unit.

The control function circuit must provide a loop stabilizing term. A term proportional to error, ϕ, is sufficient for this purpose.

An integral control term must also be provided for the following reason. A change in mains frequency will finally result in a corresponding change in the frequency of the relaxation oscillator and in the control voltage. This changed control voltage must be maintained even when the system has settled and the error in phase difference between mains and frequency divided output is again zero. Consequently one integration must exist between the phase comparator and the relaxation oscillator. An integrator will give an output varying with time unless its input is exactly zero. Therefore when conditions in the loop have become steady the output of the integrator is constant, and so its input, i.e. the current i and phase error ϕ must be zero. The control function amplifier must provide this term.

Circuit of P.C.O.

The electrical circuit is shown in Fig. 2. The phase comparator consists of a pair of short suppressor base pentodes V_1 and V_2. Their control grids are fed in anti-phase through current limiting resistors from the mains frequency f_i. For 50c/s working this is conveniently obtained from a transformer connected to the filament supply. The voltage swing is large enough to ensure rapid current turn-on and turn-off.

The suppressor grids are fed from the last multi-vibrator of the frequency divider with voltages of opposite polarity, diodes preventing them from being driven positive to cathode. Again the voltage is large enough to obtain switching action so that anode current is allowed to flow only during overlap of the two positive half-cycles as shown in Fig. 3. The peak value I_1 of the current is set by the screen potentiometer.

The mean anode current is thus a linear function of the duration T of the overlap which may be taken to vary with the phase difference ϕ as the two frequencies are, even under transient conditions, very nearly equal. Normally when the phase error ϕ is zero and conditions are steady the overlap is a quarter of a cycle.

The current passed by V_1 and V_2 is thus in the form of pulses of peak value I_1 and mean value depending upon T. This current is provided by a constant component I_0 through the anode resistor R_3, plus a current i from the cathode circuit of V_6 containing an alternating component at a frequency $2f_i$ and, possibly a more slowly varying transient term.

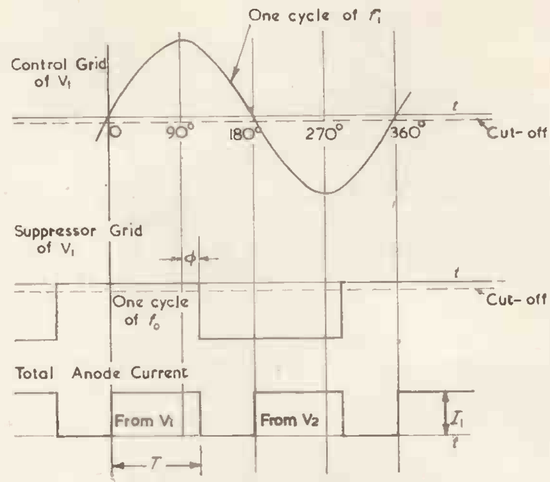


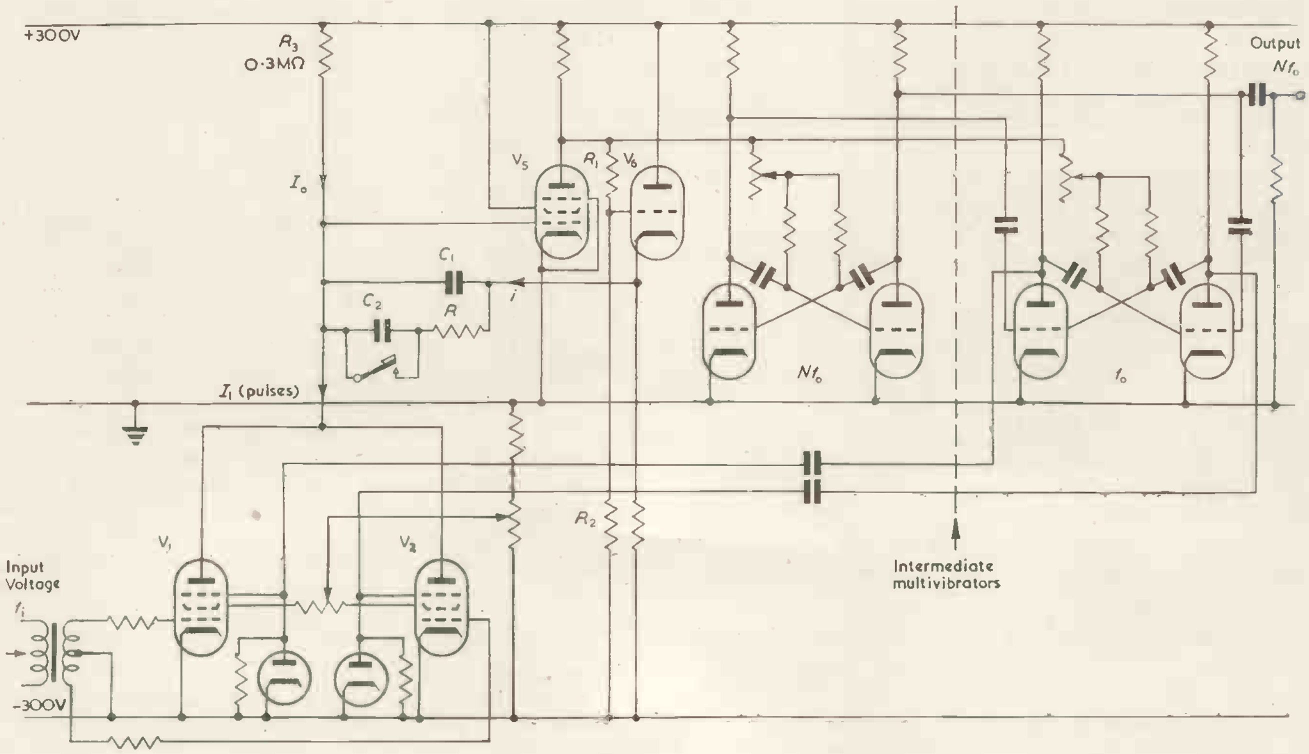
Fig. 3. Operation of the phase comparator

An alteration to the setting of the phase comparator screen potential divider will alter I_1. Suppose the screen voltage is reduced, thus reducing I_1. The circuit will reset itself so that when steady the mean anode current is again equal to I_0 so that the overlap must now be greater than 90°. The multivibrator waveforms are now retarded in phase. Thus the screen potential divider serves as a manual setting control for the phase of the marker wave.

The value of the smoothing capacitor C_1 is calculated on the assumption of equal peak currents from V_1 and V_2. A screen balancing potentiometer is provided to ensure this condition may be obtained without using matched valves.

The control function amplifier consists of a pentode V_5, d.c. coupled to a cathode-follower V_6. This pentode always works in its linear range, its grid excursion being small compared with the h.t. voltage, I_0 is therefore very nearly constant, thus providing a reference current with which the current of V_1 and V_2 is compared.

Fig. 2. The circuit of the p.c.o.



Any transient difference i between total valve current and I_0 is now supplied by the feedback amplifier through the circuit C_1, C_2 and R. C_1 is the smoothing capacitor, C_2 is the integral capacitor and R provides the proportional or damping term.

The anode of the pentode amplifier V_5 is connected to the ends of the charging resistors of the multivibrators remote from the grids so that a change in the anode voltage of V_5 will vary their frequency. Any number of stages of frequency division may be inserted, so that N may be made as large as required. For a large number of marking points to be useful the cyclic phase error due to the comparator ripple must be reduced to a sufficiently low value.

The cathode-follower V_6 provides a low impedance feedback point for the control circuit. The potential divider R_1 and R_2 enables the control circuit components to work with zero d.c. potential difference across them. This is necessary as the integral capacitor must be shorted out on starting. No other starting arrangement is required. Since the comparator current increases and then decreases with change of phase in alternate half-cycles, no attempt need be made to determine polarities in the loop. The circuit

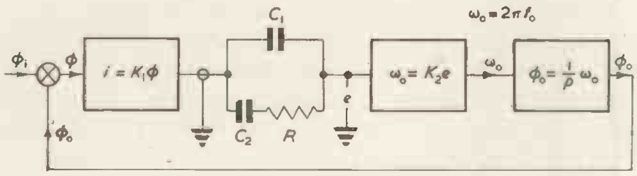


Fig. 4. Analytical diagram

synchronizes automatically on the appropriate half-cycle. C_2 may then be unshorted.

Conventional multivibrators have been used for the whole of the frequency division chain as they are easily synchronized and their waveform is acceptable both for the markers and for the comparator suppressor grids.

If stabilized power supplies are not used it is necessary to stabilize I_0 and the screen rheostat supply of the phase comparator.

Analysis of the Circuit

The control loop block diagram is shown in Fig. 4. The voltage at the cathode of V_6 is given the symbol e, but due to the presence of the potential divider R_1 and R_2 (Fig. 3) the multivibrators are controlled by a voltage greater than e, by the factor $(R_1 + R_2)/R_2$, approximately 4/3. This factor may be included in the calculation by defining K_2 as the frequency control sensitivity referred to the voltage at the cathode of V_6. The frequency division can also be included in K_2 which is thus defined as the change in angular frequency of the last multivibrator ω_0 per unit change of feedback voltage e.

Since for a step change in e the change in frequency of the first multivibrator is completed within half a cycle of its waveform the relationship between frequency and voltage is assumed instantaneous in calculating the dynamic performance of the loop.

The sensitivity of the phase comparator is K_1, where K_1 is defined as the change in mean anode current i per unit change of phase difference ϕ.

The phase of the last multivibrator output ϕ_0 is defined as the time integral of its frequency ω_0.

The circuit equations are now as follows:

$$\phi = \phi_i - \phi_0 \quad \ldots\ldots\ldots\ldots \quad (1)$$

$$i = K_1\phi \quad \ldots\ldots\ldots\ldots \quad (2)$$

$$e = \frac{i}{p(C_1 + C_2)} \cdot \frac{1 + pC_2R}{1 + pRC_1C_2/(C_1 + C_2)} \quad \ldots \quad (3)$$

$$\omega_0 = K_2 e \quad \ldots\ldots\ldots\ldots \quad (4)$$

Note $\omega_0 = 2\pi f_0$

$$\phi_0 = (1/p)\omega_0 \quad \ldots\ldots\ldots\ldots \quad (5)$$

Combining equations (2) to (5) the open loop transfer function is obtained:

$$\phi_0 = \phi\,\frac{K_1K_2}{p^2(C_1 + C_2)} \cdot \frac{1 + pC_2R}{1 + pRC_1C_2/(C_1 + C_2)} \cdot \quad (6)$$

The system is thus analogous to a position control servo with viscous friction and proportional plus integral control.

The time-constant in the denominator $RC_1C_2/(C_1 + C_2)$ is shorter than that in the numerator RC_2 so that the break frequency of the denominator term is higher than that of the numerator. The loop can thus be stabilized by suitable choice of component values. The Bode diagram is shown in Fig. 5.

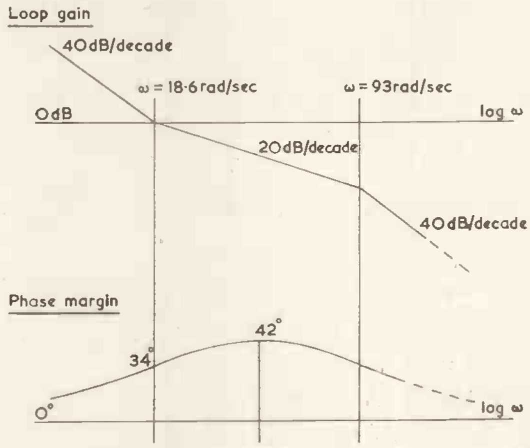


Fig. 5. Bode diagram

Design of Control Circuit

Figures are given below for an oscillator working at a multiple of 50c/s. The value chosen for N would be about 10 or 12 giving nominal 1msec or 15° marking points respectively. A different value of N would only affect the design by altering K_2 slightly and possibly altering the design specification limit for cyclic error.

Value of K_2

The value of K_2 depends upon the d.c. working voltage at the anode of V_5. At 200 volts to earth $K_2 = 1{\cdot}65$ rad/volt.sec for the multivibrator shown, while at 100V to earth, the value used in the present design, $K_2 = 2{\cdot}2$ rad/volt.sec. The value of K_2 also varies slightly with N and with the setting of the multivibrator trimming potentiometer.

Value of K_1

The value of K_1 depends upon I_0 and the normal overlap angle which together determine I_1.

$$i = I_1(\tfrac{1}{2} + \phi/\pi) - I_0 \quad \ldots\ldots\ldots \quad (2A)$$

Making $I_0 = 1$mA and $I_1 = 2$mA, $K_1 = 0{\cdot}637 \times 10^{-3}$A/rad.

Value of C_1

C_1 is determined by the values chosen for I_1, I_0 and the smoothing requirement. Fig. 6 shows the relevant waveforms drawn for the condition of 90° overlap. The symbols e', ω_0' and ϕ_0' represent the respective peak-to-peak values of the ripple on the control voltage, frequency and phase.

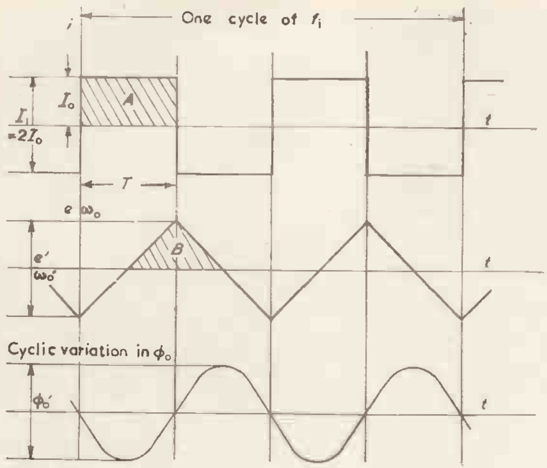


Fig. 6. Calculation of cyclic error

The presence of R and C_2 in parallel with C_1 will only affect the smoothing by a minute amount. They are ignored in the calculation.

$$e' = (1/C_1)\int idt = (1/C_1) \times \text{Area } A = (I_0T/C_1) \quad \ldots\ldots (7)$$

$$\omega_0' = K_2e' = \frac{K_2I_0T}{C_1} \quad \ldots\ldots\ldots\ldots\ldots\ldots (8)$$

$$\phi_0' = \int \omega_0 dt = \text{Area } B = \tfrac{1}{2}T \times \tfrac{1}{2}\frac{K_2I_0T}{C_1} = \frac{K_2I_0T^2}{4C_1} \ldots (9)$$

$$\therefore C_1 = \frac{K_2I_0T^2}{4\phi_0'} \ldots\ldots\ldots\ldots\ldots\ldots (10)$$

$\phi_0' = 1^\circ = 17{\cdot}45 \times 10^{-3}$rad

$T = 5 \times 10^{-3}$sec

$I_0 = 1 \times 10^{-3}$A

$K_2 = 2{\cdot}2$ rad/V.sec

$\therefore C_1 = 0{\cdot}79\mu$F.

Working with a larger value of I_1 will give increased ripple, but it can be seen by redrawing Fig. 6 that for the same value of I_0, ϕ_0' cannot be increased by a factor greater than 2. Since I_1 is altered for zero adjustment of the markers it is necessary to include a safety factor of 2 which has been done in the above calculation by choosing 1° for the peak-to-peak value of ϕ_0' instead of 2° as in the specification.

Value of R

If the integral control is removed, as would be done by short-circuiting C_2 (i.e. $C_2 \to \infty$), a second degree transfer function is obtained:

$$\phi_0 = \phi \frac{K_1K_2R}{p(1 + pC_1R)} \quad \ldots\ldots (11)$$

giving:

$$\phi_0 = \phi_i \frac{K_1K_2R}{K_1K_2R + p(1 + pC_1R)}$$

$$= \phi_i \frac{1}{1 + p/K_1K_2R + p^2C_1/K_1K_2} \quad \ldots\ldots (12)$$

R may now be found to give critical damping, a purely arbitrary choice knowing that restoring the integral term will make the system somewhat more oscillatory.

$$\frac{1}{K_1^2K_2^2 \cdot R^2} = (4C_1/K_1K_2) \quad \therefore R = \frac{1}{\sqrt{4K_1K_2C_1}} \quad \ldots\ldots (13)$$

Using the values previously assumed $R = 17\text{k}\Omega$.

Value of C_2

A value for C_2 can be estimated from the frequency response. Since in normal operation the system is never called upon to follow abrupt or even rapid changes in mains frequency the degree of damping of the response is not important.

Separating the time-constants and consequently the break frequencies in the numerator and denominator by a factor of 5 enables adequate damping to be retained.

$$C_1/(C_1 + C_2) = 1/5 \quad \therefore C_2/C_1 = 4 \quad \therefore C_2 = 3{\cdot}16\mu\text{F}$$

The Bode diagram for these values is drawn roughly to scale in Fig. 5.

Change in Marker Phase with Change in Mains Frequency

The ideal requirement that no change in marker phase should result from a change in mains frequency demands the existence of a pure integral control term which needs infinite gain in the control function amplifier. The following calculation shows that adequate accuracy is obtained with a gain of 50.

Change of mains frequency (49 to 51c/s) $= 2\text{c/s} = 12{\cdot}6$rad/sec

Change of voltage at the cathode of V_6 $= 12{\cdot}6/2{\cdot}2 = 5{\cdot}7$V

Change in grid voltage of V_5 (gain of 50) $= 5{\cdot}7/50 = 0{\cdot}114$V

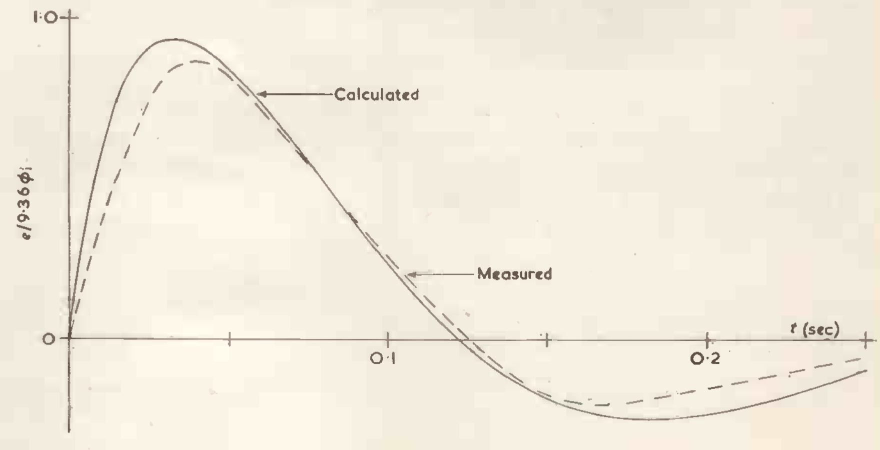
Change in mean anode current I_0 ($R_3 = 0{\cdot}3\text{M}\Omega$) $= \dfrac{0{\cdot}114}{0{\cdot}3 \times 10^6} = 0{\cdot}38 \times 10^{-6}$A

Change in ϕ_0 $= \dfrac{0{\cdot}38 \times 10^{-6}}{0{\cdot}637 \times 10^{-3}} = 0{\cdot}6 \times 10^{-3}$rad $= 0{\cdot}034^\circ$

Equivalent total leak across capacitors $= 5{\cdot}7/0{\cdot}38 \times 10^{-6} = 15\text{M}\Omega$

Thus the accuracy is not impaired by finite amplifier gain or by slight leakage in the capacitors.

Fig. 7. Comparison of measured and calculated transient



Experimental Check

To verify the analysis by the transient response the quantity most easily measured is the output voltage of the control function amplifier for a step change of phase comparator screen voltage corresponding with a step change of ϕ_i. Measured and calculated transients are compared in Fig. 7. The calculation is given below:

From equations (1) and (6):

$$\phi_o = \phi_i \frac{(1+pC_2R)K_1K_2/C_1C_2R}{p^3+p^2(C_1+C_2)/RC_1C_2+pK_1K_2/C_1+K_1K_2/C_1C_2R} \qquad (14)$$

From equations (4) and (5):

$$e = \phi_o p/K_2$$

$$\therefore e = \phi_i \frac{(1+pC_2R)pK_1/C_1C_2R}{p^3+p^2(C_1+C_2)/RC_1C_2+pK_1K_2/C_1+K_1K_2/C_1C_2R} \qquad (15)$$

Substituting the values:

$$e = \phi_i \frac{1{\cdot}5\times10^4p + 7{\cdot}94\times10^2p^2}{p^3+93p^2+1{\cdot}78\times10^3p+3{\cdot}3\times10^4} \qquad (16)$$

$$\therefore e = 9{\cdot}36\,\phi_i[e^{-8.9t}(\cos 19t + 0{\cdot}964\sin 19t) - e^{-7.5t}] \qquad (17)$$

OTHER QUANTITIES ($N = 10$)	ASSUMED OR CALCULATED	MEASURED
K_2 (rad/V.sec)	2·2	2·67
I_1 (mA)	2·0	2·27
Peak-to-Peak ripple voltage e'	6·33	7·04
Cyclic error in markers ϕ', degrees peak to peak ..	1·0	1·3

Acknowledgements

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A Multi-Trace Cathode-ray Tube Display

By K. E. Wood* and T. C. Keenan†, B.Sc.

A description is given of a nine trace cathode-ray tube display equipment operating with sequential switching of the nine input signals. The required switching waveform is derived from a neon counter. Input frequencies up to 100c/s can be displayed.

During the construction of a large analogue computer for the solution of aircraft flutter problems, the need arose for a simultaneous display of nine quantities generated in the computer. The display accuracy was not required to be better than 5 per cent in amplitude. Equipment was developed to display nine simultaneous traces initially on a standard long persistence Cossor 1049 oscilloscope. The sweep speed had to be varied between 0·5 and 4·0c/s and on account of this the sequential switching had to be done on the Y scan. A block diagram of the equipment is shown in Fig. 1.

In this, each of the nine inputs is applied through an adjustable gain amplifier and then to a diode switch. These switches allow each applied input to be switched in sequence to the oscilloscope during the period of an applied pulse which is derived from the cathodes 1 to 9 of a high speed neon counter. A Y-scan synchronized from the tenth cathode is also applied so that when a diode switch is operated the instantaneous input amplitude and the instantaneous scan waveform amplitude are applied to the Y plates.

Brightness pulses are also applied to the gun of the tube thus producing a series of dots vertically on the screen. Each channel is sampled at 1 000c/s so that for an input frequency of 100c/s 10 dots per cycle are used to produce the trace. A typical set of traces is shown in Fig. 2.

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A slight angular displacement of the raster results which with an X scan of 4c/s and a Y scan of 1 000c/s is given by:

$$\theta = \tan^{-1}\frac{4}{1\,000} \simeq 0{\cdot}02^\circ$$

where the X and Y scans are of equal amplitude, which is negligible.

Diode Switch

This is shown in Fig. 3.

Under quiescent conditions, but with a sine wave input (V_{in}) applied to P_1 diodes V_{2a} and V_{3b} alternately conduct on negative and positive excursions. The signal at point Q is approximately given by:

$$V_{in}R_1/R_2$$

Where V_{in} is the applied signal and the diode forward resistance R_f is much less than R_1. When a positive pulse is applied to the grid of V_{1a}, negative and positive pulses are applied to V_{2a} and V_{2b} respectively. If these amplitudes are greater than V_{in} peak then both diodes will be cut-off. The amplitude at point Q will be given by:

$$\frac{V_{in}R_3}{R_2 + R_3}$$

$R_3 \ll$ than the diode back resistance R_b.

Point Q in each diode switch is then fed into the common summation amplifier.

