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Commentary

THE Computer Exhibition will be held at Olympia, London, from 28 November to 4 December. This exhibition, which promises to be, probably, the most comprehensive display of computers and allied equipment that has ever been held and which will be certainly the first event of its kind in Europe, has been organized by the Electronic Engineering Association and the Office Appliance and Business Equipment Trades' Association at the instigation of the National Research Development Corporation.

A Business Computer Symposium will be held concurrently with the exhibition and at its six sessions executives from a wide range of concerns, of very diverse nature and size, will detail their own experience of computers as an aid to business organization.

The equipment on display at this exhibition will be of very varied nature and will represent 'electronic computers' in the widest possible meaning of the term, including the large high-speed type of digital machine together with its medium and desk size companions, complete data processing systems built around these machines and also the analogue type of computer of both the special and general-purpose variety. In addition, the application of digital techniques to problems such as machine tool control and curve plotting, will be demonstrated and components for incorporation in computers will be displayed.

The modern computer, be it analogue or digital is, in general, a well-engineered product and is one of the finest examples of electronic engineering; the most noticeable features being the reduction of bulk and the ease of accessibility compared with its prototype predecessors. In the case of the digital machine this has been brought about largely by the advent of suitable transistors and the development of efficient and reliable circuits for their use and, perhaps even more important, the evolution of more compact memory and data storage devices. A great deal of effort has gone into the development of the present-day computer and there can be no doubt that the integration of these machines into industrial and commercial organizations will be of ultimate benefit to the country, yet it does seem that the techniques of producing these machines is perhaps outstripping the research into their most economical and beneficial employment.

As far as the application of the computer to scientific work is concerned there is no doubt as to its effectiveness; the analogue and the digital machine have their own particular niches and they enable computations which are obviously necessary, or at least highly desirable, to be carried out speedily and accurately. It is, however, in the application of the digital computer to such matters as accounting and production control in large commercial undertakings that the greatest benefits are likely to accrue; it is also the application that poses the greatest problems. The reasons for this are manifold. There is the difficulty, common to many problems where different types of personnel are involved, of evolving a 'common language' and of understanding one another's problems—the electronic engineer on the one hand and the commercial systems planner on the other. There is the reluctance of the commercial organization to completely jettison their existing system and to plan from scratch, so to speak, for it would appear to be essential to plan the whole system around the computer and not to try to adapt an existing system to include a computer. There is also the danger, which one may charitably ascribe to the enthusiasm and exuberance of the computer builders, of overselling the computer and of advising its use in cases where conventional punched card accounting equipment would be equally or even more effective.

According to a recent lecture given by John Diebold, an American, to the British Institute of Management, American experience in the exploitation of computers for commercial purposes has not been very happy. According to him, experience generally has been that so far it has been difficult to justify the cost of a computer and that operating costs have been much higher than expected. Mr. Diebold thought that most of the faults lay with managements, who had frequently allowed scientists to 'run loose'. He gave the impression that many firms had bought computers merely to be fashionable and that the principle seemed to be to buy and install a machine and to try and justify its use afterwards, with the inevitable result that computers were being used for tasks for which they were not suitable.

It may well be thought that top American management is not as simple as all that and that some dramatic licence must be allowed to Mr. Diebold in his remarks. Both of these assumptions may be correct but, nevertheless, there does appear to be a very definite problem and it is one which could have a very lucrative solution, both for the computer manufacturer and for this country's industry and commerce as a whole. The present exhibition and, more especially, the Business Computer Symposium which is to be held in association with it, could play a valuable part in providing some of the answers or, at least, in clarifying the problem.

Ferroelectric Storage Devices

By S. Morleigh*, A.M.I.E.E., A.M.Brit.I.R.E.

An important requirement of the modern digital computer is an efficient memory device to enable the high speed storage and recovery of data at will. This has led to a critical examination of the properties of all media possessing two stable states. Among the most important characteristics of a digital storage system are economy in power requirements for the processes of reading and writing information, speed of operation, size and cost per unit of stored data, simplicity of construction, and overall reliability.

In the present article an investigation is carried out of the storage properties of ferroelectric materials, particularly with reference to barium titanate single crystal capacitors as this appears at present to be the most suitable of ferroelectrics for the purposes of digital storage. A careful examination is made of switching characteristics and other relevant factors, including the limitations imposed by the present state of development of these materials.

THEORIES associated with ferroelectricity have been closely examined within recent years[1,2,2a], and considerable research is going on at the present time into the properties of newly discovered classes of ferroelectric crystals and ceramics[3,4,5]. On these grounds a comprehensive theory of this subject would not lie within the scope of the present article.

Ferroelectric capacitors may be formed from dielectric materials which are dependent upon the process of internal polarization, rather than upon surface charge, for purposes of data storage. These materials have bistable properties as the polarization is reversible by the application of an electric field, in a manner similar to that by which the stable remanence states of magnetic materials may be reversed when they are subjected to magnetic fields of opposite polarity. This form of remanent charge of a ferroelectric capacitor will be retained for long periods even if the capacitor terminals are short-circuited.

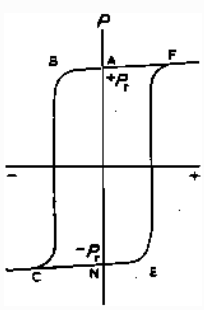

Fig. 1. Hysteresis loop obtained from barium-titanate capacitor at 500c/s

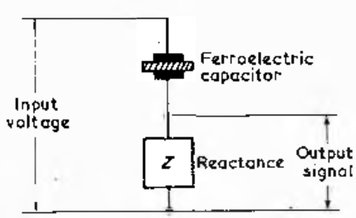

Fig. 2. Read-out circuit *(Switching current transients observed by replacing Z with a 10Ω resistance)*

In order for a ferroelectric material to be suitable for storage applications it should have a nearly rectangular hysteresis loop. A typical hysteresis characteristic for a good barium titanate single crystal capacitor is illustrated in Fig. 1. The vertical axis represents the degree of polarization P of the crystal, and the horizontal axis represents the applied voltage V, which is equal to the applied electric field E multiplied by the dielectric thickness d.

The application of a polarizing field will switch the ferroelectric to one or other of its stables states, defined by points A and N, and it will remain in this condition, with remanent polarization P_r, after the removal of this electric field. Hence, if the stable state indicated by point A represents the condition in which a binary 'zero' is stored, then that indicated by point N on the hysteresis loop corresponds to a binary 'one'. Assume initially that the ferroelectric is in state A, then the application of a negative electric field across the capacitor will cause the switching path A B C to be followed, where point C represents the saturated condition. Removal of the applied field then causes the loop to be traversed from C to the stable state N, which represents a stored 'one'. Similarly, the application and subsequent removal of a positive field will then cause the state of the ferroelectric to change in a manner indicated by the path N E F A. However, the effect of a positive voltage pulse applied to the ferroelectric capacitor when in state A will simply cause the path A F A to be traversed.

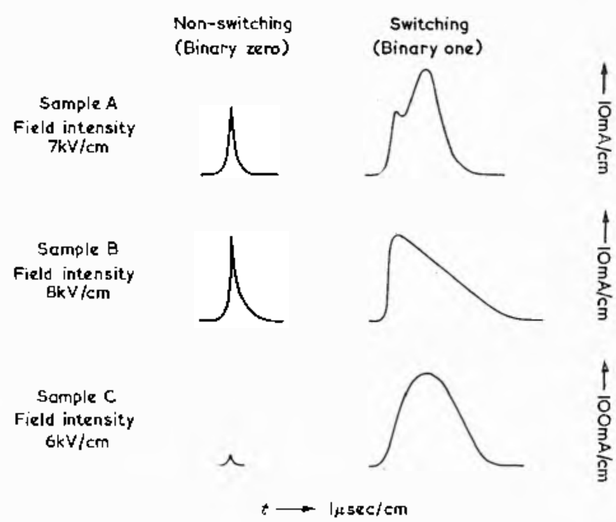

Fig. 3. Current Transients. *Transient pulse shapes of samples B and D are similar*

Dimensions of sample elements

SAMPLE	AREA (mm²)	CHARGE (COULOMBS)
A	0·03	$0{\cdot}025\times10^{-6}$
B	0·06	$0{\cdot}035\times10^{-6}$
C	0·67	$0{\cdot}43\times10^{-6}$
D	0·50	$0{\cdot}30\times10^{-6}$

Dielectric thickness approx. 0·1mm

The process of read-out, or of sensing the stored information, is carried out by the application of a positive pulse to the storage cell and, since $i = dq/dt$, the changes of charge can be observed by examining the flow of current to the capacitor via a series impedance. A typical read-out circuit is illustrated in Fig. 2, and the 'switching current' transients obtained for a number of ferroelectric crystal capacitors have been plotted in Fig. 3.

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A mathematical analysis of the switching process, and a description of the experiments carried out, are given under separate headings.

Ferroelectric Materials

Some useful tables have recently been published[6,7] of dielectric materials which have ferroelectric properties. Unfortunately, most of these materials are unsuitable for use as storage media owing to one or more of the following reasons:—

(a) The Curie point, or the upper limit to the temperature range over which the material remains ferroelectric, is in many cases less than 50°C.

(b) Hysteresis loop is insufficiently rectangular.

(c) High electric field strengths are required for polarization.

The forerunner of these materials was Rochelle salt[8], but this is only ferroelectric for temperatures between

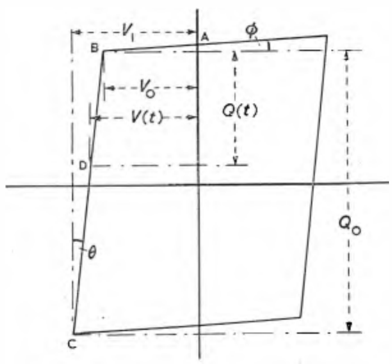

Fig. 4. Hysteresis loop for analysis

−18°C and +20°C. Other dielectrics such as potassium niobate, potassium dihydrogen phosphate, barium titanate, guanidine aluminum sulphate hexahydrate and triglycine sulphate also have ferroelectric properties. Of these, barium titanate appears at present to be the most useful for storage purposes, particularly in view of the low electric field strengths required for polarization, the rectangularity of its hysteresis loops, and its Curie point of 120°C. Experiments have disclosed, however, that in common with certain other ferroelectrics[9,10] it has the disadvantage of appearing to have no limiting field below which switching is impossible.

Triglycine sulphate, guanidine aluminum sulphate hexahydrate, and various niobates are among the ferroelectrics undergoing investigation, but up to the present none of the materials examined appears to be without some particular disadvantage.

Mathematical Analysis of the Switching Process

A basic appreciation of the switching characteristic of a ferroelectric store can be made by a consideration of the hysteresis loop of the material.

In order to simplify the analysis, it is assumed that the loop is rhomboid in shape, with a slope of tan ϕ at saturation and a slope of tan θ during switching, see Fig. 4. It will be seen later that a departure from this idealized form of loop, involves only minor modifications in the theory.

Let it be assumed that the ferroelectric has been polarized and that its state is defined by the point A. A step function voltage V is applied to a circuit in which the ferroelectric is in series with a resistance R.

The loop has been described by the variation of polarization charge Q against voltage V, rather than by charge density D against field strength E.

When the switching voltage V is applied the charge is varied along the path A B D C.

As is shown by further analysis, the change of Q from A to B occurs in a short time interval following the leading edge of the voltage pulse.

At any time t later than this interval let the voltage across the ferroelectric be $V(t)$, where $V_1 > V(t) > V_0$.

The current flowing in the circuit at this time can be expressed as: $i = \dfrac{V - V(t)}{R}$

and $Q(t)$, the change of polarization from point B to D, is given by—

$$Q(t) = \int_0^t i\,dt = \int_0^t \left(\frac{V - V(t)}{R}\right) dt.$$

Now, between B and D the rate of change of charge with voltage

$$dQ/dV = 1/\tan\theta$$

or

$$Q(t) = \frac{V(t) - V_0}{\tan\theta}$$

Therefore

$$\frac{V(t) - V_0}{\tan\theta} = \int_0^t \left(\frac{V - V(t)}{R}\right) dt \;\ldots\ldots\; (1)$$

The solution for $V(t)$ is in the form:

$$V(t) = A\,e^{\alpha t} + B.$$

Substituting in equation (1) this gives:—

$$\frac{Ae^{\alpha t}}{\tan\theta} + \frac{B}{\tan\theta} - \frac{V_0}{\tan\theta} = \frac{Vt}{R} - \frac{Ae^{\alpha t}}{\alpha R} - \frac{Bt}{R} + \frac{A}{\alpha R}$$

Comparing coefficients:—

$$\alpha R = -\tan\theta$$

$$B = V$$

and

$$A/\alpha R = \frac{B - V_0}{\tan\theta}$$

from which $\alpha = -\tan\theta/R$, $B = V$ and $A = -(V - V_0)$

Therefore

$$V(t) = -(V - V_0)\exp-\left(\frac{t\tan\theta}{R}\right) + V. \;. (2)$$

The maximum value of $V(t)$ during switching is V_1. Substituting V_1 for $V(t)$ in equation (2)

$$(V - V_0)\exp-\left(\frac{t\tan\theta}{R}\right) = V - V_1.$$

or

$$-t(\tan\theta)/R = \ln\left(\frac{V - V_1}{V - V_0}\right)$$

Therefore

$$t = \frac{R}{\tan\theta} \times \ln\left(\frac{V - V_0}{V - V_1}\right)$$

From this, the following conditions may be deduced:—

(a) For complete switching V must be $>V_1$.

(b) For no disturbance of polarization V must be $<V_0$.

(c) Since pulses applied to unused elements of a matrix are usually equal to $V/3$, then $V>V_1$ and $V/3 < V_0$.

Therefore, storage in a matrix under these conditions is impossible unless $V_0 > V_1/3$.

(d) Since tan θ is a function both of the D-E hysteresis loop and the physical dimensions of the storage element, it may be expressed as

$$\tan\theta = dV/dQ = d/A\,.\,dE/dD = d/AK.$$

Where d = thickness of the element

A = electrode area

K = incremental permittivity

Then the switching time t

$$= RA/d\,.\,K\ln\left(\frac{V - V_0}{V - V_1}\right)$$

That is, t is directly proportional to series resistance R and to electrode area A and inversely proportional to the thickness d.

(e) The dependence of t upon the applied voltage V may be clarified by expansion of $\ln\left(\frac{V - V_0}{V - V_1}\right)$

Putting $\ln\left(\frac{V - V_0}{V - V_1}\right) = \ln\left(\frac{1 - (V_0/V)}{1 - (V_1/V)}\right)$

$$= \ln(1 - V_0/V) - \ln(1 - V_1/V)$$

the expansion becomes:—

$$\ln\left(\frac{V - V_0}{V - V_1}\right) = \left(\frac{V_1 - V_0}{V}\right) + \left(\frac{V_1^2 - V_0^2}{2V^2}\right) + \left(\frac{V_1^3 - V_0^3}{3V^3}\right) + \ldots\ldots$$

If V is $\gg V_1 - V_0$, which is usually the case, then a first order approximation gives t inversely proportional to V.

If the loop is curved at B and C then two modifications should be made:—

(1) The points B and C are no longer clearly defined and should be taken further along the saturation line.

(2) The simple exponential solution of equation (1) is not strictly true.

However, it is felt that, despite these inaccuracies, the agreement between the preceding analysis and a more rigorous one would be so close that a more careful treatment involving an extremely difficult expression is unjustified.

Material Considerations Involved in Forming the Basis of a Ferroelectric Storage System

For purposes of binary digital storage the change of state of the dielectric necessary to indicate a 'one' or a 'zero' is, as has already been stated, defined by internal polarization. This change may be detected by the variation of charge flowing to the capacitor. Hence it is important at this stage to consider theoretical values of the change in capacitance of a storage cell under these differing conditions.

Since polarization is effected by domain orientation[11], in a similar manner to that operating in the ferromagnetic case, it would seem to be desirable that the electric field perpendicular to the electrodes should be much larger than the fringing fields, otherwise the domain orientation would not be clearly defined. Moreover, the field in one storage cell should have a minimum effect on the polarization of an adjacent cell.

Let the packing density required be N bits (or binary digits) per square centimetre, and let the dielectric thickness be d centimetres.

Hence, if the unit storage cell for one bit is formed of square electrodes, of length $b\,d$ centimetres, and if the minimum spacing possible between adjacent cells to satisfy the above conditions is also $b\,d$ centimetres

then $4\,b^2\,d^2 = 1/N$ cm^2,

or $$d = \frac{1}{2b\sqrt{N}}\ \text{cm}$$

If b could be specified, N could then be defined by the dielectric thickness d.

The incremental permittivity of a barium titanate crystal may vary from as much as 10^5 to as little as 200. This seems to indicate a ratio of capacitance of the order of 500:1, but this ratio would inevitably be reduced by the effects of fringing, imperfect contact, etc.

The permittivity K of the most suitable barium titanate ceramics now available is of the order of 5 000, and the maximum capacitance ratio obtainable for a ceramic dielectric of this type may possibly lie between 50:1 and 100:1. In addition to this, the thinnest samples of this ceramic at present obtainable by moulding techniques have a thickness of about 1mm, compared with 0·05mm, which has been obtained for barium titanate crystals.

However, assuming a maximum value of permittivity of 10^5, the maximum capacitance of a single storage cell becomes:

$$C = \frac{1/4N \times 10^5}{4\pi + 1/2b\sqrt{N} \times 9 \times 10^{11}}\ \text{farads}$$

$$= \frac{b\sqrt{N}}{72\,\pi N}\mu\text{F} = \frac{0{\cdot}0044b}{\sqrt{N}}\mu\text{F}$$

Suppose, for example, that $b = 3$ and $N = 100$, then the thickness of the dielectric is given by:

$$d = 1/60\ \text{cm} = 0{\cdot}0066\text{in}$$

$$\text{also, } C = \frac{0{\cdot}0044 \times 3}{10}\mu\text{F} = 0{\cdot}00132\mu\text{F},$$

and when the dielectric is polarized in the opposite direction, C becomes equal to 2·5pF approximately.

Factors Affecting Switching Operation

The switching operation of a ferroelectric capacitor can be examined by investigating the variation of response time and maximum current flow with field intensity. These characteristics are bound up with the rectangularity of the hysteresis loop. The analysis of the hysteresis curve already given shows that switching time t_s is directly proportional to dielectric thickness d. There is also a logarithmic relationship between t_s and field strength, t_s being approximately inversely proportional to the applied electric field for field intensities greater than about 2·5kV/cm.

The experimental results illustrated by Fig. 5 confirm the latter relationship, and show how switching time for a sample barium titanate crystal capacitor, having a dielectric thickness of 0·1mm, is increased from a value of 1·3μsec when a potential of 50V is applied across the crystal, up to 250μsec as the applied potential is reduced to 9V. The inverse switching time characteristic is linear for applied voltages greater than about 15V and switching times below 10μsec. It was not possible to plot t_s as a function of d, as all the available samples were of the same thickness, but this relationship has been verified elsewhere[9,12].

From Fig. 6 it can be seen that for field intensities higher than about 3kV/cm (or 30V applied across the 0·1mm thick dielectric), the maximum switching current I_s is almost directly proportional to field strength, and also increases with electrode area. Thus for a sample capacitor having an area of 0·03mm^2, the values of I_s, for applied fields ranging from 2·5kV/cm to 5kV/cm, increased from 5mA to 14mA, while for a capacitor of area 0·67mm² I_s increased from 53mA to 210mA for a similar variation of applied potential.

The operational usefulness of a barium titanate crystal memory store is limited by a number of other factors

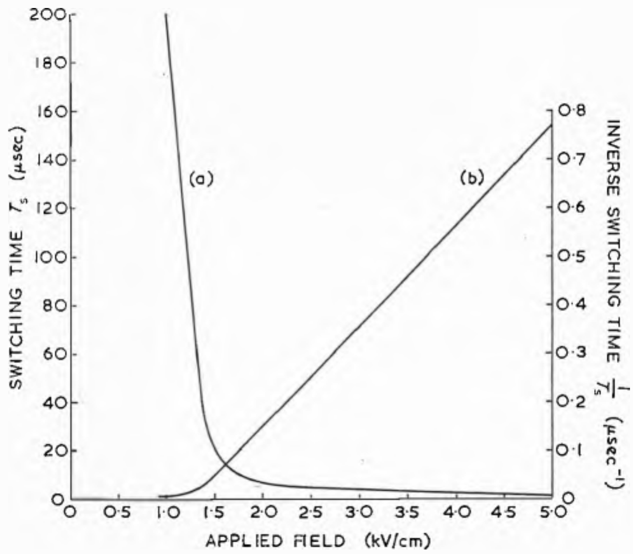

Fig. 5. Functions of applied field, at room temperature. *(a) Switching time T_s (b) Inverse switching time $1/T_s$*

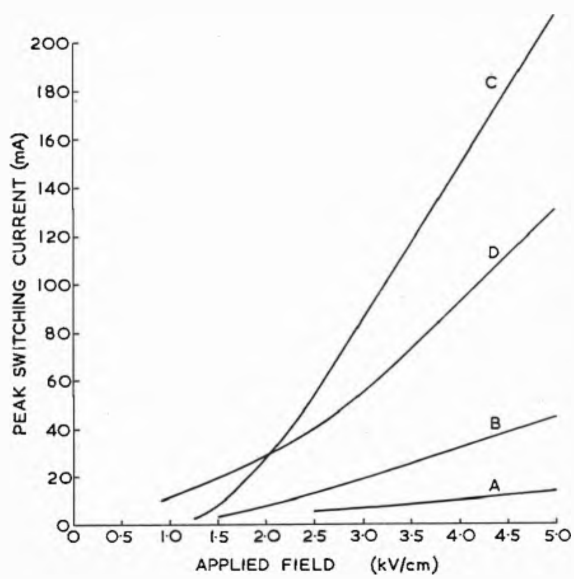

Fig. 6. Peak switching current as a function of applied field (room temperature)

which include effects due to: temperature variations, mechanical constraints, source impedance, fringing and the disturbances caused by fractional voltage pulses. These effects have also been investigated, and are discussed in subsequent sections.

Methods of Measurement

The first series of measurements taken (Figs. 5, 6 and 9, 10) were of switching time and the maximum switching current:

(a) As functions of applied field at room temperatures.

(b) As functions of temperature with applied field constant.

For these tests polarization of the sample crystal was reversed by the application of alternate positive and negative square pulses having a rise time of 0·1μsec, and readings were taken (by means of a Tektronix oscilloscope) of the current flow through a 10Ω series resistor (the output impedance of the signal generator was 550Ω). A simplified diagram of the circuit used is given in Fig. 7. The switching current transients (corresponding to binary 'ones') of some sample capacitors, together with non-switching transients (corresponding to binary 'zeros') obtained by passing a series of unidirectional square pulses through these samples, have been reproduced in Fig. 3.

The charge held by each of these samples has been determined by measuring the difference in the areas covered by its respective transient curves. From examination of these transients the most suitable of these capacitors for storage purposes would appear to be sample C. (This sample also gave the best hysteresis loop). Unfortunately, this was also the largest area capacitor (0·67mm²) and it required a far larger switching current than any other of the samples.

The next series of measurements taken were of (1) output voltage and (2) signal-to-noise ratio (i.e. the ratio of read-out 'one' to read-out 'zero') as functions of applied field

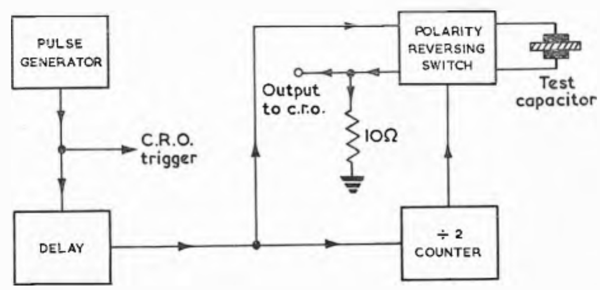

Fig. 7. Simplified diagram of switching circuit for observations of transient current

at room temperature. A typical example of the latter series of results has been plotted in Fig. 11. These tests were carried out in a similar manner to those of the earlier series, except that in this case alternate pairs of positive and negative square pulses were applied to the sample crystal, while readings were taken of output voltage across a series reactance consisting of a 0·1μF capacitor shunted by a 1kΩ resistor (see Fig. 2). By this means it was possible to record the output voltage for '1' and '0'. These measurements were also used for examining the variation of output voltage and signal-to-noise ratio with electrode area.

Some of the results of these experiments have already been discussed, the remainder are dealt with under the relevant headings in the next section, in which two further series of tests:—

(a) Hysteresis loop measurements taken over a range of frequencies for different values of source impedance.

(b) Application of fractional voltage pulses to a sample ferroelectric capacitor,

are also discussed.

Additional Limiting Factors Found by Experiment

Effect of Mechanical Constraints, Source Impedance and Frequency Variations on Hysteresis Loops

Observations of the dielectric hysteresis loops were carried out by applying potentials proportional to electric field strength E_s and polarization P_r to the X and Y plates respectively of an oscilloscope, while a sinusoidal electric field was applied across the ferroelectric capacitor under test. The circuit used for this purpose was a modified form of De Sauty capacitance bridge.

For this series of tests the frequency of the applied sinusoidal field was increased from 25c/s to 5kc/s in steps, and for each value of frequency, the field strength was adjusted to that necessary just to maintain the hysteresis loop. Readings were taken of V_0, V_1 and Q_0 (see Fig. 4) in each case. Separate sets of results (all at room temperature) were taken with the source impedance set at values ranging from 15Ω to 600Ω.

(a) *Mechanical Constraints*

A constant mechanical stress maintained on the crystal may produce a fixed piezoelectric polarization on which the polarization due to the applied field becomes superimposed.

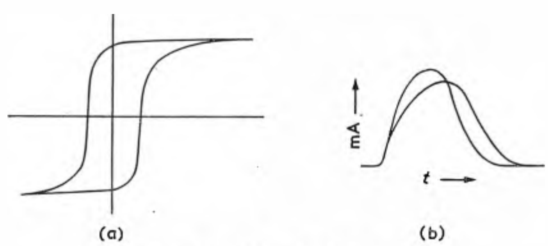

Fig. 8. Effects of mechanical constraints

(a) Asymmetrical hysteresis loop
(b) Dissimilarity in switching transients between the two polarized states

This would cause the hysteresis loop to become distorted and asymmetrical, so that the times taken for switching and reversal between the two stable states would be dissimilar. These effects are illustrated in Fig. 8, and have been examined in detail by other investigators[6,13]. This emphasizes the care that is essential in the design of suitable crystal mountings.

(b) *Source Impedance*

The source impedance is effectively in series with the capacitor under test, and the analysis of the hysteresis loop given in a previous section shows that switching time is inversely proportional to series resistance.

Examination of the hysteresis loops of the sample capacitors, taken at a fixed frequency (500c/s) for different values of source impedance Z_s, shows the loop to be squarest in shape for the lowest obtainable value of Z_s, i.e. 15Ω.

(c) *Frequency Variations*

The shape of the hysteresis loop is reasonably square even at low frequencies, but the corners of the loop become gradually squarer with increase of frequency, reaching an optimum at 500 to 600c/s, after which the remanent polarization and the slope of the loop gradually decrease with further increase of frequency.

For different values of source impedance the charge, Q_0, remained constant, and on the largest available sample it was 0·40μC for frequencies up to about 600c/s, slowly decreasing in value after that with increase of frequency. The energy dissipated in switching reached a maximum over the range 500c/s to 1·5kc/s, and for this sample it was 5·5μJ for a Z_s of 15Ω, increasing slightly with increase of Z_s. The value of Q_0 as measured from the hysteresis loop is very similar to the value obtained in the earlier experiments by plotting the area under the switching transients (Fig. 3).

The effects caused by frequency variation show that although an estimation of the value of particular storage elements is obtainable from single frequency loops, complete information cannot be gained by this method as under operating conditions rectangular pulses, containing in effect a wide band of frequencies, will be applied to these materials.

Effect of Temperature Variation

Two series of tests were carried out in order to examine the effect of temperature variations on ferroelectric capacitors. The first series, see Figs. 9 and 10, has already been

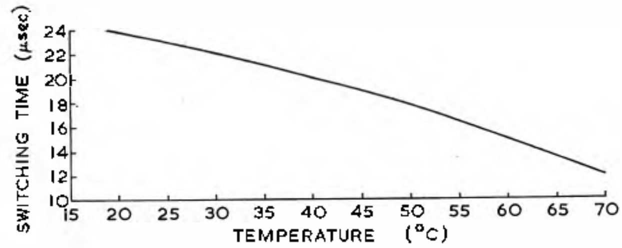

Fig. 9. Switching time as a function of temperature (field constant)

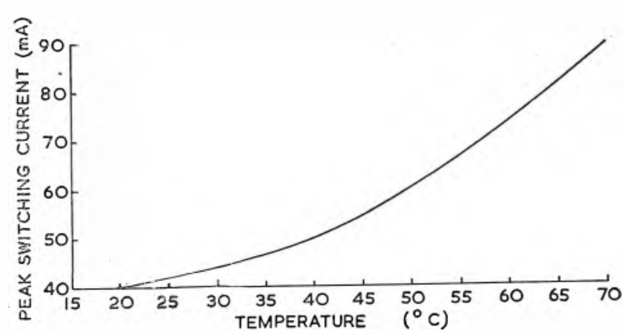

Fig. 10. Peak switching current as a function of temperature (with constant field)

described, and consisted of measurements of switching time and maximum switching current as functions of temperature at a constant applied field.

From these results it is clear that an increase in temperature from ambient by 45°C causes:

(a) A reduction of about 50 per cent in switching time.

(b) An increase of roughly 100 per cent in maximum switching current.

The second series of observations was made on the hysteresis loop of a sample capacitor in a similar manner to that described in this section, but in this case the frequency of the applied sinusoidal field was kept constant at 500c/s while temperature was increased up to 65°C. Readings were taken at intervals of about 10°C. These results indicated that:

(a) The applied field necessary just to maintain the hysteresis loop decreases appreciably with increase of temperature.

(b) The loop narrows with increase of temperature.

Other investigators[15] have shown that the storage properties of barium titanate crystals begin to diminish rapidly at temperatures above 80°C or thereabouts, and the hysteresis loop collapses at 120°C (i.e. the Curie point).

It has also been pointed out[14] that due to heating effects the pulse repetition frequency used to operate a practical size barium titanate crystal storage unit should not exceed about 20kc/s.

EFFECTS OF FRINGING

Tests were carried out on a number of sample capacitors, of the same dielectric thickness but having different surface areas, in order to examine the relationship of:

(a) Output voltage.

(b) Signal-to-noise ratio or discrimination, to applied field at room temperature in each case.

A description of the methods employed has already been given, and Fig. 11 illustrates a typical example of the latter series of tests. The results were also used for an investigation of the variations of output voltage and signal-to-noise ratio with surface area at a constant applied field.

These results indicate that:

(a) With samples having comparatively large surface area the output voltage increases rapidly with applied potential for field intensities up to about 3·5kV/cm, above which the rate of increase of output signal falls rapidly.

(b) For samples having very small surface area (i.e. about 0·04mm^2) and hence a low ratio of surface area to dielectric thickness, the rate of increase of output voltage with applied field does not fall appreciably, even for high intensity fields.

(c) For most of the samples the discrimination, or ratio of read-out 'one' to read-out 'zero', measured against applied field shows a peak at field intensities of the order of 3·5kV/cm.

(d) Both output voltage and discrimination show a marked rise with increase of surface area.