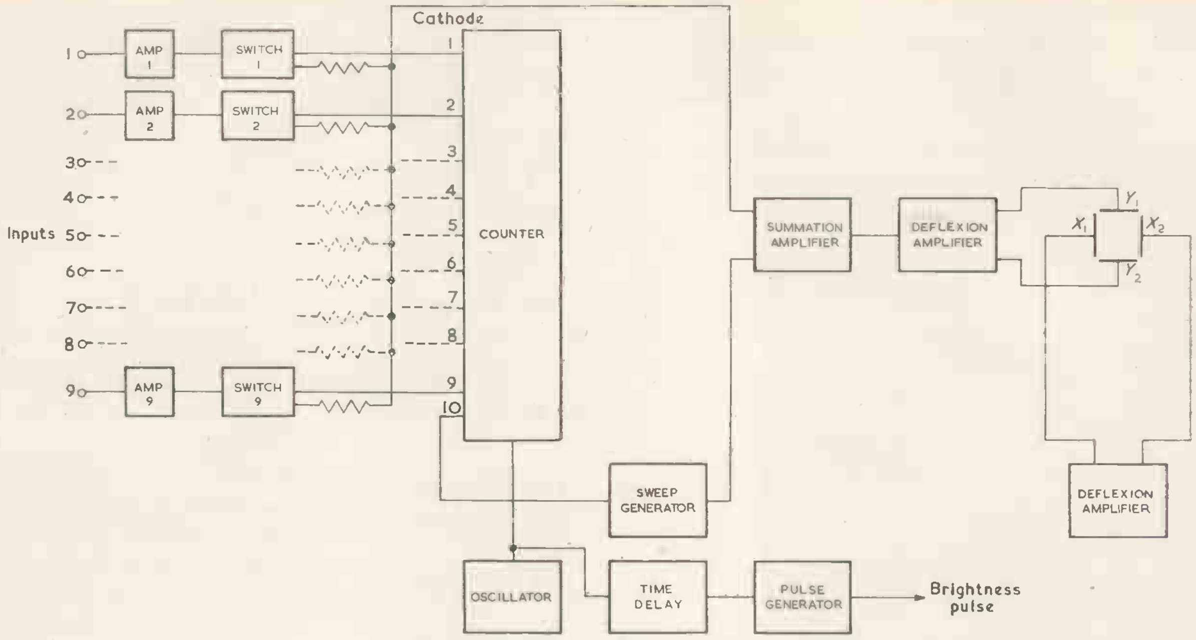


Fig. 1. General arrangement of the equipment

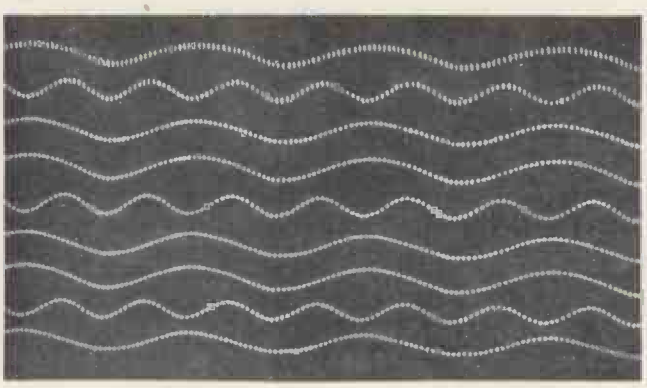
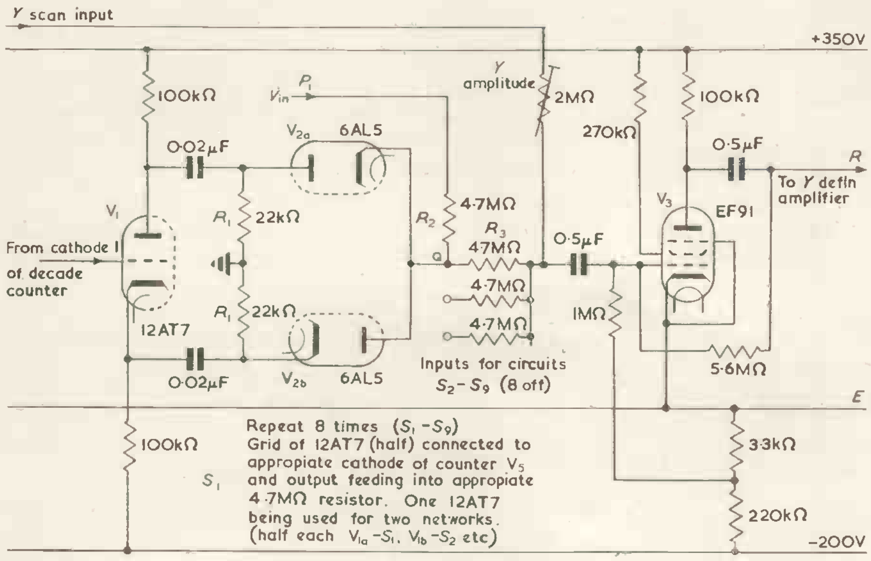


Fig. 2. A typical set of traces

Fig. 3. The diode switch



Sequential Switching (Fig. 4)

Valve V_1 is connected as a Wien bridge oscillator producing a frequency of 10kc/s. This is fed into a pulse forming valve V_{2a} which in turn applies a positive pulse to the gate valve V_3 and an additional pulse forming valve V_4. The neon counter V_5 (G10/241E) then produces pulses sequentially on its ten cathodes, nine of which are applied to the nine diode switches and the tenth cathode produces a pulse which is applied simultaneously to a pulse amplifier V_{2b} and the sweep generator V_6. The output from V_{2b} is used to cut off the brightness pulse at the anode of V_3 and hence blanks out the flyback trace. An exponential waveform is produced at the anode of V_6, but only the initial portion is used. It is fed into the summation amplifier together with the diode switch outputs and hence into the oscilloscope deflexion amplifier. A typical Y-scan waveform is shown in Fig. 5.

A simple X scan is provided in the normal day to deflect the tube. It is interesting to apply one of the nine traces to the X scan when nine Lissajous figures may be obtained.

Although a small tube was used initially in the equipment, it is proposed to apply the switching portion to a 12in long persistence electrostatic tube with only slight modifications to the more conventional deflexion amplifiers. A higher sampling rate is considered possible with the hard valve decade scaler RYG10 and silicon diodes in the switches.

Conclusion

The equipment described has satisfactorily displayed the nine input quantities on a small screen, with an accuracy within the 5 per cent required. Although the input fre-

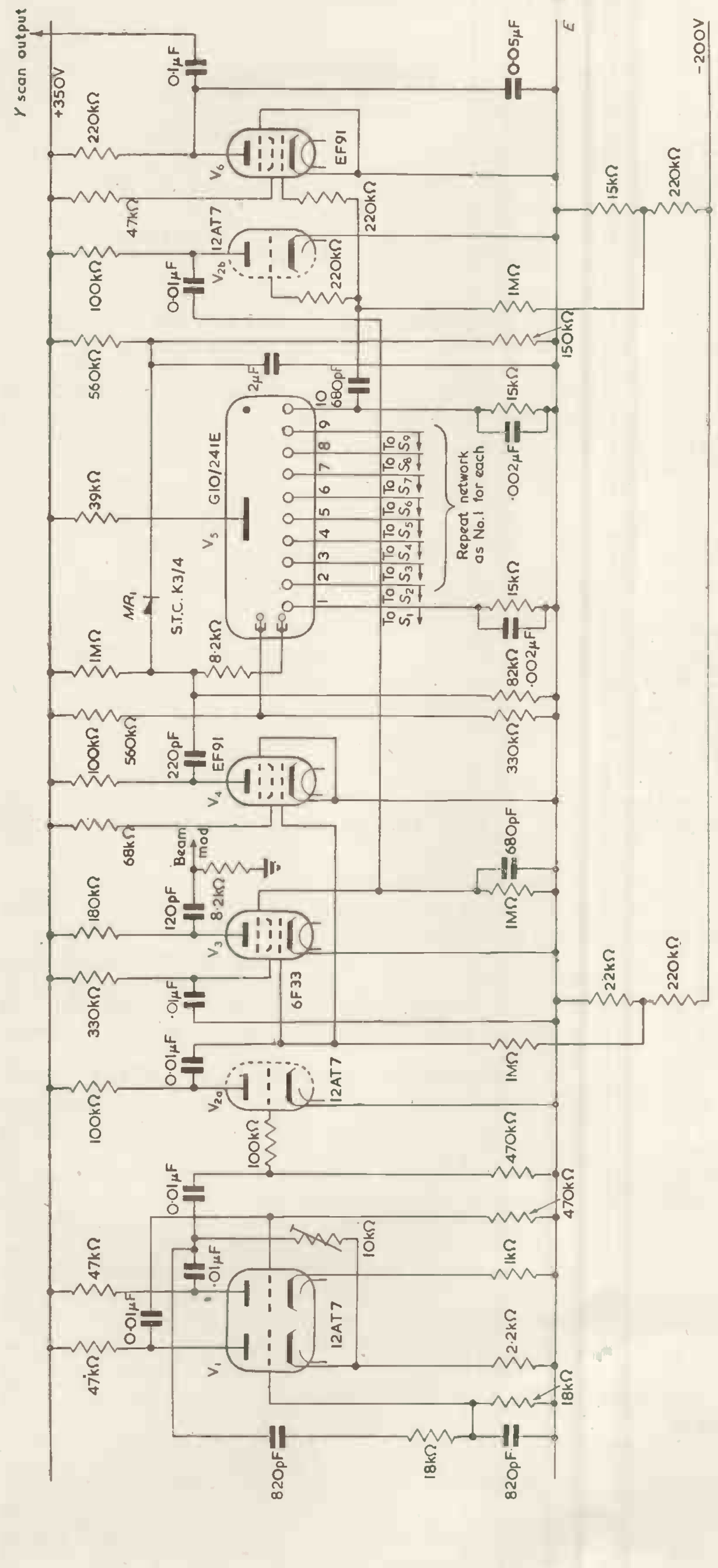


Fig. 4. Circuit for sequential switching

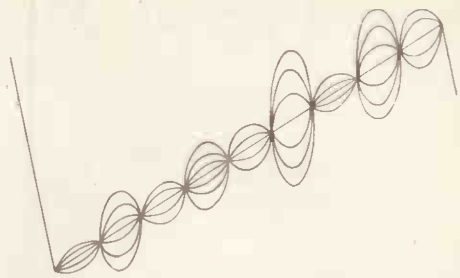


Fig. 5. A typical Y-scan waveform

quency is at present limited to about 100c/s there seem good possibilities of improving this.

Acknowledgments

The authors wish to acknowledge the Chief Electronic Engineer, Vickers-Armstrongs Aircraft Ltd. for permission to publish this article and to the Ediswan Electric Co. Ltd. for their co-operation in the construction of a large cathode-ray tube.

Addendum

Since this article was written the equipment has been modified for use with a twelve inch electrostatic cathode-ray tube. The tube uses an e.h.t. of 6kV and it's associated deflexion amplifiers require an h.t. of 1·5kV. The modifications to the actual switching unit include using a bootstrap circuit to produce the Y-scan waveform and increasing the amplitude of the brightening pulses. It was also found desirable to insert a pulse shaping amplifier between each cathode of the counter and it's associated switch, since on the large display a small amount of coupling was visible between traces, this being due to the lack of squareness of the pulses from the counter tube not producing a definite switching action. In this modified form the display has proved completely satisfactory.

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Industrial Particle Accelerator

British Insulated Callender's Cables Ltd are now installing a 2MV particle accelerator at their Research Laboratories. This is one of the first particle accelerators to be installed in the United Kingdom for purely industrial researches.

The accelerator is of the Van de Graaf type, housed within a pressure container operating at about 300lb/in^2. It is a universal apparatus which can be adapted to supply high energy beams of electrons, positive ions, neutrons or gamma rays.

The installation will be used for general research into the effects of high energy irradiation on materials used in cable and capacitor manufacture. In particular, it will be employed for continuing the researches on the effect of irradiation on polyethylene, on which the Company has carried out considerable work during the last few years.

Digital Methods in Control Systems

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A great deal is nowadays heard of the so called automatic factory and although very few completely automatic plants exist, there is a tendency in most industries to mechanize manufacturing processes of a repetitive nature. As the complexity or accuracy of the processes increases, more and more data of its parameters has to be stored or manipulated and this can be most conveniently done in digital form.

This article briefly describes and classifies some of the mechanisms required in a digital control scheme and discusses the merits of the mechanisms for various applications.

Particular attention has been paid to the problem of superimposing digital control schemes on existing plant and reference is made to the present developments in this field.

Considerable attention is now being directed toward the possibilities of the so called automatic factory in which entire manufacturing processes are performed without human intervention. Much has been achieved to this end in the control of continuous processes, such as are found in the chemical industry, with the greater understanding of closed loop systems that emerged during the war. The requirements here, however, are often to regulate the process conditions over long periods rather than to perform large numbers of intricate operations. Interest in the latter problem has been stimulated with the advent of the electronic computer and its immense capacity for storing information in a rapidly accessible form.

The large quantities of data necessary to programme complex machine operations are most conveniently stored in digital form and although certain open loop, pulse operated control systems have been successful, notably the automatic telephone, the self-correcting features of error actuated systems are desirable in manufacturing applications.

Servo mechanism practice on the other hand has been largely developed for continuously varying or analogue quantities and the fundamental equation.

$$\theta_e = \theta_i - \theta_o$$

where θ_i = input

θ_o = output, and

θ_e = error,

is solved by analogy using potential dividers, pneumatic balances or valves and linkages.

In order to combine the advantages of closed loop controls with the storage capacity and accuracy of digital systems the solution of the equation and hence the quantities θ_i, θ_o and θ_e must be in digital form.

In this article the application of digital methods in control systems is discussed and details of suitable elements for the realization of these systems are described. Reference is also made to current developments in this field.

Open Loop Systems

A simple device for controlling the position of a shaft is shown in Fig. 1. The displacement of the shaft is controlled by the number of pulses transmitted to the impulse motor by the contact *A*.

Elementary as this device may appear it is worthy of mention since it is used in the most extensive pulse operated control system in existence: the automatic telephone exchange.

The dial telephone is a fairly reliable piece of equipment but even so wrong numbers are occasionally obtained. A subscriber obtaining 6 000 instead of 7 000 may be aggravated or even embarrassed but little material damage is likely to be done. In a system controlling a manufacturing process, however, the error of one digit in a significant place may be disastrous. For this reason the self-correcting features of a closed loop system are to be preferred.

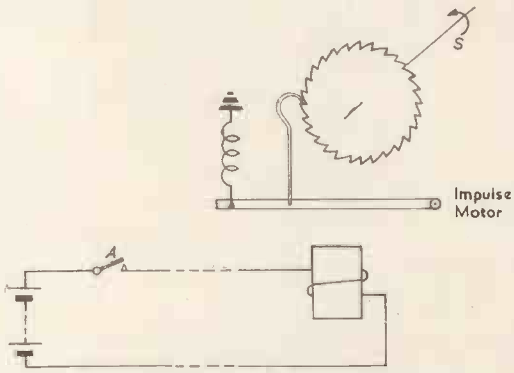

Fig. 1. Elementary pulse operated position control

Closed Loop Systems

Digital computers can be classified according to whether they perform arithmetical operations on numbers digit by digit (serial type) or on all digits together (parallel type). In the same way digital control systems can be divided into the type in which the error quantity, θ_e, is represented by a train of pulses or in which it is dealt with as one whole number.

A further subdivision can be made of the latter type into the systems which do arithmetic entirely in digital form and those which do some of the arithmetic by converting the digital quantities into analogue form.

These systems are now dealt with in more detail.

Pulse Counting Systems

This is probably the most widely used system at present and is illustrated by the simple closed loop system shown in Fig. 2. Here the number of pulses representing the desired movement of the shaft is stored in the input counter θ_i. The shaft is rotated by a servo motor and a contact system coupled to the shaft feeds pulses into an error counter until the number of pulses in each counter is the same, whereupon the on-off controller stops the servo motor. Serious errors in the shaft position should be eliminated since the number of revolutions made by the shaft have been counted and compared with the desired number.

A system working on this principle has recently been brought out commercially for filling containers with a given quantity of material under the name of Plastab[1]. A

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standard mechanical weighing machine is used with an optical commutator fitted to the pointer staff. The pulsing contact is a glass disk presenting alternate transparent and opaque segments to a beam of light falling on a photo-electric cell. As the staff revolves with increasing weight on the platform the photo-electric cell transmits pulses to an electronic counter until the number of pulses is equal to the pre-set weight whereupon the delivery of material is cut off.

The equipment has a number of interesting features. It will take into account the tare weight of the container as soon as it is placed on the platform of the machine, thereby ensuring that the net weight of the material filled is always the same.

Filling is allowed to proceed quickly to begin with but as the contents of the container approach the correct weight the rate of filling is reduced. This is a form of anticipatory control which is discussed in a later section.

The weight of the container and its contents appear in

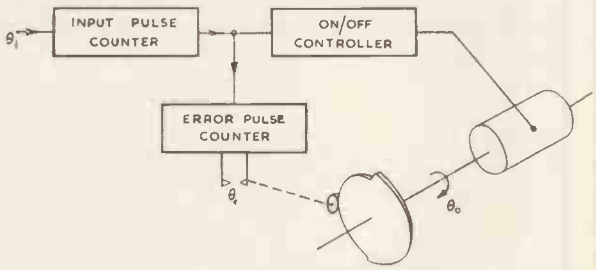

Fig. 2. Pulse counting position control

digital form and this fact is made use of to print out details of each container using an electric typewriter or to record this information on punched cards.

Pulse counting techniques can be used for measuring length and an elegant method for detecting small displacements has recently been developed at the N.P.L. using Merton diffraction gratings[2]. The movement of the interference patterns between two finely divided glass plates is counted photo-electrially as the plates move relatively to one another. Movements of ·0001in can be detected with plates divided into 10 000 lines per inch.

Diffraction gratings have been applied by Ferranti Ltd to the control of machine tools[3]. Due to the backlash that may exist in lead screws it is not considered sufficiently accurate to measure the rotation of the lead screw. The work table being controlled is therefore fitted with one diffraction grating which moves relative to another fixed to the bed of the machine. The movement of the work table causes a train of pulses to be transmitted to an electronic adder on which the required movements have been preset. Various contours can be machined from cartesian coordinates worked out by a computer.

Pulse counting systems are essentially serial in nature and are therefore economical if the control point is remote from the controlled shaft and they are relatively simple in construction. They are however open to the objection that an accidental loss of a number of pulses due to local fault conditions are not corrected for automatically.

Special precautions must also be taken if the output shaft is likely to overshoot whereupon the pulse counter must subtract instead of add the pulses emitted until the overshoot has been corrected. They are therefore particularly suitable where the load is a frictional resistance rather than an inertia or where the output mechanism is such that the system can be brought to rest within one digit of coincidence being obtained.

Commutator Systems

So far the assumption has been made that the output shaft will stop rotating as soon as the error signal becomes zero and the conditions of overshoot due to the inertia of the load have only been touched on briefly.

The output can be stopped at the correct digit provided an impulsing output mechanism is used. These mechanisms, which are dealt with in a later section, are generally only suitable for light output torques or for low response speeds and, due to their pulsating nature, are not suitable for power systems.

If an orthodox motor is used to drive the output shaft and its load, overshoot is likely to occur and indeed is often tolerated in order to obtain a high speed of response. It is then necessary to construct the digital mechanism, as in continuous servos, so that the relative positions of input and output are compared and to use the difference between them, whether due to under- or overshoot, to drive the output into coincidence with the input.

This may be done in several ways[4]:

(1) By converting the digital information into analogue form and applying the well known continuous servo techniques.

(2) By measuring the output position in digital form, calculating $\theta_e = \theta_i - \theta_o$ by digital methods and applying a function of θ_e, modified if necessary by stabilizing signals also in digital form, to the output mechanism.

(3) By a combination of these two methods.

These methods are now examined briefly.

Digital-to-analogue Conversion

A simple servo control using potential dividers is shown in Fig. 3(a) and it is evident that the input potential divider could be replaced by stepped resistors as shown in Fig. 3(b). The input is then dependant on which one of the contacts *A* to *J* is closed. The servomotor will drive the output until it coincides with the numerical value of the input.

Fig. 3(b) has been drawn with only ten discrete stages but this can be extended to further decades by the use of an appropriate input resistor network such as the Kelvin-Varley shunt. Convenient networks for binary input have also been described[5,19].

It is desirable that the error signal should be isolated from the detector while the input setting is being changed from one value to another otherwise the slightly different times of closure of the contacts may cause spurious error signals to be sent to the detector and amplifier.

Digital-to-analogue conversion is often convenient because, generally speaking, stabilizing signals are more easily produced and manipulated in continuous than in digital form. It is also convenient as it allows preset or other digital input schemes to be used on existing analogue servo controlled installations.

The overall accuracy of digital to analogue conversion systems is limited by the precision with which the components can be constructed. This can be improved to some extent by the use of coarse and fine control, but if this is done the inherent accuracy of digital systems is not made use of to its best advantage.

Digital indication of the output

It has been mentioned above that the actuating signal for the system can be obtained by comparing the input with the output when both quantities are represented in digital form. A prerequisite, obviously, is a means for representing the position of the output shaft digitally and this is referred to later.

Assuming that both input and output are available as binary numbers it would be sufficient, for an on-off servo, to decide which of the two numbers is the larger and use this information to drive the servo motor to rotate the output shaft in the direction which makes both numbers equal.

This can be done by means of a circuit of the type shown in Fig. 4.

The A contacts are set by the input and the B contacts by the output. Two circuits run through the contact net-

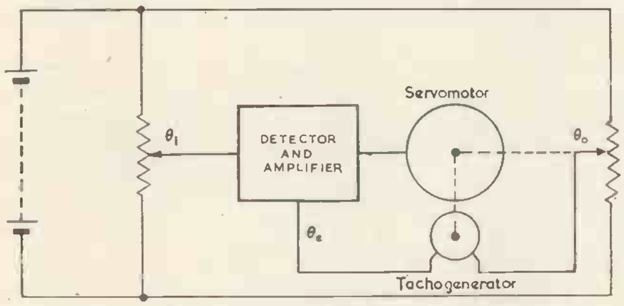

Fig. 3(a). Simple servo with tachogenerator in feedback

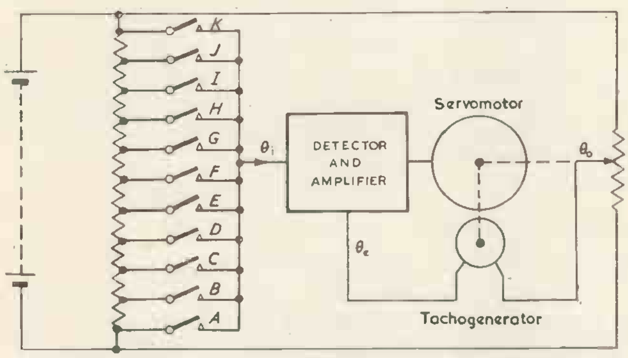

Fig. 3(b). Simple servo with digital input and stabilizing feedback

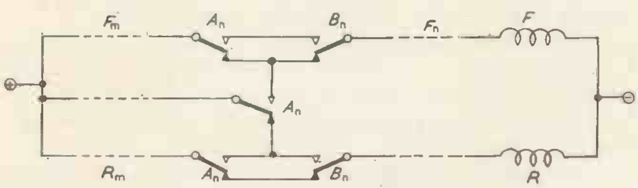

Fig. 4. Binary coincidence detector

work and are connected to the coils F and R which can be two windings of a polarized relay or the forward and reverse fields of a split field motor. If positive is connected to the coil F the shaft rotates to increase the output and if positive is connected to R the output is decreased. When both F and R are energized the polarized relay or servo motor does not operate and the movement of the output shaft stops.

The greater of two binary numbers is that having a 1 in a more significant place than the other. The circuit in Fig. 4 makes use of this fact. The most significant stage in which input (A_n) and output (B_n) differ is examined. If $A_n > B_n$ i.e. if $A_n = 1$, $B_n = 0$ positive is connected to coil F whereas if $B_n > A_n$ i.e. $A_n = 0$, $B_n = 1$ positive is connected to R. Note that while $A_n \neq B_n$ the stages of lesser significance than A_n and B_n have no influence on which coil of the polarized relay or servo motor is energized.