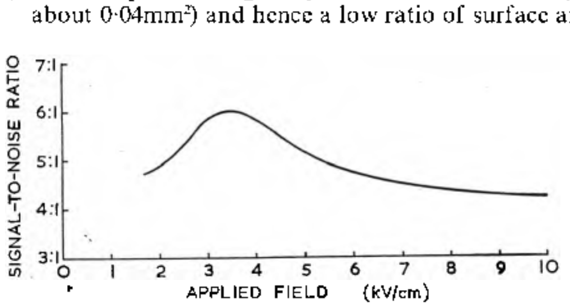

Fig. 11. Signal-noise ratio as a function of applied field

Results of this type have been commented upon elsewhere[14], and must be attributed mainly to the fringing effect. In the earlier analysis it was shown that the signal-to-noise ratio for a barium titanate crystal storage cell can be higher than 100:1. It was also deduced that effects such as fringing of the electric field beyond the electrodes and imperfect contact, may seriously reduce this value. The units most adversely affected by these difficulties and by stray capacitance must obviously be those having the smallest area electrodes. Hence the minimum surface area of a ferroelectric capacitor is clearly limited, as it would be difficult to operate a practical storage unit having a signal-to-noise ratio less than say 3:1.

DISTURBANCES DUE TO FRACTIONAL VOLTAGE PULSES

A series of tests was carried out in order to determine the effect of fractional voltage pulses on sample capacitors. The method of testing was developed from that used for the last investigations.

The pulse width necessary for switching the sample capacitor at a particular field intensity V was determined from previous results, and pulses of amplitude $V/2$ were then applied to the crystal. Read-out voltages for the disturbed 'zero' were tabulated against the number of applied pulses until the ferroelectric was completely switched. The test was then repeated with the pulse amplitude reduced to $V/3$. Pulses were counted by means of a series of binary counters.

These results have shown conclusively that the effect of such fractional voltages is cumulative, and that a finite number of fractional pulses will switch any crystal of this type. For example, consider sample D, a 27V pulse (field intensity 2·7kV/cm) would reverse the polarity of the ferroelectric in 4μsec, while with a 9V pulse applied the switching would take approximately 250μsec. The application of 80 pulses, each 9V in amplitude and 4μsec in duration, was sufficient to switch the crystal.

The Application of Switching to a Matrix Formation

ESTIMATED EFFECT OF DISTURBANCES ON UNUSED ELEMENTS

Consider a matrix store comprising $n \times n$ similar elements, each having capacitance C. Then the input circuit of any storage cell is equivalent to a capacitor of value

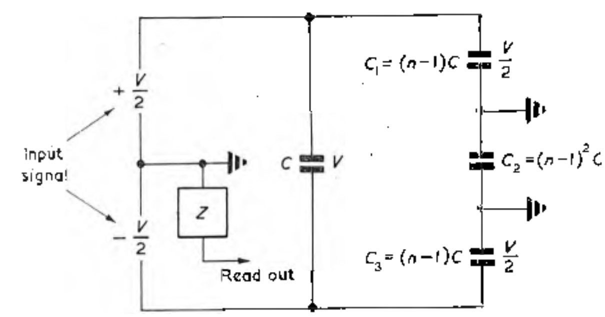

Fig. 12. Simplified equivalent circuit of matrix store

C being shunted by a series combination C_1, C_2 and C_3. Of these C_1 and C_3 each consists of a parallel arrangement having total capacitance $(n-1)c$, while C_2 is a similar array equivalent in value to $(n-1)^2c$. Let the binary data to be recorded on a particular element be applied to the corresponding row and column of the matrix in the form of negative and positive half pulses respectively, the remaining rows and columns being earthed. Hence there will be a voltage V applied across the selected element, and voltages equal to $V/2$ acting upon the arrays C_1 and C_3, while C_2 is maintained at earth potential, see Fig. 12.

The selection ratio, i.e. the ratio between the voltages acting on the selected and unselected elements of the store, obtained by this method is thus 2:1. This ratio may be increased to 3:1 if a bias is applied to the unused elements, the value being chosen to ensure that the voltage distribution across C_1, C_2 and C_3 is equal. With the latter system in operation, the applied signal should be increased in amplitude by $V/3$ to compensate for the bias.

On this basis it becomes necessary to consider the effect of fractional voltages equal to $V/3$ on unselected elements. The tests discussed in the previous section have proved the effect of these disturbances to be cumulative, so that the sample capacitor was switched by the application of about 80 fractional pulses when the applied signal was 27V in amplitude. The maximum number of fractional pulses which may be tolerated is, however, limited to that number which will reduce the signal-to-noise ratio to say 3:1. With the example quoted this took place after only about 20 of these disturbances had been applied.

It was not possible to measure the switching time t_s of sample D for values of t_s greater than 250μsec, but for low intensity fields, where only a few volts would be applied across the crystal, t_s would be very large. With 15V applied, the switching time is about 20μsec, so that it is probable that several hundred disturbances would then be necessary to reverse polarity, and under these conditions regeneration may be required after about every 100 disturbances.

Physical Dimensions

Owing to the adverse effect on discrimination caused by such factors as fringing and stray capacitance, there are practical limitations to the packing densities attainable with a single crystal ferroelectric matrix store. The signal-to-noise ratio of the small area sample capacitors was very low, although this would probably be a little higher with preselected units. Hence packing densities of 1 000 bits/cm^2, which have been envisaged[16] for dielectric thicknesses of about 0·1mm, seem to be much too high for storage purposes, and values of the order 200 to 300 bits/cm^2 may be more practicable. Due to the capacitive effects of unselected elements previously mentioned, the maximum number of storage cells which may be included in any matrix is also limited.

Non-Destructive Read-out

The previously described method of sensing recorded information automatically returns the storage capacitor to its binary 'zero' state, so that a regenerative process is necessary if data is to be retained after read-out. It would be desirable for many computing applications to have a satisfactory system of non-destructive read-out.

Polarized ferroelectric materials have other inherent properties including pyroelectric and piezoelectric characteristics. The former property has been employed[9] for carrying out observations of switching at low intensity fields, but its effects are too slow to be of value for computing purposes. Piezoelectric effects, however, have already been briefly discussed in the section dealing with mechanical constraints, and have been utilized in devices for non-destructive sensing. This is possible because the application of ultrasonic vibrations to a storage cell would produce an output at the applied frequency, having phase relationships dependent upon the state of polarization of the ferroelectric.

A device employing similar principles that has been described is the Mullenbach Capabit[17]. This consists essentially of two small ceramic disks, each conductively coated on both sides, cemented together to give mechanical coupling. Storage is carried out by the application of a polarizing pulse to one of these disks, and for recovery of information a signal is applied to the sensing circuit, which by causing mechanical distortion of the second disk, produces an output potential from the storage element.

Other modes of non-destructive read-out have also been examined[18], but no really satisfactory system appears to have yet been devised. This is because the process of recording data on selected elements of a ferroelectric matrix appears of necessity to involve the application of disturbances to unused cells, with the inevitable switching of these elements within a finite time. Hence some form of regeneration of stored data is necessary at intervals irrespective of the method of read-out that may be employed.

Further Applications of Ferroelectrics

In addition to their utilization for information storage, other possible applications of ferroelectric materials for digital computing which have been investigated[14,16,19] include their incorporation in switching and trigger circuits and in shift registers.

Conclusions

The sample barium titanate storage capacitors suffer from a disadvantage caused by a gradual variation in their switching characteristics, this change being almost proportional to the number of times that switching is carried out. This has been remarked upon elsewhere[14], where mention is made that due to progress in manufacturing and processing techniques it may be now possible to produce units relatively free of this defect.

With barium titanate, as with certain other ferroelectrics, there appears to be no threshold field below which switching cannot take place. Because of this factor, any matrix store constructed from available materials should be designed for switching potentials not exceeding about 1·5kV/cm, and hence pulse durations greater than 20μsec. Under these conditions a store may readily be operated by transistor drive circuits because of the low switching currents and voltages required. It is essential for stored information to be regenerated at intervals.

Packing densities greater than about 200 to 300 bits/cm^2 may not be practicable for storage purposes, due mainly to fringing effects.

Also, methods of mounting and heat dissipation must be carefully examined. Due to heating of the dielectric during switching, the maximum pulse repetition frequency used to operate any practicable size crystal storage capacitor is limited to 20kc/s.

It may be hoped that some of the difficulties encountered in the use of ferroelectrics will be overcome as the chemical and physical development of these materials becomes advanced, and when their peculiarities are more fully understood.

Acknowledgments

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A Sampling Comparator

By Arieh Fischmann-Arbel*, B.E.E., M.E.E., A.M.Brit.I.R.E.

A sampling comparator compares the instantaneous value of two waveforms at constant intervals of time. The response of the comparator described in this article extends from d.c. to a practical limit of half the sampling frequency. Applications of the comparator to a delta modulator, and to a binary quantizer are given.

THE comparator described in this article compares two voltages at constant intervals of time, a process required for example in voltage encoders[1]. The output of such a comparator is a positive, negative, or zero amplitude pulse proportional to the difference between the two input voltages, and in the unit described two outputs in anti-phase are provided.

A block diagram of the complete comparator is shown in Fig. 1. The pulsed comparator V_1 compares the two voltages v_1 and v_2 each time a clock pulse arrives from the clock pulse generator. A negative pulse appears at each of its two outputs, the difference between these representing $(v_1 - v_2)$ in amplitude and polarity. Consequently, the push-pull output of the following differential amplifier provides pulses whose amplitudes are proportional to $(v_1 - v_2)$, and $(v_2 - v_1)$ respectively.

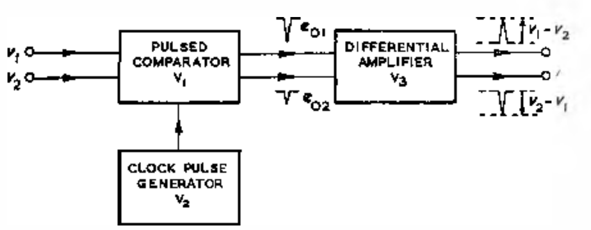

Fig. 1. General arrangement of the comparator

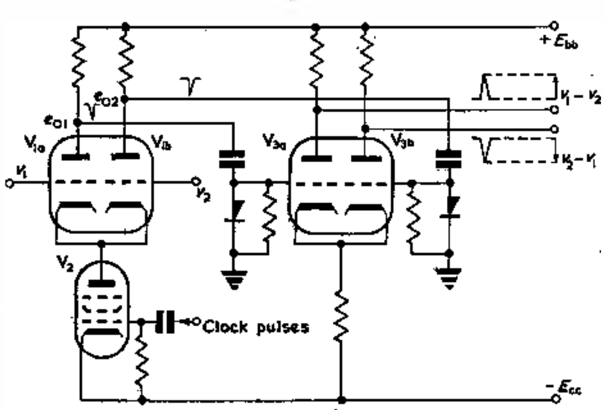

Fig. 2. The sampling comparator

Fig. 2 shows the circuit diagram of the sampling comparator. The voltages v_1 and v_2 are connected to the grids of V_{1a} and V_{1b}. Positive clock-pulses are connected to the grid of V_2 which acts as common cathode load for V_{1a} and V_{1b}. In the absence of any clock-pulses V_2 is cut off, and consequently V_{1a} and V_{1b} are cut off too.

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It may be shown that, V_2 being assumed to act as an ideal constant current source for the duration of an idealized rectangular clock pulse appearing at its grid, $(e_{o1} - e_{o2}) = \mu R_L(v_1 - v_2)/(r_a + R_L)$ in the usual notation referred to V_1.

If positive pulses of constant amplitude are applied to the grid of V_2, negative pulses of amplitude e_{o1} and e_{o2} will therefore appear at the anode of V_{1a} and V_{1b} respectively. In accordance with the above expression, the difference between these two pulses will be a pulse of an amplitude proportional to $v_1 - v_2$. This subtraction is accomplished in the following differential amplifier[2] consisting

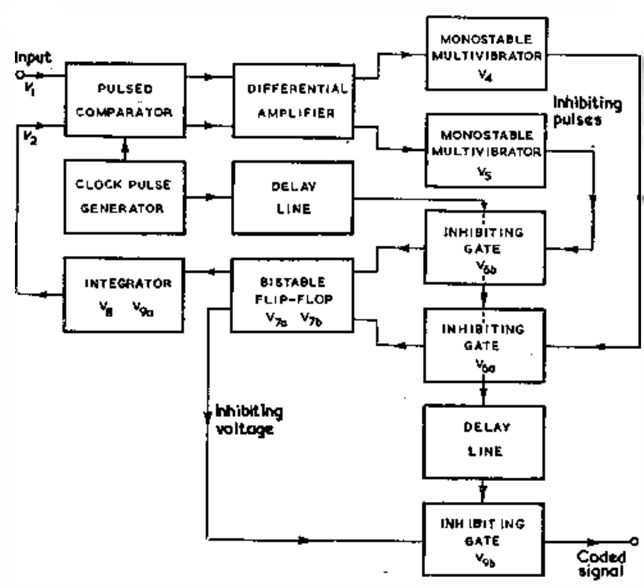


Fig. 3. The encoder

of V_{3a} and V_{3b}. The waveforms shown in Fig. 2 apply to the case $v_1 > v_2$.

The grids of V_1 should be connected to fairly low impedance voltage sources (v_1, v_2), as the considerable voltage swing at the V_1 cathode will otherwise be transmitted to them through the grid-cathode capacitance.

It is an important feature of this comparator that, although its response includes d.c., the first stage converts the difference between v_1 and v_2 into pulses; these pulses may be amplified in cases where the sensitivity should be increased, and the total drift will be determined solely by the first stage. The highest frequency of response equals half the sampling frequency. The maximum sampling frequency, and the minimum width of the sampling pulses are determined by the pulse response of V_1 and V_3, whose anode voltage should reach a steady state during the

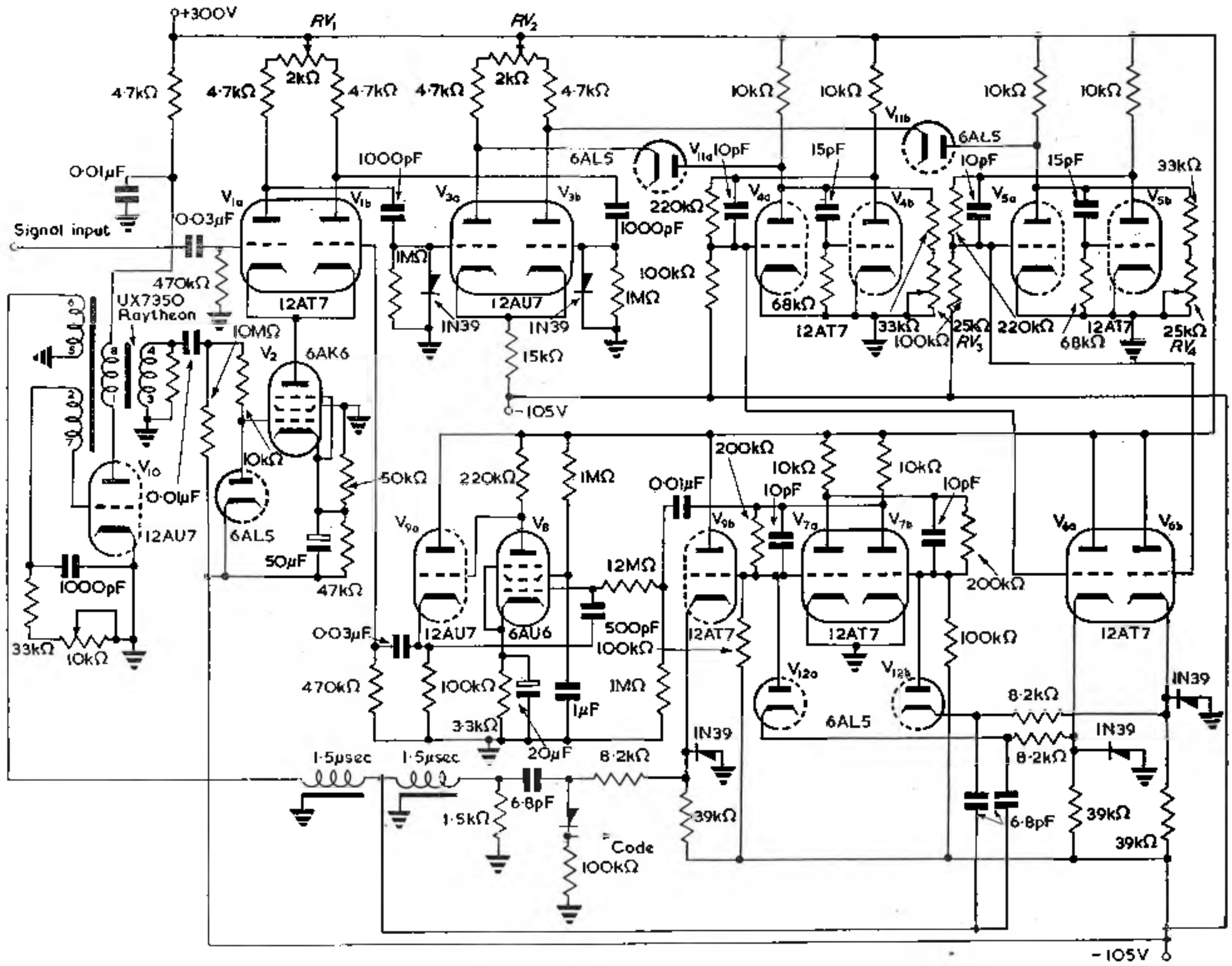
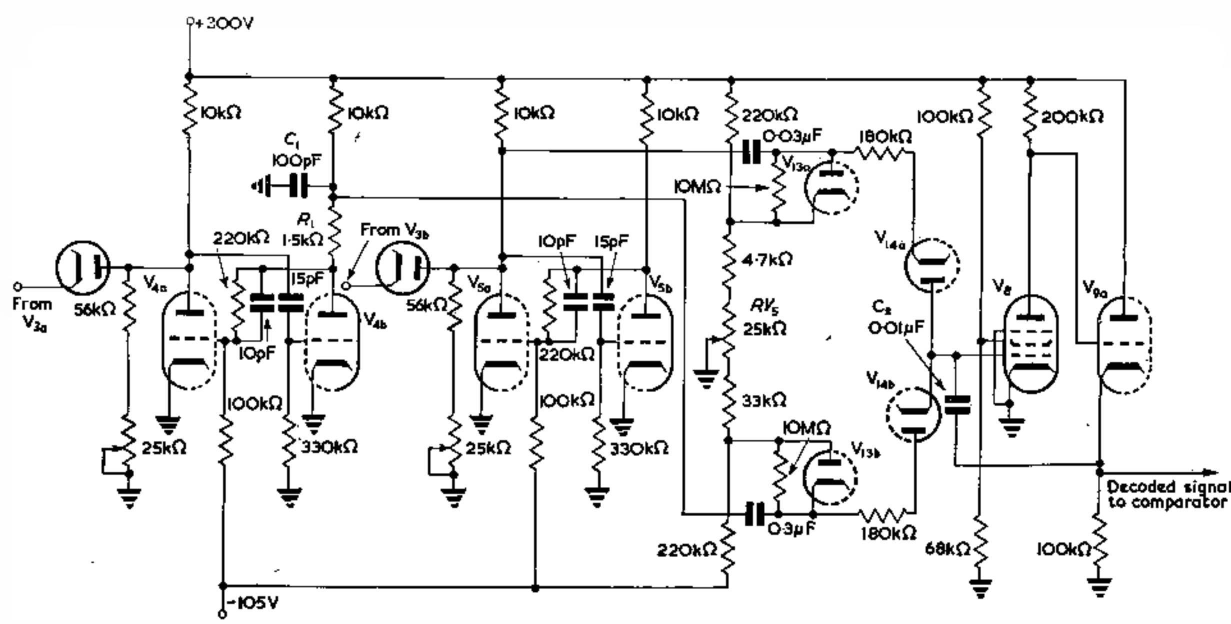


Fig. 4. The decoder

Fig. 5. Modified decoder



sampling pulses as well as during the interval between them. Conventional pulse techniques may be applied in the design of these circuits.

In applications where the comparator is used in a feedback loop, it is of importance that any transients which might appear at v_1 or v_2 during the interval between the sampling pulses should not appear at the anodes of V_3. As that valve is cut off between the sampling pulses, this condition is automatically fulfilled.

This comparator may be employed advantageously in an encoder for 'delta modulation'. The 'delta' code has been described elsewhere[3,4] in detail. In its simplest form, the presence of a pulse indicates an increase of the signal by one quantum, whereas the absence of a pulse indicates a decrease of the signal by the same amount. A single quantum will further be called q. A d.c. signal is encoded into a pulse train of half the sampling frequency, which if decoded yields a triangular waveform having an amplitude equal to q, and oscillating around a certain value at half the sampling frequency.

In an encoder for such a code recently described[5] the comparison between the original and decoded waveform, and the synchronization in time are accomplished in two different units. It was felt that these two operations might with obvious advantage be combined in a single unit, the sampling comparator. The block diagram of the resulting encoder is shown in Fig. 3. It is an 'inverted' (feedback) decoder[1] performing essentially three functions: (a) it decodes the encoded signal in an integrator; (b) it compares this decoded signal with the original one in the sampling comparator; (c) it produces the code in accordance with the result of the comparison.

The decoder consists of an integrator (V_8, V_{9a} in Fig. 4) whose input is connected through a capacitor to one anode of the bistable flip-flop V_7. V_{7b} conducts as long as there are pulses present in the coded signal, causing the output of the integrator to rise at a constant rate; it is cut off in absence of any pulses in the code, causing the output of the integrator to drop at the same rate. Consequently, the integrator output represents the decoded waveform.

The sampling comparator compares v_2, the output from the decoder, with v_1, the input voltage. As described before, the pulses at the anodes of V_3 are of opposite polarity and their amplitude is proportional to $v_1 - v_2$. Each anode of V_3 is connected through the diodes V_{11a} and V_{11b} to the monostable multivibrators V_4 and V_5, which will be triggered by only such pulses that exceed the negative threshold amplitude. This threshold may be adjusted for each multivibrator individually by RV_3 and RV_4 respectively. The short pulses from these multivibrators represent the output from the sampling comparator, and three cases must be considered.

(1) If $|v_1 - v_2| < \frac{1}{2}q$, no pulse will be transmitted.

(2) If $v_2 - v_1 \geqslant \frac{1}{2}q$, a pulse will be transmitted from V_4.

(3) If $v_1 - v_2 \geqslant \frac{1}{2}q$, a pulse will be transmitted from V_5.

In case (1), the encoded signal should be a pulse train of half the sampling frequency. The block diagram (Fig. 3) shows that in this case no inhibiting pulse will reach either gate (V_{6a}, V_{6b}) and the trigger pulses will reach the grids of the bistable flip-flop V_7. The latter acts as a frequency divider, and the inhibiting voltage derived from it will prevent every second pulse from appearing at the output of the inhibiting gate V_{9b}. Therefore, the coded signal is obtained directly from the output of this gate.

In case (2) a pulse will be transmitted by V_4 inhibiting the gate V_{6a}. A negative trigger pulse will therefore appear at the grid of V_{7b} only. This pulse will either change the state of the flip-flop if V_{7b} was previously conducting, or leave the circuit unchanged if V_{7b} was cut off already. In neither case will a pulse appear at the output of the inhibiting gate V_{9b}.

In case three, a pulse will be transmitted from V_5. A negative trigger will now appear at the grid of V_{7a} only. Applying the same reasoning as in the preceding paragraph, it may be seen that a pulse train with a repetition rate equal to the sampling frequency will appear at the output of the inhibiting gate V_{9b}.

In all these cases, the coded signal conforms with the rules previously laid down for the delta modulation code. The input to the decoder—integrator V_8, V_{9a} is obtained from the anode of V_{7b}.

The low frequency response of this encoder is limited by the capacitive coupling between V_{9a} and V_{1b}. It may be extended to very low frequencies if this capacitor is eliminated, and if the input voltage v_1 is connected directly to the grid of V_{1a}. D.C. cannot be included because V_8, V_{9a} do not form an ideal integrator.

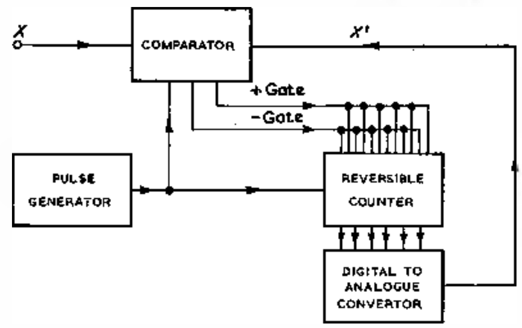


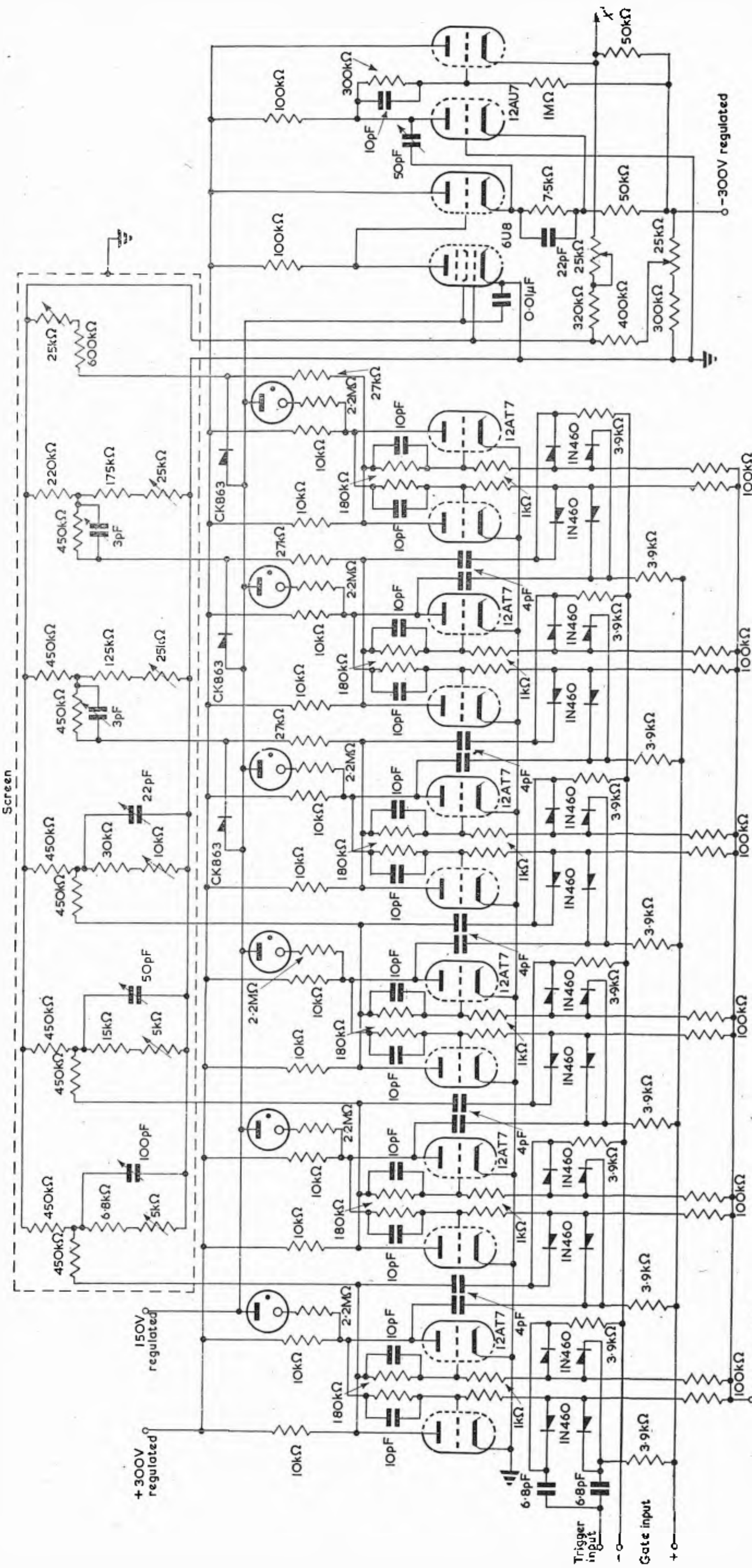
Fig. 6. Arrangement for binary quantizer

The circuit has been designed for the quantization of speech. The sampling frequency is 50kc/s, and the input waveform of a maximum peak-to-peak amplitude of 30V will be quantized into 20 steps of 1·5V each. The maximum slope which may be handled by the quantizer is therefore 1·5V/20μsec, and the frequency response up to 800c/s at maximum amplitude. Above 800c/s, the maximum amplitude of the input signal which may be handled without distortion drops by 6dB/octave, in accordance with the maximum slope condition.

A different kind of delta modulation has been described by Maas[6]. There, a pulse is transmitted for each increase of the waveform by q, and a different pulse for each increase by $-q$. The two kinds of pulses may be transmitted through different channels, or any other way of distinguishing between them in a single channel may be utilized.

A suitable modulator may be readily constructed using the comparator shown in Fig. 4, but a different decoder whose diagram is shown in Fig. 5. The pulses indicating a positive and negative increase of the waveform are obtained from the multivibrators V_4 and V_5 whose time-constant has been changed. For the purpose of decoding, these pulses are fed into the integrator with opposite polarity. If no pulses are present (no change of the encoded waveform), the grid of the integrator is floating. The d.c. performance of this encoder will therefore be determined by the drift of the integrator with disconnected input, which will be considerable unless special precautions are taken. The present design was required to operate over periods of 10sec only, and no difficulty was experienced during such short times of operation.

For each pulse in the code, the capacitor C_3 of the inte-



(a)

Fig. 7. (a) Circuit of quantizer

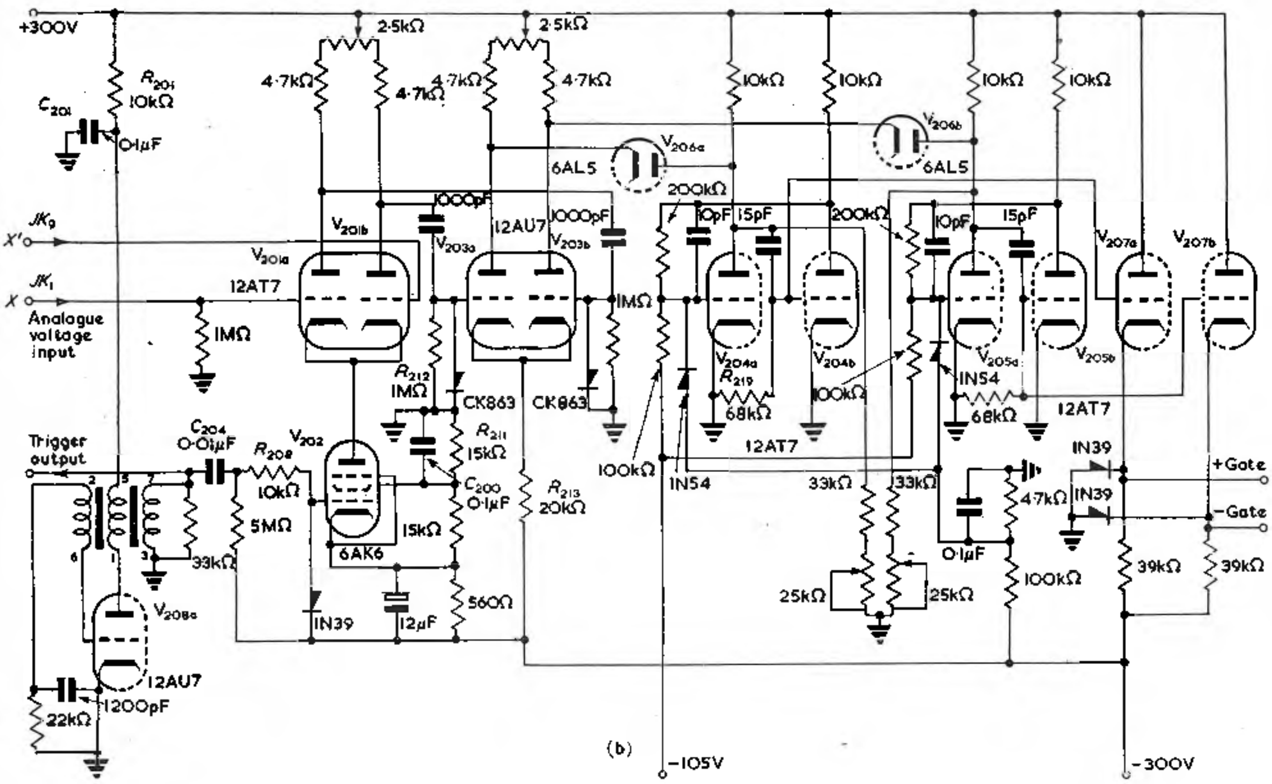


Fig. 7. (b) Circuit of quantizer (continued)

grator should receive or loose an equal amount of charge, depending on whether the coded pulse indicates an increase or decrease of the input voltage. Each time the multivibrator V_4 or V_5 is triggered, a charge transfer takes place. The negative square wave at the anode of V_{5a}, or the positive one at the anode of V_{4b} is utilized to charge, or discharge C_2.

These square waves are clamped by V_{13} to a suitable level, and fed into the integrator through V_{14}. The d.c. restorers V_{13} apply a back bias to the diodes V_{14}, thereby disconnecting the grid of V_8 from the input as long as the multivibrators V_4 and V_5 are in the quiescent state. In addition, this back bias prevents any trigger pulses which are too small to trigger V_4 or V_5, from being transmitted to the grid of V_8.

When the multivibrator V_4 returns to its quiescent state after having been triggered, a negative overshoot appears at the anode of V_{4b}. This overshoot would prevent the d.c. restorer V_{13b} from operating properly, and it is therefore eliminated by the integrating network R_1C_1. No integration is necessary in the anode circuit of V_{5a}, as no overshoot occurs when V_{5a} is cut off.

It should be noted that in order to charge and discharge C_2 by the same amounts, the shape of the negative and the positive square waves need not necessarily be identical, only the area under them should be the same. FV_5 serves to adjust these two areas for equality, its correct position being determined by observing the amplitude of the positive and negative steps at the output of the decoder, and making them equal.

The screen grid of V_8 is not by-passed. A capacitor at this point would introduce transients and seriously impair the d.c. response of the decoder. It would be possible to stabilize the screen voltage with a cold-cathode stabilizer valve, but this was not considered to be necessary.

Another application of this comparator is its use in a binary quantizer, an encoder converting an analogue voltage into a parallel binary code. This is once again an 'inverted' (feedback) decoder[1]. According to the author's knowledge, Barney[7] was the first to build such a binary quantizer. His block diagram is identical with the one employed in this encoder, but the circut techniques used here are completely different from his.

The block diagram is shown for completeness in Fig. 6, and the circuit diagram in Fig. 7. The same comparator and gating circuits as in the delta modulator are employed, and all the other circuits are designed along conventional lines. No further explanation is therefore necessary.

One quantizer built from conventional components was designed for a sampling frequency of 200kc/s, with an input amplitude from +50 to −50V; as there are 6 stages in the counter, $q = 1{\cdot}6$V. The frequency response extends from zero to 1kc/s, at 100V peak-to-peak.

Acknowledgments

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Light Intensity Meters for Local and Remote Indication

By E. F. Hasler* and G. Spurr*, B.Sc., A.Inst.P., F.R.Met.S.

Measurement of daylight illumination is necessary to the electricity supply industry for the prediction of the demand for artificial lighting. This article describes a method of measurement using a photocell as the grid leak of a blocking oscillator. Emphasis is placed on the simplicity, reliability and versatility of the principle.

THE economic operation of the business of the Central Electricity Generating Board is largely dependent upon the degree of accuracy with which the fluctuations in demand for electric power can be assessed in advance. These demands vary from hour to hour and from day to day, and in an interconnected Grid System the amount of available generating plant must follow closely the varying load. It has long been realized that the day-to-day variations of peak loads are largely due to the variations in the weather, but it is only in recent years that detailed routine weather forecasts have been made available to Control Engineers by the Meteorological Office.

There is a high degree of correlation between air temperature and electricity demand and forecasts of temperature can be converted into megawatts of electricity to be expected. On the other hand, the prediction of the demand for artificial lighting is far from simple and only limited success has yet been achieved in relating lighting demand with the variation of daylight illumination. This has been largely due to the lack of suitable instrumentation, but the apparatus described herein which is now being installed on a pilot scale at selected points throughout the country, represents an attempt to remedy this deficiency. The apparatus was originally designed for remote reading, but modified instruments suitable for local indication or recording, and for integration over any required period, both for this and other purposes, are also described.

Design Requirements

The Remote Instrument

In many instances, for the indicated or recorded value of daylight intensity to be of maximum use, the presentation must be made at a point remote from the actual measurement site. For instance, the Control Engineer is normally the person most immediately concerned with an indication of daylight value in the region or regions of high population density; such regions may well be remote from the Control Centre and from each other, requiring more than one instrument. In this respect it should be realized that whereas temperature is usually comparatively slow to change, variations in daylight often occur rapidly and are frequently much more localized. Thunderstorm clouds or fog conditions often cover a limited area and when affecting a compact, densely populated city have caused sudden and very serious increases in the local demand.

The fact that presentation is required at a distance of many miles from the point of measurement, immediately suggests the use of the existing system of telemetering. This system, as it is applied to the electricity supply industry in this country, has been fully described in the literature[1,2,3,4]; briefly, the quantity to be transmitted, whether it be voltage, power, position indication or any similar function, is first coded as a pulse rate in the range 0 to 66·66 operations per minute. To operate the receiving unit satisfactorily a mark-to-space ratio of substantially unity is required, and these pulses are used to switch the output of a voice-frequency oscillator feeding a telephone line. Since the pulses now consist of bursts of an audio-frequency tone, they will operate over standard repeaters and there is virtually no limit to the distance over which satisfactory operation can occur. At the receiving terminal the pulses are usually arranged to operate a 'cup and bucket' system[2] to produce a steady meter reading.

This system, however, is based upon a pulse rate, and the conventional methods of light measurement using barrier layer cells are not readily adaptable to this service. Furthermore, such cells are temperature dependent and age fairly rapidly, particularly in a wet climate. Although it would no doubt be possible to design a coding unit to translate the information from a conventional light meter into a suitable pulse rate and to provide compensation for other limitations, such an equipment would be complex and expensive. It was therefore decided to develop a new instrument to fulfil the functions of measurement and transmission, and to work into a conventional telemeter link.

Apart from the provision of a pulse rate matching the existing standards for telemetering, certain other requirements had to be met for this particular application. It is known that the human eye can adapt itself to a wide range of illumination intensity—it is only necessary to remember the insipid look of artificial light even in poor daylight to appreciate this fact. In any particular instance, the actual level in outside daylight at which artificial lighting is required will depend on many factors, such as the relative size and disposition of the windows, the class of activity involved, the age of the people concerned, and so on. Nevertheless, most of these factors will average out in any community large enough to influence the loading of the Grid System, and the average daylight level at which switch-on occurs is likely to be substantially the same throughout the country. It has been estimated that this level is about 10 kilolux; by comparison, the level of illumination from direct sunlight in summer is 250 kilolux or more. To the Control Engineer, of course, the higher intensities are of little interest, and it is of most value to him for the 'switch-on' value to appear at about half scale on his indicating instrument with the higher levels compressed at the full scale end.

Since the instrument is required to measure the exterior value of daylight, it must obviously be mounted in a shadow-free location with the measurement surface in a horizontal plane. To be properly shadow-free throughout the year, this plane must be above the level of any obstruction, i.e., the measurement surface must have an uninterrupted view of a complete hemisphere of sky. The

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instrument must therefore be mounted at the highest available point on the site, which will usually mean on the highest part of the roof of the power station or office block nearest to the centre of the district concerned. Such a location not only means that the instrument will be fully exposed to all weathers but also that only simple maintenance can be carried out *in situ*. Maximum reliability of components and of construction is therefore essential. Furthermore, complete calibration can only be carried out on an optical bench, which means that the instrument must be removed from the site. To maintain the intervals between successive complete calibrations as long as possible, the design must be such that all component drift, particularly the effects of valve ageing, should be compensated, but with as little additional complexity as is practicable.

Summarizing these requirements, therefore, the design of the remote instrument shall be such that:—

(a) It will provide an output pulse rate within the range of existing telemetering apparatus (0 to 66·66 pulse/min) dependent upon the intensity of illumination.

(b) The scale is compressed at the high intensity and substantially linear at the low intensity levels.

(c) Any ageing effect, such as that associated with barrier layer photocells, and the temperature effect, are both reduced or eliminated.

(d) It will operate reliably for long periods and with a minimum of skilled maintenance.

(e) It will be virtually independent of valve ageing or drift.

(f) It will withstand exposure to all weather conditions.

(g) It should be relatively inexpensive.

The Local Instrument

There are certain applications of a light intensity meter for which the point of presentation is within a few hundred yards of the measurement site, i.e., within a convenient distance for a local wiring run without recourse to Post Office lines. In such situations it is obviously uneconomic to have to use the voice-frequency oscillator and the receiver unit of a conventional telemeter link. On the assumption that a method of measurement could be evolved for the telemeter transmitter requirements previously tabulated, the use of a modification of that system, retaining the advantages over conventional light meters, could be of value for local indication. A pulse rate may be used to give a meter reading except that without considerable damping and consequently a sluggish meter movement the range of pulse rates from 0 to 66·66 pulse/min would tend to give an unsteady meter reading, particularly at low rates. For convenience, therefore, the pulse rate range in this application should be several decades higher.

Where the local instrument is to be used as an alternative to the telemeter instrument, it is obviously advisable that the same scale shape should apply. However, there are a large number of applications of such an instrument in which a linear scale shape over the whole range of daylight illumination is desirable; this could also apply to the range covered by other forms of illumination including ultra-violet and infra-red. The basic design principles should therefore be sufficiently flexible to cover a wide range of applications.

The Integrating Instrument

One of the electricity supply industry's interests is the use of electricity in horticulture. One example in this field is the beneficial effect of electric lighting in supplementing natural daylight to stimulate germination and growth[5]. Although the subject of photo-synthesis in plants is still under investigation at agricultural and other research establishments, the principle is broadly that germination and growth rates may be related to the total amount of light received by a plant during the various stages, in conjunction with the duration of each lighting period. There is a requirement, therefore, both in production processes and in research, for an instrument capable of integrating the total light received over a period of time. (This requirement is not, of course, confined to horticulture, but is applicable to industrial processes involving photo-synthesis or photo-catalysis).

The integrated value of light can be obtained from any light recording instrument by analysis of the chart. This is a tedious method unless complicated and expensive chart-reading apparatus is available. Consideration of the pulse-rate system proposed in the previous sections, however, suggests that a very simple and direct indication of the integral may be obtained from an instrument employing this system merely by counting the pulses. For such an integration system to be accurate, it is necessary that the pulse-rate should be linearly related to illumination intensity; i.e., that each pulse should represent a unit quantity of light.

The obvious way of counting pulses is by the use of an electromagnetic counter such as a Post Office message register. In the applications considered for an integrating instrument it is unlikely that the distance between measurement and presentation points would be sufficient to warrant the use of a telemeter link, although such a link could readily be adapted if required. An integrating meter is more likely to be required for use either with a local wiring run from measurement to presentation, or with presentation actually on the instrument itself. In greenhouse service, the instrument should be constructed to withstand sub-tropical conditions.

The Radio Link

An application has arisen for the measurement of an illumination intensity as part of another apparatus in a balloon-borne equipment. A chemical method[6,7,8] of gas analysis involving decolouration of a solution has been adapted for this purpose, and a continuous reading of the degree of decolouration obtained was required at a ground receiving station. Stringent weight considerations for the transmitting equipment demanded the use of a radio link. Again, a pulse-rate system of measurement is ideally suited to the problem of operating a radio-frequency transmitter with a maximum peak-to-mean power ratio, necessary to obtain maximum range for minimum weight and power consumption.

Circuit Details

Briefly, the circuit used[9] consists of a blocking oscillator specially designed and constructed for reliable operation, with a photo-emissive cell connected as the oscillator grid leak.

A blocking (or squegging) oscillator[10] is similar in general principle to a continuous-wave oscillator except that it is over-coupled. This results in a burst of oscillation which may be one cycle (blocking oscillator) or a train of a few cycles (squegging oscillator) during which the oscillator is overdriven and passes excessive grid current; the negative charge thus collected on the grid circuit capacitance cuts off the valve. This charge then leaks away through the grid leak resistance during the quiescent period until the valve can again conduct and a further burst of oscillation occurs. The duration of the quiescent period, and therefore the pulse recurrence frequency, may be controlled over a very wide range by the grid leak resistance. In general, the blocking oscillator has higher efficiency and stability than the squegging oscillator.

The vacuum type of photo-emissive cell, as used in these instruments, depends for its operation upon the release of electrons from the active cathode material by the incidence of light, and photo-electrons so emitted are collected by the application of a suitable potential across the cell.

Since the variation of the grid leak in the blocking oscillator serves to vary the discharge current from the grid capacitance, the photocell just described may be used to replace a conventional resistor as this grid leak. The The whole assembly is surmounted by a hemispherical cover of clear glass for weatherproofing purposes.

The Remote Instrument

The circuit diagram is shown in Fig. 1, and views of the instrument both with and without the cover are shown in Figs. 2 and 3 respectively.

For the purpose of this instrument, variation in illumination at high levels of intensity is of little interest and a limiting resistor has therefore been included in series with the photocell governing the maximum rate. (During calibration, adjustment of this maximum rate is carried out by a slight variation, not of the limiting resistor itself, but of the potential to which it is returned). At low levels of illumination, when the effective resistance of the cell is large compared to that of the limiting resistor, the instrument is most sensitive to changes in the illumination level. With a suitable adjustment of the distance between the cell and the opal glass screen, and using an aperture of about 0·75in², half the maximum pulse rate is produced by an illumination intensity of just over 10 kilolux. An average calibration curve is shown in Fig. 4.

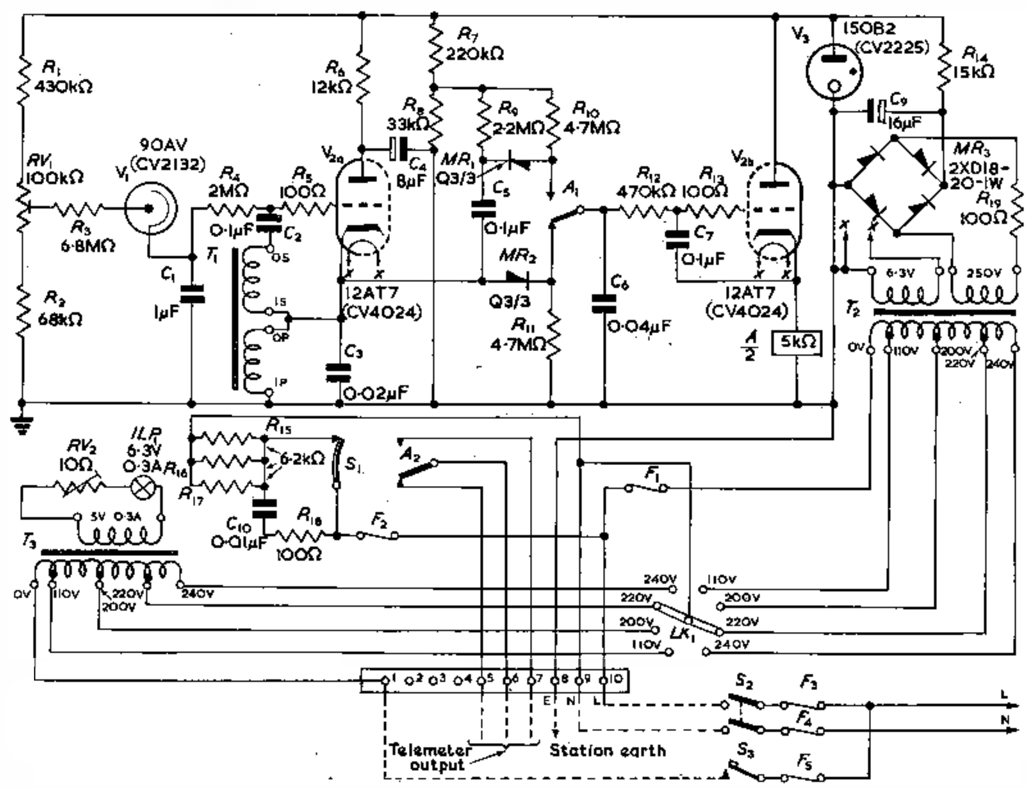


Fig. 1. Daylight telemeter transmitter

pulse recurrence frequency of the blocking oscillator is therefore controlled by the illumination intensity at the photocell such that maximum illumination will provide the highest pulse recurrence frequency.

For most purposes, the intensity of the light required by the photocell is far less than that which is normally available, particularly in the case of natural daylight. Usually, also, the measuring instrument is required to be omni-directional, at least so far as the hemisphere of the sky is concerned, and the reference plane for most purposes is the effective illumination on a horizontal surface.

These requirements are met by mounting, at the top of each of the instruments for use in such service, a horizontal screen of pot-opal glass, sand-blasted on the upper surface to avoid reflexion and to provide an approximate cosine correction to the incident light. This is viewed from beneath by the photocell. To correct the spectral response of the photocell to the required form, normally that of the human eye, a colour filter is interposed between the opal glass and the cell, together with a preset aperture to control the maximum illumination.

For telemetering purposes the output contacts require to have a mark-to-space ratio of approximately unity. Since the duration of the blocking oscillator pulse is fixed and the pulse recurrence frequency is variable, it is necessary to follow the blocking oscillator stage by a binary divider. Referring to Fig. 1, the initial blocking oscillator pulse is rectified positively by MR_2 and fed to the grid of the second half of the double-triode valve used as a cathode-follower. This causes an operation of relay A. One set of changeover contacts A_1 on the relay is included in this grid circuit and serves to transfer the grid to a hold-on potential and to a second rectifier MR_1 connected in the opposite sense. The second pulse from the blocking oscil-

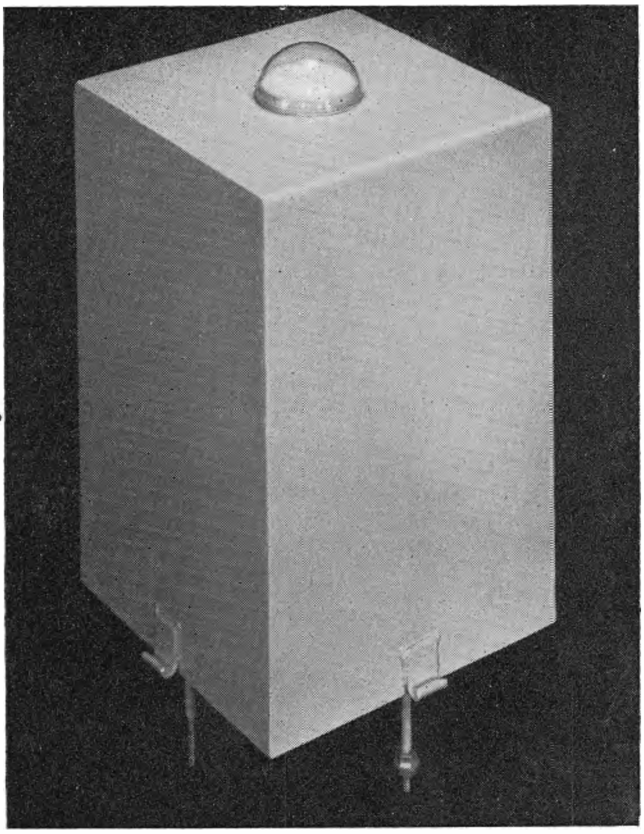


Fig. 2. The remote instrument with cover

lator is thus rectified negatively and again applied to the grid, causing the release of relay A and a second change-over operation back to the initial condition. The output contact set A_2 on the relay therefore produces one change-over operation for each blocking oscillator pulse.

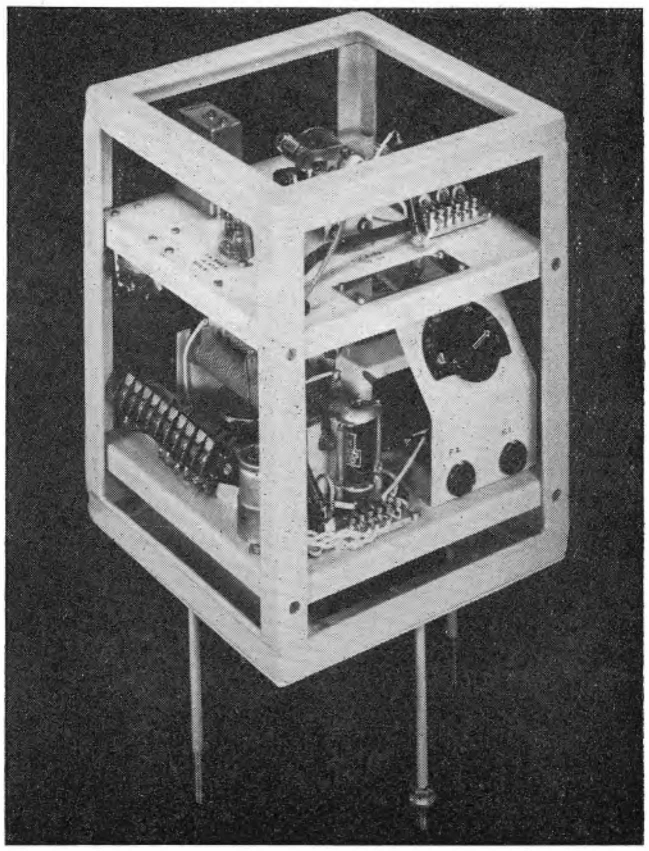
The blocking oscillator circuit designed for this instrument provides an inherently high degree of accuracy and stability without the use of negative feedback. During the peak of the conducting period of the valve, a voltage almost equal to the anode supply voltage will be developed across the cathode impedance, which is the primary winding of the blocking oscillator transformer, T_1. This transformer is basically similar to a conventional television frame blocking oscillator transformer, and has a primary inductance of the order of 40H. The mean current in this circuit is such that the voltage drop across the decoupling resistor R_6 is negligible, while the duration of the pulse is insufficient to produce any significant variation of the potential on the decoupling capacitor C_4 despite the high peak current drawn during oscillation. Within close limits the anode potential of V_{2a} is that of the h.t. supply line, which is stabilized by V_3 at 150V. The grid capacitance C_2 will therefore be charged to the same voltage from each successive pulse since the ratio of T_1 is fixed, and under constant illumination of the photocell will require precisely the same time to discharge back to the oscillation point.

The circuit is largely self-compensating for changes in valve characteristics; for instance, since during oscillation the voltage drop across the valve is small compared to that across the cathode load, a large percentage change in the emission characteristic of the valve, and therefore of the anode-cathode voltage drop, will produce only a small percentage variation in the voltage developed across the cathode load and in consequence only a small variation in the charge on the grid capacitance C_2.

This small change is even further compensated by two more factors. In the first place, reduction of the valve emission, which will be accompanied by a reduction in the slope of the I_a/V_g curve, will produce a longer rise-time for the cathode voltage, with a correspondingly longer period during which grid current may be drawn through the time-constant of C_2 and the internal valve grid-cathode (diode) forward resistance. This internal resistance also stabilizes the degree of positive grid drive required during oscillation. Secondly, the small deviation from the true charge value still present after the operation of the previous factors is further compensated by the slight raising of the point in the characteristic curve at which the gain of the valve is just sufficient for oscillation to occur. This shortening of the grid base, which follows automatically on a reduction in valve emission, requires that the discharge of C_2 through the photocell circuit must proceed for a slightly longer period before the oscillation point is reached.

So far as valve characteristics are concerned, therefore, the circuit carries a high degree of immunity from drift, and the component values have been chosen to take as much advantage as possible of this inherent compensation. Since a variation in heater voltage to the valve causes a change in the emission and slope characteristics, it follows that it is unnecessary to stabilize the heater supply against normal mains voltage fluctuations. Stabilization of the h.t. supply, however, is essential since this is the main determinant of the charge developed on C_2 during oscillation, and the degree of stabilization of this supply is one of the limiting factors in the accuracy of the instrument as a whole. Although simple, the shunt neon type of stabilizer employed has, within the load current variations occurring in this instrument, the same degree of accuracy (within ± 1 per cent) as the limits quoted for the conventional telemeter receiver unit.

Fig. 3. The remote instrument with cover removed



The stability of the blocking oscillator circuit is such that replacement of the CV4024 (12AT7) valve for V_2 by a CV4004 (12AX7) type results in an error of less than 0·5 per cent of full scale. This replacement represents a change in valve characteristics of the order of 300 per cent. The CV4004 type, however, cannot be used as an alternative in service since the other half of the valve is incapable of operating the relay satisfactorily with the present circuit values. It may therefore be assumed that, so far as V_2 is concerned, under conditions of normal valve deterioration the instrument will maintain its calibration until such time as the output relay circuit fails to operate.

The remaining source of error lies in the photocell. The particular type employed in this instrument, however, is a vacuum photo-emissive cell, which is considerably less sensitive to temperature change than the barrier-layer type used in the majority of conventional instruments and which is indifferent by comparison to changes in humidity. Furthermore, it appears to have a much better long-term stability than the barrier-layer cells, which show pro-

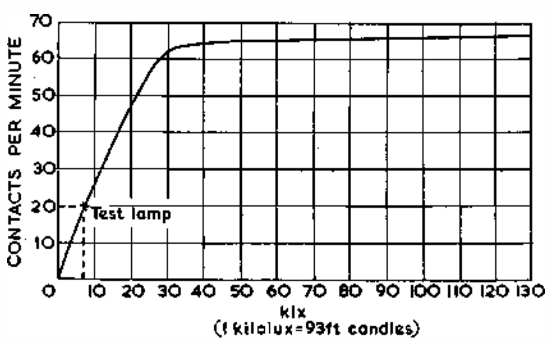


Fig. 4. Daylight telemeter transmitter general calibration curve

nounced ageing. In this circuit it was not possible to compensate for the ageing of the photo-emissive cell used, but in the experimental model the overall change of calibration from all sources including the photocell produced an error of 5 per cent at full scale after one year. Replacement of the photocell restored calibration to within 1 per cent.

However, although the importance of circuit design cannot be overstressed where high reliability and stability of calibration are required, it is essential that components of good quality, and suitable for the worst conditions likely to be experienced in service, must be employed if full advantage is to be taken of the design. In this case, the instrument is intended for continuous, but non-vital, operation completely exposed to all weather. Although under normal working conditions the atmosphere inside the instrument cover will be maintained dry and warm, there will be periods when the instrument is switched off, for example during installation or power shut-down, when condensation can occur internally.

All components, therefore, should carry a minimum humidity classification[11] of H2 and preferably of H1. Furthermore, many of the components, particularly on the blocking oscillator and the photocell circuits, must have a reasonable degree of stability if accuracy is to be maintained. Not all the components could be compensated as in the case of the valve V_2, nor should it be necessary to do so with the more reliable types of component now available. With a very few exceptions therefore, all the components have been selected from the Inter-Services catalogue, since components which conform to the appropriate Radio Component Specifications (RCS) of the Ministry of Supply have excellent reliability characteristics when used under proper conditions of circuit design and rating.

Other sources of error have been minimized as far as possible by careful choice of the circuit conditions under which the photocell is operated and by a thermostatically controlled heater in the photocell compartment. The remaining possible sources of error are ageing of the photocell and reduction of the light intensity by dirt in the optical path. To minimize the latter effect, the protective glass dome will require cleaning at regular intervals, depending upon the degree of atmospheric pollution at the site. It may also be necessary in some situations to clean the underside of the filter and the upper surface of the photocell occasionally.

The distance of the photocell from the opal glass screen and the potential to which the cell is returned are both set up during calibration. Under normal conditions, recalibration due to ageing of the photocell should not be required at intervals of less than one year.

An internal test lamp has been incorporated for the purpose of checking the instrument; this lamp is operated externally by a remote push-button and draws its power from the 230V auxiliary supply via a separate internal transformer. Since the lamp is a tungsten filament one, the illumination effect on the photocell is dependent on the colour temperature which is in turn dependent upon the supply voltage. Without excessive precautions in the stabilization of the supply voltage, it is not practicable to use this lamp for an exact check of the instrument calibration and it has been included merely to indicate that the instrument is operating. It is used for occasional test purposes at night or in the event of the instrument reading low or zero in daylight.

The internal power pack for the instrument is fed from the normal station auxiliary supply (usually 230V) which is intended to be fused and switched externally to the instrument. A double-wound transformer provides an h.t. supply line which is rectified by a bridge-connected selenium rectifier and stabilized at the required level of 150V by a neon tube V_3. The valve heater supply is taken from a tertiary winding on the transformer. The test lamp supply is taken through a separate transformer from the same auxiliary supply via a further fuse and push-button (or biased switch) external to the equipment. The test lamp current is adjusted during calibration by RV_2 to give the response indicated on the instrument calibration graph at the appropriate value of supply voltage.

The thermostat S_1 has been preset at 45°C and the three heater elements R_{15}, R_{16} and R_{17}, each 6·2kΩ, are intended for use in the range of auxiliary supply voltage of 200 to 250V. Where 100 to 110V supplies are employed, these should be replaced by five resistors each of 3kΩ. The remaining voltage selection is carried out on the one selector panel, LK_1, for both transformers.

The instrument is constructed on a small angle-iron framework carrying the individual circuit components on two platforms. A weatherproof cover carrying the optical filter is fitted overall by means of spring retainers. The instrument is cooled mainly by conduction through the cover, since apart from the thermostat-controlled heater the heat generated by the instrument is small. A few ventilation apertures are incorporated in the cable gland plate to assist the air circulation and avoid condensation in the base of the instrument. These holes are covered by fine wire mesh to prevent the infiltration of insects.

The Local Instrument

The circuit diagram is shown in Fig. 5.

It will be observed that the blocking oscillator section of this instrument is in the same form as that of Fig. 1, but that the values of C_1 and C_2 have been reduced by a factor of a hundred and that T_1 has been replaced by a transformer of the television line blocking oscillator type having a primary inductance of about 0·5H and tuned by its stray capacitance only. The mean pulse rate is further increased by raising the value of the grid stopper resistor to 10kΩ to restrict the charging rate of C_2.

For this instrument the precise value of the pulse recurrence frequency for a given illumination intensity, provided it is repeatable, is not particularly important, since calibration adjustments other than the position of the photocell relative to the opal glass screen can be carried out more readily on the indicating or recording meter circuit. This is in contrast to the strict requirements imposed on the pulse rate of the remote instrument by the use of the conventional telemeter link. The photocell anode is therefore returned directly to earth through the limiting resistor instead of to an adjustable potentiometer tapping.

The pulse output from the cathode is rectified and smoothed, using a discharge time-constant in the smoothing circuit of such a value that the output voltage increases linearly with pulse rate up to the saturation value. This voltage is applied to the grid of the cathode-follower V_{2b} which is used to drive a suitable indicating or recording instrument.

The stability of the blocking oscillator circuit for this instrument will be substantially as described for the remote instrument. The stability of the meter circuit will be decided mainly by V_{2b}; since this is used as a cathode-follower, i.e., with some 95 per cent negative current feedback, ageing of this valve involving, say, a 20 per cent reduction in both emission and slope characteristics will produce a change in zero setting of the meter of the order of 1 per cent of full scale, together with a reading which is low by about 1 per cent of the indicated deflexion. For most requirements this stability is adequate, but it is worth noting that the substitution of a White cathode-follower[12] for the conventional cathode-follower used in the circuit of Fig. 5 can improve these figures considerably, and may in some instances be worth the additional complexity.