As soon as the output has moved so as to bring $A_n = B_n$ tht circuit is set up through this stage whether $A = B = 0$ or 1 to the next most significant stage. Ultimately all stages will coincide and the system will stop. It should be noted that there will be no intermediate reversals of motion while the output is moving into coincidence with the input.

Such a circuit will only indicate the sign of the error and no indication is given of its magnitude. Since stabilizing devices depend on the modification of the error magnitude, the arrangement is not suitable where conditions of severe instability might arise.

It should be noted that this system is virtually a bang-bang servo having a single digit dead zone.

Circuits which will give the magnitude and sign of the digital error can be constructed and, for the sake of completeness, the brief details of one stage of such a circuit is shown in Fig. 5 but it should be noted that unless the magnitude of the error itself is used to prevent overshoot (as in Plastab) it will not be easy to stabilize a system because of the complication of generating functions of error and its derivatives in digital form.

The result of the subtraction process at each stage is indicated at the terminals $D(1)$ and $D(0)$. If $A_n - B_n = 1$, $D(1)$ is energized; while if $A_n - B_n = 0$, $D(0)$ is energized. $C(1)$ and $C(0)$ are the carry-over to the next stage. If the input $(A) <$ output (B) a carry-over will appear at the $C(1)$ out terminal of the last stage indicating that $(A - B)$

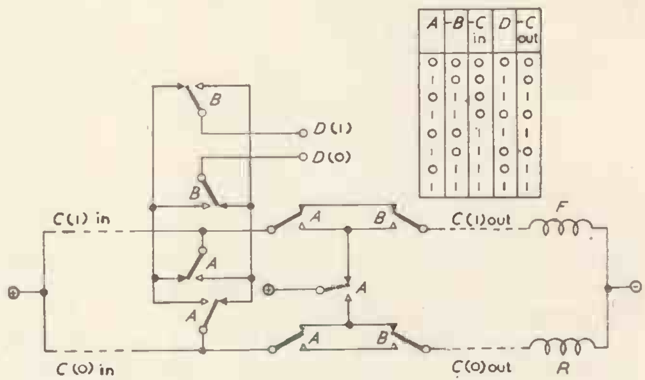

Fig. 5. Binary subtractor

is negative. This can be made to operate the servo in the correct direction.

The magnitude of the error appears at the D terminals. If $A > B$, the $D(1)$ terminals give the magnitude direct. If $A < B$, the complement must be taken by referring to the $D(0)$ terminals and adding 1 to the first stage.

A simple circuit that will indicate the order of the error in binary terms has recently been described by Foss[19]. In this circuit the most significant digit in the binary number representing the error is obtained directly although the less significant digits are not manifest as they are in the circuit shown in Fig. 5. In certain circumstances the order of magnitude of the error is as useful as its full numerical value for reducing the gain in stages as coincidence is reached. The circuit has, moreover, the advantage that with a slight rearrangement of contacts the output from a progressively coded commutator can be compared directly with a pure binary input.

In the preceding descriptions contact operated devices have been used as examples for the sake of simplicity. In a fast system the dead zone due to the transit time of the contacts, even using high speed relays, may have a deleterious effect in which case electronic circuits should be used. Valve circuits for performing the simple mathematical operations required in comparing the digital input and output are now commonplace and details will be found in most text books on electronic digital computers[6,7,8,9]. Photo-electric commutators suitable for the output indication have also been developed[10,11]. The design of commutators is referred to later.

A servo for controlling the position of a shaft accurately has been described by Thomassen[12]. Although an electronic adder was used with end-round-carry, the servo was a small one and no attempt was made to introduce digital stabilizing feedback. The output shaft was controlled by a simple mechanical star wheel so that it would stop dead within one digit of the error signal reaching zero.

Combined analogue and digital schemes

It has been seen that analogue servos are easier to construct than digital systems, particularly when dealing with the stability problem at high power levels, but have limited accuracy, while pure digital systems, although of virtually limitless accuracy can become exceedingly complex if they are not inherently stable.

Adequate stability can usually be obtained in analogue systems and it would therefore appear unwise to use digital methods for stabilizing purposes except in certain special cases. It is suggested therefore that a composite system is the best.

The system should be designed to be inherently stable within certain limits, for example, by using different gains in the analogue and digital domains. Within the prescribed limits, the output can be controlled digitally by a commutator type system. Beyond these limits, that is during the transient response of the system, it functions as an analogue type and provision is made for stability in the feedback. The limits of action of the digital part form the dead zone of the analogue part. A criterion of the design therefore is that the analogue part should bring the output to rest in the dead zone in the shortest possible time.

When discussing simple commutator schemes a circuit was developed (Fig. 4) which dealt with the most significant level of the error. To construct a composite servo, an additional stage can be added to the circuit and this serves to connect the analogue part of the servo when the dead zone limit is exceeded.

As soon as the analogue part positions the servo so that the error falls in the dead zone the digital part is brought into circuit and brings the output into exact coincidence.

This arrangement applied to a series motor drive is shown in Fig. 6. It is assumed that the motor is equipped with a suitable contactor control gear and a magnetic brake.

The sign of the error is detected in the digital domain by the polarized relay W comprising forward and reverse coils X and Y, but there is an overriding control of W by the analogue detector relays P and Q if the error exceeds a certain value.

The polarized relay W sets the power circuit to the motor to forward or reverse depending on whether coil X or coil Y is energized but the operation of polarized relay W alone does not permit the control gear to cut out the accelerating resistors R_1, R_2 and R_3. These are only allowed to close if either relay P or Q is energized. If the error is sufficiently large therefore the motor will run up to full speed.

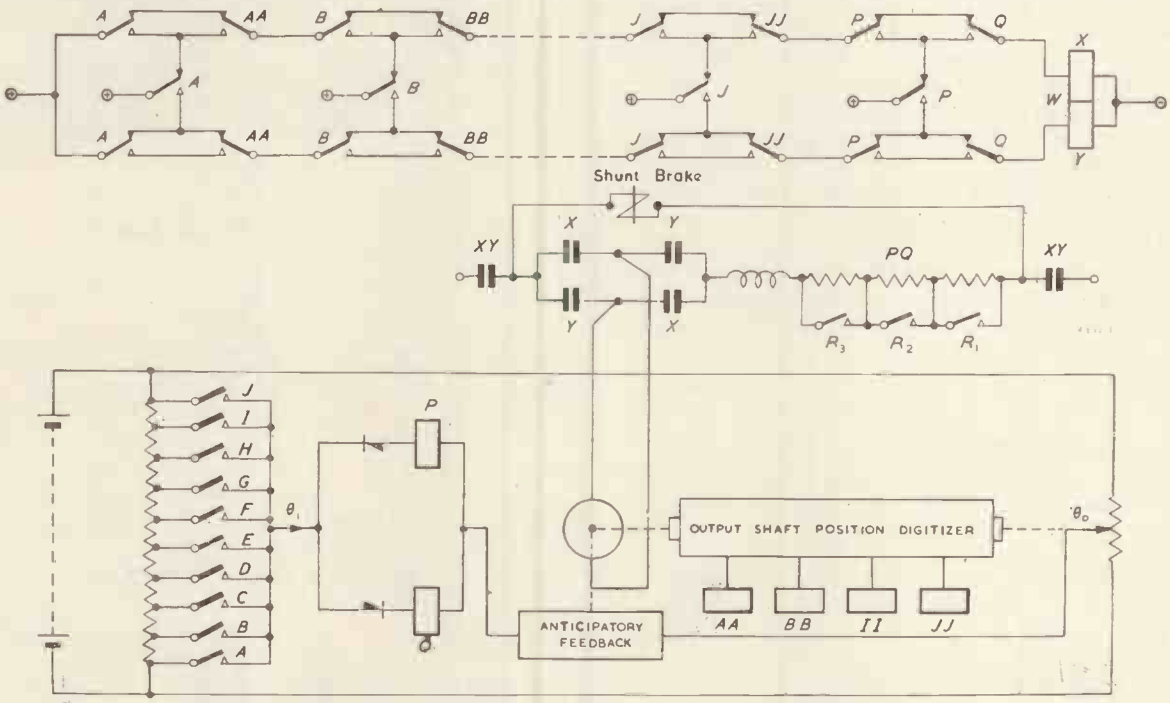

Fig. 6. Composite analogue and digital control system

Anticipatory control of the $V|V|$ type can be provided in the feedback path of the analogue circuit and under its influence relay P (or Q if the motor is running in reverse) will be de-energized before the required output setting is reached and a little later relay Q (or P for the reverse direction) is energized as the net error changes sign. This reverses X and Y and the motor is plugged. Eventually the analogue error signal is insufficient to maintain Q and the control is transferred to the digital domain which brings the system into coincidence at reduced speeds. When the digital error is zero the system is brought to rest by the magnetic brake.

To prevent hunting in the digital domain the brake must be capable of bringing the system to rest at the correct digit and a suitable value of the total starting resistance must be chosen so that the motor does not run too fast in this domain.

It is essential that the magnetic brake should be off while the polarized relay W is energized but by means of a plugging detector circuit the assistance of the brake can be provided during the plugging period.

An interesting servo of the pulse counting type for machine tool applications has recently been described[13], in which the output speed (i.e. the machine feed rate) is controlled by comparing a voltage proportional to the pulse repetition frequency with a tachogenerator on the output.

Output Mechanisms

Impulse Motors

The impulse motor shown in Fig. 1 is a suitable source of output power where the load is not too great. The ratchet wheel is moved forward one tooth for each impulse received by the electromagnet. Each movement of the armature or each tooth advanced by the ratchet wheel is therefore the equivalent of one 'bit' and by virtue of the pulsing and locking nature of the pawl and ratchet no difficulty is experienced with overshoot. Speeds of 50 to 100 pulses per second are normal but it is difficult to construct a ratchet with a small number of teeth. Impulse motors are therefore used to their best advantage where the output is light and has to move only a few degrees per step. Each revolution can be accurately divided into a large number of parts, and for these reasons they have been applied to machine tool table drives[14].

The uniselector used in the automatic telephone exchange is an example of impulse motor output and the design of the mechanism has been carefully studied[15]. Two-directional impulse motors have been made having two magnets one for forward motion and one for reverse.

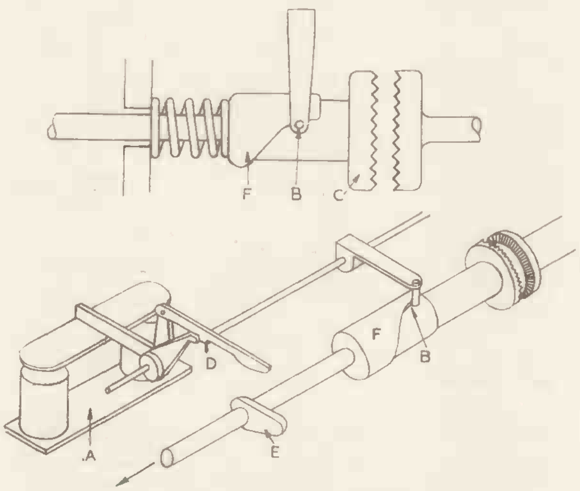

Fig. 7. Clutch operated impulsing drive

Rather more torque and a larger stroke can be obtained by the use of rotary actuators[16]. These are similar in construction to a d.c. motor but, because of the limited angular travel, the commutators are replaced by flexible leads. High torques can also be obtained by the use of a massive ratchet drive powered by a thrustor.

One complete revolution of the output shaft per bit can be obtained from the device shown in Fig. 7. The output shaft can be connected by a clutch to a motor running at constant speed. After making one revolution the mechanism disengages the clutch until a further impulse is received. The operation can be followed from the diagram. When the electro-magnet (*A*) is energized a spigot (*B*) holding one half of the clutch (*C*) against a spring is lifted and the clutch is engaged. After a few degrees rotation the linkage (*D*) holding the spigot arm is released by cam (*E*) and the spigot falls into a cylindrical cam (*F*) which disengages the clutch at the end of one revolution. The magnet must be de-energized and re-energized before the next revolution can be imparted to the output.

Two-directional operations can be obtained by using two clutch mechanisms working into a mechanical differential gear as shown in Fig. 8.

Impulse motors are convenient mechanisms to use with pulse counting systems since a local contact on the pulsing mechanism can be used to supply the counter with the number of pulses equivalent to the displacement of the output.

Aggregate Motion Mechanisms

Output movement proportional to a binary number can be obtained directly from an interesting group of devices known as aggregate motion mechanisms.

The principle is illustrated in Fig. 9, which shows a number of pin-jointed links connected to the plungers of solenoids *W*, *X*, *Y* and *Z* .

Each plunger moves a fixed distance (l) when its coil is energized but the effect of this movement on the output (g) depends upon the configuration of the linkage. The pro-

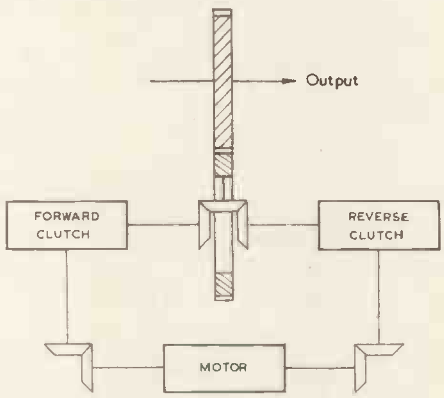

Fig. 8. Two directional impulsing drive

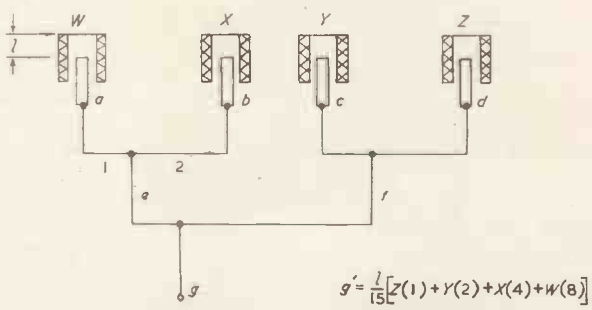

Fig. 9. Rectilinear aggregate motion system

portions of the linkage are adjusted so that the effects of the respective solenoids on the output bear a 2:1 ratio to one another. Thus in Fig. 9 the movement of $g = (l/15)(d + 2c + 4d + 8a)$.

A similar device having rotary inputs and output is shown in Fig. 10. The links for halving and adding the motions are replaced by differential gears which have the same properties.

Both the mechanisms described are constructed so that the output movement is less than each separate input. Linkages which double the sum of the two inputs can be constructed, but due to inertia and friction they become difficult to work when more than a few stages are used.

A similar limitation in the number of useful stages applies to the halving linkage (Fig. 9) for the throw of the plunger is limited and the effect of say the tenth stage on the output is $l/1023$ which may be lost in the clearances required in the pin joints or which may be too small to use.

Aggregate motion schemes are very useful for a limited number of stages particularly for actuating the valves in hydraulic servos. They are fast in operation as they are parallel in form, all input levels operating simultaneously.

Commutators and Digitizers

If a fast response and high torque is a requirement, pulsating outputs cannot be tolerated and a composite system is often suitable. The position of the output shaft must then be converted to digital form for comparison with the input. There are many ingenious ways of performing the analogue to digital conversion and 17 different methods have been described in a recent survey[17]. Not all of these methods are convenient and the commutator and digitizer types are probably the most useful, as, by suitable design, the position of the output shaft in digital form can be made almost instantaneously available using simple and robust equipment.

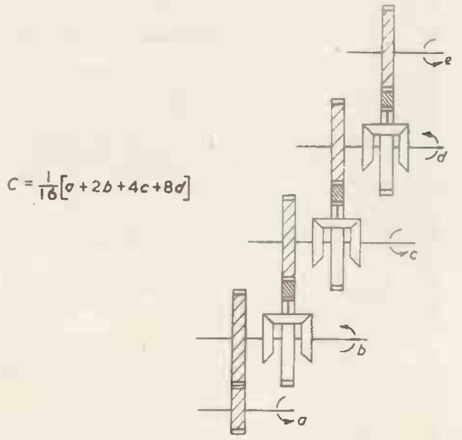

Fig. 10. Rotary aggregate motion system

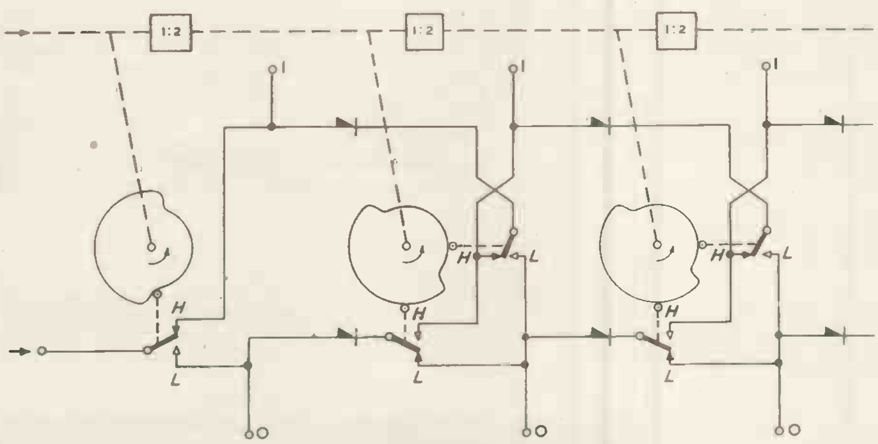

Fig. 11. Binary shaft position digitizers

A major problem with commutators and digitizers is avoiding the ambiguities that can occur in a chain carry (i.e. when 999 changes to 1000). A number of ways exist for getting round this difficulty and an interesting one is that described by Scarborough[18]. This is illustrated in Fig. 11. If the switching operations are traced through for continuously varying inputs it will be found that no ambiguities in the output occur. Although the ambiguities are avoided the electrical output will register zero during the transit of the changeover contact on the first stage. It is therefore desirable to use contacts having a toggle action and as low a transit time as possible. Suitable shaft digitizers can also be constructed on the same principles for decimal operation.

Digitizers are not always convenient mechanisms because an increment of one digit requires an appreciable rotation of the input shaft; nor are they suitable for linear measurements.

Coded commutators are far more suitable devices because of the more compact storage of information. Commutators are available having over 8 000 different combinations in a 10in disk. The problem of ambiguity can be overcome by using a progressive code in which adjacent combinations do not differ by more than one element[10], or by the use of two commutators displaced by $\frac{1}{2}$ a digit in which the selection of an unambiguous position is determined by a guard ring[11].

Input Media

Digital controllers can be operated either by hand or automatically and in certain cases it may be desirable to construct the circuits so that the system can be switched from one to the other. In Fig. 6 the input signal is made to operate a number of contacts and this can easily be done manually by the use of push buttons or selector switches.

If an automatic system is to be used the input may either be derived from a digital computer or from a store. Many forms of storage are available including punched paper tape[3], punched cards[20], magnetic drum or tape[21], photographic film[3] or, in simple cases, uniselectors. Any of these storage media may or may not be used in conjunction with a computer and the choice of the store must be made in accordance with the requirements of any particular application.

The choice of decimal or binary scale depends on circumstances. The decimal scale is probably the best if provision is being made for manual control since it will be more familiar to the operators. Punched card systems and certain electronic computers perform their functions in decimals. Magnetic and punched paper tape and the majority of electronic computers work in binaries but sometimes facilities for binary-decimal or decimal-binary conversion are available.

Generally speaking binary systems are simpler and more economical in construction but difficulties may be encountered if the human operator has to intervene.

For simple repetitive processes there is little difficulty in extracting the required settings of the output from the store in sequence at the correct moment and the relatively simple electromagnetic elements described are suitable. This does however imply that the process being controlled follows a definite sequence. It may however be a requirement that the machine being controlled should make a choice of which of several courses should be followed at a certain point. For example a rolling mill may have to make further but lighter passes if the temperature of the ingot is low. In such cases the magnetic drum store with its associated computer elements would afford a solution since the complete contents of the store are accessible at short notice and the correct course to follow can be picked out.

Conclusions

It is apparent that digital methods can and are being used to good effect in control systems and they are particularly useful where a manufacturing process involves a high degree of accuracy or consists of a large number of repetitive operations.

As these are precisely the conditions that apply to the industries engaged in making a few finished products in large quantities it seems likely that the requirements for digital methods will increase. The changeover from manual to automatic control will in all probability be a gradual one and emphasis has therefore been placed in this article on the problem of superimposing digital methods on existing analogue installations. Ultimately there is certain to be more use of digital electronic and magnetic devices for control purposes but a good deal of development in computer techniques will have to be done before their use is widespread. In the meanwhile the simple electromagnetic devices described should be useful for *ad hoc* conversion schemes and thus prepare the way for complete automation.

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A Radio Whaling Buoy

A whaling buoy incorporating a small radio transmitter has recently been designed by Venner Electronics Ltd of New Malden, Surrey, to enable factory ships to recover and process whales killed by the catcher ship.

A number of radio buoys are held by each Catcher, and once a whale is harpooned the buoy is secured to it by means of a twenty fathom nylon line attached to a barb; this is long enough to ensure that should the whale roll over, the bouy is not dragged with it.

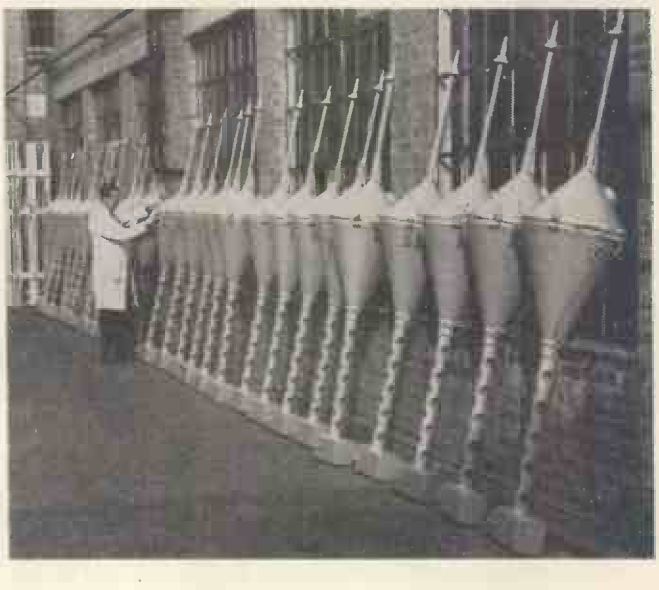
The buoy is switched on by means of an external control on the casing and left floating beside the whale. The equipment within the buoy ensures that the transmitter operates for two minutes in every ten, radiating between 1·7 and 2·1Mc/s a call sign in morse (each buoy has an individual call sign) and a long continuous signal for direction finding purposes. The call sign and d.f. signal are repeated four times. These periods of transmission are broken after two minutes to ensure that the buoys will radiate in this way for 22 hours on one battery charge. It is usual at this point for the catcher to radio the factory ship that a buoy has been secured to a whale, and to give the rough position and the call sign of the buoy. The catcher will then proceed to find further whales. During this time the factory ship has been processing other whales and before getting under way a d.f. bearing is taken on the buoy attached to the next whale.

The functioning of the transmitter mechanism is entirely automatic, being controlled by a time switch. This time switch is wound before the buoy enters the water and is capable of running on one winding for between 7 and 10 days. A clockwork motor drives a pair of cams, one of which switches on the filaments of the valves approximately half a minute before switching on the h.t. voltage. At the same time the coding mechanism commences turning and contacts open and close as the specially cut coding cam radiates the assigned call sign. This repeats four times. Should the time switch open the cam contacts before the completion of the fourth d.f. signal further self-holding contacts will ensure that the period is completed, the mechanism being left in the correct condition so that the next ON period will begin with the call sign.

The transmitter portion of the equipment contains two valves and is crystal controlled, the whole assembly being mounted on rubber supports to ensure that valves and components are protected from sudden shocks. A meter is fitted on the main top plate to facilitate tuning the transmitter. This is normally done aboard the factory ship but may be done aboard the catcher if required. The adjustment is simple—a knob on the transmitter chassis is turned until a minimum reading is obtained on the meter and then locked in position. The transmitter power needs are supplied by an accumulator housed at the foot of the buoy in which position its weight adds to the stability of the buoy.

A range of up to 150 miles has been obtained in the Antarctic due principally to the absence of interference on the 2Mc/s band in that region.

Radio whaling buoys ready for despatch. The transmitting equipment is located in the inverted cones, the upper cone being hinged for inspection and adjustment purposes. The box at the foot of the buoy houses the accumulator. The aerial is mounted on the insulator at the top of the buoy at the time of launching.



Silicone Insulants

Their Properties and Applications

(Part 2)

By B. F. W. Hayward*, D.F.H., Grad.I.E.E., Assoc.Mem.A.I.E.E.

IN part 1 of this article the various forms in which silicones are produced were described and details given of their unique combination of properties. It was seen that these properties stem principally from that part of the chemical structure which is common to all silicones, namely the alternate silicon-oxygen linkage. Silicones, therefore, possess the good dielectric properties, chemical inertness, and much of the thermal stability of glass, quartz and mica to which they are chemically related. They acquire flexibility, ease-of-handling and water-repellency from organic groups attached to their inorganic " back-bone ". The wide versatility of silicones arises from these basic properties, together with the fact that they are produced in a variety of forms including resins, elastomers, fluids, compounds and greases. Their uses in the electronic and electrical industries are inevitably growing rapidly and this article reviews some of the forms available together with applications of particular interest to the electronic engineer.

The outstanding feature of silicones to the electronic engineer is that they maintain good dielectric properties over a temperature range from $-100°C$ to $+250°C$, but there are many instances where silicones are used in electronic equipment because one or more of their many unusual physical properties is required. Usually, it is a combination of the properties of silicones both dielectric and physical, which by making possible simplification of design, justifies their use.

Silicone-Varnished Glasscloth and Tape

These products are made by applying several coats of a flexible silicone varnish to form a continuous film on glasscloth. Fig. 9 shows the thermal endurance of silicone coating varnish MS994 applied to glasscloth. For maximum flexibility and ease of application these materials are usually supplied with the varnish in a slightly under-cured condition.

The optimum physical and electrical properties are developed by further heat treatment, during which bonding of adjacent layers occurs.

Typical properties of a silicone coating varnish on 3 mil base glasscloth are:—

Electric strength, 1½in diameter electrodes, measured at 90°C, (min value) after ageing for 12 hours at 300°C	1 000V/mil
Permittivity, at 10^6c/s	2·8
Power factor, at 10^6c/s	·001
Water absorption, 24 hours immersion, per cent	0·5 to 1·0

Experiments carried out on the heat-ageing of silicone varnished glasscloth in vacuo, would indicate that there is a considerable future for the material in vacuum devices. Varnished glasscloth was heated in vacuo for one hour by which time gas development had stopped almost completely. Subsequently, the temperature was raised to 400°C, and at this temperature the gas development also dropped after an initial rise, but it remained appreciable, indicating progressive decomposition. The surface insulating properties were shown to be unimpaired. A gap of 0·15in between surface electrodes did not break down at 6kV applied for several minutes, and insulation resistance exceeded $5 \times 10^{11}\Omega$/sq.

Silicone-bonded mica and mica/glass

Silicone varnishes are particularly suitable for bonding mica, and the bond remains flexible throughout the working life of the material. This is an important feature in many

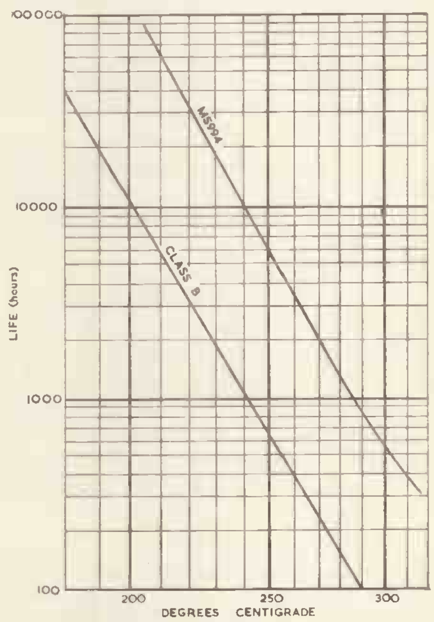


Fig. 9. Comparison of the thermal endurance of silicone coating varnish MS994 applied to glasscloth with an organic (Class B) varnish

The life is based on the hours necessary to reduce the electric strength of varnished glasscloth to half its initial value.

applications such as layer insulation in transformers. The high electric strength is especially important where space is at a premium.

Silicone resin-bonded glasscloth laminate

Rigid insulating materials are made from woven glasscloth bonded under heat and pressure with a silicone resin. They are produced as flat sheets in thicknesses up to 1in and as tubes. Complicated shapes can also be made by contact pressure techniques. Resin pre-impregnated glasscloth is also available and laminates can be prepared without the necessity for impregnating equipment.

Some properties of silicone resin-bonded glasscloth laminates and organic resin-bonded laminates are compared in Table 4.

* *Midland Silicones Ltd.*

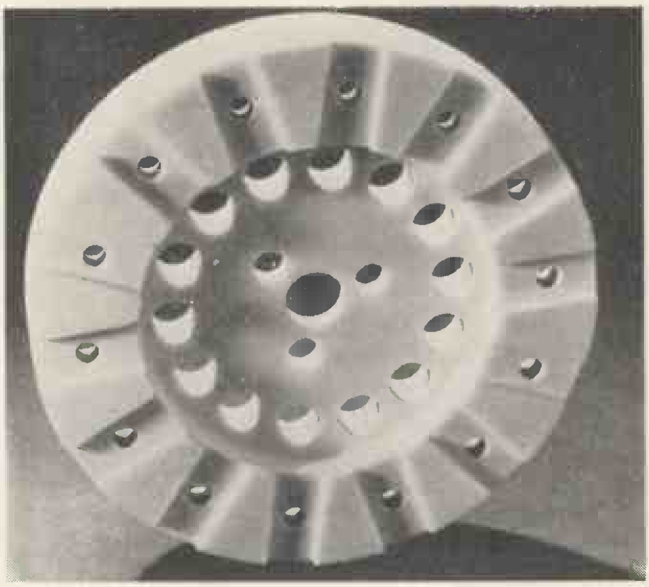


Fig. 10. A machined laminate of 1in thickness
[*Photograph by courtesy of Micanite & Insulators Co. Ltd.*

The variation in dielectric properties of one grade of silicone resin-bonded glasscloth laminate with temperature, frequency and humidity have been evaluated by Hyde[4], and some of the results obtained are given in Tables 5 to 9.

It will be seen that there is some power factor drift after exposure to high humidity, also a drop in insulation resistance. The material rapidly regains its original dielectric properties, however, once removed from the wet condition.

Silicone Impregnating Varnish

Silicone impregnating varnish MS997 is used to bond and impregnate electrical apparatus which is to operate at high temperatures or where a high order of reliability is required. Temperatures as high as 250°C can be tolerated for many hundreds of hours. The varnish would be expected to have at least 10 times the "life" cf the best organic varnish when operating at the same temperature. MS997 may be applied by vacuum impregnation, but excellent penetration can be

TABLE 4.
Comparison of silicone and organic resin-bonded laminates.

PROPERTY	PHENOLIC	MELAMINE	SILICONE
Power factor at 1Mc/s	0·06	0·08	0·002
Electric strength (V/mil)	150–200	150–200	250–300
Laminae breakdown (kV/in)	15–20	10–15	50
Insulation resistance (MΩ)			
(1) dry	10 000	20 000	500 000
(2) after immersion in water for 24 hours at 20°C	10	10	10 000
Water absorption (mg)	40	20	5–10
Tensile strength (lb/in².)	20–25 000	20–25 000	15–20 000
Shear strength, (lb/in².)	20–25 000	15 000	14 000
Impact strength (ft.lb) ($\frac{1}{2}$in × $\frac{1}{2}$in notched samples)			
(1) flatwise	5·0	6·3	5·6
(2) edgewise	2·0	1·8	2·0
Effective temperature resistance, °C	150	150	250

The test figures given in this table were obtained on $\frac{1}{8}$ in laminates in accordance with B.S.1137 where applicable.

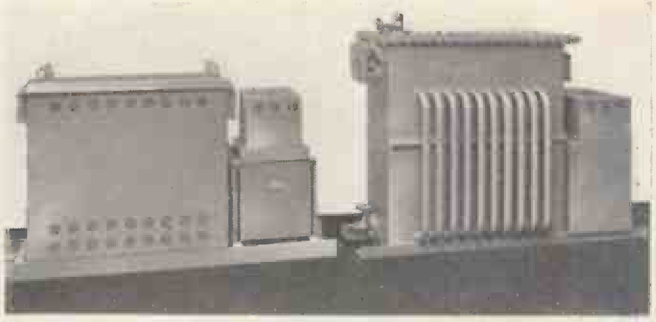


Fig. 11. These transformers are both of 100kVA rating. The larger is oil-filled and the smaller is a dry-type employing silicone insulating materials and silicone impregnating varnish MS997. The dry-type shows a saving in weight of 680lb
[*By courtesy of Marconi's Wireless Telegraph Co. Ltd.*

TABLES 5-9
Properties of one grade of silicone resin-bonded glasscloth laminate

Table 5. Variation of power factor and permittivity with temperature and frequency.

	TEMPERATURE	TAN δ	K
At 50c/s	20°C	0·006	3·6
	100°C	0·005	3·5
	250°C		
	250°C	0·005	3·5
At 1kc/s	20°C	0·001	3·4
	100°C	0·001	3·4
	250°C	0·001	3·4
at 100Mc/s	20°C	0·002	3·5

Table 6. Variation of power factor with heat ageing at 250°C.

LENGTH OF EXPOSURE AT 250°C (days)	TAN δ AT 50c/s	TAN δ AT 1kc/s
As produced	·006	·0011
3	·003	·0009
7	·002	·008
14	·002	·0007
17	·002	·0007
21	·002	·0007

Table 7. Change in insulation resistance under continuous exposure to 100 per cent relative humidity.

LENGTH OF EXPOSURE (days)	INSULATION RESISTANCE (MΩ)
7	300 000
14	110 000
62	80 000
70	75 000
85	45 000

Table 8. Rate of recovery of insulation resistance after removal from wet condition.

CONDITION	INSULATION RESISTANCE (MΩ)
Dry, as produced	7 000 000
Immediately after removal from water and wiping	175 000
2 minutes later	500 000
3 minutes later	1 000 000
45 minutes later	7 000 000

Table 9. Change in power factor on exposure to moisture.

CONDITION	TAN δ AT 1kc/s
As produced	·0006
After 24 hours at 100% R.H. and 40°C	·0091
After 24 hours in water at 25°C	·0127

obtained by dipping. Heat curing facilities are needed within the temperature range 200° to 250°C.

Silicone Enamelled Wire

Silicone resins do not have the properties of a good wire enamel. Synthetic wire enamels of the vinyl acetal type, as covered by B.S.1844, set a very high standard with regard to abrasion and solvent resistance and the ability to withstand "heat shock". Enamels of this type are, however, limited to a maximum long-term operating temperature of about 120°C. By combining a silicone with an organic resin, a modified silicone wire enamel with superior thermal stability, and having other characteristics approaching the requirements of B.S.1844, has been developed, and is now used by the principal wire coaters. Tests indicate long-term service for this enamel at temperatures up to

Fig. 12. The coils of these 50A and 100A relays for use in high speed aircraft at ambient temperatures of −75°C to +150°C, are wound with silicone enamelled wire

[By courtesy of Thorne Electrical Industries Ltd.

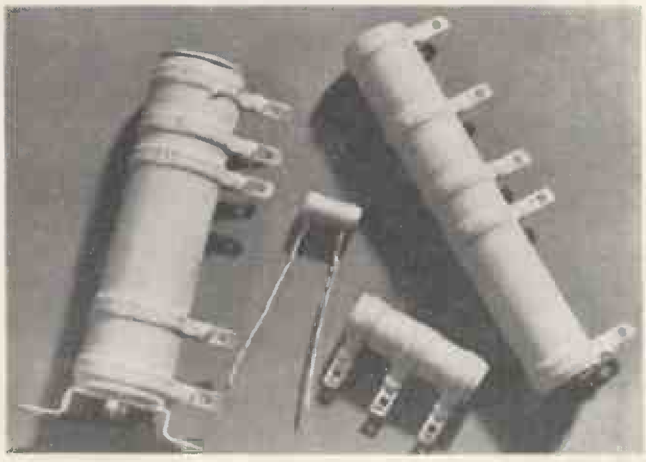


Fig. 13. Wire-wound resistors coated with silicone resin based cement

[By courtesy of J.L. Goldsman Ltd.

180°C when used within a silicone insulated system. The enamelled wire can be wound around itself without damage at room temperature, but to withstand "heat shock" the elongation must be reduced. Thus, when wound around 4 to 5 times its own diameter the enamelled wire withstands a "heat shock" test at 225°C. The thermo-plasticity of a wire enamel is important, and tests at 225°C have shown that the modified silicone enamel is satisfactory in this respect.

Resistor Coatings

Silicone based cement coatings have been developed to simulate the advantages of glazed coatings without the attendant processing difficulties. These coatings consist of thermo-setting silicone resins mixed with finely divided mineral fillers. Applied by dipping, they may be cured at temperatures below 250°C.

On cracked carbon resistors, pigmented silicone varnishes provide a heat resistant and water repellent coating.

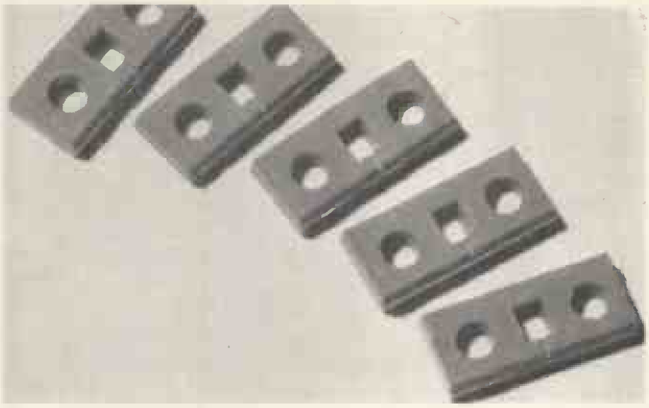


Fig. 14. Silastomer pick-up damping pads made by Dunlop Rubber Co. Ltd.

[By courtesy of Electrical & Musical Industries Ltd.

Lamp Capping Cements

Silicone resins are added to the cements used to bond the metal holders of lamps and valves to the glass envelopes. This is done to improve the high temperature bond and maintain good bond strength under wet or humid conditions.

Silicone Rubber

Silicone rubber, also known by its trade name—Silastomer, is an elastic material containing only silicone polymer and heat stable fillers.

Silicone Rubber Mouldings

There are a wide range of Silastomer mouldings, principally for use as gaskets, seals, diaphragms, etc. Special grades of Silastomer have a compression set, under constant deformation at 150°C and measured in accordance with B.S.903, of the order of 5 to 8 per cent of the original thickness. Reference is made here to three uses of Silastomer mouldings in the electronic industry each exploiting a different property. Illustrated in Fig. 14 are Silastomer

Fig. 15. Klystron tube employing silicone rubber moulding

[By courtesy of Varian Associates.

damping pads as used on pick-ups. Silastomer is used here as it is the only elastic material which shows virtually no change in modulus from arctic cold to tropical heat. Silastomer end caps on tubular capacitors ensure that high surface resistivity is maintained even under humid conditions. Silastomer mouldings seal capacitors in which chlorinated di-phenyls are the filling medium.

Illustrated in Fig. 15 is a klystron unit used in high-flying aircraft and guided missiles. The tube itself reaches temperatures as high as 150°C. One method of sealing out moisture and overcoming the corona discharge problem encountered at high altitudes, is to enclose the unit in a heavy and space-consuming pressure system. Equally satisfactory results have been achieved, however, by moulding silicone rubber around the connexions and lead wires. Flexible, light-weight and no larger than the tube itself, the silicone rubber withstands high operating temperatures and mechanical abuse. Of equal importance in this application is the fact that silicone rubber has very high resistance to the effects of corona.

Silicone Rubber Insulated Wires, Cables and Sleeving

Silastomer insulated connector wire and Silastomer sleeving is produced in various colours. The insulation is unaffected by soldering irons or by the curing temperatures associated with varnish impregnation and resin potting. Equipment wire is specified for the temperature range −75°C to +150°C in Defence Specification D.E.F.12.

Aircraft cables employ silicone rubber insulation with a glass braid overall. In the event of fire the silicone rubber will decompose, but the residue, held in position by the glass braid, is itself a good insulator, and the cable will continue to function. Precautions are necessary as certain oils and aircraft fluids have a deleterious affect on silicone rubber.

Silicone Rubber Pastes for Caulking and Filling

Silastomer pastes are solvent free materials available in various consistencies. They may be extruded on to winding wire, as a coil is being wound, to produce a "filled" coil or may be used as caulking and sealing compounds. After application they are cured by heating to form a resilient dielectric seal that excludes dirt and moisture and minimizes damage by vibration. Because of its relatively high thermal conductivity, coils filled with Silastomer can carry about 20 per cent more load and still operate at the same temperature as coils employing resinous insulating materials.

Silastomer paste may be applied to delicate components which are subsequently to be potted in resin. This provides protection against shrinkage of the potting resin.

Room Temperature Vulcanizing Silicone Rubber

The high curing temperature required by most grades of Silastomer has until recently limited its use in delicate electronic components. The recently introduced silicone rubber compounds that vulcanize at room temperature are, therefore, of particular interest. These compounds are supplied in fluid, medium and heavy consistencies, each comprising a two part system. When the two components are mixed in equal proportions, combination of the catalyst present in each part promotes vulcanization without additional heat or pressure. The material changes consistency after approximately four hours at room temperature, becomes a rubbery, non-tacky mass in about 24 hours, and develops optimum physical properties within 4 to 7 days.

The Doelcam Division of Minneapolis Honeywell, Bost n, have found R.T.V. silicone rubber the only material with all the essential properties for potting a very sensitive d.c. amplifier. The small amplifier illustrated (Fig. 16) has to withstand hard service and still amplify signals as low as 6×10^{-4} volts to a full scale output of 100 volts.

Silicone Rubber Coated Glasscloth

This is produced as semi-cured and fully cured material. The former comprises a layer of semi-vulcanized silicone rubber on one side of a glasscloth backing. When used as tape and half-lapped, it vulcanizes on heating to form a homogeneous, resilient and moisture proof jacket. The fully cured material is made using a grade of Silastomer which has been specially developed for coating woven fabrics from solvent dispersion. It is available from 0·004in thickness, and some of its properties were given in Table 3.

Glass braided sleeving coated with Silastomer is also available.

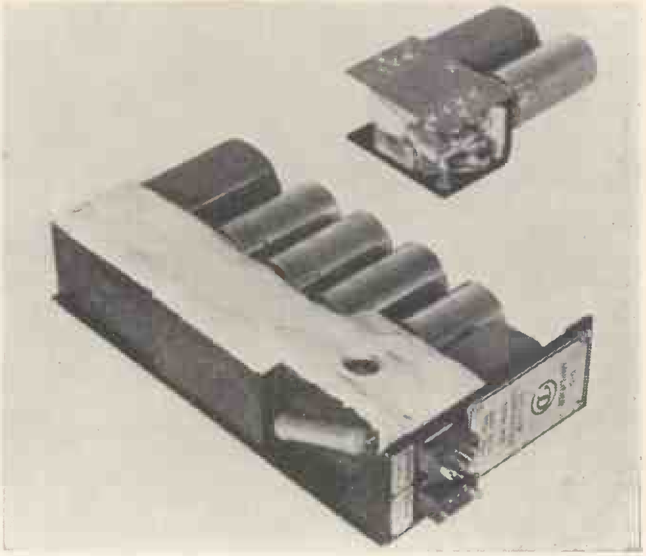


Fig. 16. A d.c. amplifier potted in R.T.V. silicone rubber
[By courtesy of The Doelcam Division of Minneapolis Honeywell, Boston.

Silicone Fluids

The properties of silicone fluids, which are supplied in a range of viscosities from 0·65cS to 100 000cS, were tabulated in Part 1. It was shown that these fluids are unique among liquid dielectrics in having low loss factors at high temperatures and frequencies, resistance to heat and oxidation, relatively slight change in viscosity with temperature change and high water repellency. They are also chemically inert and do not affect—and are not affected by—most other materials. These fluids are the basis of silicone compounds and greases, and find their main direct electrical applications as dielectrics in transformers and capacitors, in time delay mechanisms such as dash pots and as water repellent treatments.

In a study of small power transformers for aircraft, Morris[5] reached the conclusion that in view of the high premium on weight, the use of expensive silicone materials is justifiable. He pointed out that for climatic reasons the transformers should be sealed in containers, and while air-filling is simpler, a fluid filling results in lower thermal gradients, and also avoids corona trouble. A 500VA, 1600c/s, 5kV, silicone fluid filled transformer is the same size and weight as conventionally insulated, oil-filled types of only 200VA.

Supplementing, but not replacing, established types, silicone treatment of plastic film capacitors enables higher voltage stressings. The improvements resulting from the use of silicones have been thoroughly evaluated by The Telephone Manufacturing Co. Ltd. For high voltage d.c. and high frequency a.c. applications considerable size reductions over equivalent paper capacitors now become possible. The power factor of impregnated plastic film capacitors is so much lower in comparison with the best paper capacitors that heavy current power factor correction, for example, can be achieved with extremely small heat dissipation. As a consequence designs based on this new medium can be encompassed in a container having a heat dissipating surface area of dimensions which hitherto would have been thought quite impracticably small.

Glass or ceramic surfaces when treated with silicones will maintain high surface resistivity even under 100 per cent relative humidity. The moisture condenses in discrete droplets which remain insulated from one another. The efficacy of the treatment is well established, a well known application being the treatment of fluorescent tubes for instant start circuits.

It has been found that a very thin layer of MS550 fluid on the surface of solder baths will help to prevent the usual formation of scum.

Silicone Insulating Compound MS4

This is a smooth silicone grease with the consistency of petroleum jelly which it retains essentially from $-50°C$ to $+200°C$. Its good dielectric properties, high order of water-repellency, oxidation resistance, and non-melting property make it suitable for many applications outside the field usually associated with such compounds. It was originally developed, over ten years ago, to exclude moisture from aircraft electrical systems, and continues to be used for this purpose by the world's principal airlines. It is also used in the ignition systems of fighting vehicles.

Some of the properties of MS4 are listed in Table 10. The applications of MS4 in electronic equipment stem from these properties, together with its ability to exclude moisture and its extreme resistance to tracking and the effects of corona discharge. Television set manufacturers have found that an application of MS4 around the anode terminals of c.r.t's and associated high voltage components will prevent leakage due to moisture. It is also a sealing and potting compound for small aircraft transformers, capacitors, aerial connectors, coaxial cables, plugs and sockets and a damping medium in gramophone pick-ups.

The weatherproofing properties of MS4 have proved themselves in tests carried out by Post Office Engineers. Starting in February, 1954, nearly 4 000 terminal blocks were treated, the lids being deliberately left off in some cases. No fault due to low insulation has occurred so far. The coaxial plugs and sockets of a four stack v.h.f. turnstile antenna, have been treated with a thin film of MS4, and have given trouble-free service for 18 months.

The British Broadcasting Corporation use MS4 to help preserve the resilience and flexibility of rubber sealing washers on high power transmitting valves. The treatment also helps to prevent the washers sticking to the anode flanges. Outside broadcasting equipment used in connexion with the recent television broadcast from H.M.S. "Bulwark" was treated with MS4 to help keep out moisture.