It will be obvious that the meter itself can be situated at any reasonable distance from the remainder of the instrument, being limited only by the convenience of the wiring run involved, since the resistance of that wiring run can be compensated on RV_1. If desired, RV_1 and/or RV_2 may be mounted at the meter, provided there is no risk of unauthorized interference.

The design shown in Fig. 5 is for a direct-reading light intensity meter having a response characteristic identical with that of the remote instrument described previously. There are applications of either instrument, however, which can demand a response curve which is linear with illumination intensity over the whole range—this may readily be accomplished in either simply by reducing the size of the aperture by which the cell is illuminated together with the removal of the 6·8MΩ limiting resistor. Similarly, considerably lower levels of illumination intensity may be dealt with by increasing the aperture and if necessary introducing an optical lens system directed at the cell. Radiation at other wavelengths such as infra-red and ultraviolet may be accommodated by the use of a photocell of the appropriate response, of either the photo-emissive or photo-conductive pattern, in conjunction with a suitable optical system for the radiation concerned.

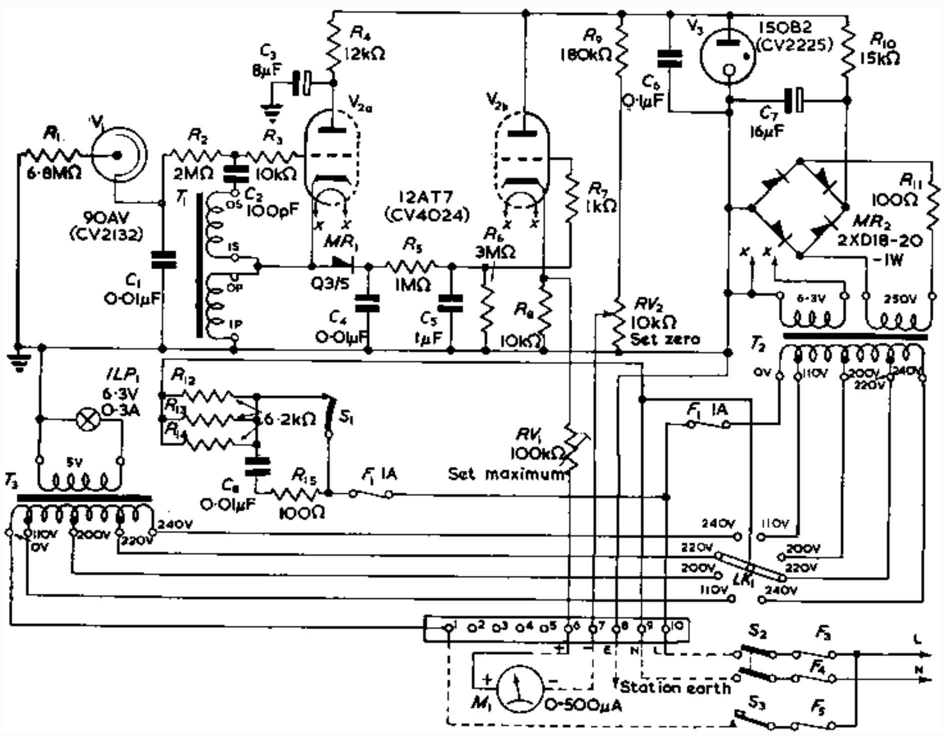


Fig. 5. Light meter type 3a

The Integrating Instrument

The circuit diagram is shown in Fig. 6.

Apart from the removal of the 6·8MΩ limiting resistor in the photocell anode lead and the reduction in value of the filter resistor R_3 from 2MΩ to 1MΩ, the blocking oscillator circuit is identical with that of Fig. 1. The optical system is also similar to that of the remote instrument, except that the aperture is suitably reduced. This procedure has been mentioned in connexion with the provision of a linear response curve, which is obviously essential in the case of an integrating instrument. In the case of natural daylight, adjustment of the pulse rate to a value of one pulse per second at an illumination intensity of, say, 100 kilolux will provide an integrated light value in units of 100klx-sec; any other convenient unit may, of course, be set up by appropriate adjustment of the optical arrangements.

The output at the cathode of the blocking oscillator is coupled through a rectifier to a pulse-lengthening time-constant circuit feeding the grid of V_{2b}. Each pulse is lengthened sufficiently for the counter in the cathode of V_{2b} to make one complete operation. This counter is shown mounted externally to the instrument, with a 22kΩ resistor internally from cathode to earth to guard against damage due to inadvertent disconnexion of the counter. In some instances it may be desirable for the counter to be integral, when R_9 may be omitted.

Discussion

Consideration of the three instruments described in detail in the preceding section, and the mention of the radio link instrument, will indicate that the basic principle is sufficiently versatile to cover a number of allied applications in the field of natural and artificial illumination. There are many more applications in which incorporation of this principle in instrumentation may be of value. The continuous monitoring and recording of solid particle content in flue gases, the remote measurement of the temperature of materials above red heat by their light emission, the continuous examination of boiler feed water for suspended solids by reflected light, the provision of a low-level alarm for opaque liquids or finely divided solids, and a repeater system for the remote observation in meter form of water levels utilizing the reflective or refractive properties of the water in the gauge glass, illustrate some of the possible applications which may be of interest to the electricity supply industry. In other fields, agricultural research, atmospheric pollution research, holiday and health resort statistics are a few examples in which the instrument could have valuable application.

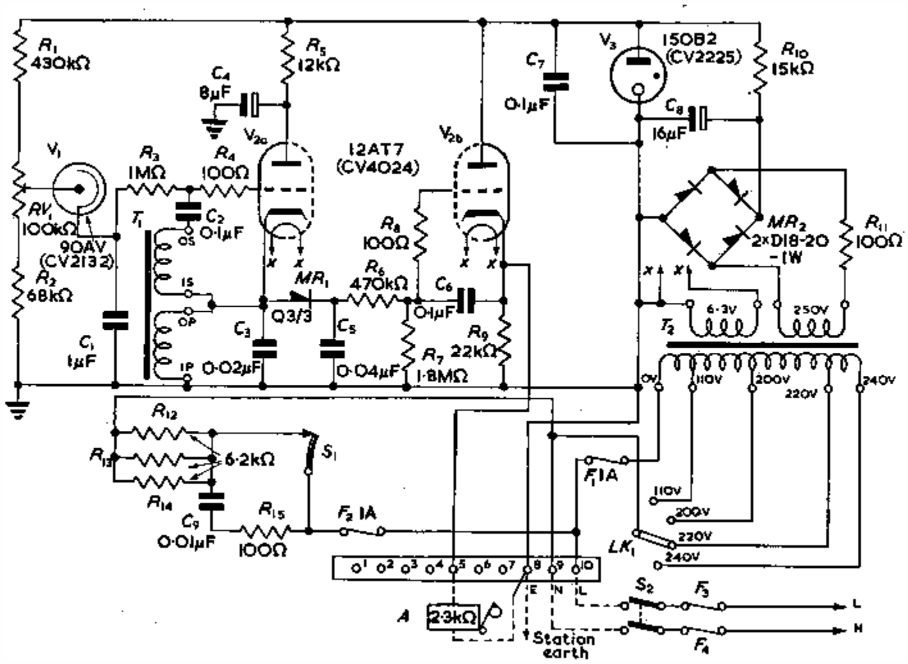


Fig. 6. Light integrator

The present trend in design for improvement of accuracy and reliability in electronic equipment is generally for an increase in the number of valves and components employed. This is often necessitated by the use of negative feedback and duplication techniques; the additional complexity of such equipments, and the increased number of components open to failure, offsets to some extent the initial apparent gain in reliability and accuracy. Such methods, however, have been employed with considerable advantage in many instruments, although to be successful a high order of design ability is required, and where this is not present many promising instruments fail to provide the performance expected of them.

All too rarely, unfortunately, is it possible to achieve high orders of accuracy and reliability by the opposite process, the drastic simplification of an instrument to little more than a basic stage of intrinsically high stability. The circuit analysis, however, suggests that the use of a properly designed blocking oscillator in instruments where it is applicable can achieve just such a result.

A small number of remote instruments for telemetering service are being installed in the new Area Control Centres now being commissioned. Subject to satisfactory performance, some hundreds of such instruments are to be installed in the next few years to give adequate coverage for the assessment of lighting demand. A local instrument has been in operation for some time on the roof of the Air Ministry Meteorological Office, for direct comparison with their apparatus, and an integrating instrument is on loan to the Scottish Horticultural Research Institute at Invergowrie.

Conclusions

A number of instruments have been described, employing the principle of a photocell as the grid leak of a blocking oscillator. This principle has made possible a range of instruments of extreme simplicity, high accuracy and reliability. Because of this simplicity, both initial cost and maintenance charges are small. The versatility of the system can be employed to advantage in a variety of instruments, using photo-emissive or photo-conductive cells appropriate to the wavelength of the radiation concerned. Consideration could also be given to the employment of the blocking oscillator principle in conjunction with other means of controlling the discharge rate of the grid capacitance. Cold cathode and transistor blocking oscillator circuits have been developed[9] and may possess advantages in specific applications.

Acknowledgments

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Some Novel Circuits Employing Cold-Cathode Tubes

(Part 2)

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

A Staircase Waveform Generator

The circuit of the device is shown in Fig. 13. It consists of nine triodes, type XC18, which are triggered successively by means of external pulses. The anode supply for the triodes is taken from point X of the univibrator of Fig. 12, in which the tube V_4 is normally non-conducting. Each triode of the generator is provided with an anode resistor and an adjustable cathode resistor R_k. The anode resistors (see Fig. 13) are chosen so as to compensate for the voltage drop across the 5kΩ resistor in the circuit of Fig. 12, and thus ensure that the voltage drops across the cathode resistors are approximately equal. The cathode resistors are connected to a common resistance (R_o = 1·3kΩ) whose lower terminal is grounded.

In the 'zero' state all the tubes of the generator are non-conducting, and only V_1 is biased. When a positive pulse is applied to the triggers through the triggering capacitors, C, the first tube is ignited and a voltage step appears across the common resistor, R_o. The following input pulses ignite further tubes, until after nine pulses all the tubes are conducting. Cathode potentiometers, R_k, of the tubes are adjusted so that each conducting tube adds a 1·0V step to the potential across R_o. After nine pulses the output voltage from R_o is 9·0V and V_4 of the univibrator of Fig. 12 is biased to about 45V. The tenth pulse ignites V_4 and causes the extinction of V_3 and of all the nine small tubes. The output voltage (across R_o) is therefore reduced back to zero. After a time t_1 (about 4msec) the univibrator of Fig. 12 returns to the stable state and the generator is ready to produce another staircase waveform.

The generator (together with the univibrator) is designed to produce 100 steps per sec and it requires driving pulses of 100μsec duration and about 55 ± 5V amplitude. The resistor R_o is shunted by a 0·1μF capacitor. In spite of the fact that the capacitor reduces the steepness of the steps, it is essential to include it in the circuit in order to suppress the noise (oscillations) generated by tubes V_1, $V_2 \ldots V_9$. Furthermore, the capacitor facilitates the extinction of V_1, $V_2 \ldots V_9$ when V_4 (Fig. 12) becomes conducting. The generator can produce any number of steps between 1 and 9. For this purpose, the 'number-of-steps selector is set so as to bias V_4 (Fig. 12) from an appropriate tube. If, for example, as shown in Fig. 13, the bias to V_4 is taken from V_5, the circuit will generate five steps.

It may be worth mentioning that the circuit of Fig. 13 can also be used as a counter (up to a decade), a square-wave generator (if V_1 alone is used) or as a source of various time-division waveforms. It is also possible to modify it into a monostable circuit, suitable for the generation of sharp rectangular pulses.

A Rectangular Pulse Generator

The device (see Fig. 14) employs a univibrator similar to that of Fig. 12 except that it is provided with an additional timing circuit (R_2C_2). Normally, V_3 is conducting and V_4 and V_7 are off, so that their anodes are at a potential V_b. V_7 can be struck by an external pulse, whereupon the anode voltage of V_5 is, after a time τ_1, raised to a value V_{as}; at this point V_5 is struck and produces a positive pulse across its cathode resistor. The pulse ignites V_4 and results in the extinction of V_3. V_7 is either extinguished due to the fall of the anode voltage of V_4 or it continues to conduct, but at a very low current level.

When V_3 is extinguished, the voltage at the anode of V_2 rises and, after a time t_1, V_2 produces a positive pulse which re-ignites V_3. This results in the extinction of V_4. Also, if V_7 was not extinguished when V_4 became conducting, it would finally be deionized during the extinction of V_4. The cycle is therefore completed and a rectangular pulse having a duration τ_1 and a comparatively short rise-time is obtained at the cathode of V_7.

The generator may seem unduly complicated, since it employs seven tubes. It may be borne in mind, however, that it produces fairly sharp rectangular pulses whose duration τ_1 can be controlled independently. In practical circuits, it should be possible to obtain pulse amplitudes of up to 100V (across 10kΩ) and durations down to about 200μsec. The difficulty in extinguishing V_7 at the instant when V_4 is ignited, can be overcome by choosing the tubes in such a manner that the maintaining potential of V_4 is slightly lower than that of V_7. The generator can be based entirely on the small triodes, type XC18, but this results in a ten-fold increase of its output resistance.

A Fast Decade Counter

The most common cold-cathode tube counter is a chain circuit[1,3,13,14] based on the arrangement of Fig. 7. The circuit contains n tubes and is capable of recording n pulses. Each tube is provided with a cathode resistor connected in parallel with a suitable capacitor. The anodes of all the tubes are connected to a common resistor. Only one tube of the chain is conducting and it provides a bias voltage for the trigger of the next tube. Trigger circuits of all the tubes are connected through suitable capacitors to a common pulse source. Application of a pulse to the triggers results in the ignition of the biased-trigger tube and in the extinction of the preceding tube.

The chain counter is simple and reliable in operation, but it has a rather limited counting speed. This limitation is due both to the tubes employed and to the circuit. If the counter is based on the triodes, type XC18, it should have a cathode time-constant of about 1·0msec. Normally, its trigger time-constant is about 0·5msec. The trigger time-constant can be reduced to perhaps 0·2msec, but it is impossible to decrease the cathode time-constant, since the tubes may fail to extinguish. The maximum counting speed of the circuit is, therefore, primarily determined by its cathode time-constants.

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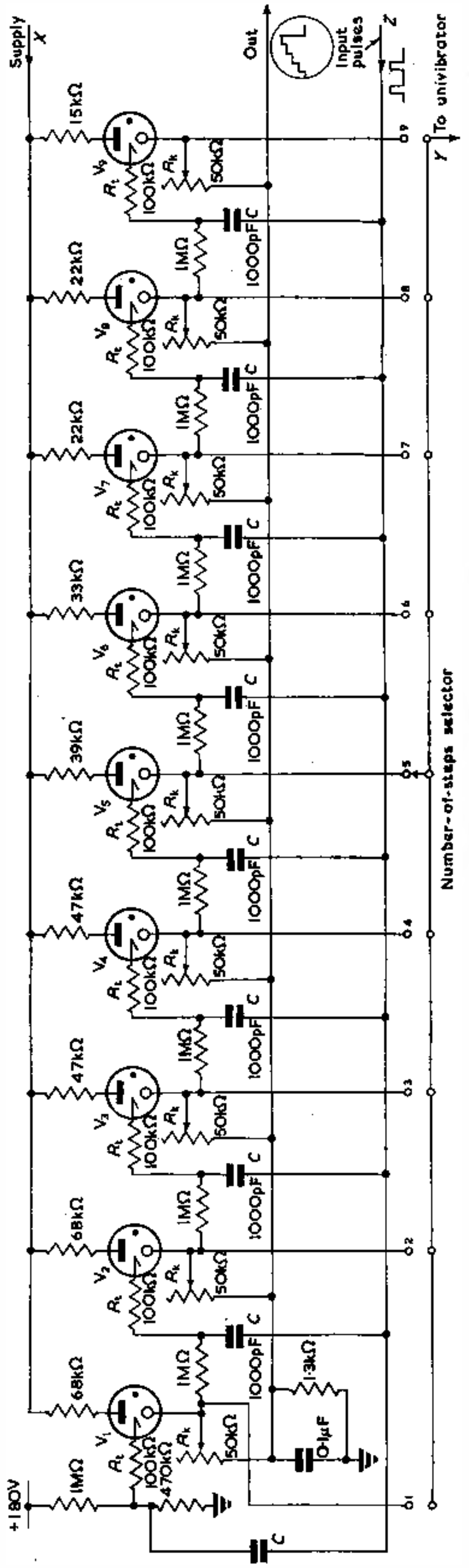


Fig. 13. Staircase waveform generator producing 1·0V steps

Attempts have been made by various authors[1] to increase the counting speed by modifying the circuit. It appears that a very effective method[1] of achieving this is to employ the circuit shown in Fig. 15. A number of triodes are connected in a chain, but their cathode resistors are not shunted by capacitors. The trigger of a following tube is biased from the cathode of a preceding tube. After N pulses, all the tubes are made conducting. If the next pulse is made to trigger some device, such as an Eccles-Jordan circuit and a suitable amplifier valve, the tubes can be extinguished by a negative pulse, superimposed on their voltage supply V_b. In the meantime, a further set of N pulses can be counted by means of another chain of N tubes. At the end of $2N$ pulses the anode voltage of the first chain is returned to V_b. The next pulse will trigger V_1 again and, by means of a suitable circuit, will extinguish the second chain of tubes. Theoretically, the maximum speed of the arrangement of Fig. 15 can be about N times higher than that of the standard ring counter. If it is assumed, for example, that the de-ionization time of the tubes in a standard ring counter is $t_x = 1{\cdot}0$msec, the maximum speed is of the order of 1kc/s; if the tubes are connected as in Fig. 15, it should be possible with $N = 5$ to achieve a counting rate of about 5kc/s, since the tubes are allowed to deionize for the duration of 5 cycles. In practice, however, it is impossible to achieve an N-fold increase in speed due to the presence of the trigger time-constants.

One of the disadvantages of the arrangement of Fig. 15 is that it is usually controlled by circuits consisting of vacuum valves. It appears possible, however, to generate a suitable control waveform by means of cold-cathode triodes, provided the tubes have a current rating about an order higher than that of the controlled tubes. Consider the binary circuit of Fig. 16 and assume that V_1 is conducting and V_2 is non-conducting. If V_2 is now ignited (by a positive pulse applied to its trigger), the voltage waveforms at the two anodes will be as shown in Fig. 16, provided C is sufficiently large to extinguish V_7. The two waveforms can be combined by means of two rectifiers, as shown in Fig. 16. The resulting waveform (see Fig. 16) has a steep negative leading edge, an amplitude equal to $(V_b - V_m)$, a flat portion having a duration $t_0 = RC \ln 2$ and an exponentially rising trailing edge. The waveform is suitable for the control of the circuit of Fig. 15.

A decade counter employing two control circuits of the type shown in Fig. 16, is given in Fig. 17. The ten small tubes, V_1, $V_2 \ldots V_{10}$, of the counter are divided into two groups. In the zero state V_1 and two of the control tubes (V_{11} and V_{13}) are conducting. The first pulse from the common line will ignite V_2 and after the 4th pulse all the five tubes of the first group will be ignited. The 5th pulse will ignite V_6 and V_{11}; consequently, a waveform of the type shown in Fig. 16 will be generated at the output of the rectifiers of the first control pair (V_{11} and V_{12}) and V_1, V_2, V_3, V_4 and V_5 will be extinguished. After the 9th pulse, all the tubes of the second group will be ignited. The 10th pulse re-ignites V_1 and, by triggering V_{14}, extinguishes V_6, V_7, V_8, V_9 and V_{10}.

With the component values indicated in Fig. 17, the counter is capable of operating at speeds of up to 2·5kc/s when it is driven by positive rectangular pulses having a duration of 50μsec and an amplitude of 60 ± 4V. It will operate even at 3kc/s, but the driving pulse amplitude becomes critical. The maximum operating speed can probably be increased to about 3·5kc/s by reducing the trigger capacitors to 300pF.

It is worth mentioning that the anode resistors of the counting tubes are chosen in such a way that the voltages

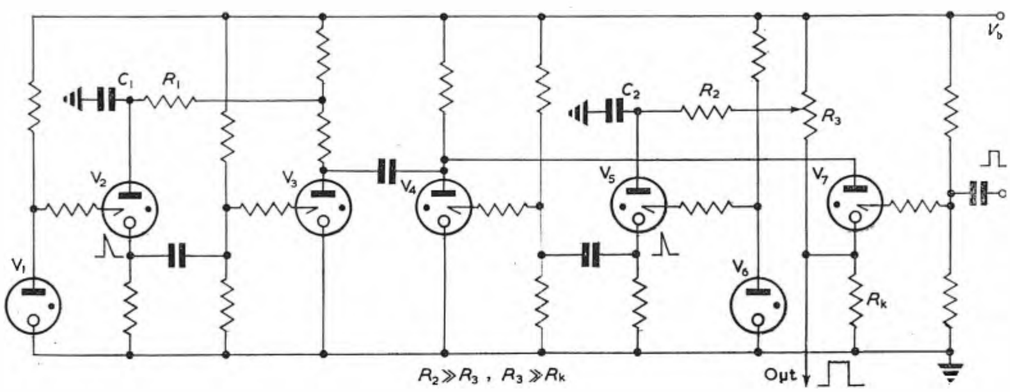


Fig. 14. A rectangular pulse generator

at their cathodes are approximately equal. The total 'parallel' value of all the anode plus cathode resistances, for each group, is such that the supply voltage at the output of the rectifiers is not less than 140V. The rectifiers

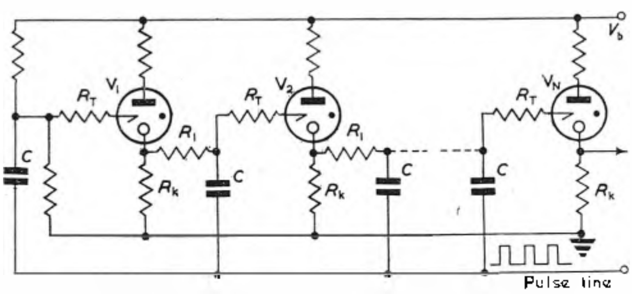


Fig. 15. A fast counting circuit

employed in the control pairs should have a peak-inverse voltage of over 120V, but do not have to be very efficient; a pair of selenium rectifiers connected in series provides suitable rectifying units. The zero-setting or re-setting of the counter can be done by means of a triple changeover switch. Two poles of the switch would be connected to the outputs of the rectifiers of the control pairs, while the third pole would be joined directly to V_b. In the zero or re-set position, the switch would break the supply lines to $V_2 \ldots V_4$ and to $V_6 \ldots V_{10}$, and connect the triggers of V_{11} and V_{13} to the outputs of their rectifiers by means of suitable resistors; the supply line of V_1 would not be broken and its trigger would be connected to V_b through a suitable resistor. In the normal position, the switch would 'release' the triggers of V_{11}, V_{13} and V_1 and restore the supply lines.

A Six-Tube Decade Counter

The arrangement of Fig. 15 and the control pair of Fig. 16 can be combined in a circuit which would be capable of recording ten pulses by means of six tubes. The resulting counter is shown in Fig. 18. It consists of four small triodes, type XC18, and two large tubes, type XC23. In the zero position only V_5 is conducting, and the trigger of V_1 is biased positively from the supply line. The first four pulses successively ignite V_1, V_2, V_3 and V_4; the next pulse triggers V_6 and results in the extinction of the small tubes. The further four pulses successively re-ignite V_1, V_2, V_3 and V_4, while the tenth pulse re-strikes V_5 and results in the extinction of all the remaining tubes. It is seen, therefore, that the circuit is re-set to zero after accepting ten pulses.

In comparison with the standard counting chain the counter of Fig. 18 has no advantages as regards its maximum operating speed, but it uses a smaller number of components (resistors and capacitors). The principle underlying the counter can be employed to devise circuits which would count to any required scale. For example, a scale-of-nine circuit would employ six tubes, as shown in Fig. 18, but V_6 would be biased from the cathode of V_3 instead of V_4; a scale-of-eight would be based on three small

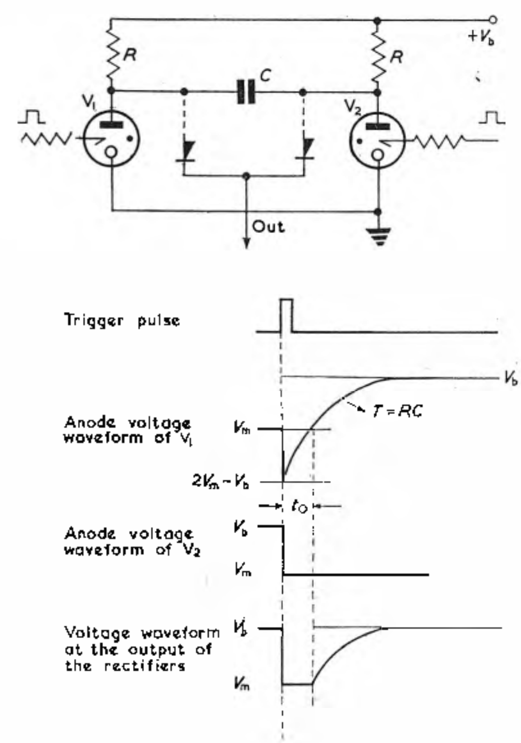


Fig. 16. Binary pair with a rectifier output

and two large tubes, while a scale-of-eleven or a scale-of-twelve would use five small and two large tubes. It may be interesting to mention that the counter can easily be modified into a shift register[3]. The resulting circuit is shown in Fig. 19. If it is assumed, for example, that V_1, V_2 and V_3 are conducting, the trigger of V_4 is biased positively from the cathode of V_3. The tubes can be extinguished by a negative pulse from a control pair, but provided the constants R_1C_1 are suitably chosen, the bias voltages of V_2, V_3 and V_4 will not change appreciably during the extinguishing pulse. If after a suitable interval, the negative pulse is followed by a triggering pulse, tubes V_2, V_3 and V_4 will be ignited. In this manner, the pattern of conducting tubes is stepped on one position. A futher pair of pulses, applied after a suitable interval, will move the pattern another step forward.

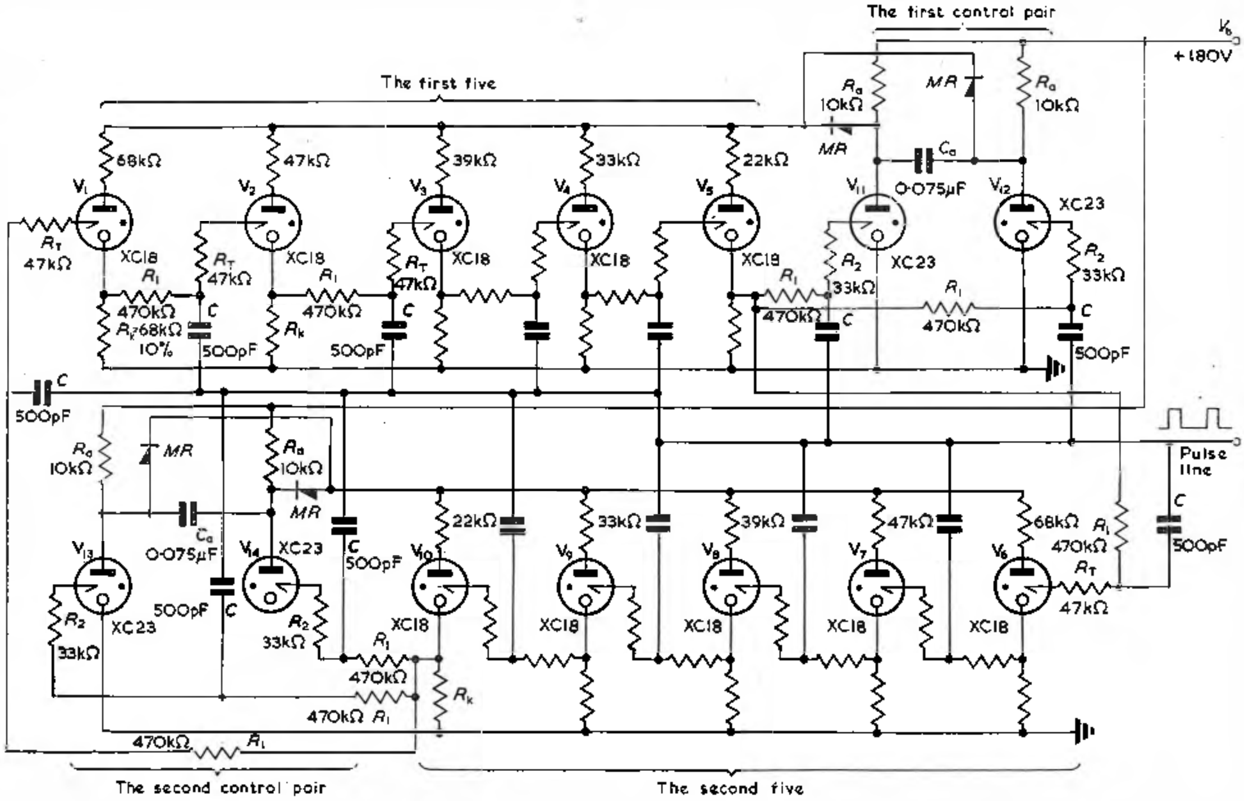


Fig. 17. Fast counting decade employing ten small and four large tubes

MR = Two rectifiers, type M3, in series
Drive pulses: Duration 50μsec, amplitude 60 ± 4V

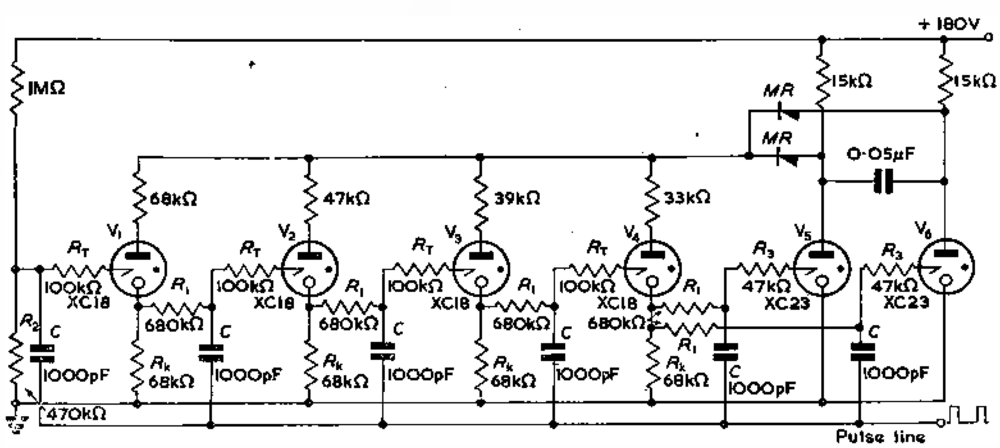
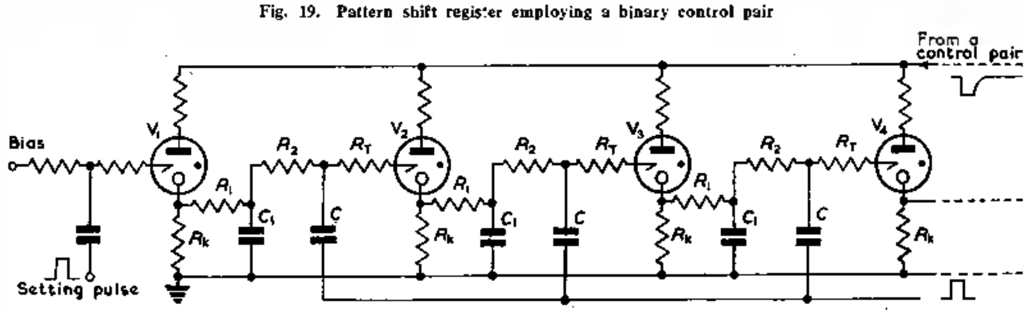


Fig. 18. Decade counter employing six tubes

MR = Two rectifiers, type M3, in series
Drive pulses: Duration 100μsec, amplitude 50 ± 5V (up to 500c/s)

Fig. 19. Pattern shift register employing a binary control pair



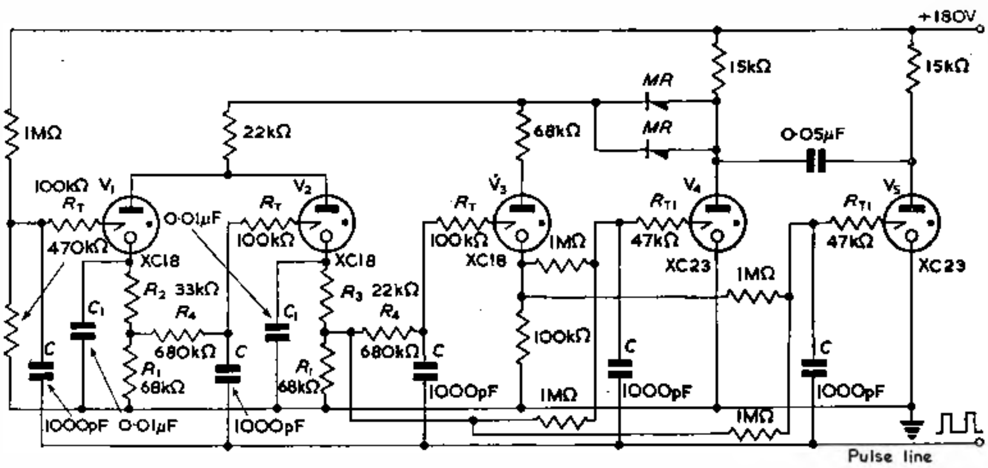


Fig. 20. Decade counter employing five tubes
MR = Two rectifiers, type M3, in series

the second pulse ignites V_2 and, since V_1 and V_2 form a bistable pair, results in the extinction of V_1. The next pulse ignites V_1 and V_3, and extinguishes V_2; after the fourth pulse V_2 is re-ignited and V_1 is extinguished. The fifth pulse ignites V_5, whereby all the remaining tubes are extinguished. The next sequence of five pulses re-sets the counter to zero.