TABLE 10.

Some typical properties of Silicone Insulating Compound MS4.

Volatility*, 24 hours at 200°C, (per cent)	Max.2
Volume resistivity, up to 200°C (Ω/cm³)	10^{12}
Permittivity, between 100c/s and 10Mc/s	2·8
Power Factor, between 100c/s and 10Mc/s	0·001
Electric Strength†, at 10 mil, (V/mil)	Min.500

* *M.O.S. D.T.D.*825
† *M.O.S. D.T.D.*900/4298

Core-retaining Compound

A thicker version of MS4 has been specially developed for locking iron dust cores and other electronic components. It shows little change in consistency from $-50°$ to $+200°C$, and over this range of temperature it will lock firmly, and yet allow smooth adjustment. Reports from electronic manufacturing companies confirm that the material is a considerable improvement over compounds previously used for this purpose. One major company has stated that other compounds which they had tried were so unsatisfactory that they reverted to mechanical means of locking. Their tests on the silicone core-retaining compound have caused them to reconsider the problem.

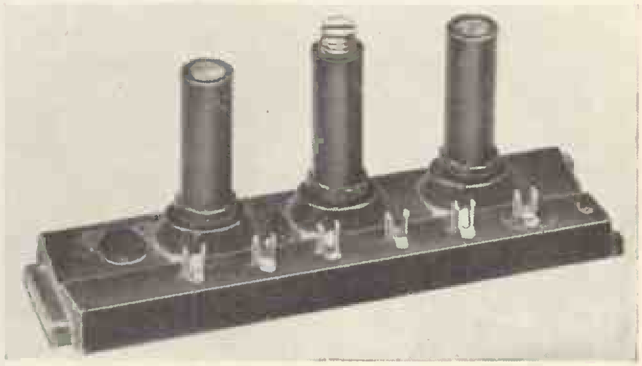


Fig. 17. Iron-dust cores employing silicone core-retaining compound
[By courtesy of E.K. Cole Ltd.

Conclusion

This article has demonstrated that silicones, being chemical hybrids that bridge the gap between organic and inorganic chemistry, possess properties, which considered individually, are associated with either the organic or inorganic field. Taken collectively these properties are unique. Furthermore, silicones are produced in a wide variety of useful forms, and the scope for further development seems unlimited.

This survey by no means tells the full story of how the electronic industry is making use of silicones. The examples chosen serve to illustrate that their significance is becoming widely appreciated at a time when the need has never been greater for equipment to meet new standards in performance and reliability.

It should not be thought that silicones can always replace conventional materials. For instance, silicone rubber does not equal organic rubbers in certain physical properties at normal temperatures. Rather, it can be regarded as an addition to the range of rubber-like materials extending the temperature limits very considerably at both ends of the scale. While the price of silicones has dropped considerably since they were first marketed in America, they are inherently expensive to produce. Often, however, their use is economically justified because simplification in design is made possible.

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Principles and Application of

Electronic Analogue Computers

(Part 1)

By P. Heggs*, A.M.I.E.E.

A series of articles designed to give an introduction to analogue computers and their use. Simple examples of types of analogues are given, together with a discussion of equipment employed. The process of scaling a problem is described and schematic diagrams are given illustrating simple computer set-ups.

Design considerations for universal analogue computers are presented.

ANALOGUE computation has been in use for a number of years, but it is only comparatively recently that analogue methods have become increasingly employed. Simple analogues were built as the need arose and if the importance of a particular problem warranted their construction. Many examples of these may be found in the literature on servo systems and electrical technology.

Solutions of the differential equations of motion of dynamical systems have always been required, but until the advent of the electrical analogue, electronic analogue, or the analogue computer, these solutions were only obtained at the expense of much computing labour.

This meant that only the most important solutions were obtained since, in most cases, variation of a single parameter in the equations meant that an entirely new set of calculations had to be undertaken.

Reference has been made above to the electrical and electronic analogue, and analogue computer. A few words here on definitions might not be entirely out of place.

An electrical analogue is considered as a system of passive networks, by which mass in a mechanical system is represented by electrical inductance, mechanical resistance (viscous friction) by electrical resistance, and mechanical compliance by electrical capacitance.

The mechanical system thus simulated may be either rectilineal, rotational or a combination of the two.

An electronic analogue may be considered as a special case of the universal analogue computer, or simulator, although this latter word is apt to be a little misleading, since analogue computer and simulator are now almost synonymous. It is the writer's opinion that the term analogue computer should refer to a machine which allows any dynamical system to be solved, either by specifying a series of differential equations, or a set of transfer functions, while the use of the word simulator should be confined to a machine built to solve a specific class of problems, e.g. aircraft wing flutter and heat conduction.

Direct electrical analogues can yield very useful results, but in general much ingenuity is required to work out their configuration, especially when the mechanical or other system is complicated. To illustrate the principle of the electrical analogue, where each quantity in a mechanical system has a direct correspondence in the electrical system, the following example will be used.

Fig. 1 shows a mechanical system having two degrees of freedom, represented by the angular rotation of shafts (1) and (2).

The system consists of inertia, viscous friction, and pendulous mass, coupled by a spring. ϕ and θ are the angular rotations of shafts (1) and (2) respectively, while the symbols I, f, k, are the inertia, torque per unit velocity, and restoring force per unit angle of twist. Masses m_1 and m_2 are at radius r_1 and r_2 from shafts (1) and (2) respectively.

It is assumed that the problem is to determine the motion of shaft (2) when a given torque is applied to shaft (1); or alternatively to rotate shafts (1) and (2) from their equilibrium position, and then determine the motion which follows when the system is free from restraint.

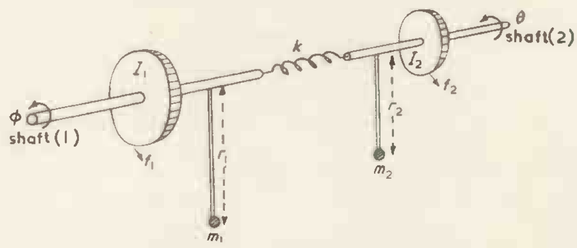

Fig. 1. Example of a coupled mechanical system

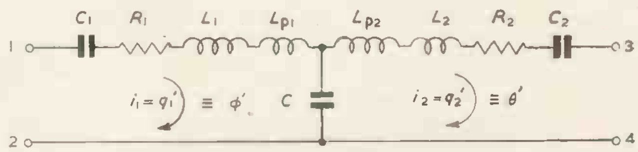

Fig. 2. Passive network analogue of mechanical system in Fig. 1

It can be shown that the equations of motion for small oscillations are:

$$(m_1r_1^2+I_1)(d^2\phi/dt^2)+f_1(d\phi/dt)+(k+m_1gr_1)\phi-k\theta=0 \quad .. \ (1)$$

$$(m_2r_2^2+I_2)(d^2\theta/dt^2)+f_2(d\theta/dt)+(k+m_2gr_1)\theta-k\phi=0 \quad .. \ (2)$$

for the torque free case; or if a torque is applied then the left hand side of equation (1) becomes $T_\phi(t)$.

The electrical analogue of this system is shown in Fig. 2.

Here the two inertias I_1 and I_2 are represented by L_1 and L_2, the spring coupling by C and the two pendulous masses m_1 and m_2, at radius r_1 and r_2, by the combination C_1L_{p1}, C_2L_{p2}. Application of Maxwells equations for cyclic currents shows that the circuit equations of Fig. 2 are:

$$(L_{p1}+L_1)\frac{d^2q_1}{dt^2}+R_1\frac{dq_1}{dt}+(S+S_1)\,q_1-Sq_2=0 \quad .... \ (3)$$

$$(L_{p2}+L_2)\frac{d^2q_2}{dt^2}+R_2\frac{dq_2}{dt}+(S+S_2)\,q_2-Sq_1=0 \quad .... \ (4)$$

if the terminals 1 and 2, 3 and 4 are considered to be shorted. In the above equations $S=(1/C)$, $S_1=(1/C_1)$, $S_2=(1/C_2)$. The initial conditions require specifying; and the other symbols have their usual electrical meaning.

Taking $q_1 \equiv \phi$, $q_2 \equiv \theta$, these equations are of identical form.

* Canadian Westinghouse Co. Ltd., formerly Saunders-Roe Ltd.

The introduction of the pendulous masses, however, has precluded a direct electrical analogue, since length is not represented. Without these masses, which were intentionally introduced into the example, the analogue would have been direct. Here the inertia $m_1 r_1^2$, represented by L_{p1} and C_1, would be chosen to give the natural frequency of the pendulum of length r_1 in conjunction with L_{p1}.

To solve the system, an e.m.f. from a low impedance source is applied to terminals 1 and 2, while 3 and 4 are shorted. The form of the e.m.f. is made equivalent to the form of the desired torque applied to the mechanical circuit. The voltages across any element in the circuit may be recorded, and a graphical integration of the voltage across R_2 would give the displacement $\theta(t)$, when divided by the value of R_2.

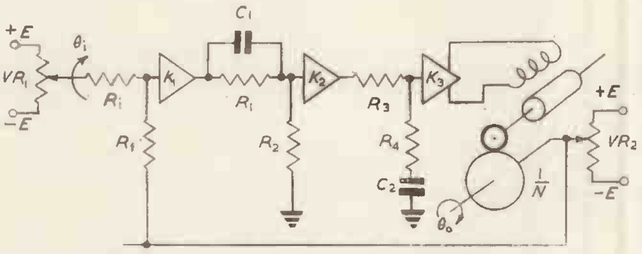

Fig. 3. Position following servo-mechanism

teristics to be varied at will, together with the constants of the phase advance and integral control circuits.

Briefly the potentiometer wiper of VR_1 is coupled to the input shaft, which is rotated through an angle θ_1 and in doing so applies a voltage increment $2E(\theta_i/\theta_m)$ to the input resistance R_i (assuming $R_i \gg R_p$ the potentiometer resistance). θ_m is the maximum angle of wiper travel.

This voltage increment at once unbalances the current in the motor field coils causing the output shaft to drive the wiper of potentiometer VR_1 through the angle θ_o which will tend to equality with θ_i provided that the feedback resistance R_f is equal to the input resistance R_i.

The limiting error due to static friction may be minimized by including the network $R_3R_4C_2$. While the system is in motion C_2 builds up a voltage which tends to keep the output shaft rotating when the input and output displacements approach equality.

The modification to the response and stability of the system due to a variation in the values of the phase advance components C_1, R_1, R_2 and the integral control components R_3, R_4, C_2 may be easily ascertained. The investigation may be carried out by the use of a pen recorder, or a cathode-ray oscilloscope, the latter method being preferable if the time scale of the simulation apparatus is a multiple of that of the physical system.

Since in general it is desirable to know both the response

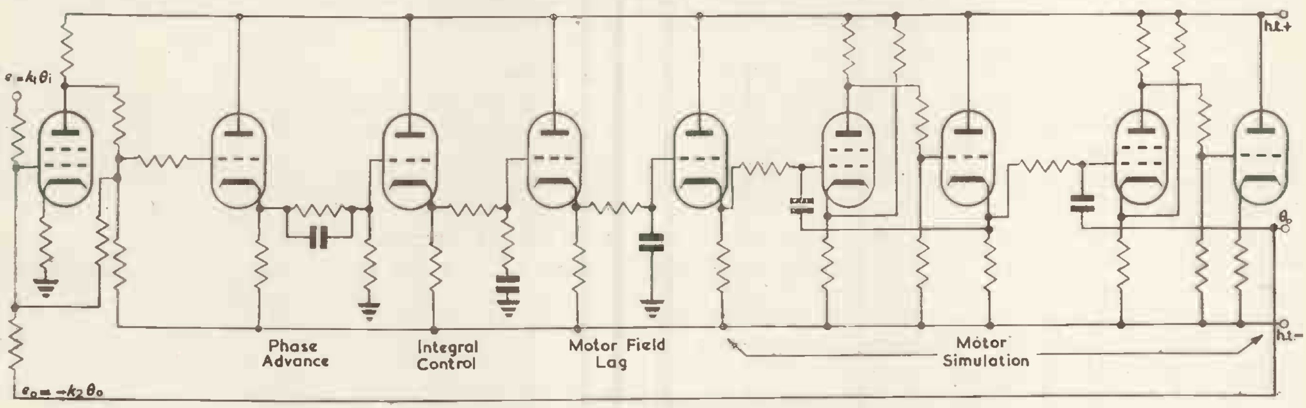

Fig. 4. Skeleton circuit of electronic simulator for the servo-mechanism of Fig. 3

Since $E_{R_2} = i_2 R_2$, $i_2 = q_2/dt = (E_{R_2}/R_2)$

E_{R_2} being the p.d. across R_2.

Hence $\theta(t) \equiv q_2 = \int E_{R_2}\, dt/R_1$ and in general a scale factor k would be involved.

It is obvious that these methods, although instructive, are not very practical.

A type of analogue which has been employed to advantage in servo-mechanism studies is shown in Fig. 4.

In its simple form single valve stages are used as integrators, which causes the simulator to be less accurate than one making use of multi-stage amplifiers.

The analogue of a position following servo-mechanism is shown.

No inductances are used, but the analogy now appears between the transfer functions of the mechanical and electrical systems.

Fig. 3 shows a position follower servo-mechanism employing phase advance stabilization and integral control, the system being driven by a split field d.c. motor M which drives the output shaft through gearing.

Here the mechanical equipment, whose characteristics will be known, is represented by the motor and gear train to the output shaft.

The simulator enables the motor and gear train charac-

of a system to a suddenly applied input or step, and the response to a constant amplitude sinusoid of varying frequency, the electronic simulator working at a fast time scale is an extremely versatile design tool.

Further the effect of torque limitations in the servo motor may be reproduced by diodes backed off via low impedances. These will be discussed later in relation to general purpose analogue machines.

The basic circuit of an electronic simulator for the position controller of Fig. 3 is shown in Fig. 4.

The equation of motion of the motor whose armature is supplied with a constant current is:

$$(I_m d^2\theta_m/dt^2) = K_m i$$

where I_m, K_m, i, are the inertia of the armature, torque per unit field current difference and field current difference, respectively.

If N is the gear ratio between the motor shaft and the potentiometer wiper $\theta_o = (\theta_m/N)$.

and:

$$I_m N(d^2\theta_o/dt^2) = K_m i, \quad I_m N p^2\theta_o = K_m i,$$

$$(\theta_o/i) = (K_m/I_m N p^2)$$

This last expression is known as the motor transfer function showing that a double integration takes place from the motor input to motor output. The differential current i is related to the input error by the electronic gain constants K_1, K_2.

Thus the overall system transfer function becomes:

$$(\theta_o/\epsilon) = \frac{K_1 K_2 K_m}{I_m N p^2} G_1(p)\ G_2(p)\ G_3(p) \quad \ldots\ldots\ (5)$$

$G_1(p)$, $G_2(p)$, $G_3(p)$ being the transfer functions of the phase advance, integral control and motor field lag, respectively. Suitable choice of constants in Fig. 4 making use of the fact that the transfer function of an integrator from the relation $e_o = (1/RC)\int e_i dt$, is:

$$(e_o/e_i) = (1/pRC) = (K/p)$$

enables an analogous equation to (5) to be obtained.

Much work using simulators of this nature was originally done by Prof. F. C. Williams and F. J. U. Ritzon. These simulators may be regarded as the forerunners of the versatile general purpose analogue computers in use today, the design and application of which will be discussed in the next two articles.

(To be continued)

An Electrostatic Pulse Generator †

By W. Woods-Hill*

This article describes a means of generating pulses locked to a drum without the need for electromagnetic pick-up heads.

In electronic calculators numerous and complex pulse patterns are required to ensure correct operation of the arithmetic units and associated circuits, and these are often generated by valve circuits and synchronized to the magnetic storage drum by means of a single "clock" track engraved or recorded on the drum. Attempts have also been made to generate these pulse patterns by having numerous magnetic pick-up heads, one for each separate pulse stream, and engraving or recording the required pattern on the drum.

The use of magnetic pick-up heads has several drawbacks, not the least being the cost of the somewhat complex and accurately machined parts required for each head. Another is that the output of the head is low, and needs several stages of high gain amplification before the pulses have sufficient amplitude to be effective. A third and sometimes the biggest drawback, is that the output depends upon the rate of change of flux so that when a long engraving is sensed an impulse is obtained only at the beginning and end. The missing d.c. component must be re-introduced with an Eccles-Jordan trigger or similar device.

Another annoying effect, but not so markedly felt in the computer field, is that the output falls off as the speed of rotation is reduced.

In an attempt to overcome these drawbacks the following means were devised and an elementary experiment carried out to prove its practicability.

Electrostatic Probe

If the metallic drum was raised to a high potential and a screened probe, such as depicted in Fig. 1, was approached to the surface, then the configurations on the drum surface would induce corresponding voltages in the probe. The output from such an arrangement is rather low and the voltage requirements somewhat high.

The impedance of the probe circuit is high, very prone to pick up interference and hum from a.c. circuits. If, however, the drum is fed with a radio frequency voltage, then a tuned circuit resonant at that radio frequency can be introduced between the probe and earth, effectively reducing this impedance to a low value for all signals except the wanted one.

Experiment

Fig. 2 shows how the drum was fed with radio frequency at 1Mc/s via a coupling loop (L) and resonant circuit (C) mounted concentrically with the shaft. The shaft itself was earthed via the electric motor bearing, but the drum was mounted on bakelite bushes and thus insulated except for the connexion via the parallel resonant coil.

When r.f. was fed to the fixed coupling loop (L) the drum assumed a high r.f. potential as it is effectively one plate of the parallel tuning capacitor, the earth shaft being the other.

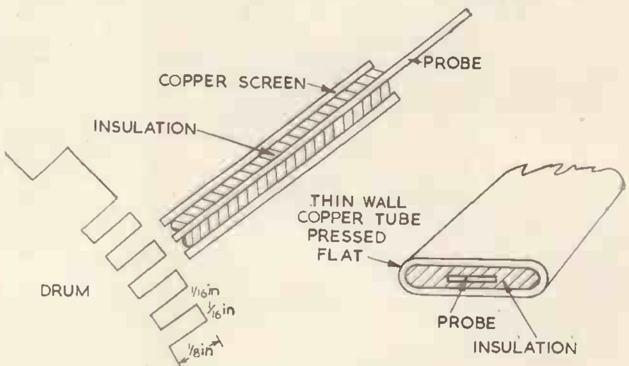

Fig. 1. The screened probe

This radio frequency was picked up by a metal probe which was approached as close as possible to the periphery, but not in contact. The probe was connected to an amplifier-detector.

Fig. 3 shows the circuit of the r.f. generator and the detector. V_1 is a 6V6 wired as a Hartley Oscillator with L_1C_1 resonating at 1Mc/s. Link coupling to the drum coil was by 80Ω coaxial cable.

V_2 is a high slope pentode wired as an amplifier detector. The probe was connected to the grid of V_2 by coaxial cable and the tuned circuit L_2C_2 was adjusted to resonate

† *Patent No. 703,038, W. Woods-Hill, Assigned to British Tabulating Machine Company Limited.*
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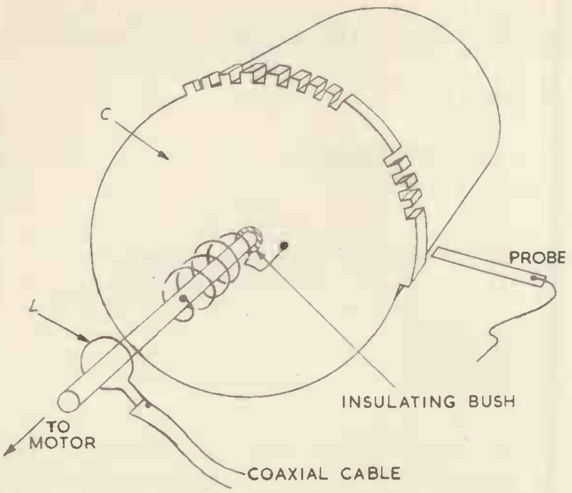


Fig. 2. Method of feeding drum with radio frequency

at 1Mc/s after the valve was plugged in and the probe connected up. This effectively "tuned out" the grid to cathode capacitances which tended to by-pass the r.f.

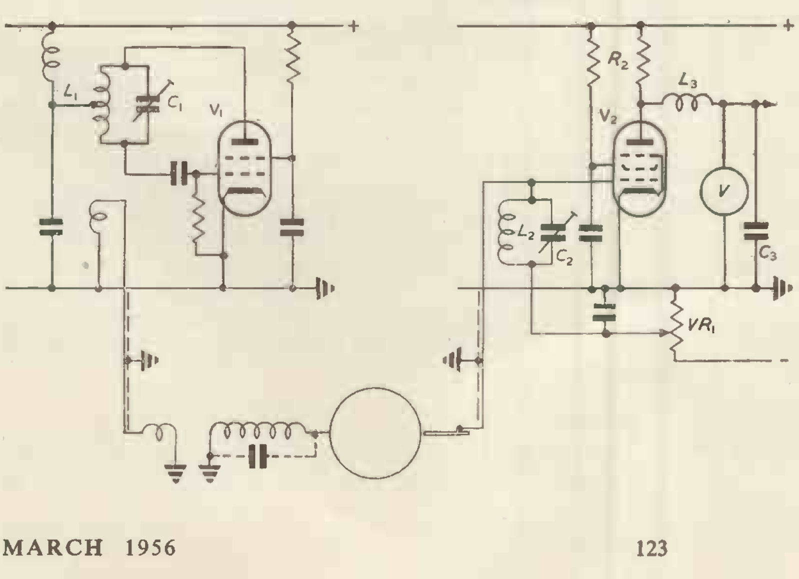
The negative bias on the grid of V_2 was adjusted by means of VR_1 in such a way that when the drum was stationary and the probe was opposite a "trough" the r.f. picked up by the probe plus any stray pick-up was just insufficient to cause the current to flow on r.f. peaks in the anode circuit of V_2. Rotation of the drum so that a "pip" was under the probe, increased the capacitive coupling and hence the amplitude of r.f. reaching the grid, and V_2 now drew half-waves of r.f. current. The voltage developed across R_2, the anode load, was filtered by C_3L_3 and the d.c. component measured at "V".

A study of the input circuit to the amplifier-detector of Fig.3 shows that the coaxial cable connecting the probe to the input of V_2 is capacitively loading this tuned circuit (L_2C_2).

This capacitance will increase with the length of coaxial cable and if made too long may force the resonant frequency below 1Mc/s, despite adjustment of C_2.

To overcome this, and allow V_2 to be at a remote point from the drum, the alternate circuit of Fig. 4 can be used, where links L_1 and L_2 provide an impedance match to the characteristic output and input impedance of the coaxial cable.

Fig. 3. The r.f. generator and detector



Results

With the motor running at 3 000 rev/min and a drum configuration consisting of teeth 1/16in wide, 1/16in thick and 1/16in gaps ⅛in deep, the pulse repetition rate was 5kc/s (see Fig. 1). The pulse amplitude measured with an oscilloscope at the anode of V_2 was 55V with an h.t. voltage supply of 200V and R_2 100kΩ. The r.f. amplitude applied to the drum was 100V peak-to-peak.

Reduction of drum speed down to near zero-revolution per minute produced no noticeable reduction in the pulse amplitude.

At these low running speeds a relay was substituted for R_2 and its armature followed the configurations on the drum down to zero revolution per minute.

Conclusion

More development is needed to establish what the limits to the definition of such a device are and how the output will fall as the pitch of the teeth is reduced, but at the p.r.f. at which it was tested the following points can be noted.

(1) Low cost of pick-up device.

(2) High output.

(3) Low impedance.