Though the counter uses less tubes than the decade of Fig. 18, it employs a slightly larger number of resistors and capacitors. Also, its maximum operating speed is lower and its reliability less satisfactory than that of the six-tube decade. The loss in the reliability is due to the fact that V_4 and V_5 should not be struck by triggering pulses, when biased from the cathode voltage of V_2 or V_3 alone; on the other hand, they should be ignited when the biases from V_2 and V_3 appear together. As a result of this bias arrangement, the counter is more sensitive to voltage variations than the six-tube decade or the standard ring counter.

It may be worth pointing out that the bottom voltage level in the extinguishing pulses produced by the control pairs of Fig. 17 and 18 is equal to the maintaining potentials of the tubes employed. In general, it appears that the pulses are quite adequate to extinguish a set of XC18 tubes. However, since the nominal maintaining voltage of the XC18 triodes is only slightly higher than that of the XC23 and since the maintaining potentials of individual tubes can differ by a few volts, a situation may arise in which a control pair will fail to extinguish some of the controlled tubes. This can be remedied by connecting the lower terminals of the cathode resistors of the small tubes to a common resistor, shunted by a large capacitor; a resistor of 3·3kΩ and a capacitor of above 0·25μF are sufficient for the circuits of Figs. 17 and 18.

A Five-Tube Decade

The reader is probably aware of the fact that, since a cold-cathode triode has two stable states ('on' and 'off'), it should be possible theoretically to count 9 to 16 pulses by means of four tubes. From this point of view it would appear that the number of tubes in the decade of Fig. 18 could be reduced to less than six. An attempt was made to devise such an 'economy' counter and the resulting circuit is shown in Fig. 20.

The circuit consists of three small tubes, and two large tubes connected in a control pair. In the zero position V_1 is biased positively from the supply line and only V_4 is conducting. The triggers of all the tubes are connected to a common pulse line through capacitors *C*. The first pulse ignites V_1, whereby the trigger of V_2 is biased positively;

TABLE 2

TYPE OF COUNTER / PARAMETERS	THE STANDARD CHAIN COUNTER (DECADE)	THE FAST DECADE	THE 6-TUBE DECADE	THE 5-TUBE DECADE
No. tubes	10	10+4	4+2	3+2
No. resistors ..	41	52	23	22
No. capacitors ..	20	16	7	8
Current consumption	1mA	20mA	8mA	8mA
Max. counting speed	500c/s to 1 000c/s	2·5kc/s	600c/s	400c/s
No. rectifiers	None	4	2	2

Conclusion

It may be useful to summarize briefly the characteristics of the three decade counters described above and to compare them with those of a standard ring counter. A number of important data are compared in Table 2.

It is seen from the table that the fast counter employs a comparatively large number of components and its current consumption is much higher than that of the standard ring counter. The six-tube and five-tube decades compare very favourably with the ring counter, except for their current consumption. The three decades are also disadvantageous in that they do not provide a simple visual indication, as in the ring counter, but rely on a combination of several conducting tubes. It cannot be expected therefore that the above counting circuits will displace the ring counter. However, in certain applications their disadvantages will be outweighed by their relative merits; in particular, they may prove useful in various time division systems and in pattern shift registers. It may also be pointed out that the control pair of Fig. 16, which is used in the counting decades, can also be employed advantageously in various other cold-cathode tube circuits.

In the above comparison it is not intended to convey the impression that the counters described are rivals to the standard ring counter. On the contrary, it is hoped that the circuits will supplement the ring counter and thereby extend the field and scope of cold-cathode tube circuits.

Acknowledgment

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A Track Switching System for a Magnetic Drum Memory

By D. D. Majumder*

An electronic track switching system using a rectifier function mesh, and magnetic gating transformers is described. The same information channel and magnetic head with one winding can be used for both recording and playing back signals. Superiority of the system with respect to cost, reliability, and performance is claimed by the author.

THIS article discusses an electronic track switching mechanism for use in reading from, or writing on to, a serial type magnetic drum memory of a digital computer.

When there is a large number of tracks on a magnetic drum, the problem of selecting and switching to a track is a serious one. The simplest solution is to use a relay tree, and the method has the advantage that in a serial machine one reading amplifier and one writing circuit only are required. Sooner or later, however, relays inevitably give trouble, and the trouble is often of an intermittent nature. An electronic switching system is considered therefore preferable. But, since several valves are required for each track, the total amount of equipment required in the case of a drum with a large number of tracks is considerable. This is one drawback. Secondly, it is difficult to provide electronic switching for writing, especially when low impedance heads are used. The advantages of low impedance heads are, however, manifold: (a) narrow pulses of high repetition rates can be used in the computer, as low impedance heads afford very fast rise and fall time for recording current pulses; (b) at the time of reading from the drum they act as voltage generators of low internal impedance capable of driving a low impedance transmission system, for example a cable connecting to an amplifier of input capacitance of the order of 10 to 100pF, without any appreciable attenuation; (c) low impedance of the heads minimizes the chances of interaction between the heads themselves while recording and the chances of stray pick-up from extraneous fields during reading from the drum.

A compromise has been made in some machines, namely to use relays for switching the writing current, and electronic switching for switching the reading head. The switching problem becomes easier if separate heads are used for reading and for writing.

In the relatively short period, since its first application as a gating device[1], the magnetic core has firmly established its usefulness in switching circuits. With the help of a low impedance winding, current operated magnetic cores can be easily matched to a low impedance magnetic head.

In the all-magnetic core selection system discussed by Gutermann and Kodis[2], two separate selection systems for recording and playback are used but its playback selection system is not suitable for a large number of tracks. Another all-magnetic core selection system suitable for both recording and playback processes was suggested by A. H. Sepahban[3]. The switching circuit described here uses rectifier switching circuits and a magnetic core gating transformer; and the same information channel can be used for the flow of information during both recording and playback processes.

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Principle of Operation

All the conventional switches operate by virtue of a change of resistance from substantially zero to infinite resistance. It is also possible to change other parameters to carry out switching operations. Control of mutual

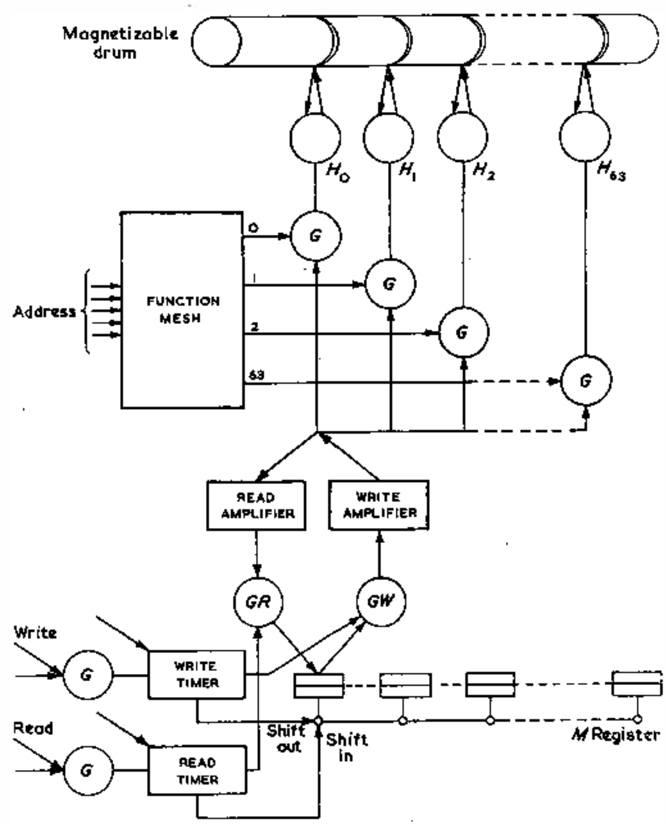

Fig. 1. Block diagram of electronic track switching in a magnetic drum memory

inductance, for example, can be achieved in a manner very similar to that used in saturable reactors. Thus a pulse transformer wound on a high permeability magnetic core is designed to couple the magnetic head of a storage drum to the read-write amplifier. If now by means of a steady large current through a third winding the core is saturated, the mutual inductance is reduced to the low value corresponding to unity permeability. Transformer action is inhibited thereby, and information can no longer pass through it. On withdrawal of the steady bias current the transformer reverts to normal operation (the coercive force of the material being low—'Supermalloy' or 'Mumetal').

Block Diagram

The block diagram of the track-switching mechanism is given in Fig. 1. It is a combination of multi-position electronic switching using germanium diodes, and magnetic gating transformers working on the principle enunciated in the previous paragraph. According to the digital configuration in the 'address' register only one output line of the function mesh marked 0 to 63 gets 'high' and all others remain 'low'. The selected (high mesh line opens the gating transformer denoted by G by unsaturating the core. All other mesh lines being low, the corresponding cores remain saturated, thus inhibiting the information channel through these gates. Digital information can pass through the switched channel in between the 'memory'-register and synchronized word location of the selected track. (Word synchronization is not shown here.)

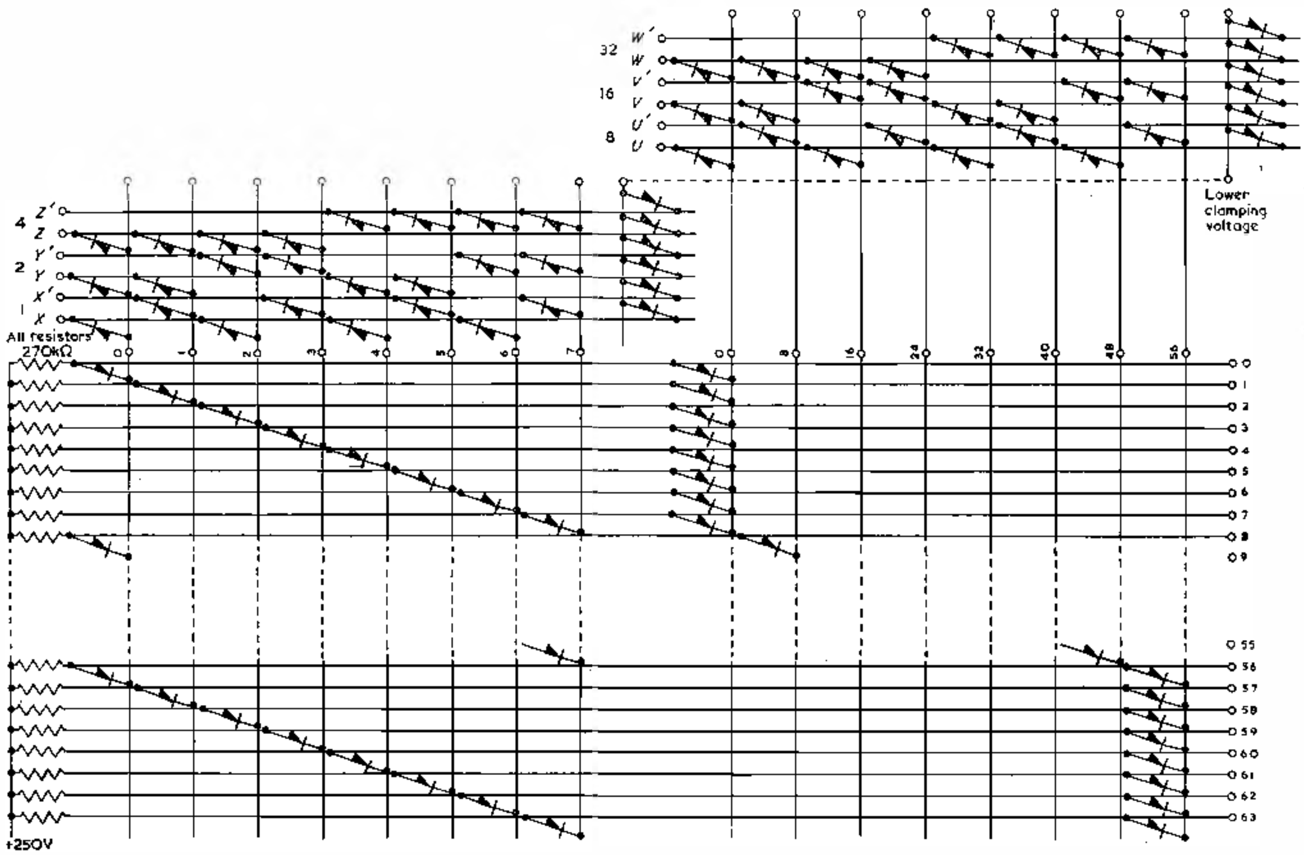

Fig. 2. Germanium diode function for 64-lines track switching circuit

Function Mesh Design

The function mesh used in this switching circuit is of six input variables and of sixty-four outputs with the requirement that for each combination of input variables one unique output line shall be 'high'.

The subject of 'switching algebra' stemmed from a paper by C. E. Shannon[4] in which Boolean Algebra was first applied to relay circuits. Minimal design of switching networks is of current interest in switching algebra. Though a lot of work has been carried out in this direction, the problem of the most economical design is still an aspiration for the future. A method has been successfully developed by Harvard Computation Laboratory[5,6] for designing certain types of switching circuits. The most economical of the designs described by the Harvard Laboratory for the generation of all n outputs in the case of a thermionic valve selection circuit is what is known as a rectangular circuit[7], and the most economical means for obtaining this is to construct the subsidiary rectangular circuits by splitting up the groups, each group having $n/2$ variables when n is even, and $(n + 1)/2$ and $(n - 1)/2$ variables when n is odd. Continued splitting until the largest group contains three variables, leads to the solution of what is termed a rectangular circuit. The most economical rectifier solution is also obtained in the same manner. Writing the output equations, in terms of rectifier operators, and following the above procedure, it is seen that the number of rectifiers required in a rectangular solution is defined by the recursion formula,

$$R_{2n} = 2R_n + 2^{2n+2}$$

$$R_{2n+1} = R_{n+1} + R_n + 2^{2n+2}$$

where R_k is defined as the number of rectifiers required in a circuit having k input variables. The number of rectifiers necessary for the function mesh described is 176 whereas in the case of the pyramidal solution and matrix solution the number would have been 248 and 384 respectively. The corresponding symbolic circuit for the rectangular solution of six input variables is shown in Fig. 2.

Magnetic Gating Transformer

The design of the gating transformer is complicated by the fact that it has got to handle both recording and playback signals. There is a 50 to 100dB difference in power levels between recording and playback signals. The recording pulse is about 1·5A peak-to-peak and of approximately 10μsec duration. The playback signals vary from 5 to 200mV peak-to-peak. The first step for such a design is to ascertain the exact conditions under which the core is working, but the magnetization curve does not obey any simple law. The mathematical treatment of relations between the applied and resulting current is complicated by the non-linearity of the core and the magnetic head. One can start with the idealized characteristic as in Fig. 3, which has been extended to cover the reverse swing, assuming that iron behaves in the same manner if the current is reversed.

In terms of the idealized characteristic of the core the working principle is shown in Fig. 4. When there is no current in the bias winding the operating point is at the origin, Fig. 4(a), and perfect transformer action is achieved. Now if the d.c. magnetizing force supplied by the steady

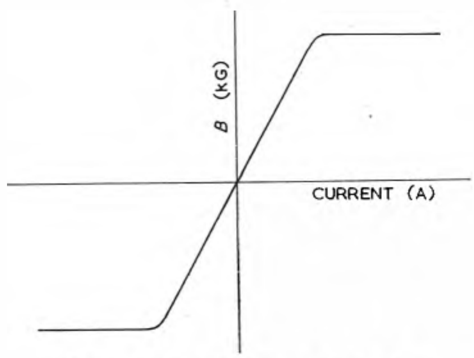

Fig. 3. Idealized flux-current relationship

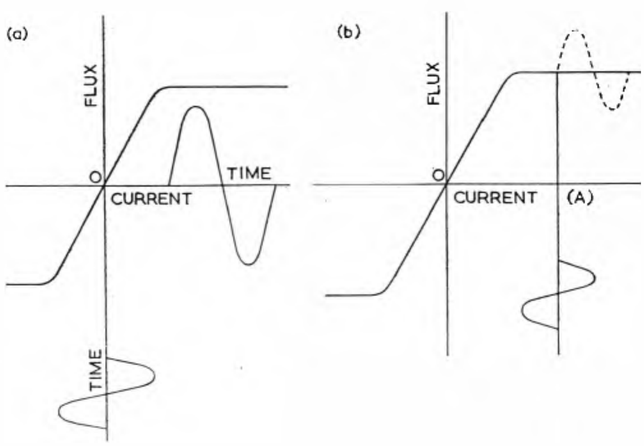

Fig. 4. Flux-current relationship in magnetic gating transformer (a) without biasing current (b) with sufficient biasing current

current in the bias winding is sufficiently increased, the operating point is shifted to the saturated region (Fig. 4(b)) and so signal transformation either way is inhibited.

The design considerations are two fold. Firstly, the design of the signal windings with an open-circuited bias winding, and, secondly the design of the bias winding. The signal windings should act as a perfect pulse transformer and with the secondary looking into the head impedance, the primary impedance should match the write amplifier. The current pulse in the head circuit during the writing operation is to be of specified amplitude, determined by the coating thickness, magnetic characteristics and thickness of the head laminations. It is necessary to have large incremental permeability in order to keep the number of turns in the coil and the cross-sectional area of the core reasonably small. Also the coercivity of the core material has got to be low enough, which is a necessary criterion for gating low level signals and also for high speed operation. The desirable B-H loop for the material is obtained by the introduction of an air-gap[3] in the magnetic circuit and thus shearing the loop as shown in Fig. 5.

The determination of the core cross-section and the number of turns in the two signal windings have been determined by following an approximate method[2], and also practical considerations of available cores, driving circuits, and economics of fabrication. The number of turns in the bias winding is known from the considerations of ampere-turns necessary to keep the core in the completely saturated condition. A small number of turns requires large currents to provide the necessary ampere-turns, and so large power tubes are needed. On the other hand smaller windings have the faster switching. So the number of turns in the bias winding has to be a compromise value. This winding places virtually no load on the driving stage, since no flux change occurs in the inhibited core.

Description of the Circuit

The block diagram (Fig. 1) of the switching system has been explained in an earlier section. The circuit of the

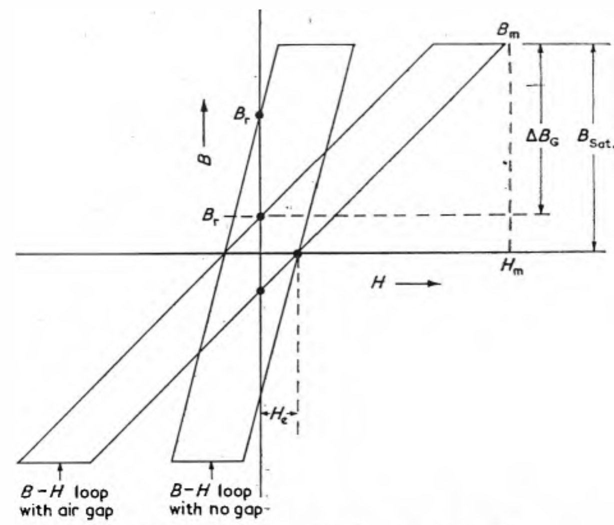

Fig. 5. Effect of air-gap in the magnetic circuit

diode function mesh is shown in Fig. 2. The address register output is fed to the function mesh input through a cathode-follower stage, because for good results the diode-mesh should be fed from a low impedance source, and the mesh output should be connected to a high impedance destination, e.g., the grid of a valve. The voltage discrimination between the selected line and others is of the same order as that of the input voltage discrimination of the function mesh. The complete switching circuit for track selection is shown in Fig. 6. Each of the function mesh output lines is fed to the input of a triode (6SN7), the anode of which is directly coupled to the input of a power valve. The cathodes of the triodes are kept at a fixed potential such that when the mesh lines are 'low' the triodes are driven to cut-off. And so the power valves operate at zero bias, the anode current of which saturates the cores of the magnetic gating transformers. When any of the mesh lines is 'high', the corresponding gating transformer reverts to normal operation. In the left-half of Fig. 6 is shown the read-write transformer, one high impedance winding for read amplifier input with a false centre tap, another centre tap winding that provides a balanced output to the main write amplifier and also one output winding are wound on the same core. This is possible because, during writing extremely large current pulses are being induced in the read-winding, as a result of which the coup-

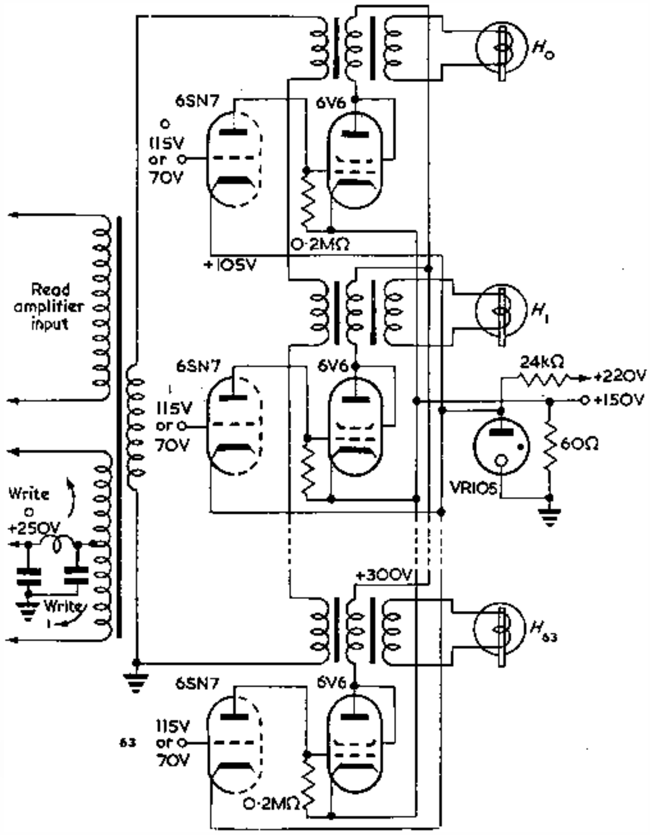

Fig. 6. Complete switching circuit with the magnetic gating transformers

ling capacitors in the read amplifier acquire quite a large charge which drives all the stages of the amplifier to cut-off, the read amplifier remains paralysed thereby. During reading there is no pulse at the write amplifier input. The read-out pulse pattern, of course, induces some voltage in the turns across the writing amplifier, which is open-circuit at that time; so the voltage pulses induced in the read-amplifier input winding are only effective during that period.

Conclusion

The track selection system for a magnetic drum memory suitable for both recording and playback processes described in this article has several advantages. These are:

(1) The ultimate switching is done by magnetic cores which have an exceptionally long life.

(2) It can be implemented for any number of tracks.

(3) Only one amplifier for recording, and, another for reading is necessary for all the tracks.

(4) The amount of associated electronic equipment is small.

(5) The cores being low impedance devices are comparatively free from pick-up and crosstalk.

The switching system, however, has some limitations:

(1) Its reliability is limited by the reliability of associated electronic devices, e.g., germanium diodes and power valves.

(2) For a large number of tracks good layout is very important to avoid cross coupling.

(3) The design of the magnetic circuits is rather critical.

Acknowledgments

The author wishes to thank S. K. Mitra, Head of the Department of Electronic Computers, I.S.I. for his interest and general supervision. Thanks are also due to J. S. Chatterjee, Head of the Electrical Communication Engineering Department of Jadavpur University for his helpful discussions. The author is grateful to P. C. Mahalanobis, the Director of the Institute, for providing funds to carry on this work.

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Production Control by Computer

Probably the first attempt in this country to achieve integrated electronic computer control of production planning processes and plant loading as a means to the more effective deployment of manufacturing resources is now in progress in one of the British Tabulating Machine Company's main Letchworth factories.

Today there is a much wider conception of what constitutes production control than in the past. The piecemeal approach to production is outmoded, the need to recognize the critical interrelation between all the contributing factors to efficient manufacturing and sales organization has been generally accepted.

Production control has been defined as "a complexity of interrelated functions, exercised by Production Management, for organizing and communicating information upon which action may be taken to manufacture products to pre-determined delivery dates."

It includes:

Programme breakdown needed for the translation of a sales programme of end product requirements into a production plan for their manufacture.

Plant loading to determine the total load per machine group available in each manufacturing period and to allocate time of specific machines and groups for carrying out specific operations.

Progress control in order to compare actual and projected performance and to make the necessary adjustments for discrepancies between them.

Stock control to ensure the availability of material—raw material and piece parts—at the times and in the quantities required at minimum inventory cost.

Bearing in mind that the actual operation of the Hollerith Computer installation is yet only to be measured in weeks and extends only to a part of the B.T.M. production in Letchworth, the positive results so far achieved point to the ultimate benefit.

These may be summarized as:

(1) The ability more rapidly to give effect to varying sales demands and in consequence enhanced service to the customers.

(2) A much more rapid production programme changeover than had hitherto been achieved.

(3) Production time—within cycle time—is reduced to a realistic minimum.

(4) The production of more accurate and more timely management information, which includes some significant data and a number of illuminating comparisons hitherto not available. One effect of this has been a review of the volume and deployment of subcontracting commitments.

The exercise now in progress has established basic principles for the employment of electronic data processing techniques in all types of manufacturing. The British Tabulating Machine Company Ltd proposes to extend the scope of its computer controlled production to embrace all its manufacturing activities.

A LOW CAPACITANCE INPUT CIRCUIT

By J. C. S. Richards*, B.Sc., Ph.D.

It is customary to connect an electronic measuring instrument to an apparatus under test by means of several feet of coaxial cable, having a capacitance of 50 to 100pF. Various methods of reducing the effect of this cable capacitance are discussed. A circuit is described in which a compact probe containing only passive elements is used in conjunction with a single valve feedback amplifier to give an input capacitance of less than 5pF, and a gain of unity over a bandwidth of 2Mc/s.

ONE of the desirable features of a general purpose electronic measuring instrument (e.g. an oscilloscope) is a high input impedance. The input impedance of almost all instruments can be represented as a resistance and a capacitance in parallel. It is not difficult to make the input resistance sufficiently large (>1MΩ, say); it would not be difficult to make the input capacitance sufficiently small (<10pF, say), were it not for the necessity of using several feet of cable (generally coaxial) to connect the measuring instrument to the apparatus under test.

Various schemes have been proposed for reducing the effect of cable capacitance. After a brief discussion of these, the bulk of the present article is concerned with a circuit, which, while not in itself novel, does not seem to have been used previously for this purpose.

It uses a single triode-pentode valve; it gives a stabilized gain which may be made exactly unity; the input capacitance may be chosen from the range 0 to 5pF depending on the length of cable, input resistance, and high frequency response required. A rise time as small as 0·2μsec can be obtained, corresponding to a bandwidth of 2Mc/s.

Neutralization Circuits

The use of feedback to neutralize unwanted capacitances dates from the early days of radio. The basic circuit is shown in Fig. 1. C_c is the capacitance to be neutralized. C_n is the neutralizing capacitance. The effective input capacitance is $C_c + C_n(1 - A)$. If A is real and positive the input capacitance may be reduced to zero. The disadvantage of the circuit is that A can be regarded as real only over a limited band of frequencies. At high frequencies phase changes in the amplifier tend to give the input impedance a negative resistance component. A carefully designed amplifier with stabilized gain is necessary if neutralization is to be effective over a wide band of frequencies, and if oscillation is to be avoided. Bell[1] describes an arrangement which is effective over a frequency range of about 0·7Mc/s. Attree[2] gives a simple circuit suitable for signals whose frequency is less than 1kc/s.

A modified form of circuit is often employed when the capacitance of screened cable must be neutralized. A double screened cable is used. The outer screen is earthed and the capacitance between the inner screen and the core of the cable takes the place of C_n in Fig. 1. The gain of the amplifier need only be a little greater than unity for complete neutralization. In a system designed for use at frequencies less than 10kc/s, Attree[3] uses a cathode-follower as amplifier. This has the merit of simplicity, but, as the gain is less than unity, only partial neutralization is achieved. Boston[4] employs a more elaborate amplifier and obtains good performance over the audio frequency range.

When efforts are made to improve the high frequency performance of the system, the main difficulty arises from the comparatively large capacitance between the inner screen and the outer earthed screen, which forms part of the load on the amplifier. Lampard[5] overcomes this difficulty by omitting the earthed screen, and points out that since his amplifier has a very low output resistance, there is little chance that unwanted e.m.f.'s will be induced on the cable. However, it is often just as important to ensure that the signals on the cable do not induce unwanted e.m.f.'s at other points in the apparatus, and obviously in a general purpose test equipment the outer earthed screen cannot be omitted.

Probes

The difficulties of neutralization techniques can be avoided by mounting the first stage of the instrument in a probe unit at the end of the cable. This is, of course,

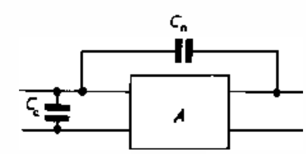

Fig. 1. Basic circuit of neutralized amplifier
The amplifier gain A must be positive

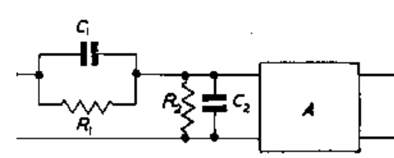

Fig. 2. Basic circuit of amplifier with compensated attenuator
R_1 and C_1 are mounted in probe unit. C_2 represents cable plus amplifier input capacitance

standard practice in valve-voltmeters in which the first valve is a diode and the 'signal' thereafter is d.c. In general, however, the valve in the probe unit acts as an amplifier, and a coaxial cable is used to carry the signal to the main instrument. A typical circuit for dealing with fast pulses is given by Reed[6].