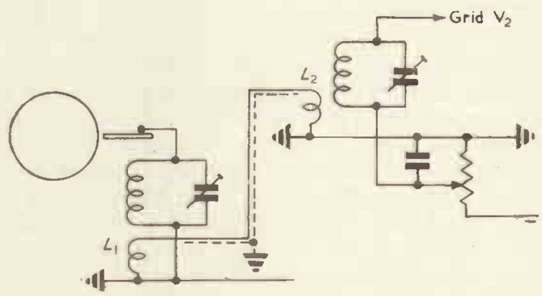


Fig. 4 Alternative coupling circuit

(4) Output constant at all drum speeds down to zero.

(5) Because of (4) the d.c. component is preserved.

Against this must be set:

(1) Drum must be insulated.

(2) Radio frequency in close proximity to computer circuits (Reading amplifiers) will mean a 1Mc/s filter required in certain circuits.

A use for such a principle could possibly be found in allied fields. For instance, capacitive microphones could be made with a low impedance. Such a microphone might be adapted comparatively easily if used with a tape recorder because a r.f. oscillator is already available in the equipment. Another use may be in electronic organs with disk type tone wheels, or, in fact, anything where electrostatic or magnetic pick-up is employed, might benefit from this lower output impedance.

Acknowledgments

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Impedance-Frequency Characteristics of some Glow-Discharge Tubes

By F. A. Benson*, M.Eng., Ph.D., A.M.I.E.E., M.I.R.E., and G. Mayo*, B.Eng.

The variations of impedance with frequency of eight types of glow-discharge, voltage-regulator tube have been determined over a fairly wide frequency range. The impedances of all the tubes increase considerably with frequency. Resistance-frequency and inductance-frequency curves have been plotted for two of the tube types. Each inductance-frequency curve shows a pronounced maximum at low frequencies. It seems that the kind of gas filling is one important factor in determining the shape of the impedance-frequency characteristic of a tube.

A GOOD deal of research has been carried out recently to determine the behaviour of glow-discharge, voltage-regulator tubes when an alternating current is superimposed on the direct tube current. The results of the work show that the a.c. impedance of a glow-discharge tube at high frequencies is considerably greater than it is at low frequencies. Andrew[1] has drawn attention to this fact and has described briefly a few measurements carried out on 7475 and 85A1 tubes in the frequency range 50c/s to 20kc/s. Benson[2] has given a graph, taken from a previous publication[3], showing that the impedance of stabilovolts increases with frequency, although the upper frequency limit of the graph is only 6kc/s. Other interesting articles on the subject have been published by Hunt[4], Williams[5] and the authors[6]. It has been shown that tubes have surprisingly high values of apparent inductance and appreciable values of effective resistance. The impedance and resistance of a given tube both increase considerably with frequency and the inductance-frequency curve of a tube shows a pronounced maximum in the region of 100c/s.

The authors have now investigated, in detail, the variation of impedance with frequency of a number of other types of glow-discharge tubes and, where possible, the nature of the impedance has also been determined. The results of the work are presented here and discussed.

Measurements

The impedance of each tube was measured over a fairly wide frequency range using the circuit of Fig. 1. The lower limit of the frequency range was 20c/s, while the upper limit varied from 10 to 50kc/s. The test circuit is identical with that used previously[6]. An audio-frequency oscillator supplies a voltage across a non-inductive resistor R_1 and the glow-discharge tube under test. The a.c. through the tube is measured by recording the voltage developed across R_1 (i.e. V_1) by means of a valve voltmeter VV_1. The value of the dropping resistor R_2 was made large so that it produced negligible shunting effect across the tube. An electronically-stabilized power supply was used to provide the d.c. supply. The d.c. through the tube could be maintained constant at any desired value by means of an output-voltage control on the power-supply unit. The alternating voltage (V_2) across the tube under test was measured with a valve millivoltmeter.

Measurements have been made on several tubes of each of the following types: 90C1, 150B2, QS70/20, QS75/60, QS83/3, QS92/10, QS105/45, and QS150/15. In each case readings of VV_1 and VV_2 (i.e. V_1 and V_2) were taken over the frequency range, with a constant alternating voltage across the tube. This voltage was chosen carefully to give a reasonable swing about the d.c. operating point which was set near the middle of the current range. The swing was such that deionization did not occur on negative half cycles. The impedance Z of a tube is calculated from the expression:

$$Z = (V_2/I) = (V_2R_1/V_1)$$

where I is the a.c. through the tube.

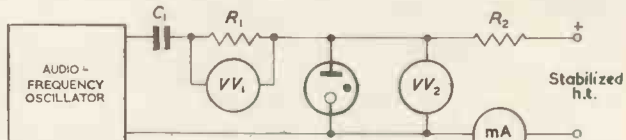


Fig. 1. Circuit used for impedance measurements

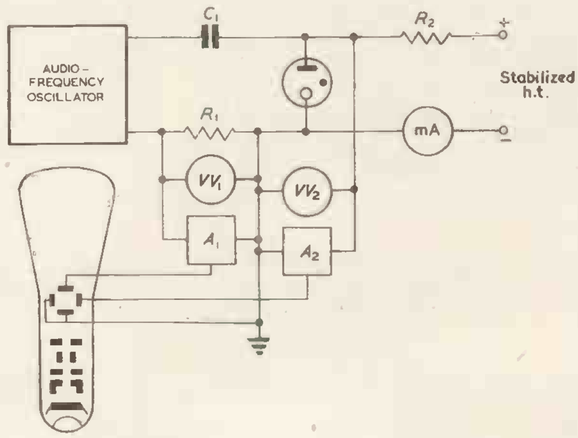


Fig. 2. Circuit used for phase-angle measurements

It was then desired to find the resistive and reactive components of the tube impedances. One suitable way of doing this, involving a measure of the phase-angle existing between voltages V_1 and V_2, has already been described[6]. The circuit used is shown again, for convenience, in Fig. 2. Essentially the method consists of applying the two voltages V_1 and V_2 to a cathode-ray oscillograph, the resulting ellipse traced out on the screen giving a measure of the phase angle[7,8]. To obtain zero phase-shift in the measuring equipment, at all frequencies used, similar amplifiers from two identical oscillographs were employed and then the X and Y plates of one oscillograph were used. In many cases it was impossible to obtain the resistive and reactive components of the impedance by this method because of severe distortions of the expected ellipses caused by rather peculiar characteristics of the tubes.

Results

Mean impedance-frequency curves for the various types of tube are shown in Figs. 3, 4, 5 and 6. Because of the

* *The University of Sheffield.*

large number of readings taken, only the mean curves have been drawn for each type of tube to avoid confusion. The mean impedance-frequency curves have all been plotted together on Fig. 7 for easy comparison. Typical inductance-frequency and resistance-frequency curves for tube types QS70/20 and 90C1 are plotted on Figs. 8 and 9.

Discussion of Results

The characteristics of Figs. 3 to 6 confirm the previous findings that the impedance of a glow-discharge tube increases a good deal with frequency. There are considerable variations in the impedance curves from tube to tube

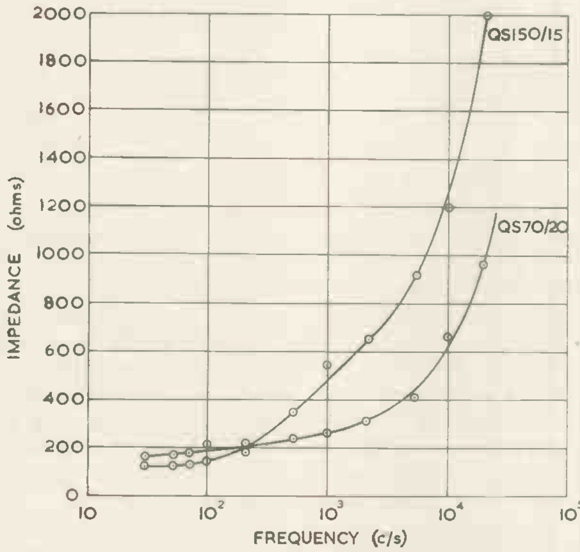


Fig. 3. Impedance-frequency curves for types QS150/15 and QS70/20

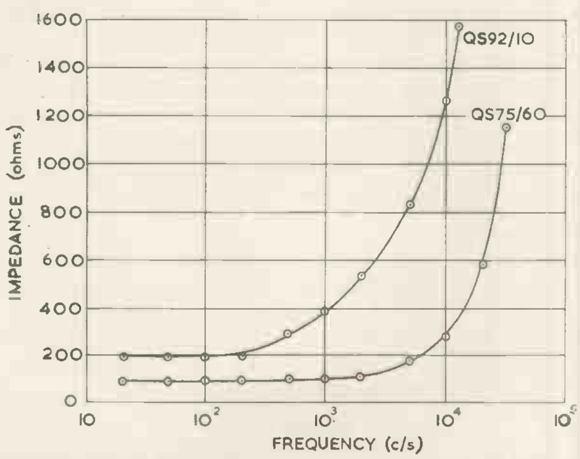


Fig. 4. Impedance-frequency curves for types QS92/10 and QS75/60

even of the same design, as might be expected from results of investigations on other tube characteristics[9].

It is interesting to see from Fig. 7 that the three types of tube QS70/20, QS75/60 and 90C1 have similar characteristics. In these cases the tube impedance does not change much in the frequency range 20 to 1 000c/s, but above 1 000c/s it increases fairly rapidly. It is known that these three types of tube all contain the same kind of gas and it seems, therefore, that the gas content influences the shape of the impedance-frequency curve. This might have been anticipated because marked differences between the running-voltage/current characteristics of tubes containing neon and

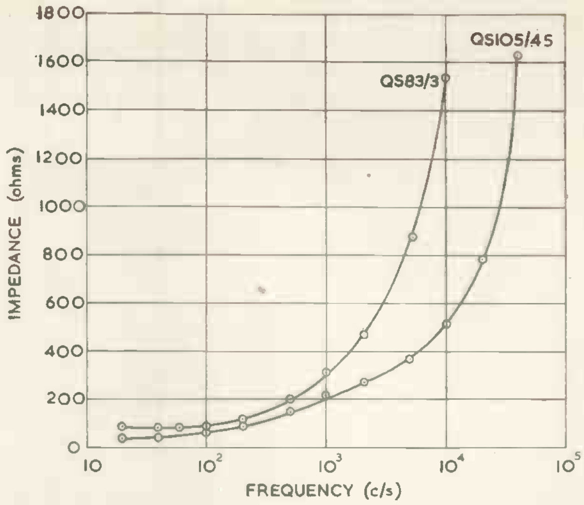
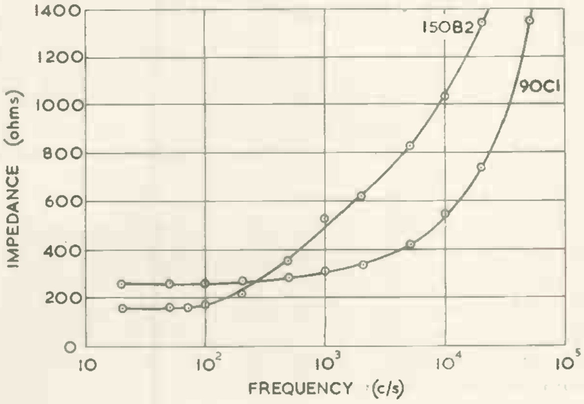


Fig. 5. Impedance-frequency curves for types QS83/3 and QS105/45

those containing helium have already been reported[10]. It has also been found[11] that the gas filling influences greatly the shape of the running-voltage/temperature curve of a tube and is an imporant factor in deciding whether a given tube will show a positive or a negative temperature coefficient of running voltage. There is also some evidence now that the kind of gas filling is one important factor in governing the duration, magnitude and direction of the initial drift of running voltage which takes place after switching on a tube. It is clear from Fig. 7, however, that the gas filling is not the only important factor which decides the shape of the impedance-frequency curve of a tube because QS83/3 tubes also contain the same kind of gas as the other three mentioned above. The impedance of the QS83/3 tubes, however, begins to increase fairly rapidly at a frequency of about 100c/s. The tube types 150B2, QS105/45 and QS150/15 all contain the same kind of gas (different from that in the tubes discussed above) and here the impedance begins to increase quite rapidly at the low end of the frequency range.

Figs. 8 and 9 show that the inductance-frequency and resistance-frequency curves for the tube types QS70/20 and 90C1 are of the same general forms and similar to those previously published[6] for tube types 7475, 85A1, 85A2 and S130. The peak values of the inductances for the 7475, 85A1 and S130 tubes were found to be 110, 70, 55 and 35mH respectively and the peaks occurred at frequencies between 80 and 100c/s[6]. The peak values of the inductances for the QS70/20 and 90C1 tubes are quite high, i.e. 185mH and 205mH respectively and the maximum in

Fig. 6. Impedance-frequency curves for types 150B2 and 90C1



the inductance-frequency curve of the 90C1 occurs at a frequency of about 40c/s which is unexpectedly low.

The inductance-frequency and resistance-frequency curves could not be obtained for all the tube types because the phase-angle method of measurement used was not suitable. The phase-angle ellipses were very distorted in many cases because of hysteresis and rapid jumps in the running-voltage/current characteristics. Previously-published oscillograms[10] showing the differences in the shapes of the running-voltage/current characteristics of the 150B2 and 90C1 tubes indicate clearly the difficulties which are encountered with this type of measurement.

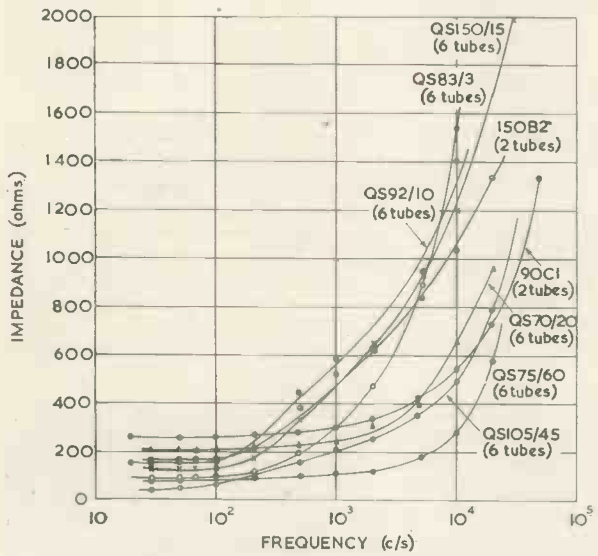


Fig. 7. Mean impedance-frequency curves for all tubes measured

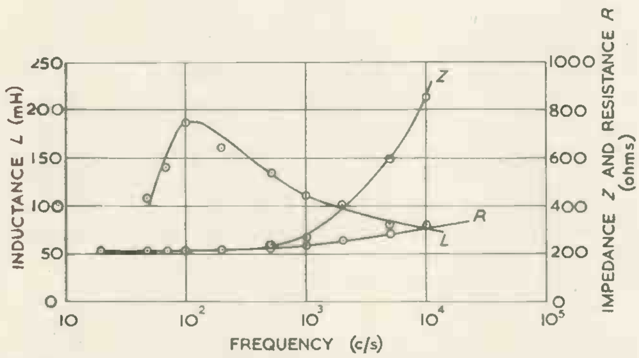


Fig. 8. Inductance, impedance and resistance curves for a QS70/20 tube with frequency

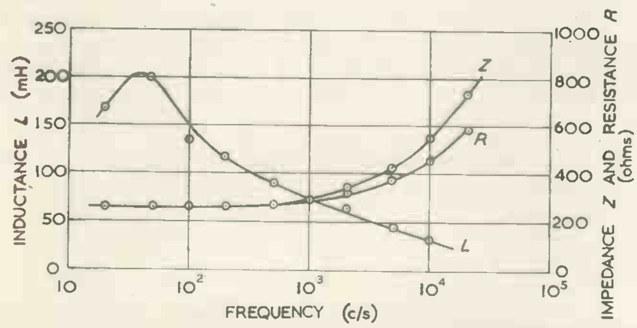


Fig. 9. Inductance, impedance and resistance curves for a 90C1 tube with frequency

For the 90C1 and QS70/20 tubes the resistance increases with frequency from a value of about 50 to 70Ω at 20c/s to several hundred ohms at 20kc/s. There are large variations in the resistance-frequency curves from type to type of tube and it is interesting to compare Figs. 8 and 9 with Table 1, which indicates typical approximate changes of resistance with frequency for the 85A1, 85A2, 7475 and S130 tubes[6].

TABLE 1.

Typical Approximate Changes of Resistance with Frequency for Various Types of Tube.

TUBE TYPE	RESISTANCE IN OHMS AT 20c/s	RESISTANCE IN OHMS AT 10kc/s
85A1	250	800
85A2	350	1 000
7475	300	1 000
S130	50	225

The inductance of a tube may be attributed to the effects of ion inertia. The fact that the inductance is a maximum at a particular frequency seems to indicate that a mechanical resonance of the ions occurs at this frequency. Williams[5] has discussed the origin of the quadrature component in a cold-cathode tube and gives an equation which illustrates a close analogy, at frequencies less than 5kc/s, between a one-dimensional electron-dynamic system and an electrical network of inductance and resistance. He goes on to say, however, that this is not a complete answer to the question of how can the quadrature current arise, but at least it gives a clue to the origin of this current.

A glow-discharge tube is commonly assumed to be equivalent to a constant voltage in series with a resistor. The work reported here emphasises the importance of ion-inertia effects in the gas and, as Hunt[4] suggested, the equivalent circuit should consist of a constant voltage in series with a resistor and an inductor.

It should be pointed out that the measurements described in this article were made prior to April 1954. Additional work concerned with the influence of some tube parameters on impedance-frequency characteristics has since been carried out[12]. The equivalent circuit of a glow-discharge tube has also been further discussed[12].

Acknowledgments

The authors wish to thank Mr. O. I. Butler, M.Sc., M.I.E.E., A.M.I.Mech.E. for laboratory facilities in the Department of Electrical Engineering at the University of Sheffield. They also wish to acknowledge the kindness of the English Electric Valve Co. Ltd, and Mullard Ltd, in supplying some of the tubes for examination.

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Batch Production of Electronic Equipment

A considerable amount of electronic equipment is produced in small batches which do not readily lend themselves to production organization on flow principles. This is particularly so in the case of chassis production, and in this connexion a turret punch press with pantograph template table is an admirable tool. Fig. 1 shows a Weidemann type RA.41.P press installed at the factory of Mullard Equipment Ltd and Fig. 2 shows parts of the chassis of the Mullard L101 oscilloscope which is produced on this machine.

The RA.41.P is a 15 ton punch in which between 16 to 20 separate punches are located in a rotating turret, each of which can be brought into the operating position as required. The metal plate to be punched is held by quick acting clamps and is moved into the correct position for each hole, by the pantograph system. The pantograph operating handle carries a stylus pin which engages in a hole in a template, (see Fig. 2) to give the correct positioning of each individual hole. By this arrangement co-ordinate dimensions for hole centres, can be accurately and quickly reproduced from the template. When the stylus pin is lowered into the appropriate hole in the template a micro-switch is actuated which in turn operates the press clutch mechanism, thereby effecting a blow of the press. The downward movement of the stylus is pre-set, to ensure that the press clutch mechanism can only be actuated when the stylus pin has been correctly located in a hole in the template, and the depth of entry has been sufficient to produce accurate location. The template is normally made from ⅛in thick steel plate and all the stylus location holes are 0·250in diameter. The accuracy of the hole centres in the template can normally be reproduced on the work piece to within ± ·003in. A simple system of colour coding on the template to coincide with the colour tabs on each of the turrret stations ensures simplicity of operation. The most efficient operation sequence is for the operator to index the turret to a particular hole size and to then pierce all such holes consecutively, by scanning the template in the direction indicated by the coloured lines connecting the template holes, also similarly coloured.

Fig. 1. A Weidemann turret punch press in use

The manufacture of templates can, if required, be performed on the press itself, with the supplementary equipment available, consisting of precision scales and micro-scopes. Alternatively, conventional equipment such as a layout drilling machine with co-ordinate table, or a jig boring machine, readily produces these. In instances where the accuracy of hole location is not so important normal marking-out methods and drilling are adequate.

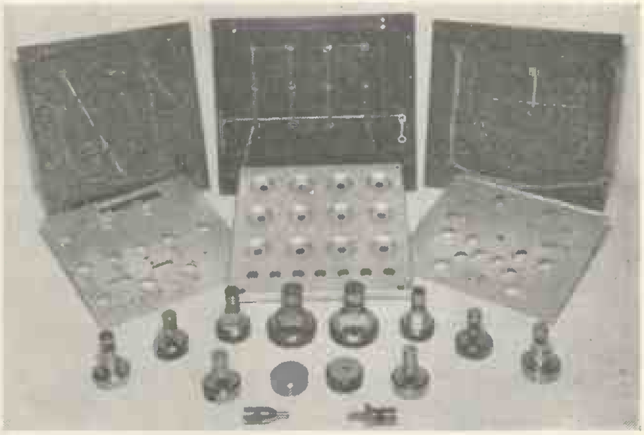

Fig. 2. Templates (top) and some of the tools (bottom) used for producing the three panels (centre)

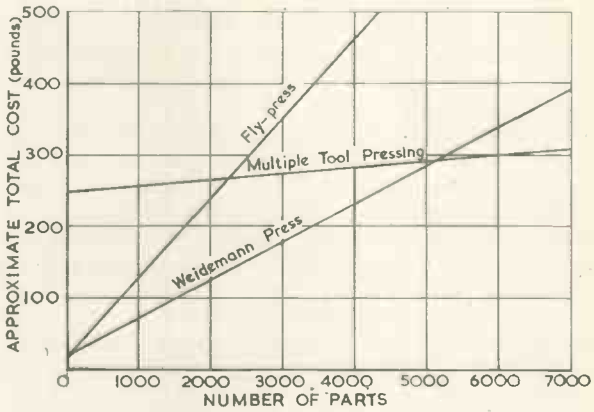

Fig. 3. Comparison of costs between turret press, fly press and multiple tooling

When the template making attachment is fitted to the machine, the manufacture of single workpieces may be performed by direct measurement of hole centres, dispensing with the need to make a template.

A comparison of costs with two other methods, namely, Fly Press and Multiple Tooling, may be of interest, and on a typical study for the manufacture of a chassis having 114 holes spread over twelve different sizes, the cost figures are as shown in Fig. 3, the part time being 4½ minutes.

From this it will be seen that the turrett press method is considerably more economical up to 5 000, with the additional important advantage—the facility to commence production of new designs with the minimum of delay.

Wiedemann Turret Punch Presses are British-built to the order of Dowding & Doll Ltd, London.

Notes from ——

NORTH AMERICA

IRE News

The IRE Board of Directors, at its January meeting, appointed six members to the Board for 1956.

Reappointed as Treasurer of the IRE was W.R.G. Baker, Vice President of Electronics, General Electric Company. Haraden Pratt was appointed to his fourteenth term as IRE Secretary. Donald G. Fink, Director of Research of the Philco Corp. and former Editor of *Electronics*, was appointed Editor of the IRE to succeed John R.Pierce, Director of Electronics Research of Bell Telephone Laboratories, Inc.

Appointed as Directors were Alfred N. Goldsmith, Consulting Engineer and Editor Emeritus of the IRE; T.A. Hunter, President of the Hunter Manufacturing Co. and Editor of the *IRE Student Quarterly*; and J.W. McRae, President of the Sandia Corp. and past President of IRE.

Advanced Studies at M.I.T.

The establishment of a school for advanced study at Massachusetts Institute of Technology has been announced recently.

The new school will provide means by which post-doctoral scholars from all over the world can join with the M.I.T. faculty in high-level theoretical studies and research. Dr. Martin J. Buerger, professor of mineralogy and crystallography, has been appointed director. In its initial embodiment the school will be merely an organizational entity, but it is hoped ultimately to provide a centre and adequate housing for fellows and guests and 'by this means to gain the advantage of cross stimulation of ideas which occurs when learned and ingenious men are brought together into close social contact'.

Vibration Reduction

A simple electronic device which can isolate vibrating machines or cut down their vibration has been built by the Radio Corporation of America.

Dr. Olson described the novel device as an electronic means of turning vibration against itself through a detecting and amplifying system which responds instantaneously to each motion of the vibrating machine, either by creating an opposite force or by absorbing the motion. He pointed out that its operating principles are similar to those of the RCA electronic sound absorber, developed earlier as a means of reducing heavy, low-frequency noises over a limited area in automobiles, aeroplanes, machine shops, and offices.

The reducer itself is a small, box-like assembly containing a lever system upon which the machine rests. In a normal installation, several of these units would be placed beneath the machine to isolate it from the floor.

The device can be used either to reduce the vibration of the machine itself, or simply to prevent the vibration from passing to the floor. In operation, each vibration of the machine produces a downward pressure on the reducer assembly. This pressure is translated into an electrical impulse which passes from the reducer through a separate amplifier unit.

In cases where the vibration of the machine itself is to be reduced, the amplified signal is used by the vibration reducer to exert an upward pressure through the lever system, substantially reducing or entirely counteracting the motion of the machine. If the purpose is only to isolate the vibration, the reducer is adjusted to absorb the motion through its lever system so that the continuing vibration of the machine is prevented from passing to the floor. Since the system is electronic, the entire process in both cases is carried out almost instantaneously, permitting the reducer to counteract or absorb vibrations at speeds of thousands of times per minute.

High Voltage Engineering

A new 3MV Van de Graaff particle accelerator, said to be the most powerful and versatile machine radiation source now commercially available, has been announced by High Voltage Engineering Corporation, Cambridge.

The 9 ton supervoltage generator, several models of which have already been sold to industrial firms in the United States and abroad, is designed to produce nearly all the fundamental radiations—electrons, X-rays, positive ions or neutrons. Conversion of the machine from production of one type of radiation to another is a relatively simple step, achieved through the use of appropriate components manufactured by High Voltage. For example, the accelerator can be changed from electron processing to X-ray production by replacing the beam scanner of the machine with a special heavy metal target 1in in diameter and ⅛in thick.

The unit is designed for either vertical or horizontal mounting, and has a power output rating of 3kW of radiation at 3MV. It is stated that this unit will produce a radiation field several hundred times more intense than the most powerful radioactive cobalt source now in industrial use.

This company has also recently announced that it has under construction a new building which will house what is claimed to be the world's largest radiation machine testing facilities.

The building, which will occupy a 50 acre site in Burlington, Mass., will contain enough radiation test vaults to allow operation of 16 high voltage units simultaneously. (With all the vaults in operation at the same time the radiation produced would be equivalent to that of over 3 million curies of cobalt 60 or more than 12 million curies of cesium 137.) The vaults will be made to accomodate both Van de Graaff particle accelerators and microwave linear accelerators.

Solar-powered Radio Receiver

A pocket-size solar-powered radio receiver, weighing only 10 ounzes and capable of working for some 500 hours in total darkness without recharging, has been developed by The General Electric Co. at Syracuse. The major advantages over solar-powered receivers previously announced are reduced size and length of operation in the absence of light.

Long operation in total darkness is made possible by the use of a miniature storage battery which is contained in a transparent plastic case along with four transistors, seven solar cells and other components. Transparency of the case permits battery charging by simply exposing the unit to the sun's rays.

The General Electric Co. point out that the present high cost of the solar cells precludes their use in commercial equipment and they have, therefore, no immediate plans for production.

Short News Items

The Television Society Annual Exhibition will be held from 6-8 March at The Royal Hotel, Woburn Place, London, W.C.1. Tickets for the use of non-members may be obtained from members of the Society, patron member firms who are exhibiting or from the Exhibition Organizers, Cinema-Television Ltd, Worsley Bridge Road, Lower Sydenham, London, S.E.26.

The Second International Instrument Show will be held at Denison House, Vauxhall Bridge Road, London, S.W.1. from 7-18 May. It will be open daily from 9 a.m. to 7 p.m. (12 to 7 p.m. on 7 May.)

The Annual National Exhibition of French Components, Valves and Test Equipment for the Radio and Electronic Industries will take place in the Parc des Expositions, Porte de Versailles, Paris, from 2-6 March and will be open daily from 9.30. a.m. to 6 p.m.

The Radio Industry Council announces that more British radio equipment was exported in 1955 than ever before. The provisional value was £32·93m, 12·8 per cent more than 1954, which itself was a record, and more than three times the value in 1947.

Ekco Airborne Search Radar is to be installed in the entire BOAC Bristol Britannia fleet and the two Britannias recently delivered to the Corporation are already fitted with this equipment. The main purpose of Ekco Search Radar is to detect turbulent cloud formations which may be in the path of an aircraft. With the aid of this equipment the pilot is given adequate warning of these clouds and is therefore able to take action to avoid them.

Fielden Electronics Ltd of Wythenshawe, Manchester, manufacturers of industrial, laboratory and research electronic instruments, announce that they are freezing both home and export prices of all their products for the next six months.

Pye Ltd, with the help of the Bristol Aeroplane Co, recently carried out a demonstration to show how a system of air-to-ground industrial television could help police to study traffic or carry out searches in difficult country. A "Sycamore" helicopter, with a miniature Pye industrial television camera mounted in its doorway, flew at heights between 500 and 1 000ft transmitting pictures of the countryside to a screen on the ground. Installations at Avonmouth docks, four miles away, were clearly visible on the miniature screen, as well as traffic movement on the Gloucestershire roads.

Pye Ltd also announce that with the installation of a miniature television camera in the Vienna State Opera House a new use has been found for industrial television. The purpose of the installation is to enable the conductor of the choir and the organist to see the conductor of the orchestra. With the aid of the Pye industrial television camera, which is fitted into the left hand side of the prompter's box and completely hidden, the choir master can now watch the conductor very carefully on a monitor screen and so lead the choir without fear of missing his cue. The camera also enables the organist to sit in the organ room and receive his cue from the conductor on a screen.

The Council of The Institution of Electrical Engineers have elected Colonel Sir Arthur Stanley Angwin, K.B.E., D.S.O., M.C., T.D., D.Sc.(Eng.), a Past President of The Institution, to Honorary Membership in recognition of his outstanding life's work in the field of telecommunication, both national and international, and of his distinguished services to The Institution. The Council have also made the thirty-fourth award of the Faraday Medal to Emeritus Professor George William O. Howe, D.Sc., LL.D., M.I.E.E., for his pioneering work in the study and analysis of high frequency oscillations and on the theory of radio propagation; and for his outstanding contributions to engineering education.

The Lord President of the Council has appointed Professor G.B.B.M. Sutherland, Sc.D., F.R.S., Professor of Physics and Director of the Biophysics Research Centre in the University of Michigan, to be Director of the National Physical Laboratory. It is expected that Professor Sutherland will take up the appointment in September next. He will succeed Sir Edward Bullard, Sc.D., F.R.S. who retired on 31 December last. The Appointment of Dr. R.L. Smith-Rose, C.B.E., D.Sc., M.I.E.E., Acting Director has already been announced; he will continue to act until Professor Sutherland takes up duty.

Sir Ben Lockspeiser, K.C.B.,F.R.S., will be retiring from the post of Secretary to the Committee of the Privy Council for Scientific and Industrial Research on 10 March. Her Majesty The Queen has been graciously pleased to approve the appointment in his place of Professor H.W. Melville, F.R.S., who is now Mason Professor of Chemistry at the University of Birmingham. Professor Melville will take up his new appointment in August.

Mr. J.E. Bolton, Chairman and Managing Director of The Solartron Electronic Group Ltd issued with his annual statement a comprehensive book surveying the past, present and future of the Group. Today the Group is trading at the rate of £1 000 000 per year, and announce that the prices of their many electronic instruments will not be increased during the next six months. Solartron Research and Development Ltd, has now completely removed to its new headquarters at Goodwyns Place, Dorking, Surrey, telephone Dorking 4661.

Editions Radio of 9 rue Jacob, Paris 6, will be pleased to despatch copies of their technical journal, *Electronique Industrielle,* free of charge on request.

Mr. J.R. Brinkley has been appointed Managing Director of Pye Telecommunications Ltd, Cambridge, and an Executive Director of Pye Ltd.

Mr. N.B. Balaam has recently been appointed to the Board of 20th Century Electronics Ltd. He joined the firm in 1949 and after two years as Production Manager became General Manager with responsibility for technical direction.

United Components Ltd, the manufacturing company for the Regentone and R.G.D. radio and television group, have appointed Mr. D.H. Fisher as Chief Engineer.

The University of Liverpool Department of Extra-Mural Studies announce post-graduate courses in science during the Easter vacation. They are on "Numerical Analysis and Digital Computing Machines (from 9-13 April) and "Fundamental Processes of Gaseous Discharges" (from 8-13 April). Further details may be obtained from the Director of Extra-Mural Studies, The University, 9 Abercromby Square, Liverpool 7.

LETTERS TO THE EDITOR

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

Relay Scale-of-Two Circuits

DEAR SIR,—Mr. R. C. M. Barnes, in the November 1954 issue, described a scale-of-two circuit using a single unpolarized relay with one changeover contact (Fig. 17, p. 495). This circuit is, however, somewhat expensive, using as it does, in the two preferred forms, a special transformer and five rectifiers.

In a letter in the March 1955 issue Mr. R. G. Wicker described a very simple circuit using a relay of similar characteristics. This latter circuit, however, has three limitations. Firstly, with only one changeover contact per relay, circuits of this type cannot drive one another directly; secondly, the supply voltage is restricted to a critical range; thirdly, the release of the relay is slow.

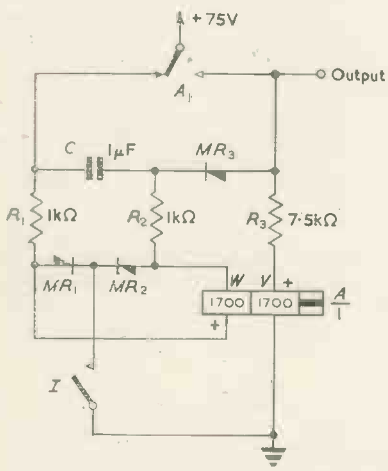

Fig.A. Relay Scale-of-Two Circuit.

The circuit shown here (Fig. A) overcomes these limitations and is cheaper than the circuits of Fig. 17 described by Mr. Barnes. If the relay is released, closure of the driving contact I results in a current through R_1 and MR_1 to earth, and the capacitor C charges to the supply voltage through R_2 and MR_2. As long as I is closed both sides of winding W are effectively at earth potential, so that the relay does not operate. When I releases the capacitor is free to discharge through R_1, W and R_2. The relay makes and holds through R_3 and winding V.

During the next operation of I current flows through MR_3 and the capacitor charges with reversed polarity by way of R_1 and MR_1. Release of I results in a pulse of current through W which neutralizes the flux and releases the relay.

The values shown in the diagram are suitable for operation of a Siemens miniature sealed high-speed relay at moderate speeds. MR_1 and MR_2 may be germanium diodes. In the general case, using this type of relay, the following data may be used as a rough design guide. Let $R_1 = R_2 = R$; relay resistance $= r + r$ kΩ; $(r + 2R)/r = p$. Then the minimum nominal supply voltage $(E) = 26p\sqrt{r}$ volts with $C = 8/p^2r$ µF and $R_3 = 5R + 1{\cdot}5r$. For maximum economy of driving power $p = 2$ with E as given, but p and E may be increased independently for higher speed.

These formulas are based on the fact that for a relay of given type the operating voltage is proportional to $\sqrt{r}$, and L/r is constant. Furthermore, for a given optimum shape of current transient the quantity $L/C\,(r + 2R)^2$ must be constant. R_3 is chosen so that the transient exactly neutralizes the holding flux. Owing to uncertain factors such as inertia and eddy currents it is not practicable to state general formulas for any type of relay.

Circuits of this type may be cascaded to form a binary scale of higher order if the supply connexions and rectifier polarities are inverted in alternate stages.

Yours faithfully,

B. D. CORBETT.

Dept. of Clinical Research.
University College Hospital
Medical School,
London, W.C.1.

The Author replies:

DEAR SIR—I agree with the advantages claimed for the circuit and, although I have not built it up, feel that it should find an application where the principal aim is to reduce the number of moving parts subject to wear.

I suspect that the performance of the circuit may be very sensitive to component tolerances if R_3 is indeed "chosen so that the transient exactly neutralizes the holding flux" of the relay. A similar situation exists in the circuit of Fig. 6(b) of my original article (p. 434, October, 1954, issue) where relay A is released by an opposing winding. The release times of this relay were measured for the cases where the windings had nominally equal turns and resistance, and where the W winding had at least 2 per cent extra turns. The ratio of release times was almost 2:1. This will also apply to Mr. Corbett's circuit and it would be advisable to check that the flux in the relay is, in fact, reversed to ensure a reasonably constant and rapid release.

Yours faithfully,

R. C. M. BARNES,

Atomic Energy Research Establishment,
Harwell,

Regulated D.C. Power Supplies at Low Voltage

DEAR SIR,—I was interested to read the article by R. K. Hayward, J. C. Jennings and R. C. Barker in the July 1955 issue of ELECTRONIC ENGINEERING and the comments on the subject by S. L. Krauss in the November issue.

Many methods of obtaining a stabilized supply whose d.c. output voltage can be varied from a few hundred volts down to zero have been suggested, but it appears that they are not well known. I have recently written an article[1], in conjunction with L. J. Bental, reviewing the various methods and giving detailed information about the performance and limitations of one form of such stabilizer. Since it will be some considerable time before this article is published it may be of interest to many of your readers to briefly mention here the stabilizers which have been developed by Working[2], Garratt[3], White and Mansford[4], Scroggie[5], Green[6], Houle[7,8], May and Skalnik[9] and Admiraal[10].

Working[2] used two stabilized supplies in parallel, one having a fixed and the other a variable output voltage. In this way any voltage between 0 and 200V could be obtained.

The fundamental principle of White and Mansford's circuit[4] is that an auxiliary source of voltage keeps the cathode of the amplifying valve (in the normal series-parallel valve circuit) negative with respect to the negative terminal. The arrangement is similar to the circuit using an additional negative supply given by S. L. Krauss. Garratt[3] also uses an additional supply which, in this case, is stabilized with a VR150 glow-discharge tube. The basic circuit of White and Mansford has frequently been employed.

Scroggie's stabilizer[5] provides a continuously variable output voltage from zero to 5000V at currents up to 1mA. Two EL37 power pentodes are used as control valves and the amplifier is an EF42 high-slope pentode. Again an auxiliary negative-supply circuit is employed; this is stabilized with an 85A1 glow-discharge tube. A supply similar to Scroggie's has been described by Green[6]. It gives an output voltage adjustable within the range 3·5 to 500V at currents up to 300mA. Here four 807 valves in parallel act as the control element and a single 6AC7 valve, operated below earth, serves as the d.c. amplifier.

Houle[7,8] has pointed out that stabilizers based on the above principle dissipate excessive power in auxiliary circuits and require many separate heater transformer windings. He has described an alternative circuit giving an output voltage variable from zero to 300V. The normal d.c. amplifier has been replaced by one based on a carrier-voltage principle. Although the various elements in this unit are simple and small the modulated high-frequency signal seems to me to introduce unnecessary complication and the principle has not been generally adopted.

May and Skalnik's stabilizer[9] not only gives an output voltage adjustable to zero, but also maintains regulation while carrying reverse currents which is important for bias supplies for Class-*C* amplifiers. A beam power valve connected in parallel with the load serves as a bleeder drawing an approximately constant current regardless of the output voltage. Thus, the supply can draw a negative current equal to the bleeder current before losing its stabilizing ability.

All the arrangements mentioned so far possess the disadvantage that the amplifier valve is controlled by only part (1/6 to 1/2) of the output-voltage variation, and this part depends on the desired output-voltage. Admiraal[10] has designed a circuit in which the variations in output voltage are applied in their entirety to the amplifier. This circuit is, therefore, useful when a very high degree of stabilization is required. The article[1], referred to above, describes a stabilizer very similar to Admiraal's with an output voltage adjustable within the range from zero to 300V. At output voltages up to 50V the maximum load current is 180mA; at higher output voltages the maximum current is 200mA.

Yours faithfully,

F. A. Benson,
Department of Electrical Engineering,
University of Sheffield.

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2. Working, E. R. An Electronic Voltage Regulator. *Ind. and Eng. Chem.* (*Anal.*) *30*, 435 (1938).
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4. White, E. L. C., Mansford, H. L. Improvements in Voltage Stabilizing Devices for Electric Supply Systems. *British Patent 515*, 435 (1939), and *U.S. Patent*, *2*, 268, 790 (1942).
5. Scroggie, M. G. Stabilized Power Supplies. *Wireless World 54*, 453 (1948).
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9. May, J. C., Skalnik, J. G. Regulated Bias Supply. *Electronics 24*, 136 (December, 1951).
10. Admiraal, D. J. H. A Stabilized d.c. Supply Unit for Output Voltages of 0 to 300V. *Philips Electronics Application Bull. 14*, 61 (May, 1953).

The authors' reply:

Dear Sir,—The correspondence seems to have missed the point we were trying to make, namely that power supplies to feed highly efficient devices such as transsistors, should themselves be efficient otherwise much of the advantage of these devices is lost.

We showed the conditions to get the best efficiency out of a valve type stabilizer but for low voltages this is not good enough. The articles given in Dr. Benson's excellent list of recent contributions to the subject describe valve stabilizers of good stability which is obtained at the expense of using high h.t. voltages.

Stability of voltage more than adequate for most practical purposes can be obtained with the new rectifiers now coming into use, providing the filter is redesigned to take advantage of the condition that in practical cases it is no longer the rectifier which sets a limit to the regulation.

Yours faithfully,

R. K. Hayward,
Post Office Engineering Department,
Research Branch.

Amplifier Loading

Dear Sir,—I have recently devised a scheme for determining the effect of a load applied to the output circuit of an amplifier, which seems much simpler than the usual method, and which is applicable to such cases as the addition of extra loudspeakers, or the maximum permissible shunt reactance for a given loss in gain.

Briefly, we consider the amplifier as a generator in series with an impedance, by Thevenins theorem, and calculate the loss due to an additional load (Fig. 1).

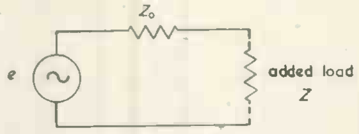


Fig. 1. Effect of additional load.

This gives a simple solution that the loss in dB is $20\log_{10}\left|\frac{Z}{Z + Z_o}\right|$.

Z_o is to be taken as the nominal output impedance of the amplifier, in parallel with original load, and Z as an additional load.

The case of a reactive Z is of interest for its simplicity as compared with the usual drudgery of calculating gains for two cases, dividing one by the other and rationalizing.

If x is a reactance, then the vector

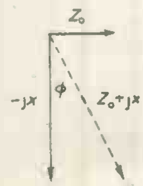


Fig. 2. Vector Diagram

diagram (Fig. 2), will demonstrate

(a) that $\tan\phi = Z_o/x$

(b) that $\cos\phi = \left|\frac{jx}{Z_o + jx}\right|$ which is identical with the ratio $\left|\frac{\text{gain }(x)}{\text{gain }(Z_o)}\right|$ where the suffixes (x) (Z_o) indicate the load conditions.

As example:

To determine the maximum capacitive shunt across a 1kΩ load in a cathode-follower, $G_m = 7{\cdot}5$ for a loss of 1dB at 10kc/s.

Since the unloaded output impedance is $1/G$, Z_o becomes $\frac{1/G}{1/G+1} = \frac{1}{1+7{\cdot}5}$ kΩ $= 117{\cdot}7\Omega$. In this case, $\cos\phi = 10^{-1/20} = 0{\cdot}891$, $\phi = 27^\circ$. Then $x = \frac{117{\cdot}7}{\tan 27} = \frac{117{\cdot}7}{0{\cdot}509} = 233\Omega$.

$$C = \frac{1}{2\pi fx} = \frac{1}{2\pi\times 10^4\times 233} = 0{\cdot}0683\mu F.$$

I think this method sufficiently simple and straightforward to merit its being brought to the attention of your readers.

To avoid recriminations, I would mention that the calculations are by slide rule, though I believe the error is well below 1 per cent.

Yours faithfully,

H. Moorby,
Farnborough,
Hants.

RECENT BRITISH STANDARDS

The Performance of Power Transformers (not exceeding 2kVA rating) for radio and allied electronic equipment (B.S. 2214:1955)

The standard covers general purpose power transformers of up to 2kVA rating, for operation at frequencies up to 3kc/s, and intended for use in radio and allied electronic equipment. It applies to open, enclosed non-sealed and hermetically sealed transformers; but vibrator and pulse types are excluded. It is intended that a future revision shall extend the standard to include chokes and other types of transformer suitable for the use of the Services.

This Standard should be read in conjunction with B.S.2011 "Climatic and durability Tests for Components for Radio and Allied Electronic Equipment.

Copies may be obtained from the British Standards Institution, Sales Branch, 2 Park Street, London, W.1. Price 3s.

Colour Codes for Connections in Radio and Allied Electronic Equipment (excluding telephone exchange and associated transmission equipment) (B.S.2311.1955)

Although these codes are not specifically intended for application to domestic radio or television receivers, they may, if desired, be used for that purpose in their present, or a simplified, form. Because of manufacturing limitations, and in view of the fact that the standard covers the use of both plastics and textile coverings for connecting wire, it is not possible to give a precise specification of the preferred colours.

Copies are obtainable from the British Standards Institution, price 2s. 6d.

BOOK REVIEWS

Automatic Feedback Control System Synthesis

By J. G. Truxal. 663 pp., 80 figs. Demy 8vo. McGraw Hill Publishing Co. Ltd. 1955. Price 89s. 6d.

SEVERAL good text books treating fundamental servo theory have been written in the years since the war. Of these, this is the first book which assumes that the reader has a certain knowledge of the subject at the outset. This book is also the most comprehensive to be published to date in one volume on basic theory. The subject is developed in a consistent and uniform manner and a philosophy is advanced that is likely to supersede that at present existing, in that newer techniques for treating the transient response of systems are given preference over the harmonic analysis procedure which has been stressed in literature hitherto.

The first three chapters adequately deal with the basic mathematical tools and network theory required in servo-system synthesis. Special emphasis is laid on complex variable theory, transformation methods for the solution of differential equations, circuit theory and system flow diagram techniques. Then follow two chapters on system synthesis by the root-locus method of approach in which pole-zero locations of the open-loop transfer function are used in a graphical procedure to determine the corresponding pole-zero location and loci for the closed loop system. Chapter VI treats systems with multiple inputs, transfer function derivation and a.c. system design. Then follows an excellent chapter dealing with the behaviour of systems subject to random disturbances and random input signals. The statistical design approach is employed. Use is made of the autocorrelation function and system spectral density in this analysis. The following chapter advances the subject further to optimization of systems and minimization of r.m.s. error. Non-linear systems are treated and the non-stationary time series approach is explained.

In Chapter IX graphical and analytical methods are presented for the analysis of sampled data systems. The book concludes with two chapters on the analysis of non-linear systems by describing function techniques (i.e. approximation of periodic non-linear oscillations to the fundamental harmonic) and on the use of phase-plane methods of the type developed by Minorsky.

The book leaves little to criticism except that perhaps rather more space than is justified is devoted to signal flow theory. On the other hand little description is given in early chapters of typical servo elements encountered in practice. A resumé of these together with types of non-linearities arising would be helpful.

Guillemin's pole-zero location methods are presented for what is probably the first time in any text book. While this topic is illustrated with suitable examples the presentation is not as direct as might be hoped. One feels the author realizes this in the section "Comments on Guillemin's method" where an attempt to clear up the preceding points is made.

The author stresses the advantages of the root-locus plot over the Nyquist diagram but this does not justify the brevity of treatment given to the latter and to the dB-log frequency methods of approach.