Although this is probably the best way to obtain both low input capacitance and good high frequency performance, both the probe and the cable (which must carry power supplies as well as signal) tend to be rather bulky.

Attenuators

An alternative method, which makes the design of a compact probe a very simple matter, is to use an attenuator and amplifier combination as shown in Fig. 2. The input impedance is greater than R_1 and C_1 in parallel and, if $R_1C_1 = R_2C_2$, the frequency response is determined solely by the properties of the amplifier, provided the length of the cable is very much less than a wavelength at the highest frequency of operation. If the amplifier gain A is made equal to $1 + R_1/R_2$, the signal is restored to its initial value.

To take a typical case, if C_2 is 100pF, and R_1 is 2MΩ,

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then R_2 may be made 100kΩ and the appropriate value of C_1 is 5pF. The amplifier is then required to have a gain of 21, and it is easy to design a single pentode stage having the required gain and a bandwidth of several megacycles per second.

If desired, the input capacitance can be reduced to zero (or, more accurately, to the self-capacitance of R_1) merely by omitting the compensating capacitor C_1. The bandwidth is then much reduced. In the example given, with C_1 omitted, the gain falls from its low frequency value by 3dB at about 17kc/s.

Feedback Systems

It has been said above that gain adequate to compensate for the attenuation together with wide bandwidth can be obtained from a single pentode valve. However, if constancy of gain is an important factor, as it generally is, a two or even three valve amplifier with negative feedback will be required. By using the circuit of Fig. 3, an overall gain of unity stabilized by negative feedback can be obtained from a single valve amplifier. The price paid for this simplicity is that the high frequency response, rather than the gain, varies with the valve characteristics. This is very often a less important consideration.

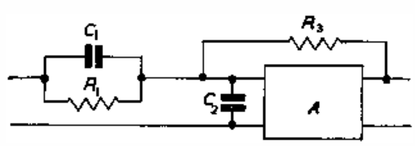

Fig. 3. Basic circuit of feedback amplifier
The amplifier gain A must be negative

The similarity, and difference, between the circuits of Fig. 2 and Fig. 3 can be appreciated by noting that there is an equivalent circuit for Fig. 3. In this, R_3 is removed and instead a resistance of magnitude $R_3/(1 - A)$ is connected across the input to the amplifier.

A More Exact Analysis

In order to predict accurately the performance of a practical circuit, it is necessary to take into account the fact that A will only be real and constant over a limited range of frequencies. Only the high frequency performance will be considered in detail. The optimum design will be taken as that which gives a minimum rise-time t_1 without overshoot on the output, when the input is a step function. The rise-time t_1 is defined as the time taken for the output signal to change from 10 to 90 per cent of its final value. The high frequency f_1 at which the response is 3dB below its mid-frequency value is then given with sufficient accuracy as $0{\cdot}4/t_1$.

The components of a practical pentode amplifier which are essential for the calculation of t_1 are shown in Fig. 4. The analysis is straightforward, and results only need be quoted.

If one assumes $R_4 \ll R_3$, then when reactances can be neglected, the overall gain A_1 is given by

$$A_1 = -(R_3/R_1)\,\frac{g_m R_4}{(1 + R_3/R_2 + R_3/R_1 + g_m R_4)}$$

$$\simeq -(R_3/R_1)$$

if $g_m R_4 \gg 1$, $g_m R_4 R_2 \gg R_3$, $g_m R_4 R_1 \gg R_3$.

Uncompensated Case

Consider now the case $C_1 = 0$. If there is to be no overshoot, then $R_3 C_2 \geqslant 4 g_m R_4^2 C_4$. For the critical value of R_3 which just fails to give an overshoot, the rise-time t_1 is about $1{\cdot}5\ R_3 C_2/g_m R_4$. If R_3 is greater than its critical value, then it is not possible to give a simple exact formula for t_1. However, when R_3 is appreciably greater than its critical value, the rise-time t_1 is approximately $2{\cdot}2\ R_3 C_2/g_m R_4$.

Inserting typical values in these expressions (g_m=6mA/V, R_4 = 5kΩ, C_4 = 15pF, C_2 = 150pF), it is found that the critical value of R_3 is 60kΩ, and the corresponding value of t_1 is about 0·5μsec. If R_3 is made 1MΩ, t_1 becomes about 11μsec.

Compensated Case

If C_1 is finite, then to avoid overshoot the conditions are $R_3(C_1 + C_2) > 4 g_m R_4^2 C_4$ and $R_1 C_1 g_m R_4 \leqslant R_3 C_2$. The second formula is a slight simplification of the exact formula, made on the assumption that $R_4 C_4 \ll R_1 C_1$. In the limiting case, when C_1 is just big enough not to give an overshoot, the rise-time is approximately $2{\cdot}2\ R_4 C_4$, provided $R_3(C_1 + C_2) \gg 4 g_m R_4^2 C_4$.

With the typical values given above for g_m, R_4, C_4 and C_2, letting $R_1 = R_3$ = 1MΩ, the critical value of C_1 is 5pF, and the rise-time is about 0·17μsec.

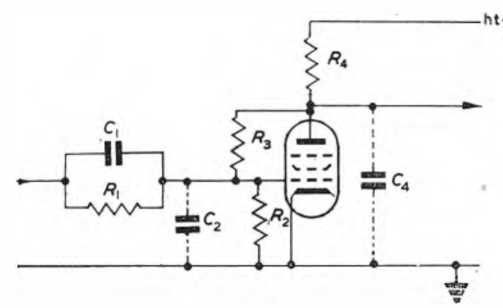

Fig. 4. Circuit for calculation of high frequency response

A Practical Circuit

It is clear from the above formulae that for good performance a valve having a large mutual conductance should be used, and that C_4, the total capacitance between its anode and earth, should be made as small as possible. C_4 is made up of valve and wiring capacitance, and the input capacitance of the main instrument. Unless the latter capacitance is very small (<10pF say), it is desirable to use a cathode-follower as a buffer stage, as shown in the practical circuit of Fig. 5. It then makes little difference whether the feedback is taken from the anode of V_1 or the cathode of V_2, provided the capacitance C_{11} is not appreciably greater than 100pF.

In order to stabilize the high frequency performance to a certain extent, the grid leak R_2 is taken to the junction of R_5 and R_6, which is about 8·7V positive with respect to earth. The relatively large auto-bias resistance ($R_9 + R_{10}$) then tends to keep the cathode current, and hence the mutual conductance, of V_1 constant despite valve replacement or ageing. In addition, this allows the value of R_2 to be made larger than the manufacturers' recommended value of 1MΩ for a normal auto-bias circuit. Although the value of R_2 is not critical, it is desirable to make it of at least the same order of magnitude as R_3.

The values of R_1, R_3, C_1 and the type of cable used are not given in the diagram; they may be chosen to give the desired values of input impedance and gain, on the basis of the discussion below. To obtain good low frequency performance, C_9 should be chosen so that $R_1 C_5 = R_3 C_9$. A suitable method of switching R_9 and C_9 is shown—the earthing pole of the switch is necessary to prevent unwanted feedback via the unused values of R_3 and the switch capacitance. When C_1 is not made zero, it is desirable to

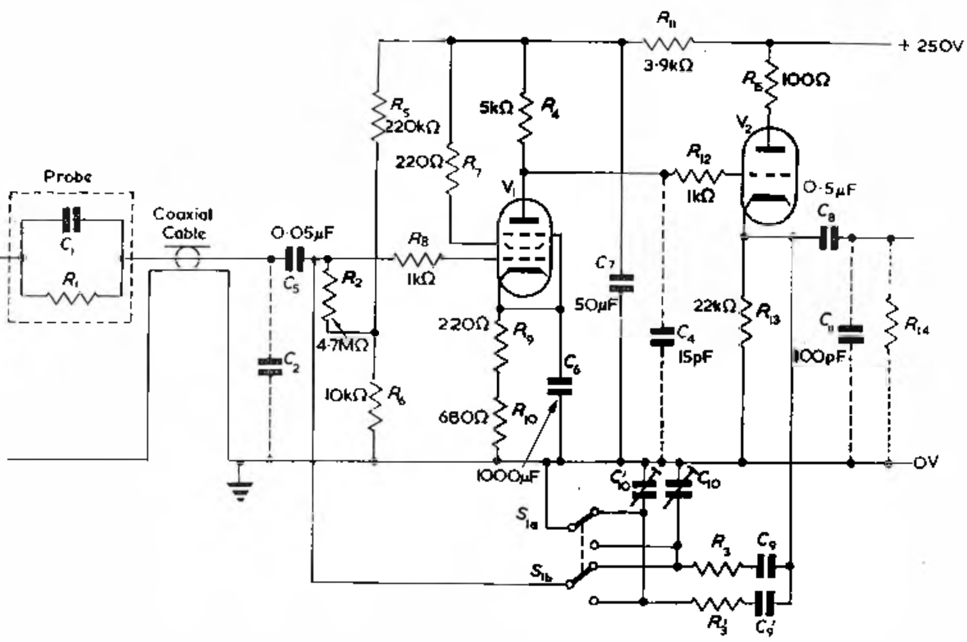

Fig. 5. Complete circuit of low capacitance input stage

C_2 represents capacitance of cable (see text) plus capacitance of input circuit (20pF)

V_1 and V_2—triode-pentode valve ECF80

Interchangeable probes or cables may be used if desired, but they have the disadvantage of being easily mislaid unless some means of stowage is provided. For most purposes the author has found it convenient to use only cable A, with $R_1 = 4{\cdot}7\text{M}\Omega$ and $C_1 = 5\text{pF}$. To enable the amplifier to handle a wide range of signals, R_3 may be chosen by means of a switch to be either $(4{\cdot}7 + 0{\cdot}47)\text{k}\Omega$ or $(470 + 22)\text{k}\Omega$. The small extra series resistors compensate for the finite values of $g_m R_4$ and R_2, since gains within 3 per cent of 1 and 1/10 are required. When the gain is unity, C_{10} is a 0 to 50pF trimming capacitor; when the gain is 1/10, C_{10} is an 0 to 500pF trimming capacitor in parallel with a fixed capacitor of 1 000pF.

have some fine adjustment to give minimum rise-time. The value of C_1 is the obvious parameter to vary, but, since the capacitor is mounted in the probe, it is better to use a small fixed component for C_1, and to set up the circuit by varying the preset capacitor C_{10} in parallel with C_2. When C_1 is omitted, C_{10} also is omitted.

The current drawn from the h.t. supply is about 20mA.

Performance

In order to check the correctness of the theory given above, the rise-time was measured using various values of R_1, R_2, C_1 and C_2. The calculated and observed values of the rise-time were within ± 20 per cent of each other. Apart from errors in measurement (± 20 per cent), discrepancies are to be expected, not only from the approximations used in the theoretical calculations, but from neglect of such factors as the distributed capacitance of the resistors, the non-linearity of the valves, and the effect of the capacitance C_{11} across the cathode-follower load.

The actual values of R_1, R_3 and C_1 may be chosen to meet the design requirements, on the basis of the data given below. Two typical cables (cable A and cable B) were used for test purposes. Cable A was 5ft long, and had a nominal capacitance of 22pF/ft; cable B was 3ft long, with a nominal capacitance of 10pF/ft.

When C_1 is zero, the rise-time is proportional to R_3, provided R_3 is greater than its critical value. R_1 is chosen to give the required value of input impedance or gain. Using cable A, the critical value of R_3 is about 60kΩ, and when R_3 is 1MΩ, the rise-time is about 8μsec. Using cable B, the critical value of R_3 is about 90kΩ, and when R_3 is 1MΩ, the rise-time is about 4μsec.

If C_1 is included in the circuit, its value is given by $R_1 C_1 g_m R_4 = R_3 C_2$. Since the overall gain A_1 is approximately equal to R_3/R_1, C_1 is approximately proportional to C_2/A_1. With a gain of unity, C_1 should be about 4·3pF with cable A, and about 1·8pF with cable B. In practice, C_1 is chosen to be a little larger than the theoretical value, and the final adjustment made by adjusting the preset capacitor C_{10} in parallel with C_2.

The low frequency performance has not been dealt with in detail. If $R_3 C_9$ is within 20 per cent of $R_1 C_5$, the 'sag' on a 10c/s square wave should be less than 5 per cent, if R_{14} is infinite. If R_{14} is appreciably less than 2MΩ, the performance is mainly determined by the time-constant of $C_8 R_{14}$. It is, of course, possible to compensate to a certain extent for the 'sag' introduced by $C_8 R_{14}$ by making $R_3 C_9 < R_1 C_5$. However, this is probably not worthwhile for most applications.

At a gain of unity the amplifier will handle square waves having a peak-to-peak amplitude of up to 20V without deterioration in performance, and an amplitude of up to 50V with only a slight increase in the 'fall time' produced by the capacitive loading on the cathode-follower.

Possible Modifications

If the input capacitance of the main instrument (C_{11} in Fig. 5) is sufficiently small, the cathode-follower stage may be omitted. If shorter rise-times are required, a factor of between two or three may be gained by using a valve of very high mutual conductance for V_1 (e.g. Mullard E180F, $g_m = 16{\cdot}5\text{mA/V}$), and a smaller value of anode load R_4. If large negative going steps are to be handled, R_{13} can be decreased in value.

The circuit lends itself to direct coupling if a negative h.t. supply is available. The mean potential of the grid of V_2 is then made a few volts negative with respect to earth by means of a compensated potential divider between the anode of V_1 and the negative h.t. line. The cathode load of V_2 is taken to the negative h.t. line, and the cathode of V_2 is then approximately at earth potential. With this arrangement R_2, R_5, R_6, C_5, C_8 and C_9 may be omitted, and R_9, R_{10} and C_6 either omitted or reduced in value.

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New Types of Germanium Diodes and their Circuit Applications

By G. Grimsdell*

The properties of the recently introduced gold-bonded diodes and small-area junction diodes are compared with those of the well-known point-contact types, and the relative advantages and disadvantages are outlined.

The requirements of controlled hole-storage and controlled turn-on time are discussed with particular reference to the new types.

THE germanium point contact diode has been extensively used for the last 10 years, and although its reliability and stability have been steadily improved, its basic construction of a pointed tungsten wire on a germanium crystal, has remained unchanged.

There are two main families of point-contact diodes; those made from a relatively impure germanium and having a low maximum reverse voltage, but capable of highly efficient operation as radio detectors up to 100Mc/s or more; and those made from a purer germanium with which reverse voltages of 80V, 100V, or even 150V, are achievable but with a less desirable performance at high frequency.

The two types are typified by the Services types CV442 and CV448 respectively, which can be taken as representative of the vast majority of germanium point contact diodes on sale today, although, of course, for some purposes, types made from germanium of a purity intermediate between the two offer a compromise of circuit performance better suited to some applications.

It is into this hitherto relatively stable situation that the new types of germanium diodes will find their way.

Silicon diodes will, of course, also enter the field, but their properties are quite different from those of germanium types and are not necessarily as good for all purposes.

In particular, silicon has very low reverse leakage, but a poor forward conductance at low forward voltages and higher capacitance, while the new germanium devices have a good forward conductance, a low capacitance and a very good hole storage performance.

The New Types of Germanium Diodes

Two new methods of making germanium diodes are of immediate interest. They provide two diode types which differ sufficiently from one another for them both to be considered for any particular application.

The Small Area Junction Diode

Until recently, germanium junction diodes were targeted on their power rectification applications and had large junctions giving good forward current characteristic but a poor reverse resistance, a very large reverse capacitance (hundreds of picofarads) and a hole storage characteristic which prevented their operation above a few hundred cycles.

Now, however, by reducing the junction area, choosing a suitable degree of impure germanium, and by adopting suitable after-treatments, it has proved possible to make a germanium diode capable of operating at a much higher frequency, having a capacitance of only 5 or 6pF, a reverse resistance at room temperature of several megohms, and a hole storage time of a fraction of a microsecond. The Mullard junction diode type OA-10 is an example of this type (see Fig. 1).

Gold Bonded Diodes

By exchanging the original tungsten wire in a point contact diode for a gold wire and when it is in position on the germanium, discharging a large pulse of current through it, a new type of diode is made called the gold-bonded diode.

These diodes have similar properties to the small-area

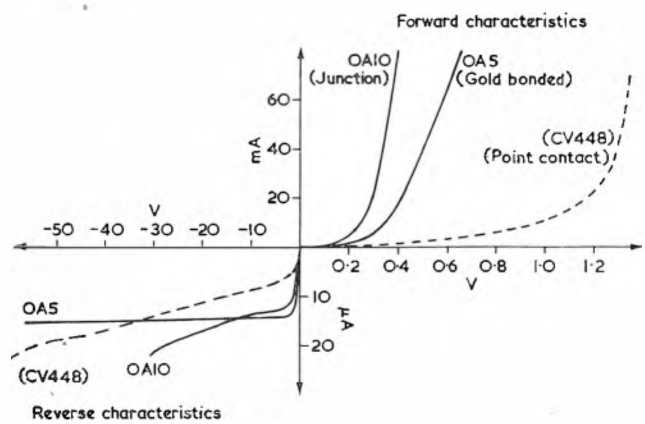

Fig. 1. Forward and reverse characteristics of germanium diodes

junction diode described above, but have even smaller reverse capacitances. Their rated peak currents are intermediate between those of the junction type and the original point contact.

Various gold-bonded diodes have been made in the U.S.A. but two types in particular appear to have found favour. They are the 1N270 and 1N277, for which the Mullard type OA-5 is an electrical replacement.

The Choice of a Suitable Diode for a Given Application

Taking point-contact diodes as the existing standard, the main improvement of the new types is in forward conductance (at, say, 1V) from the 3 or 4mA of the CV448 to 300mA or more for the OA-5 and over 1A for the OA-10.

In addition, the forward voltage at which conduction starts is also much lower in the new types (see Fig. 1).

One particular use for these new diodes will be in circuits with germanium transistors, particularly computing circuits. Here the h.t. is limited to a few volts only and the voltage drop across the diode can be important. With say, 5mA of signal current, four CV448 point-contact diodes in series will drop the full h.t. voltage, while 9 or 10 gold-bonded or junction diodes could be used for the same voltage drop.

Waveform clipping, diode pump counting circuits, diode gating, diode speech paths and diode modulators are all

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applications where the new types will be, in most cases, more useful in the circuit.

The reduction in forward impedance reduces the power loss in the diode and thus its internal heating. For the very high currents met with in the square loop ferrite memory and computing applications this is important, the small area junction diode being the only type particularly targeted towards the 1A 'full pulses' found in that application.

These high current ratings also make the small-area junction diode suitable for use as a miniature mains rectifier for transistor h.t. supplies, valve bias supplies and the like.

The low hole storage characteristic ensure a reasonable rectification efficiency at 400c/s, 2kc/s or at even higher frequencies.

Conclusion

The two new types of rectifier offer a much better diode performance in the forward current direction than the older point contact diodes; they are about the same as point contact diodes in the reverse direction.

Of the two new types, the small area junction is rather better in forward performance than the gold bonded type, but a little worse than the latter in reverse capacitance.

The small area junction can, of course, handle much larger currents and higher powers than the gold-bonded technique.

Acknowledgments

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

Filter Attenuation Characteristics

Derivation of Formulæ allowing Slide-rule Computation

By M. D. Johnson* and D. A. G. Tait*

Expressions for the attenuation of ideally-terminated m-derived filters are derived in a form that avoids the use of hyperbolic functions. It is found that the resulting expressions are quite simple, and an attenuation curve is given showing their application to a specific case.

DURING the design of a simple two-section Zobel filter, it became obvious that the commonly-quoted expressions ($\alpha = \cosh^{-1}[1 + (Z_1/2Z_2)]$ etc.) were inconvenient for the speedy computation of theoretical attenuation in the stop-band, since they involve the use of both slide-rule *and* tables of hyperbolic functions. An alternative expression has therefore been derived (equation (6)) the use of which in conjunction with a slide-rule having a scale of $\sqrt{(1-x^2)}$ avoids entirely the need for mathematical tables. As shown in the appendix, well-known simple expressions giving the salient features of m-derived sections are easily obtained from this expression. A not so well known expression is also derived, from which may be obtained the frequency of minimum attenuation, and the value of attenuation, when two m-derived sections are cascaded. It should be possible to extend the use of these equations ((10) and (11)) to filters of more than two sections by taking in pairs sections having adjacent m-values†.

Results are summarized below, and in Fig. 1 for the case of a two-section filter. It will be seen that the expressions are quite simple. In conjunction with a knowledge of the effects of dissipation and reflection they should allow a speedy calculation of the performance of simple filters in the stop-band. Moreover, those engineers who are versed in the art of mentally converting ratios to decibels should find it possible to estimate the salient features of performance of a simple filter without recourse to either tables or slide-rule. (N.B. The expressions within the modulus sign are not complex in the stop-band. The significance of 'modulus of' is merely that a resultant negative sign may be ignored.)

SYMBOLS USED

- A = attenuation of one section of a filter, in decibels (dB)
- A_T = the total attenuation of a two-section filter.
- Z_1 = total series impedance of a filter section.
- Z_2 = total shunt impedance of a filter section.
- Z_{1K} = Z_1 of a prototype (constant k) section.
- Z_{2K} = Z_2 of a prototype (constant k) section.
- L = total series or shunt inductance of a dissipationless prototype section.
- C = total series or shunt capacitance of a dissipationless prototype section.
- m_1 = the derivation factor of one of two sections of a composite filter.
- m_2 = the derivation factor of the second of two sections of a composite filter. (A prototype section is regarded as the limiting case—m = 1—of a derived section).
- x = the normalized frequency variable $\omega/\omega_c = f/f_c$ or $\omega_c/\omega = f_c/f$ for a high-pass or low-pass filter respectively.
- ω_c = $2\pi f_c$, where f_c is the frequency of transition between stop-band and pass-band.

Summary of Results

The frequency of infinite attenuation of the n^{th} section of any filter is at

$$x_n = \sqrt{(1 - m_n^2)} \quad \text{(appendix equation (7))}$$

With two sections in cascade, minimum attenuation is at

$$x = \sqrt{(1 - m_1 m_2)} \quad \text{(appendix equation (10))}$$

when the attenuation of each section is the same and given by

$$A = 20\log\left|\frac{\sqrt{(m_1/m_2)}+1}{\sqrt{(m_1/m_2)}-1}\right| \text{dB} \quad \text{(appendix equation (11))}$$

(To avoid mistakes due to change of sign, use $m_1 > m_2$)

When $x = 0$, (corresponding to infinite freqeuncy for the

* *R. B. Pullin and Co., Ltd.*

† *For example see* MOLE, J. H., Filter Design Data for Communication Engineers, p. 24 (E. and F. N. Spon, Ltd.).

low-pass filter), the attenuation per section is given by

$$A = 20\log \frac{1+m}{1-m}\,\text{dB} \quad \text{(appendix equation (8))}$$

It is interesting to note that the attenuation is only 1dB greater than this limiting case at $x = 0{\cdot}2$ for $m = 0{\cdot}8$, and much closer for lesser m values.

Where a slide- rule having a scale of $\sqrt{(1-x^2)}$ is available, the following expression can be used for plotting intermediate points:

$$A = 20\log\left|\frac{m+\sqrt{(1-x^2)}}{m-\sqrt{(1-x^2)}}\right|\text{dB (appendix equation (6))}$$

Note that if $y = \sqrt{(1-x^2)}$, then $x = \sqrt{(1-y^2)}$; thus, to maintain accuracy for values of x greater than $0{\cdot}707$, reverse the role of the two scales.

Appendix

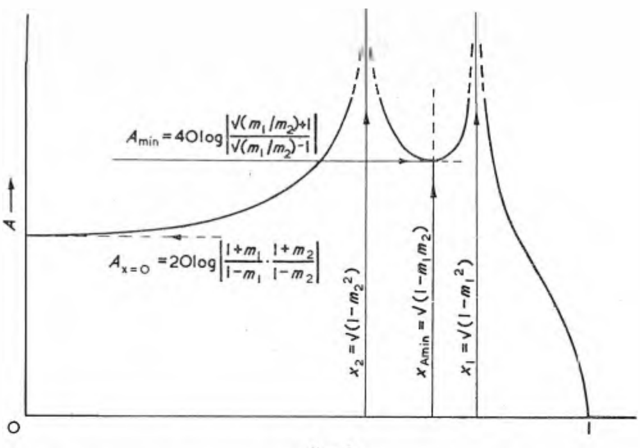

Fig. 1. Theoretical response of 2-section filter showing application of various formulae

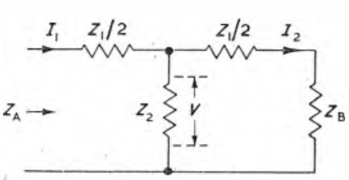

Fig. 2. Analysis of m-derived sections

Equating $Z_A = Z_B = Z_o$ yields (see Fig. 2)

$$Z_o = \sqrt{[Z_1Z_2 + (Z_1/2)^2]} \quad \dots\dots\dots\dots (1)$$

Now $V = Z_2(I_1 - I_2) = I_2(Z_1/2 + Z_B) = I_2(Z_1/2 + Z_o)$

$$\therefore I_1/I_2 = Z_1/2Z_2 + Z_o/Z_2 + 1$$

$$= \frac{1 + Z_1/2 + \sqrt{[Z_1Z_2 + (Z_1/2)^2]}}{Z_2} \quad \dots\dots\dots\dots (2)$$

By definition, the attenuation in decibels is given by

$$A = 10\log\ I_1^2R/I_2^2R$$

where R is the real part of Z_o

$$\therefore A = 20\log|I_1/I_2|$$

$$= 20\ \log\left|1 + \frac{Z_1/2 + \sqrt{[Z_1Z_2 + (Z_1/2)^2]}}{Z_2}\right| \dots\dots (3)$$

For an m-derived section,

$$Z_1 = mZ_{1K},\ Z_2 = Z_{2K}/m + \frac{Z_{1K}(1-m^2)}{4m}$$

where Z_{1K} and Z_{2K} are the total series and shunt impedances of a constant-k filter.

For a low-pass filter, $Z_{1K} = j\omega L_1\ Z_{2K} = 1/j\omega C$, and $\omega_c^2 = 4/LC$.

$$\therefore Z_{2K}/Z_{1K} = \frac{-1}{\omega^2/LC} = -\tfrac{1}{4}\,.\,\omega_c^2/\omega^2 = -x^2/4 \quad \dots\dots (4)$$

where $x = \omega_c/\omega$.

For a high-pass filter, $Z_{1K} = 1/j\omega C$, $Z_{2K} = j\omega L$, and $\omega_c^2 = 1/4LC$.

$$\therefore Z_{K2}/Z_{1K} = -\omega^2LC = -\tfrac{1}{4}\ .\ \omega^2/\omega_c^2 = -x^2/4 \quad \dots\dots (5)$$

where $x = \omega/\omega_c$.

Thus with the above definitions of x, Z_1 and Z_2, for a high-pass and a low-pass filter, and using results (4) and (5), equation (3) becomes

$$A = 20\log\left|1 + \frac{mZ_{1K}/2 + \sqrt{[Z_{1K}Z_{2K} + (Z_{1K}/2)^2]}}{Z_{2K}/m + \dfrac{Z_{1K}(1-m^2)}{4m}}\right|$$

$$= 20\log\left|1 + \frac{m + \sqrt{[(4Z_{2K}/Z_{1K}) + 1]}}{2Z_{2K}/mZ_{1K} + (1-m^2)/2m}\right|$$

$$= 20\log\left|1 + \frac{m + \sqrt{(1-x^2)}}{\dfrac{-x^2 + (1-m^2)}{2m}}\right|$$

$$= 20\log\left|\frac{1 - x^2 - m^2 + 2m^2 + 2m\sqrt{(1-x^2)}}{1 - x^2 - m^2}\right|$$

$$= 20\log\left|\frac{[m + \sqrt{(1-x^2)}]^2}{m^2 - [\sqrt{(1-x^2)}]^2}\right| = 20\log\left|\frac{m + \sqrt{(1-x^2)}}{m - \sqrt{(1-x^2)}}\right|\text{dB}$$

$$\dots\dots\dots\dots (6)$$

When $A = \infty$, $m = \sqrt{(1-x^2)}$ $\therefore x^2 = \sqrt{(1-m^2)}$ .... (7)

When $x = 0$, $A = 20\log\left|\dfrac{1+m}{1-m}\right|$ dB .... (8)

For two sections in cascade, the total attenuation is given by

$$A_T = A_{(m1)} + A_{(m2)} = 20\log\left|\frac{m_1 + y}{m_1 - y}\cdot\frac{m_2 + y}{m_2 - y}\right|$$

$$= 20\log f(y) \text{ say}$$

(from equation (6) and putting $\sqrt{(1-x^2)} = y$)

$$\therefore dA_T/dt = \frac{20\log e}{f(y)}\,.\,f'(y) = 0 \text{ for minimum attenuation}$$

$$\therefore f'(y) = 0$$

Thus

$$\frac{m_1 + y}{m_1 - y}\cdot\frac{2m_2}{(m_2 - y)^2} + \frac{m_2 + y}{m_2 - y}\cdot\frac{2m_1}{(m_1 - y)^2} = 0$$

$$\therefore m_2(m_1^2 - y^2) + m_1(m_2^2 - y^2) = 0$$

$$\therefore y^2(m_1 + m_2) = m_1m_2(m_1 + m_2)$$

$$\therefore y^2 (= 1 - x^2) = m_1m_2 \quad \dots\dots\dots\dots (9)$$

$$\text{or } x = \sqrt{(1 - m_1m_2)} \quad \dots\dots\dots\dots (10)$$

Substituting equation (9) in (6) gives the attention for each section at the frequency of minimum attenuation of the combination. Thus

$$A_{(m1)} = 20\log\left|\frac{m_1 + \sqrt{(m_1m_2)}}{m_1 - \sqrt{(m_1m_2)}}\right| = 20\log\left|\frac{\sqrt{m_1} + \sqrt{m_2}}{\sqrt{m_1} - \sqrt{m_2}}\right|\text{ dB}$$

which is symmetrical,

i.e.,

$$A_{(m1)} = A_{(m2)} = 20\log\left|\frac{\sqrt{(m_1/m_2)} + 1}{\sqrt{(m_1/m_2)} - 1}\right|\text{ dB} \dots (11)$$

The Influence of the Output Time-Constant of a Cathode-Follower

By C. Edwards

The purpose of this article is to examine the influence of the output time-constant, λ, on the maximum amplitude of a positive pulse that may be applied to a cathode-follower without causing positive grid current to flow. The method used is to develop an equivalent generator circuit for the simple cathode-follower and to determine the response of this circuit to a pulse of rise-time T_r. This analysis yields a Reduction Factor that, when plotted in the form of a nomogram enables the peak amplitude of pulse input to be determined graphically over a wide range of T_r/λ.

FIG. 1 shows a simple cathode-follower circuit having an output time-constant λ. If a positive signal of amplitude E and rise-time T_r is to be applied to the grid and if no grid current is to flow then the maximum permissible value of E is a function of the ratio T_r/λ.

$$\text{i.e. } E_{max} = f\left\{ T_r/\lambda \right\} \quad \dots\dots\dots \quad (1)$$

When the output time-constant, λ, is negligible by comparison with the rise-time, T_r, of the signal then the value of E_{max} may be obtained from the anode characteristic curves and the load-line as in Fig. 2 which shows the I_a E_{ak} curves,

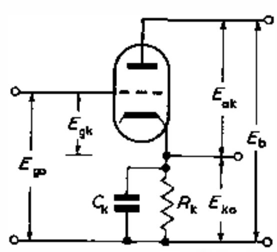

Fig. 1. The cathode-follower $\lambda = \dfrac{C_k R_k}{1 + g_m R_k}$

of the valve of Fig. 1, for several values of E_{gk}, superimposed on these are the curves of I_a E_{ko} for several values of E_{go}; these curves have been obtained from the I_a/E_{ak} curves and the equations

$$E_{ko} = E_b - E_{ak} \quad \dots\dots\dots \quad (2)$$

$$E_{go} = E_{ko} + E_{gk} \quad \dots\dots\dots \quad (3)$$

The load line has been drawn for the d.c. load R_k. Q is the quiescent point and E_{max} is that value of input signal amplitude which would carry the operating point from Q to point P where $E_{gk} = 0$. The value of E_{max} is given by the intercept PQ on the E_{go} curves or if the I_a E_{ak} curves and load line only are used then E_{max} is the value of the intercept pq on the E_{ak} scale plus the standing grid to cathode bias.