The book is a valuable contribution to existing literature on servo-system analysis and will find favour in all quarters where there is real interest in this subject.

R. D. LANGMAN,
J. P. CORBETT.

Reports on Progress in Physics

Edited by A. C. Stickland. 447 pp., 137 figs. Demy 8vo. Vol. XVIII. The Physical Society. 1955. Price 50s.

LIKE its predecessors, the present volume of authoritative survey papers contains something of interest for almost anyone whose field borders on that of physics. There is in this issue a strong emphasis on "particles". No fewer than ten of the fifteen papers have to do with one aspect or another of this subject, ranging from the highly mathematical to the severely empirical. Four of the remaining five are concerned with magnetic phenomena and one with electrostriction. But despite this apparent concentration of attention on a small front (in itself no demerit) the treatments range sufficiently widely to make contact with many other fields.

The first paper, by G. H. Kinchin and R. S. Pease, covers a particularly wide area. On "Displacement of Atoms in Solids by Radiation", it offers some valuable information on almost every aspect of the solid state. The author's documentary thoroughness is attested by the inclusion of H. G. Wells' *Tono Bungay* among the historical references to the possibility of mechanical failure under intense irradiation.

Readers of this journal will find the section on Semiconductors of special interest. A normal semiconductor may have only one free electron per 10^7 atoms. Since however one free electron is released on the average for each atom displaced by radiation, the effect of only moderate doses is considerable. It is quite common for an n-type semiconductor to be converted into a p-type by such irradiation.

Forty eight pages of lucid exposition are supplemented by four pages of references in this painstaking review. Indeed all the authors supply a large number of references; and one feels reason to be grateful to those, such as E. W. Lee, who have "gone the second mile" and done some judicious weeding. His paper on "Magnetostriction and Magnetomechanical Effects" is the first review of its kind since 1926. It does not however deal with dynamic magnetostriction nor with magnetomechanical effects at other than zero frequency, so that its relevance to electronic applications of these effects is limited.

H. F. Kay's treatment of 'Electrostriction" covers many recent developments in the rapidly expanding field of ferroelectricity. The remarkable and technically important properties of $BaTiO_3$ and other perovskite-type ceramics are discussed in some detail, chiefly at an empirical level. The practical value is enhanced by the inclusion of some helpful recipes for growing perovskite-type crystals. On the theoretical side the distinction between piezostrictive (linear) and electrostrictive (square-law, etc.) response to applied field is dicussed in detail. A brief but interesting section indicates how complex hysteresis loops can result from—and throw light on—the various types of domain rotation possible in single crystals.

J. L. Symonds' survey of "Methods of Measuring Strong Magnetic Fields" is again the first of its kind to appear for two decades, and reveals remarkable advances both in the accuracy and the diversity of methods now available. One of the best depends on the absorption of r.f. energy due to magnetic resonance in electrons up to 300 gauss) and protons (up to 20 000 gauss). "For the sake of progress in physics" this author also supplies a chemical recipe, for the preparation of 1 : 1-diphenyl-2 picryl hydrazyl, in which electron resonance is particularly sharp. Protons are conveniently available in water or mineral oil.

Some ingenious solutions have been found to problems in electronics raised by this and other methods of field-measurement, which include the use of high-permeability "peaking strips", vibrating coils, and novel forms of drift-corrected fluxmeter. One of these uses the familiar Miller electronic integrating circuit (to which only Swedish and American references are given), in order to secure a rapid response.

Methods of attaining temperatures as low as 0·001°K by the cyclic magnetization and demagnetization of paramagnetic salts, are described in the article by E. Ambler and R. P. Hudson on "Magnetic Cooling". The increase in resistivity shown by many metals below 1°K is apparently not yet understood, but is now comparatively easy to study. Because of its abrupt change in thermal conductivity with temperature, lead, in the form of wire, is now extensively used as a thermal "valve" or "switch" to control the flow of heat in such experiments.

It is possible only to mention some of the other titles, but the paper by J. D. Craggs and C. A. McDowell on "Ionization and dissociation of complex molecules by electron impact" has considerable bearing on the design of vacuum-electronic devices. Paramagnetic Resonance and the Theory of Radiation receive lengthy theoretical treatment; and the volume ends with six papers on Field Theory originally presented at the Autumn meeting of the Physical Society in the University of Birmingham, in December 1954.

Professor Peierls in his introductory paper has had a difficult task in outlining this abstruse subject for the non-specialist, and (on page 425 and the foot of page 430 for example) he may not always carry his reader with him. On page 432 however the balance is redressed by a happy qualitative exposition of Yukawa's Meson-field concept. We are given also considerable insight into the current solution of the classical problem of the self-energy of the electron by "re-normalization".

The notation used is often less happy, and it seems a pity at this date that one of the authors should still define $\psi\psi^*$ as the probability of finding a particle *at a point* (which is of course zero) rather than the probability, per unit volume, of finding the particle in a given volume-element.

The book as a whole is well produced. The lack of an alphabetical index, (which would presumably have reduced its news value by adding delays) is largely off-set by the inclusion of a logically organized list of contents at the head of each paper. As a work of reference this eighteenth volume will undoubtedly prove as indispensable as its predecessors.

D. M. MacKay.

BBC Handbook

288 pp. Demy 8vo. The British Broadcasting Corporation. 1955. Price 5s.

THIS handbook is a guide to the organization and work of the BBC with up to date statistics. The obligations of the BBC under its Charter as a public service broadcasting organization and the role of the BBC in "acting as a trustee for the national interest" in political, religious, controversial and educational broadcasting, as well as in the field of entertainment, are explained. Writers, composers and those wishing to apply for auditions will find information they require in a section "Practical Advice", which also explains the BBC rules for SOS messages and appeals. Another section deals with the administration of the staff of over 13 500 and the Corporation's relations with outside unions and similar bodies. The index has been enlarged, cross references having been made wherever possible.

ELECTRONIC EQUIPMENT

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

Communication Receiver

(Illustrated below)

THE Airmec receiver type C.864 is a high grade general purpose communication receiver operating over the frequency range 15kc/s-45kc/s, 100kc/s-30Mc/s.

Although the specification is in general based on the recommendations given in the G.P.O. publication 'Radio for Merchant Ships 1947', it is superior in certain important features such as sensitivity, signal-to-noise ratio and second channel rejection.

The receiver incorporates a tuning unit with a 90:1 slow motion drive and a film scale giving an actual total length of 23ft of direct frequency calibration. An entirely separate directly calibrated incremental tuning control operating over a constant range of 100kc/s above 1Mc/s is also fitted, and these two controls when used in conjunction with the built in crystal calibrator enable the receiver to be set directly to any frequency with an accuracy of better than 1kc/s.

A double superheterodyne circuit is employed above 1Mc/s, the first and second intermediate frequencies being 800kc/s and 85kc/s respectively. Below 1Mc/s an intermediate frequency of 85kc/s only is used. Special precautions, which include neon stabilized h.t. supplies for the local oscillators and the b.f.o., are taken to ensure good stability.

A built in 2½in monitor speaker operates from a reduced output of 50mW. The full output of 2 watts at an impedance of 3Ω is available from a jack on the front panel. A telephone output of 20mW into an impedance of 600Ω is obtainable from the same jack and the internal speaker is muted automatically when the jack is in use. Negative feedback keeps the output voltage substantially constant over a wide range of load impedance from 3Ω upwards without excessive distortion.

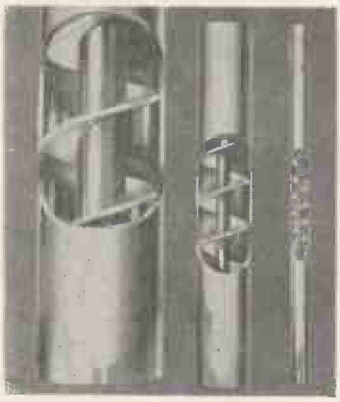
Two aerial inputs are provided, one at approximately 75Ω impedance unbalanced for a coaxial line, and the other at approximately 400Ω impedance unbalanced for high impedance aerials. An aerial circuit trimmer is fitted on the front panel to correct for unmatched aerial inputs.

Airmec Ltd,
High Wycombe,
Buckinghamshire.

Helical Membrane Cable

(Illustrated below)

A NEW high-power feeder capable of carrying 20kW at 100Mc/s has now been added to the range of helical membrane cables manufactured by The Telegraph Construction and Maintenance Co. The new cable, known as HM.10.A1, has a diameter of 3⅛in over the aluminium tube outer conductor, and a characteristic impedance of 50Ω. Intended for applications such as television transmitter aerial feeders where previously built-up feeders would have been used the new cable gives the advantages of high electrical uniformity with easier and more economical installation.



In common with the six smaller helical membrane types already on the market, in impedances of 50 and 75Ω, HM.10.A1 is constructed with a central copper conductor supported within a tubular outer conductor of aluminium by a helical edge-on tape of polythene.

The Telegraph Maintenance & Construction Co Ltd,
Mercury House,
Theobald's Road,
London, W.C.1.

Field Ratemeter

(Illustrated above right)

THIS transistorized battery operated instrument has been developed in conjunction with the A.E.R.E. (Harwell) for geological surveying and mapping, prospecting for radio-active ores and for the examination of mine workings. When used with the beta probe unit, assay work can be carried out in the field.

The ratemeter employs a junction type transistor and four g.m. tubes. The count rate is indicated on a 50μA meter calibrated in milli-Rontgens per hour. Five ranges are provided covering count rates of 0 to 25mR/h. The accuracy is selected by means of a switch and is either 10 per cent or 2 per cent. Speed of response depends on the accuracy selected. Audible indication of the count rate is available, if required.

All power supplies are derived from three 1·5V torch cells, such as the Ever Ready Type U2, and a corona stabilizer ensures that the instrument is not affected by changes in battery voltage over the range 4·5V to 3V.

The ratemeter weighs 8½lb and is built to withstand the rough treatment which is inevitable under field conditions.

The beta probe unit, which obtains its power supplies from the ratemeter via a 6ft cable is used for making assays of radio-active ores as well as for the detailed examination of rock faces and mine workings which are inaccessible to the ratemeter operator.

Ericsson Telephones Ltd,
Beeston,
Nottingham.

Lightweight Headphone/Microphone Set

(Illustrated below)

THESE headsets weigh only about 7oz complete with microphone. The headphones are constructed from plastic moulded parts fitted to adjustable sliding extensions on a flat stainless steel p.v.c. covered headband. Perspex ear-disks are attached to the earphone units by flexible rubber diaphragms of original design. The diaphragms distribute the pressure of the disks lightly and evenly over the outer ear avoiding contact with any part of the bone structure.

The insert units are of high quality electromagnetic pattern, acoustically equalized to give a response which is level from 200c/s to 3·3kc/s ± 3dB.

The absence of any marked peaks in the response not only results in good intelligibility but permits the use of higher listening levels without the discomfort and annoyance of resonances, which are characteristic of earlier types of headphone.

Three alternative types of microphone, including a noise cancelling type, are available.

Amplivox Ltd,
2, Bentinck Street,
London, W.1.

Automatic Dip-Soldering Bath

(Illustrated below)

THE introduction of an automatic dip-solder bath, for soldering printed circuits and turret terminal panels, is announced by the Reid Manufacturing Company. The bath itself is electrically heated with a capacity of approximately 35lb of solder. The maximum size of panel which can be accepted by the adjustable work holder is 11in by 10in. Soldering is accomplished by an automatic device which first cleans off surface dross from the solder and then lowers the board for soldering with a rocking motion which eliminates the trapping of air or extraneous gases. Other refinements include full thermostatic control, adjustable control of cycle times, automatic level indication, and drain-off plug.

A. Reid Manufacturing Co Ltd,
48, Station Road,
Harrow,
Middlesex.

Marine Echo-Sounder

(Illustrated above right)

THE Pye Leadsman works on exactly the same principle as elaborate recording echometers, utilizing the Sonar system used in the Royal Navy for submarine detection. It is essentially an accurate electronic timing machine wherein an electrical impulse is made in the transmitter and sent via the coaxial cable to the transducer. This single unit converts this impulse into an ultrasonic beam projected downwards towards the seabed. When this beam hits the bottom or other reflecting surfaces such as obstructions or wrecks, it is reflected back to the transducer which sends it back to the receiver. Here the signals are amplified and applied to a neon-indicator on a synchronous rotating arm which flashes once when the pulse is transmitted (zero position) and again when received back (against the depth sounded on the scale) having measured

the time taken for the whole cycle. The whole operation is automatic and is repeated hundreds of times a minute.

This unit will indicate depths between 3ft and 270ft (45 fathoms).

Pye Marine Ltd,
Lowestoft,
Suffolk.

Hermetically Sealed Rechargeable Cells

(Illustrated below)

DEAC Perma-Seal cells are permanently sealed rechargeable electric storage cells of the nickel-cadmium type.

They require no topping-up or other maintenance (other than recharging). The cells are constructed so that they can be permanently wired into a circuit. The present range includes disk-type cells of 60 to 150mAh capacity, cylindrical cells of 125mAh and over and rectangular prismatic cells of from 1·7Ah to about 20Ah.

Distributed by **G. A. Stanley Palmer Ltd,**
Maxwell House,
Arundel Street,
London, W.C.2.

Miniature Solenoid

(Illustrated below)

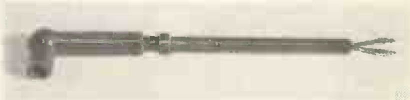
THE Varley S.M.O. solenoid is the first of a range of miniature solenoids being developed for the electronic and accounting machine industries. The length of the solenoid, excluding the plunger and fixing boss is 1¼in. The solenoids may be used on a.c. or d.c., and can operate at approximately 300 operations per minute dependent on the particular application. They are available in both pull and thrust types.

The Varley Magnet Company,
Cambridge Row,
Woolwich,
London, S.E.18.

Convertible Test Lead

(Illustrated below)

THE 'Prodclip' is a test lead termination which can be used as either a prod or a clip. In the normal position the spring loaded jaws protrude about 3/16 in from the plastic body and form a pointed prod. Slight pressure on the remote end of the prod causes the jaws to open and they may then be clipped on to a wire or terminal. The overall length of the Prodclip in 5½ in and the weight ½oz; they are available coloured red and black.



Saco (Electronics) Ltd,
1a, Norfolk Place,
London, W.2.

Microphone Amplifier

(Illustrated below)

THE Brüel and Kjaer microphone amplifier type 2602 is designed for amplification and measurement of small voltages, for example from microphones. The total amplification is 100dB and elaborate stabilizing maintains constant amplification. The total voltage amplification can be set to known values by means of two knobs, so that the apparatus forms a measuring amplifier. A switch changes the straight-line frequency response characteristic of the amplifier to the 3 internationally standardized antisonoric characteristics which make the amplifier suitable for use in many acoustic investigations. Provision is made for inserting external filters, such as the 1/3 Octave Filter Set Type 1609, between the amplifier stages.

Distributed by

B & K Laboratories Ltd,
59-61, Union Street,
London, S.E.1.

Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 28 March. Time: 6.30. p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Lecture: Some Problems of Secondary Surveillance Radar.
By: K.E. Harris.

Merseyside Section

Date: 14 March. Time: 7 p.m.
Held at: The Council Room, Chamber of Commerce, 1 Old Hall Street, Liverpool 3.
Lecture: Underwater Television.
By: R. Swinden.

North-Western Section

Date: 1 March. Time: 6.30. p.m.
Held at: The Reynolds Hall, College of Technology, Sackville Street, Manchester.
Lecture: The Design and Application of Quartz Crystals.
By: R.A. Spears.

West Midlands Section

Date: 14 March. Time: 7.15. p.m.
Held at: Wolverhampton Technical College, Wulfruna Street, Wolverhampton.
Lecture: The Application of Transistors to Radio Rectification.
By: E. Wolfendale.

Scottish Section

Date: 12 March. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders. Elmbank Crescent, Glasgow.
Lecture: Principles of the Transistor and Some Important Applications.
By: M.H.N. Potok.

South Wales Section

Date: 21 March. Time: 6.30. p.m.
Lecture: Electronic Servo-Mechanism.
By: J.L. Russell.

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings, unless otherwise stated, will be held at the Institution, commencing at 5.30. p.m.

Radio and Telecommunication Section

Date: 7 March.
Lecture: Frequency-Modulation Radar for use in the Mercantile Marine.
By: D.N. Keep.
Date: 19 March.
Informal talk: Electronics and Automation.
By: H.A. Thomas (with films and demonstrations).

Measurement and Control Section

Date: 13 March.
Lecture: The Dual Input Describing Function and its Use in the Analysis of Non-Linear Feedback Systems.
By: J.C. West, J.L. Douce and R.K. Livesley.

Measurement and Radio Sections (Joint meeting)

Date: 27 March.
Lecture: Radiation Monitors Using Transistors (and other associated papers).
By: E. Franklin and J.B. James.

East Midland Centre

Date: 14 March. Time: 7.30. p.m.
Held at: The George Hotel, Kettering.
Lecture: Colour Television.
By: J.A. Darbyshire.
Date: 15 March. Time: 7.30. p.m.
Held at: The Masonic Hall, Pinchbeck Street, Spalding.
Repeat of above lecture.

Cambridge Radio and Telecommunication Group

Date: 13 March. Time: 8.15. p.m.
Held at: The Cavendish Laboratory.
Lecture: Television Methods in Astronomy.
By: P.G. Fellgett.

Mersey and North Wales Centre

Date: 19 March. Time: 6.30. p.m.
Held at: The Town Hall, Chester.
Lecture: Automatic Circuit Reclosers.
By: G.F. Peirson, A.J. Pollard and N. Care.

North Eastern Centre

Date: 2 March. Time: 7 p.m.
Held at: Carlisle Technical College.
Lecture: Colour Television.
By: G.N. Patchett.

North-Eastern Radio and Measurement Group

Date: 5 March. Time: 6.15. p.m.
Held at: King's College, Newcastle-upon-Tyne.
Lecture: Maintenance Principles for Automatic Telephone Exchange Plant.
By: R.W. Palmer.

North-Western Centre

Date: 19 March. Time: 6.30. p.m.
Held at: The Town Hall, Chester. (Joint meeting with the Mersey and North Wales Centre).
Lecture: Automatic Circuit Reclosers.
By: G.F. Peirson, A.H. Pollard and N. Care.

North-Western Radio and Telecommunication Group

Date: 14 March. Time: 6.45. p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Lecture: Frequency-Modulation Radar for use in the Mercantile Marine.
By: D.N. Keep.

South-East Scotland Sub-Centre

Date: 6 March. Time: 7 p.m.
Held at: The Carlton Hotel, North Bridge, Edinburgh.
Lecture: Nuclear Reactors and Power Production.
By: F.M. Greenlees.
Date: 20 March (Time and place as above).
Lecture: Equipment of Instrumental Accuracy for Recording and Reproduction of Electrical Signals using Cinematograph Film.
By: H. McGregor Ross.

South-West Scotland Sub-Centre

Date: 21 March. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39, Elmbank Crescent, Glasgow.
Lecture: Colour Television.
By: G.N. Patchett.

South Midland Radio and Telecommunication Group

Date: 26 March. Time: 6 p.m.
Held at: The James Watt Memorial Institute. Great Charles Street, Birmingham.
Lecture: TRIDAC A Large Analogue Computing Machine.
By: F.R.J. Spearman, J.J. Gait, A.V. Hemingway and R.W. Hynes.

Southern Centre

Date: 9 March. Time: 6.30. p.m.
Held at: The South Dorset Technical College, Weymouth.
Lecture: Transistor Power Amplifiers.
By: D.D. Jones.
Date: 14 March. Time: 7.30. p.m.
Held at: R.A.E. Technical College Farnborough.
Lecture: Germanium and Silicon Power Rectifiers.
By: T.H. Kinman, G.A. Carrick, R.G. Hibberd and A.J. Blundell.

Western Centre

Date: 12 March. Time: 6 p.m.
Held at: The Lecture Hall, University Engineering Laboratories, Bistol.
Lectures: The Application of Symmetrical Components to the Measurement of Phase Difference in Single Phase Circuits.
By: R.L. Russell.
and: The Design of Semi-Conductor Wattmeters for Power-Frequency and Audio-Frequency Applications.
By: H.E.M. Barlow.

Maidstone District

Date: 6 March. Time: 7 p.m.
Held at: The Prince of Wales Hotel, Railway Street, Chatham.
Lecture: Radio Aids to Marine Navigation.
By: F.J. Wylie.

Hatfield District

Date: 20 March. Time: 7 p.m.
Held at: Hatfield Technical College.
Lecture: Information Theory.
By: E.C. Cherry.

Oxford District

Date: 14 March. Time: 7 p.m.
Held at: District Office, Southern Electricity Board, 37 George Street.
Lecture: New Tools in Industry—Ultrasonics.
By: W. Owen Roberts.

Reading District

Date: 19 March. Time: 7 p.m.
Held at: The George V Room, George Hotel, King Street.
Lecture: The Nuclear Power Programme.
By: B.L. Goodlet.

THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Date: 6 March. Time: 5 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2.
Lecture: Review of Basic Factors in Local Line Planning.
By: R. MacWhirter and H.J.S. Mason.
Date: 21 March. Time. 5 p.m.
Held at: The Conference Room, 4th Floor, Waterloo Bridge House, London, S.E.1.
Lecture: External Construction and Maintenance —Are our methods out of date?
By: R.H. de Wardt and S.T.E. Kent.

THE SOCIETY OF INSTRUMENT TECHNOLOGY

Date: 21 March. Time: 7 p.m.
Lecture: A Novel Relay Amplifier.
By: J.H. Askew.
Held at: Manson House, Portland Place, London, W.1.
Date: 27 March (Time and place as above).
Lecture: Atomic Energy Plant Instrumentation.
By: R.K. Sandiford.

THE TELEVISION SOCIETY

Date: 9 March. Time: 7 p.m.
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue. London, W.C.2.
Lecture: Properties and Problems of Bands IV and V.
By: R.L. Smith-Rose.

PUBLICATIONS RECEIVED

ATOMIC ENERGY CAREER FOR YOUTH is the subject of a booklet describing the United Kingdom Atomic Energy Authority apprenticeship scheme for training in engineering (electrical and electronic, or mechanical or chemical, or physics), and in metallurgy. United Kingdom Atomic Energy Authority, Bedford Chambers, Covent Garden, London, W.C.2.

SERVOMEX RANGE OF ELECTRICAL CONTROL INSTRUMENTS is a new general catalogue with current price list of instruments produced by Servomex Controls Ltd of Crowborough Hill, Jarvis Brook, Sussex, who would be pleased to add new names to their mailing list.

TELEVISION SERVICE COURSE LABORATORY MANUAL by Matthew Mandl is a workbook of experiments that are designed to give students practical experience in applying trouble-shooting and servicing procedures to television receivers. The Macmillan Company. New York and London. 10 South Audley Street, London, W.1. Price 24s.

MOULAGE SOUS PRESSION DU ZINC ET DES METAUX LEGERS is a book on pressure diecasting of zinc and light metals produced by the European Productivity Agency of the Organization of European Economic Cooperation, 2 Rue Andre Pascal, Paris 16e.

WIGGIN NICKEL ALLOYS NO.37 is a special issue of this publication devoted to the nickel copper alloy Monel, which was first produced commercially nearly half a century ago. Monel has since been used for a very wide variety of applications requiring a combination of mechanical strength and resistance to corrosion. Copies are obtainable from the Publicity Manager, Henry Wiggin & Co Ltd, Thames House, Millbank, London, S.W.1.

RSGB AMATEUR RADIO CALL BOOK 1956 edition is now available from booksellers and radio dealers or direct from the Radio Society of Great Britain, New Ruskin House. Little Russell Street, London, W.C.1. Price 2s. 6d. postage 3d.

LIMITERS AND CLIPPERS and A-M DETECTORS are two further recent publications of John F. Rider Publisher, Inc. 480 Canal Street, New York 13. N.Y. Price $1.25 each.