If λ is not negligible by comparison with T_r the cathode potential is no longer able to change at the same rate as the grid potential; then the value of E_{max} given by the I_a E_{ak} curves and load line must be reduced in order that the change in grid-cathode potential, δE_{gk}, may not exceed the standing grid-cathode bias on the valve and cause grid current to flow.

The purpose of this article is to develop a reduction factor that may be applied to the value of E_{max} given by the graphical method of Fig. 2 in order to extend the results of this particular case (i.e. $\lambda = 0$) to the general case $0 < \lambda$. A nomogram is given for the evaluation of the reduction factor over the range of values $0{\cdot}01 \leqslant T_r/\lambda \leqslant 100$. In this article it is assumed that grid current flows only when the grid is positive with respect to the cathode, it is also assumed that the input signal, δE_{go}, is a linear rise in e.m.f. from zero at time $t = 0$ to the value E at time $t = T_r$ and is constant afterwards; i.e. when $0 \leqslant t \leqslant T_r$, Fig. 3.

$$\delta E_{go} \Big|_{t=0}^{t=T_r} = (E \,.\, t/T_r) \quad \dots\dots\dots \quad (4)$$

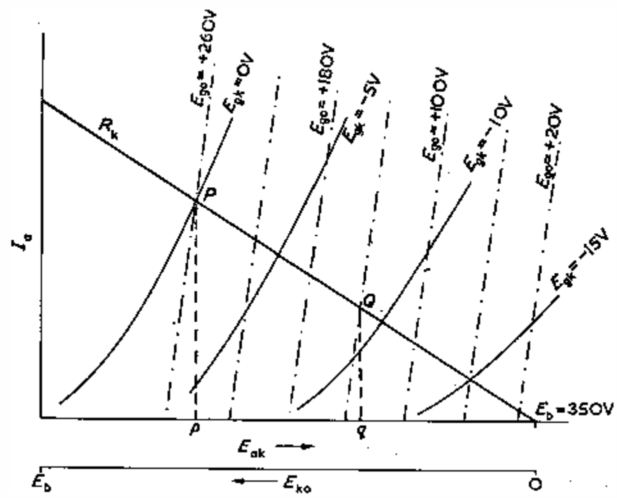

Fig. 2. Curves of Fig. 1. $C_k = 0$

In order to employ mathematical methods in the development of the required reduction factor an equivalent generator circuit for the valve of Fig. 1 will first be obtained. This is derived as follows:—

By Taylor's theorem:—

$$I_a + \delta I_a = \phi\left\{E_{go} + \delta E_{go},\ E_{ko} + \delta E_{ko}\right\}$$

$$= \left\{ \exp\left(\delta E_{go}\frac{\partial}{\partial E_{go}} + \delta E_{ko}\frac{\partial}{\partial E_{ko}}\right)\right\} \phi\left\{ E_{go}, E_{ko}\right\} \dots \quad (5)$$

By the 3/2 power law:—

$$I_a = k\left\{ E_{gk} + (E_{ak}/\mu)\right\}^{3/2}$$

$$= k\left\{E_{go} + (E_b/\mu) - \frac{\mu+1}{\mu}\, E_{ko}\right\}^{3/2} \quad \dots \quad (6)$$

Neglecting differential coefficients beyond the first, then from equations (5) and (6)

$$\frac{\partial I_a}{\partial E_{go}} = 3/2\, k\left\{E_{go} + (E_b/\mu) - \frac{\mu+1}{\mu}\, E_{ko}\right\}^{1/2}$$

$$= 3/2\, k^{2/3} I_a^{1/3} \quad \dots\dots\dots \quad (7)$$

$$\frac{\partial I_a}{\partial E_{ko}} = -\,3/2\, k\left(\frac{\mu+1}{\mu}\right)\left\{E_{go} + (E_b/\mu) - \left(\frac{\mu+1}{\mu}\right)E_{ko}\right\}^{\frac{1}{2}}$$

$$= -\,3/2\, k^{2/3}\left(\frac{\mu+1}{\mu}\right) \; I_a^{1/3} \quad \dots\dots\dots \quad (8)$$

$$\therefore \frac{\partial I_a}{\partial E_{go}} = \frac{\partial I_a}{\partial E_{gk}} = g_m \quad \ldots\ldots\ldots \quad (9)$$

and

$$\frac{\partial I_a}{\partial E_{ko}} = (\mu+1)\frac{\partial I_a}{\partial E_{ak}} = \frac{\mu+1}{r_a} \quad \ldots\ldots \quad (10)$$

$$\therefore \delta I_a = g_m\,\delta E_{go} - \left(\frac{\mu+1}{\mu}\right) g_m\,\delta E_{ko} \quad \ldots\ldots \quad (11)$$

The logical interpretation of equation (11) is that the valve of Fig. 1 acts as a current generator having an internal impedance equal to $\left\{\left(\frac{\mu+1}{\mu}\right) g_m\right\}^{-1}$ and generating a

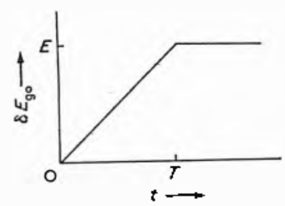

Fig. 3. The input signal

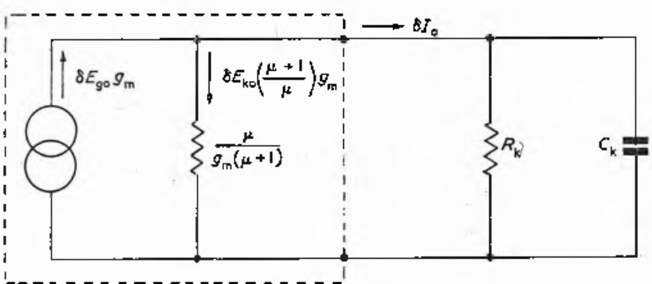

Fig. 4. Equivalent circuit to Fig. 1

current equal to $g_m \delta E_{go}$, Fig. 4; therefore for the present purposes the circuit of Fig. 4 may be taken as the equivalent of that of Fig. 1. The analysis may now proceed as follows:

From Figs. 4 and 3

$$g_m \frac{E\,.\,t}{T_r} = I\{R_k\} + I\left\{\frac{\mu}{(\mu+1)g_m}\right\} + I\{C_k\} \quad \ldots\ldots \quad (12)$$

But the output signal is the potential drop across R_k; let this p.d. $= v$

then

$$v = R_k I\{R_k\} = \frac{\mu}{(\mu+1)\,g_m} \,.\, I\left\{\frac{\mu}{(\mu+1)g_m}\right\}$$
$$= 1/C_k \int I\{C_k\}\,dt \quad \ldots \quad (13)$$

therefore

$$dv/dt + (1/\lambda)\,v = g_m/C_k \;\; E\,.\,t/T_r \quad \ldots\ldots \quad (14)$$

The integrating factor is exp $(1/\lambda \int dt)$ and integration of equation (14) gives

$$v e^{t/\lambda} = g_m/C_k \;\; E/T_r\{t - \lambda\}\lambda e^{t/\lambda} + K \quad \ldots\ldots \quad (15)$$

from the boundary conditions ($t = 0$, $v = 0$).

$$K = g_m\lambda/C_k \,.\, E\lambda/T_r \quad \ldots\ldots\ldots\ldots \quad (16)$$

from equations (15) and (16)

$$v = \frac{\mu\,(R_k/r_a)}{(\mu+1)\,(R_k/r_a)+1} \,.\, Et/T_r \,.\, \left\{1 - \lambda/t\,(1 - e^{-t/\lambda})\right\} \quad \ldots\ldots\ldots\ldots \quad (17)$$

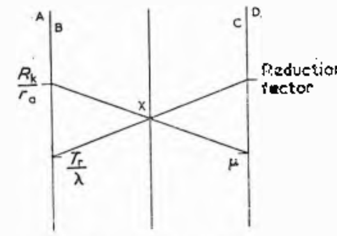

Fig. 5. Key to Fig. 6

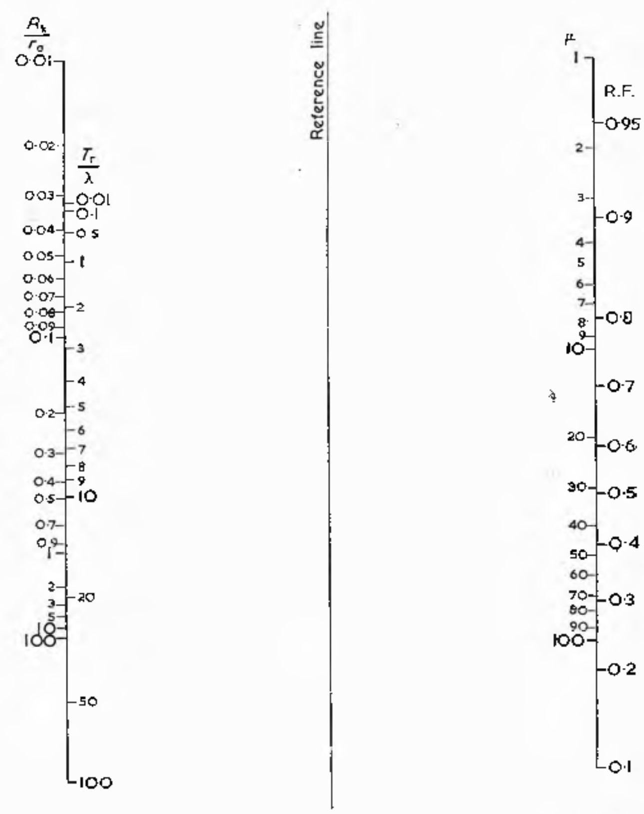

Fig. 6. Nomogram Reduction Factor $= \left[1 + \left(\frac{\mu(R_k/r_a)}{(R_k/r_a)+1}\right)\frac{1 - e^{-T_r/\lambda}}{T_r/\lambda}\right]^{-1}$

and the voltage 'gain' of the circuit is

$$A = \frac{\mu\,(R_k/r_a)}{(\mu+1)\,(R_k/r_a)+1}\left\{1 - \lambda/t\,(1 - e^{-t/\lambda})\right\} \quad \ldots \quad (18)$$

For no grid current to flow δE_{gk} must not be greater than the standing grid/cathode bias. δE_{gk} is a maximum when $t = T_r$ therefore

$$E_{max} = \frac{\text{Grid-cathode Bias}}{1 - \frac{\mu\,(R_k/r_a)}{(\mu+1)\,(R_k/r_a)+1}\left\{1 - \lambda/T_r\,(1 - e^{-T_r/\lambda})\right\}} \quad \ldots\ldots\ldots\ldots \quad (19)$$

If λ is negligible by comparison with T_r then equation (19) becomes

$$\{E_{max}\}_{C_k=0} = \frac{\text{Grid-cathode Bias}}{1 - \dfrac{\mu\,(R_k/r_a)}{(\mu+1)\,(R_k/r_a)+1}} \quad \ldots\ldots\ldots\ldots \quad (20)$$

This is the value of E_{max} found by the graphical method of Fig. 2. The reduction factor required to modify the value of $E_{max\ C_k=0}$ to cover the general case $0<\lambda$ is given by $\dfrac{E_{max}}{E_{max\ C_k=0}}$ therefore from equations (19) and (20):—

$$\text{Reduction Factor} = \left\{1 + \left(\frac{\mu\,(R_k/r_a)}{(R_k/r_a)+1}\right)\frac{1 - e^{-T_r/\lambda}}{T_r/\lambda}\right\}^{-1} \quad (21)$$

This equation is plotted in the form of a nomogram in Fig. 6; the range of values covered is $0{\cdot}01 \leqslant T_r/\lambda \leqslant 100$. Fig. 5 is the key to the use of the nomogram; a straight line joining the appropriate values of R_k/r_a, scale A, and μ, scale C, cuts the reference line at a point X. A second straight line from the appropriate value of T_r/λ, scale B, and passing through the point X intercepts scale D at the required value of the reduction factor.

Surface Barrier Transistor Manufacture

A new factory for the production of surface barrier transistors has recently been brought into operation by Semiconductors Ltd. a company which has been formed jointly by The Plessey Co. Ltd. and the Philco Corporation of America.

The factory, which is at Swindon, is windowless and is completely insulated from the outside atmosphere. An air conditioning plant controls the ambient temperature to 2°F, humidity to ±5 per cent and, by electrostatic and mechanical filtration, controls ambient dust to less than one cubic micron per centimetre of air. Such safeguards are indispensable for the production of a highly reliable transistor.

Facilities are provided in the factory for supplying deionized water of 18MΩ/cm resistivity. Ultra pure gases are piped through the plant and are available at strategically located points. To control dust, the whole factory is free from overhead piping, all services being brought up through the floor. All internal surfaces are p.v.c. coated to be smooth and non-dust-catching. Thirteen separate air conditioning units are set up so that separate zones can be controlled and maintained and sufficient flexibility is built into the plant so that the latest developments and techniques can be incorporated as they come along with a minimum of disruption.

Transistors are manufactured by electro-chemical techniques and a large degree of automation has been incorporated in the production lines giving exceptionally close control of tolerances and purity.

The production process is briefly as follows:

The basic material is germanium which has been purified by zone-refining to reduce the impurity content to below 1 part in 10^9. This is melted in a crucible by radio-frequency heating, the crucible being contained in a silica tube filled with argon. A measured amount of antimony is added to obtain the required electrical characteristics in the finished crystal. A temperature detector and sensitive servo-amplifier maintain the germanium at a constant temperature just above its freezing point (about 900°C). A small 'seed' crystal is dipped into the molten germanium, and as it is withdrawn, the germanium crystallizes on to the seed to form a single crystal. Both the 'seed' and the crucible rotate to obtain a uniform temperature distribution.

After electrical tests to check the quality, the crystal, typically 8in long and 1in in diameter, is sawn into thin disks by a high-speed diamond-impregnated wheel and a planetary lapping machine is used to obtain a uniform thickness over the whole diameter of each of the germanium disks. Since the abrasive used in the lapping machine tends to disrupt the crystal structure on the surface of the germanium, the disks are then chemically etched in a nest of rotating plastic beakers.

The disk now has a high polish and is ready for conversion into the small rectangular blanks required in the transistor. A wafer scriber scratches the surface of the glass-hard germanium with a diamond tool, and a vacuum pad then gently cracks the germanium along the scratch to convert the disk into a series of strips. A second similar machine then breaks the strips into a number of small rectangular blanks.

Since it is vital to the accuracy of the later processes that the blanks should be sorted into groups of similar thickness, they are passed to a roller thickness gauge, which automatically sorts them into small bins with an accuracy of ±·00005in.

Coming next to the assembly stage, the transistor stem, consisting of three wires sealed in glass inside a metal ring, is mounted in a work carrier, which is used as an indexing jig for the later precision operations. A small rectangular nickel tab is welded to the centre wire of the three in the stem, the function of the tab being to support the germanium blank.

A germanium blank is now soldered to the nickel tab, and the assembly is ready to be fed into the automatic transfer machine, after checking in a dual beam comparator which projects a magnified plan and elevation of the assembly on to a ground-class screen to ensure that its dimensions and location relative to the work carrier are within the necessary limits.

An automatic transfer machine automatically performs the three key operations in the production of the surface barrier transistors, as well as intermediate washing, drying and testing. After a preliminary wash in deionized water, the assembly is fed to the rough etch position, where two jets of electrolyte impinging on opposite sides of the germanium wafer produce concentric pits. Next follows the precision etch position, which has a single jet carrying an infra-red monitoring beam. Since the intensity of the infra-red transmission is determined by the thickness of the germanium at the bottom of the pit, a detector on the other side can feed a 'thickness' signal to a servo-amplifier which stops the etching process when the required thickness is reached. A pen-recorder allows the progress of each etching cycle to be observed. This precision etching is the most important process of all; it gives a precise control of the thickness of the germanium, and this thickness determines the high-frequency performance of the transistor. Another washing cycle follows, and the assembly is then fed to the precision plate position. Two jets of electrolyte are used, the electrical bias and electrolyte composition causing indium electrodes of the required diameter to be plated into the etch pits. After washing and drying with hot nitrogen, the transistors next go to a position where contact is made to the indium electrodes and an electrical check is made. This measurement is also recorded on a chart.

An automatic whisker attacher prepares thin nickel wire fed from a spool by plating a measured amount of special solder on the end, bending to the required shape, and cutting the wire from the spool. When the transistor is fed into the machine, an optical servo system locates the wire precisely on the centre of the transistor electrode and solders it on by radiant heat. It now remains for the free end of the whisker wire to be attached to one of the stem wires by the whisker welding equipment. Since this is a simple non-critical operation it is performed manually under a magnifying glass.

The transistor is now cleaned chemically and rinsed with deionized water. A baking cycle in a vacuum oven removes all traces of moisture before the transistors are passed directly into a nitrogen-filled 'dry-box'. Here the top caps, which have previously been silicone grease filled, are also passed into the 'dry-box' via the vacuum ovens and welded on, the transistors therefore being hermetically sealed before they are again exposed to the atmosphere. The completed transistors now go to the final test bay for comprehensive checking of their electrical characteristics. A proportion of each batch goes into life test equipment where some transistors are run at their maximum ratings and others at elevated temperature to simulate, in a comparatively short time, the effect of a very long period of operation under normal conditions. Provided that these tests are successful, the remainder of the batch has its type number printed on and is ready for sale.

The transistors manufactured by this process have very good high frequency characteristics (oscillation frequencies up to 100Mc/s) and they may be used in r.f., i.f. and vidieo circuits or in high speed switching circuits. Other types of transistors will be made at this factory in due course.

Linear Thyratron Control Circuits

By G. G. E. Low*, M.Sc., Ph.D.

A description is given of two simple circuits in which thyratrons are used in such a way as to provide a linear relationship between a slowly varying d.c. signal voltage at the grid and the average anode current. The anodes of the thyratrons are supplied from an a.c. source in the usual manner and clearly the circuits are limited to applications in which the variations in the signal voltage are slow compared with the frequency of the supply.

THYRATRONS are frequently used to control small electric motors and other electromechanical devices, but such applications do not usually call for linearity between the output current and the input control signal. However, if required a reasonably linear relationship between a d.c. signal voltage at the grid of a thyratron and the average anode current may be achieved in a simple manner. For example, such a relationship is provided by the method of control in which a small fixed a.c. voltage, 90° out of phase with the anode supply, is applied to the grid. That this is so may be seen by considering an idealized thyratron which, for a positive anode potential,

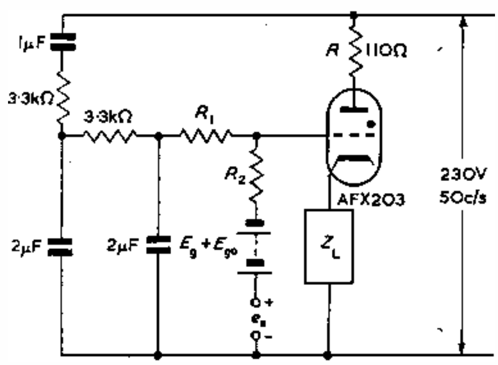

Fig. 1. The thyratron control circuit

conducts if the grid potential exceeds a fixed value of $-E_{go}$. Suppose that the anode and the grid respectively of this valve are supplied with voltages of

$$e_a = E_a \cos \omega t$$

and

$$e_g = e_s + E_g (\sin \omega t - 1) - E_{go}$$

where the input signal voltage e_s varies slowly compared with $\cos \omega t$ and has a magnitude less than $2E_g$. If a resistor R is included in the anode circuit, then, on the assumption of negligible voltage drop across the thyratron, an anode current of

$$i_a = e_a/R = E_a \cos \omega t/R$$

will flow during each conduction period. It follows that the total charge flow per cycle is

$$\int_{e_g=-E_{go}}^{e_a=0} i_a dt = E_a/R \int_{\sin \omega t = 1 - e_s/E_g}^{\sin \omega t = 1} \cos \omega t \, dt = \frac{E_a}{\omega R E_g} \, e_s$$

and that the average anode current is given by

$$i_a' = \frac{E_a}{2\pi R E_g} \, e_s$$

Thus, there is a linear relationship between i_a' and e_s and from the point of view of these two quantities the system as a whole has a mutual conductance of $E_a/2\pi R E_g$.

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The above analysis is based on an idealized thyratron. It is to be expected that any actual circuit will not follow completely the behaviour predicted because of departures from the assumptions that $-E_{go}$ is a fixed quantity independent of anode potential and that the voltage dropped across the valve during conduction is negligible. However, reason-

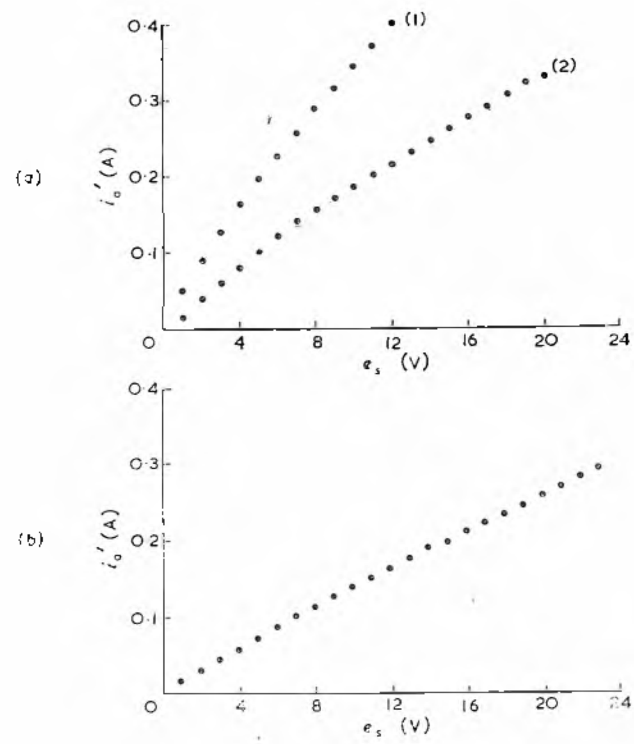

Fig. 2. (a) Relationship between average anode current and d.c. signal voltage for the circuit shown in Fig. 1.

When Z_L = 0. Curves (1) and (2) correspond to grid resistor values (R_2) of 10Ω and 5kΩ respectively ($R_1 + R_2$ = 20kΩ)

(b) Average anode current plotted against d.c. signal voltage for the circuit shown in Fig. 1

When Z_L has a resistance of 50Ω and a time-constant of about ½sec

able linearity is achieved with real thyratrons as may be seen from the curves in Fig. 2(a). These results were obtained using the simple circuit given in Fig. 1 with the cathode load Z_L equal to zero. (The components in the grid phase shift network are adjusted so as to give a 90° phase shift, the values shown being only approximate.) The departures from linearity found for low anode currents arise partly from the breakdown in this region of the assumption of negligible voltage drop across the conducting thyratron. An additional complication is the change in grid potential resulting from grid current flowing through the resistor in the grid-cathode circuit. The effect on linearity of changes in this resistor is shown in Fig. 2(a). Curve (1) refers to a 10kΩ resistor, while curve (2) gives results for a 5kΩ resistor. (The difference in average slope between the two curves arises as a consequence of a change in E_g. In each case the slope is in agreement with the mutual conductance calculated from the expression given above).

Should the device to be controlled have an electrical time-constant long compared with $2\pi/\omega$ or if its time-constant be increased to meet this condition by connecting capacitance in parallel with it, then an action analogous to that of the cathode-follower results if the device is connected into the cathode circuit of the thyratron (i.e., as Z_L is in Fig. 1). In these circumstances the average d.c. voltage established across a device Z_L of resistance R_L is given by

$$\frac{R_L E_a}{2\pi R E_g} e_i$$

where e_i is the signal voltage between grid and cathode. (Although it is necessary that the device have a time-constant long compared with $2\pi/\omega$ it must not be so long as to impair the device's response to e_s.) Thus as

$$e_i = e_s - \frac{R_L E_a}{2\pi R E_g} e_i$$

the voltage across Z_L may be expressed as

$$\left(\frac{2\pi R E_g}{R_L E_a} + 1\right)^{-1} e_s$$

and the circuit as a whole may be considered to have a mutual conductance of

$$\left(\frac{2\pi R E_g}{E_a} + R_L\right)^{-1}$$

The output resistance of the circuit driving the load Z_L is given by $2\pi R E_g/E_a$. Fig. 2(b) shows a plot of cathode current as a function of input signal voltage obtained using the circuit given in Fig. 1 with Z_L included. The improvement in linearity afforded by the negative feedback may be seen. For these measurements Z_L had a resistance of about 50Ω and a time-constant of approximately $\frac{1}{2}$sec. The grid resistors R_1 and R_2 were both 10kΩ. A measurement of the output resistance of the circuit from the point of view of Z_L gave a value of roughly 40Ω. It would appear that the armature winding of an electric motor could very well be used as a cathode load (Z_L) in the above circuit as the back e.m.f. developed would provide the reference voltage necessary for the cathode-follower action.

A Dekatron 'Add-On' Unit

By F. W. Lovick*

The circuit described in the following article is essentially a form of Dekatron Scaling Unit designed to 'add-on' to an existing valve scaler, so that the latter can operate at speeds consistent with its input resolution time, while still retaining the ability to record on its internal mechanical register. The unit obtains its power supplies from the valve scaler.

THE A.E.R.E. scaling unit type 1009, which is widely used, has an input resolution time of about 1·5μsec, but it is limited in counting rate because the mechanical register cannot operate more than approximately five times per second. This difficulty is usually overcome, when high counting rates are required, by using two or more scaling units in series. The resulting system is satisfactory in performance, but is both bulky and expensive; the unit described in the present note was designed to provide for large counting rates with greater economy.

General Description

An 'add-on' unit consisting of a Dekatron scale of 1 000, followed by a multivibrator stage operating a high speed relay, is connected between the electronic scaling circuits of the 1009 scaler and its mechanical register. The unit has been designed to fit into the front portion of the cooling unit type 1172A which is mounted on top of the scaling unit 1009. A scale of 10^5 with an input resolution of 1·5μsec and mechanical register storage is then available. Power for the 'add-on' circuit is taken from the existing 'probe power supply' socket at the rear of the 1009 scaling unit.

Circuit Description

The input to the Dekatron scaler is taken from the 'waveform out' socket of the 1009 scaling unit, where a positive pulse of about 20V amplitude and of 20 to 25μsec duration is produced for every 100 counts in the 1009 scaler. This positive pulse is applied to the grid of a pentode valve V_1 (CV138 (EF91)), which is biased almost to cut-off, producing in the anode circuit a large negative pulse of 150V amplitude.

This pulse is applied to the guide network of resistors of a single pulse Dekatron tube V_3 (GC10D)[1], where it causes the glow discharge to advance one step to the next cathode. This Dekatron counter tube is capable of operating at speeds up to 20 000 counts per second, but in this application it is not required to accept pulses spaced less than 100 × 1·5μsec apart. The diode valve V_2 (CV1092 (EA50)), is inserted to clamp the guides to their correct bias potential at the highest counting rates.

The output pulse from the first Dekatron stage is applied to the trigger electrode of a cold-cathode tetrode trigger tube V_4 (GTE175M), which is held at a positive potential of 140V with respect to its cathode by a potentiometer consisting of high stability carbon resistors. The gas in the trigger tube is partly ionized by the auxiliary cathode which is connected to 105V negative through a 10MΩ resistor to limit the current flowing. The positive output pulse of counter tube V_3 raises the potential of the trigger electrode, thus firing the tube, and the pulse produced in the anode circuit is phased by the following guide network and applied to the second Dekatron counter tube V_5 (GC10B)[2].

The output pulse from the second Dekatron stage is passed to a further cold cathode trigger tube V_6, connected in the same manner as the previous drive tube, and this operates the last Dekatron stage V_7. Dekatron tubes type GC10B are used in the last two stages, where the counting rates are comparatively low and well within the capabilities of these tubes.

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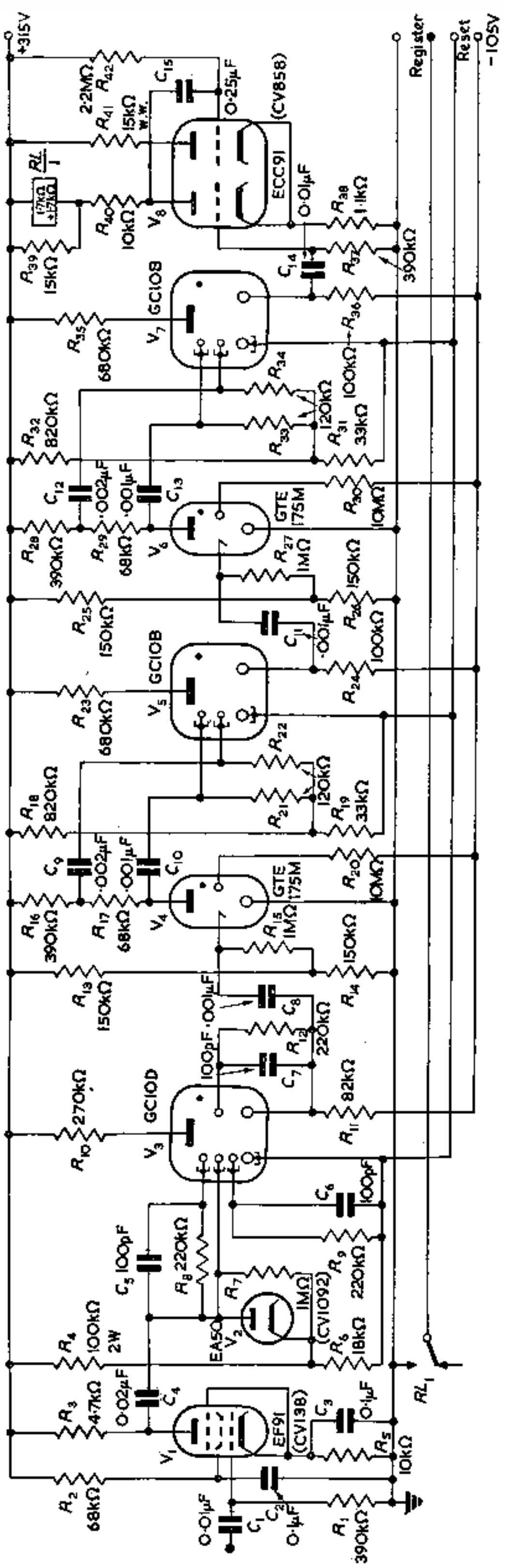

Fig. 1. Dekatron 'add-on' unit for scaling unit type 1009.

The final stage comprises a double triode valve V_8 (CV858 (ECC91)), connected as a multivibrator, and driven by the output pulse of V_7. The valve is normally held with the right-hand side conducting, and the left-hand side cut-off by reason of the cathode bias. The high speed relay forms part of the anode load of the left-hand triode. On receipt of a positive pulse from V_7, the circuit operates switching a current into the relay for about 80msec. The contacts of this relay when closed earth one side of the mechanical register in the 1009 scaling unit, thus causing it to operate in the normal manner from its own 50V supply.

In order to obtain a sufficiently high voltage to maintain the glow in the Dekatron counter tubes (the manufacturers suggest +475V) it was found desirable to operate these tubes between the h.t. +315V and the negative 105V lines of the scaling unit. Then if the common cathodes of the Dekatrons, together with their guide network potentiometers are connected to the 'reset' line of the scaler, as shown in Fig. 1, on resetting the electronic scale of 100 of the latter, the Dekatron 'add-on' scaler will be reset simultaneously.

Modifications to the 1009 Scaling Unit

The 'add-on' scaler has been designed to be operated with the minimum of alterations to the existing circuit of the A.E.R.E. scaling unit, and only the following alterations are necessary:

(1) Connect a lead from the 'reset' line to pin E of the 6 pin probe power supply socket (this pin is normally blank).

(2) Insert a socket at the rear of the scaling unit and connect a lead from this socket to the mechanical register terminal. (The other terminal of the register is already connected to the +50V supply).

(3) Switch the scaling unit to 'waveform out.'

(4) Mount the Dekatron scaling unit in the front portion of the cooling unit 1172A, inserting a 6-pin power supply socket at the rear, together with a socket for the lead to the register, and an input socket.

If it is desired to operate the scaling unit as a normal scale of 100 plus register, it is only necessary to remove the 6-pin power lead from the probe unit power supply socket, and return the front panel switch to the 'internal register' position.

Performance

A number of scaling units type 1009 with these 'add-on' units affixed have been used in this Department on experiments connected with the elastic scattering of 10MeV protons, and on a polarization direction correlation apparatus, where high counting rates were required. In the latter case, the combined unit was constantly checked for counting accuracy against an A.E.R.E. type 1009 scaling unit in series with an A.E.R.E. type 200 scaling unit, and the following results were obtained.

Duration of Count (Min.)	*Counts per Minute*	
	1009 *plus Dekatron unit*	1009 *plus* 200 *scaler*
12	237 638	237 586
25	249 162	248 996
42	255 694	255 996
50	233 669	233 406
60	243 563	243 698
90	220 765	220 833

The above table of results demonstrates that the accuracy of the combined scaler and 'add-on' unit is well within the limits of the dead-time correction, which in this case was 2 per cent.

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Data on Ferrite Core Materials

By A. C. Hudson* and E. J. Stevens*

Charts are presented which permit comparison of the permeability and loss at low levels for various ferrites, and some other core materials, over the frequency range 100kc/s to 100Mc/s. The loss is represented in terms of an inherent low-level 'Q' of the material, for convenience in design calculations. An example of a simple transformer design using the charts is given.

WHEN using ferrite core materials at low power levels the familiar concept of Q is useful.

The relative permeability of the ferrite is regarded as a complex quantity, where the imaginary component is present because of the core losses. Then Q is the ratio of μ'_s, the real (inductive) component, to μ''_s, the imaginary (lossy) component of complex permeability, where a series representation is used; or the ratio of the imaginary component, μ''_p, to the real component, μ'_p, where a parallel representation is used. It is also the Q of a resulting inductor or transformer wound on a closed core of this material, except to the extent that the Q is reduced by coil losses.

Q, and the real component of series or parallel permeability, are shown in terms of frequency in all of the accompanying figures. The ordinate is Q and the abscissa is permeability, while the values of frequency are marked on the curves.

The curves have been calculated from typical data on permeability and loss published by Mullard Ltd[1]. No doubt ferrites of other manufacture would follow similar trends.

In order to permit comparison with various metallic and dust cores, Figs. 5 and 6 include some data calculated from Figs. 5 to 8 of reference 2.

Ferrite cores are often used with a small air-gap. The effect of the gap is to raise Q and lower the apparent μ by the same factor. (In addition to the increase in Q, there are other beneficial effects.) Some typical properties, based on measurements made on a core having a gap, are shown in Fig. 7.

It is hoped that these charts will save time for component designers, and will also provide a composite picture of the low-level properties of ferrites.

As an example of the use of the charts, the design of a resonant transformer to match the output stage of a radar intermediate-frequency pre-amplifier to a 50Ω terminated cable will be considered. The output stage consists of a pentode with total capacitance, including strays, of 10pF. The bandwidth of the circuit and the centre frequency are both to be 30Mc/s.

Assuming an ideal transformer, elementary theory indicates that the 50Ω must be transformed to 530Ω, to meet the above specifications; hence the ratio of the transformer is 3·26. The Q of the circuit is about unity.

The Q of the transformer should be just high enough so that core losses contribute negligibly to the total damping, so with a circuit Q of 1, it can be decided that a core Q

* *National Research Council, Canada.*

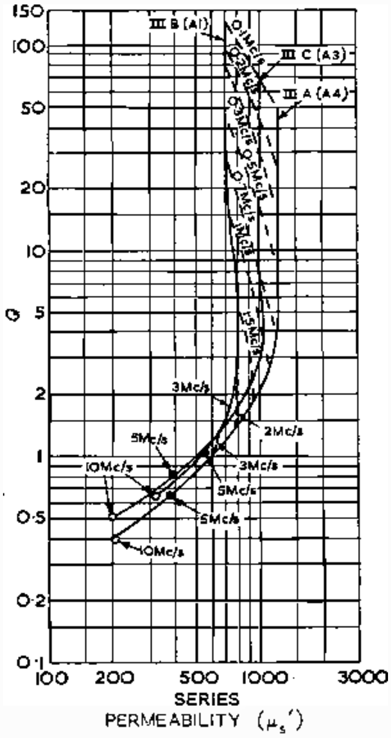

Fig. 1. Manganese-zinc ferrite series representation

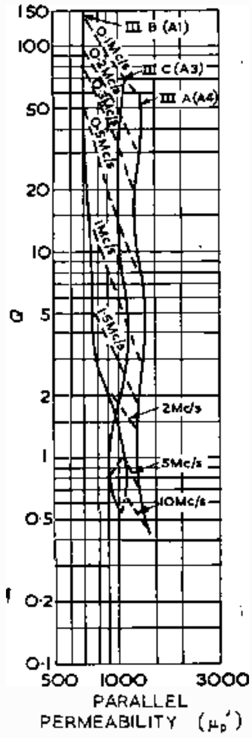

Fig. 2. Manganese-zinc ferrite parallel representation

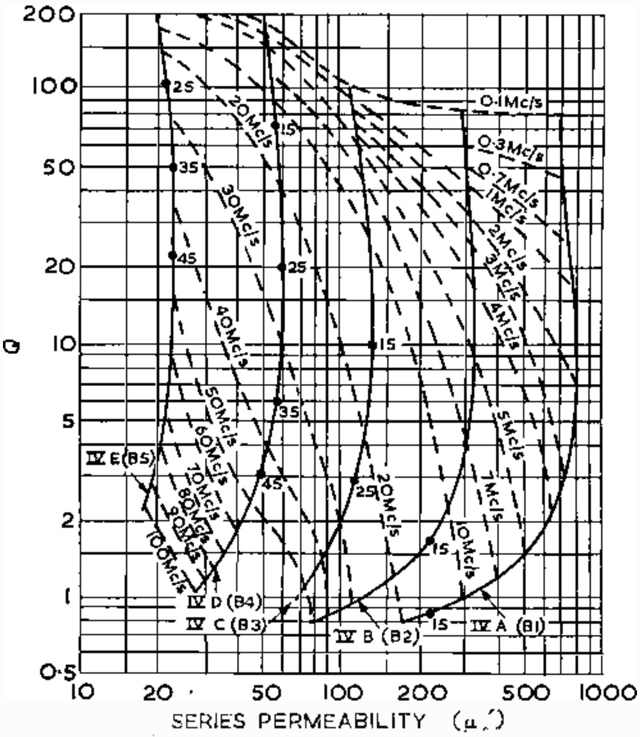

Fig. 3. Nickel-zinc ferrite series representation

of 10 is adequate. Referring now to Fig. 4, it is seen that material IV D with no gap has a Q of 10 at 30Mc/s, and a permeability of 70.

Knowing the permeability, and the geometry of the core, the number of turns required may be calculated by conventional methods.

This transformer was constructed, using a one-inch diameter pot core. Its performance, using either copper tape or 16 s.w.g. wire, was found to be equivalent, within

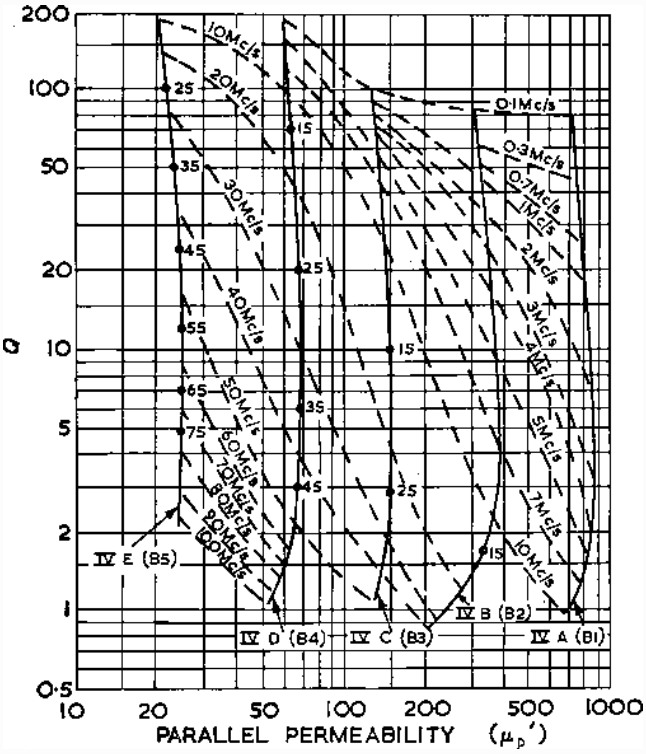

Fig. 4. Nickel-zinc ferrite parallel representation

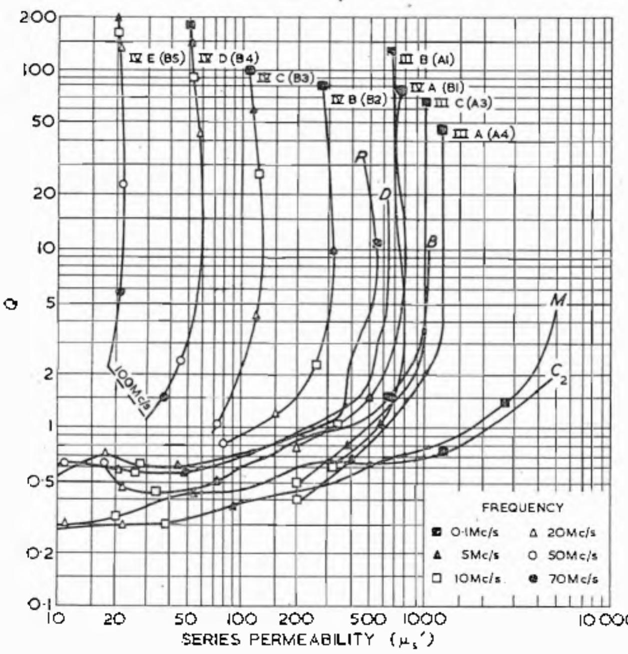

Fig. 5. Comparison of ferrite and other materials series representation
Materials Ferrites (as shown)
B — ·002*in Permalloy B* C_2 — ·002*in Permalloy C*
D — ·002*in Permalloy D* *M* — ·0015*in Mumetal*
R — ·002*in Rhometal*

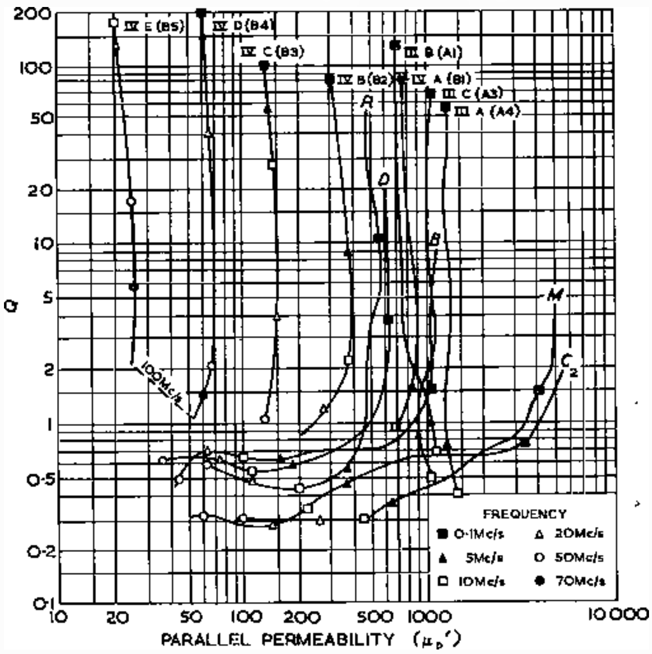

Fig. 6. Comparison of ferrite and other materials parallel representation
Materials Ferrites (as shown)
B — ·002*in Permalloy B* C_2 — ·002*in Permalloy C*
D — ·002*in Permalloy D* *M* — ·0015*in Mumetal*
R — ·002*in Rhometal*

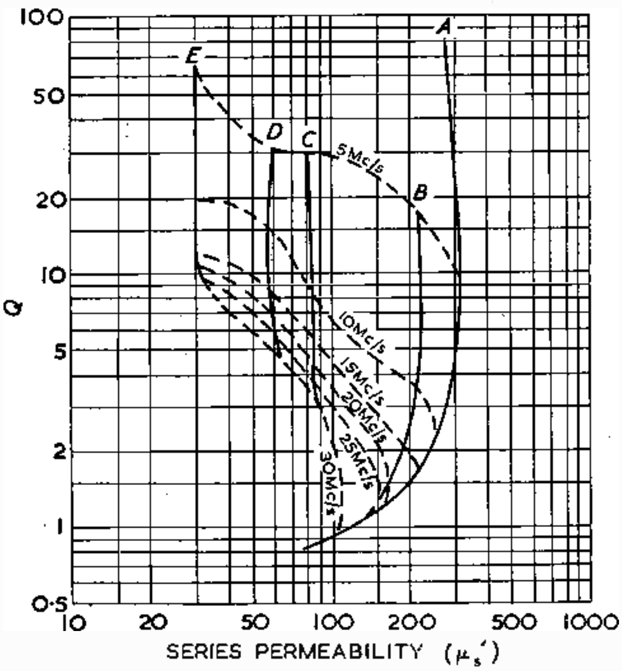

Fig. 7. Nickel-zinc ferrite (IVB(B2)) with various air gaps series representation

CURVE	GAP (PERCENT OF TOTAL FLUX PATH LENGTH)
A Published	0
B Measured	Residual gap in pot core
C Measured	0·5
D Measured	1·0
E Measured	3·5

the errors of measurement, to that of an ideal transformer.

Reference to Fig. 7 shows that IV B material will also have an adequate Q at 30Mc/s if a gap of 3·5 per cent of the total path length is used. In this case the effective permeability is seen to be 30. A transformer using these constants was built, using a ½in diameter pot core. The performance of this transformer was reasonably close to that of an ideal transformer. Strictly, parallel representation should be used, but for values of Q of 10 or higher there is negligible difference between the series and parallel permeability.

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A Three-band Aerial Combining Network

By S. L. Fife*, A.M.Brit.I.R.E.

The facility of incorporating radio-frequency channels for reception of the BBC frequency-modulated broadcasts in a multi-channel television receiver, generally requires the use of three aerials for Band 1 (BBC television), Band 2 (BBC frequency-modulated broadcasts), Band 3 (commercial television). This article describes a network utilizing three prototype filter sections of low-pass, high-pass and band-pass characteristics which are connected together to a common selective load. The filter sections have related cut-off frequencies to obtain a reciprocating 'm' half section termination for the particular operating section. The network provides for feeders from the three aerials to be combined at the masthead, and preserves the matching condition with negligible insertion loss.

The effect of off-band reactive aerials, as the source impedance of the filter sections, is examined in relation to the feeder system as an impedance transformer.

THE intermediate-frequency sound channel of a television receiver is ideally suited to an essential requirement for frequency modulation reception[1], i.e., zero phase shift of those frequencies contained in the frequency-modulated transmission. The bandwidth requirement for the BBC system[2] is approximately 180kc/s.

The television sound i.f. channel when operated within the B.R.E.M.A. recommended band of 34Mc/s to 38Mc/s provides this bandwidth in approximating the characteristic of 300kc/s at −3dB.

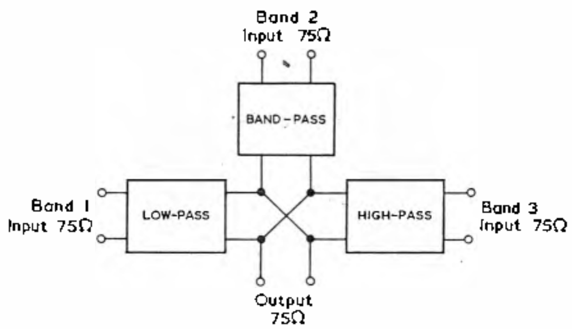

Fig. 1. Three band aerial combining unit

By suitable pre-mixer and detector switching, the additional facility of f.m. is incorporated in the television receiver which is then identified as a three band receiver, i.e., Band I, 41Mc/s to 68Mc/s (BBC television), Band 2, 87·5Mc/s to 100Mc/s (BBC f.m.), Band 3, 174Mc/s to 216Mc/s (commercial television). Generally the receiver will require an aerial installation for each band, which would involve bringing three pairs of feeders to the receiver and switching these as each band is selected[3]. The author's solution was in the design of the network shown in Fig. 1, which enabled the three aerials to be combined by short lengths of balanced feeder and a single balanced feeder brought to the receiver.

Use of Reciprocate m Half Section Termination

The initial design of the network was by conventional filter approach with due regard to the frequency spread of each band. The ideal frequencies characteristic of the network is shown in Fig. 2, where the cut-off frequency of the low-pass Band 1 filter section coincides with the lower cut-off frequency of the band-pass Band 2 filter section. Also the upper cut-off frequency of the Band 2 filter coincides with the cut-off frequency of the high-pass Band 3 filter section. By choosing this relationship of the cut-off frequencies, each filter section is terminated by the other two sections which combine to effect an m-derived half section termination. The last series reactance of each filter section is therefore modified to incorporate the prototype T section series component and the series component of the terminating half section. The prototype T sections and the m-derived half section terminations are shown in Figs. 3 and 4.

The function of the cut-off frequencies relationship is demonstrated by examining the termination of the prototype band-pass section of Fig. 3(a). The shunt arms of the terminating half section are series resonant at the edge of the pass-band, i.e. just below 80Mc/s and just above

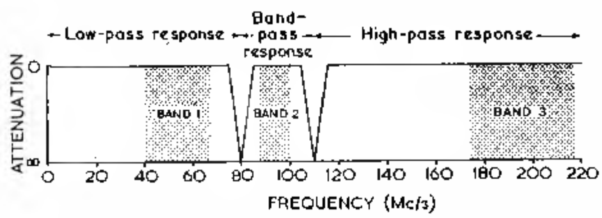

Fig. 2. Ideal frequency/attenuation characteristic

110Mc/s depending upon the m factor. In the design, $m = 0{\cdot}6$ was adopted to obtain the uniform resistive characteristic in the pass-band. These two series resonant shunt arms are provided by the low-pass and high-pass filter sections respectively, since the characteristic resistance variation with frequency falls to zero in the low-pass filter at 80Mc/s and in the high-pass filter at 110Mc/s as shown in Fig. 3(b). The reactive components in the series arm of the half-section termination are absorbed by the modification of the reactive components in the last series arm of the prototype section.

Similarly the low-pass filter m half section termination and the high-pass m half section termination have series resonance shunts at 80Mc/s and 110Mc/s respectively. The band-pass filter section provides for these cut-off points as shown in Fig. 4(b).

Reactive Off-Band Aerials

The characteristic impedance of each filter section was assumed to be 75Ω in the design of the prototype T-sections but under operating conditions the input impedance source will vary according to the aerial reactance characteristic given by:—

$$-jZ_r \cot BL$$

* *Formerly English Electric Co. Ltd.*

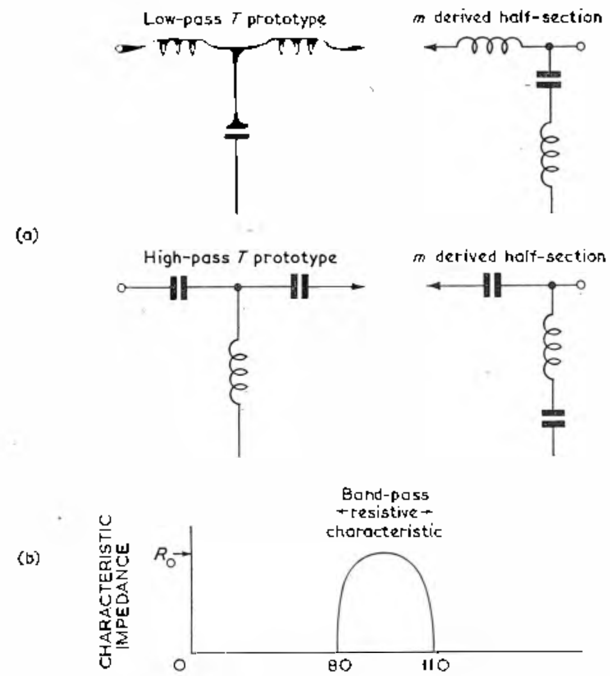

Fig. 3. (a) Band-pass prototype section terminated by an m half-section with shunt arms resonant at 80Mc/s and 110Mc/s
(b) Low-pass and high-pass resistive characteristics simulating the action of the half-section shunt arms at the cut-off frequencies . . .

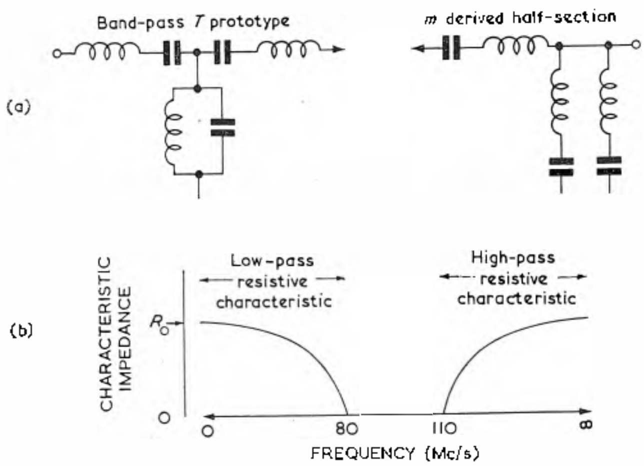

Fig. 4. (a) Low-pass and high-pass prototype sections terminated by their m half-sections with shunt arms resonant at 80Mc/s and 110Mc/s respectively
(b) Band-pass resistive characteristic simulating the action of the half-section shunt arms at the cut-off frequencies

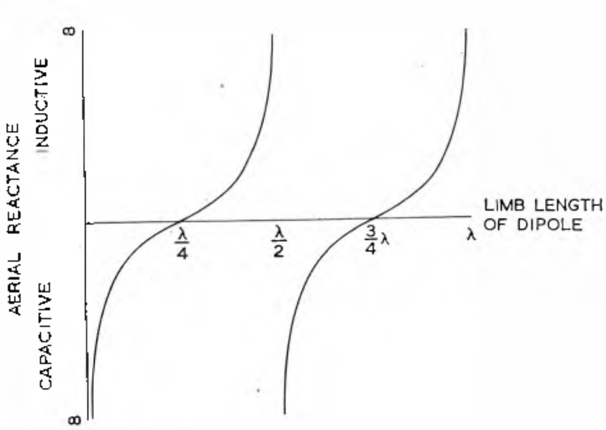

Fig. 5. Reactive characteristic of aerial

where Z_r = radiation resistance of the $\lambda/2$ dipole

L = limb length of the dipole

B = phase constant of feeder ($2\pi/\lambda$).

This is shown in Fig. 5 where the half-wave dipole is represented as an open-ended transmission line and exhibits reactive behaviour with frequency in that it may be capacitive or inductive under non-resonant conditions and approximates a resistive 75Ω source at resonance.

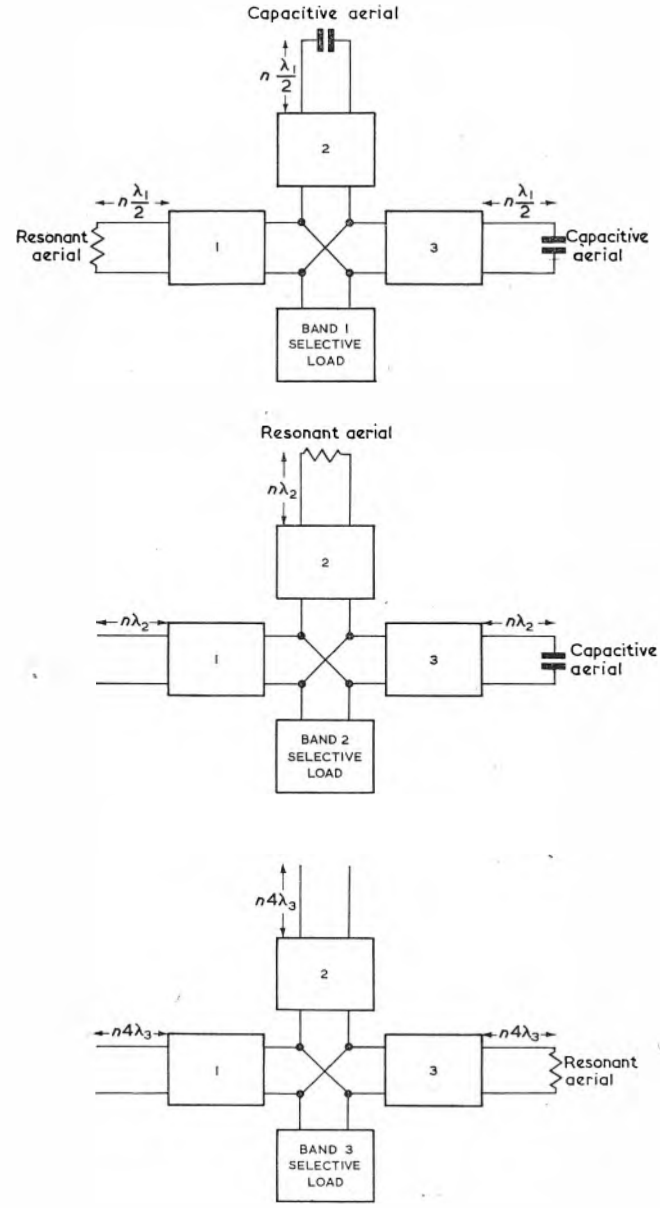

Fig. 6. Three-band working conditions, showing aerial source impedance and effective line length assuming $\lambda_1 : \lambda_2 : \lambda_3 : 4 : 2 : 1$

The effect of a reactive off-band aerial assumes importance when the feeder connecting the particular aerial to the network functions as an impedance transformer. The impedance transformation ratio when assuming negligible feeder attenuation is given by:

$$Z_s/Z_t = \frac{Z_t/Z_o + j\tan Bl}{Z_t/Z_o(1 + Z_t/Z_o\, j\tan Bl)}$$

where Z_s = source impedance

Z_o = characteristic impedance of feeder

Z_t = terminated end impedance

l = length of feeder

B = phase constant of feeder ($2\pi^2/\lambda$).

It is seen that when $l = n(\lambda/2)$ then $Z_s/Z_t = 1$, i.e. a single half-wave length of feeder effects an impedance transformation of 1:1. When $l = (2n+1)\lambda/4$ then $Z_s/Z_t = (Z_0/Z_t)^2$, i.e. an odd multiple quarter-wave length of feeder effects an impedance transformation such that $Z_s.Z_t = Z_0^2$.

The ideal installation should therefore have feeder lengths between the aerial and network of single or multiple half-wave lengths. A near approach to this ideal is achieved as shown in Fig. 6 which assumes the Band 2 carrier frequency to be $2f_1$ and the Band 3 carrier frequency to be $4f_1$, where f_1 is the band 1 carrier frequency. The reactive nature of the aerials for each band is shown for three band working.

Experimental Results

The connexion of the three filter sections into a complete combining network produced a result approximating

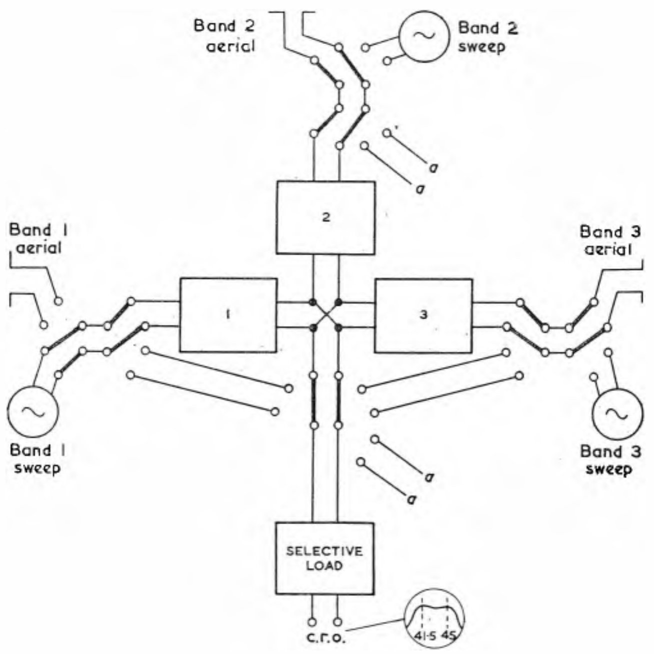

Fig. 7. Test circuit used to obtain optimum result from combined sections

the three frequency/attenuation characteristics required, but component values were then adjusted in each filter section by observing a selective response at the extremes of each of the three bands. By this means the two response curves checked in each band were made to be identical with or without the presence of the combining network. The method is shown in Fig. 7.

The fine adjustments necessary to achieve this condition were laborious, since the continuity of the feeder system was not interrupted by switching as shown diagrammatically in Fig. 7. This led to an optimum result as shown in the circuit diagram[4] of Fig. 8. The preservation of the selective load response indicated that the matching condition was identical with that when the feeder only was used for connexion. The frequency/attenuation characteristic is shown in Fig. 9. The modifications necessary to the prototype sections when connected together as a combining unit arises through the variations with frequency of the source impedance of each filter section and the reactive nature of the two off-band filters terminating the operating filter section. The desired result of combining three aerials to a common load and preserving the selective response, standing wave characteristic and zero insertion loss as with a single input is attained by experiment.

No 'ghosting' effect due to phase changes through the multiple signal path from the three aerials to the receiver has been observed. This is avoided by the small signal pick-up of the two non-resonant aerials and the severe attenuation in the two off-band signal paths. The 1·5MΩ resistors provide for aerial static discharge to conform with the requirements of a British Standards safety specification[5].

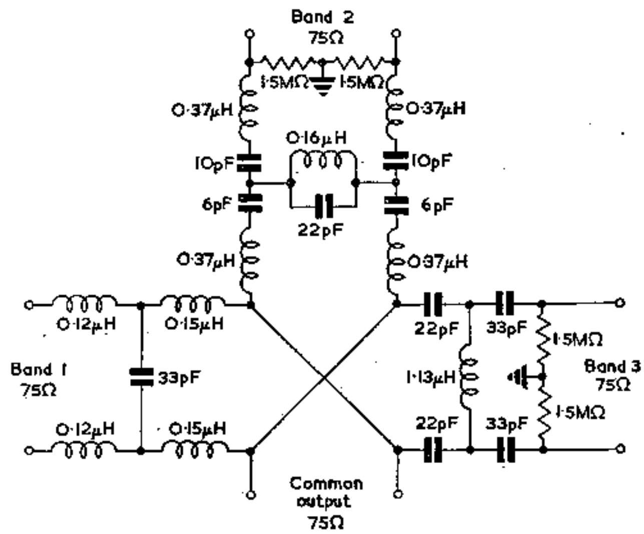

Fig. 8. The combining network

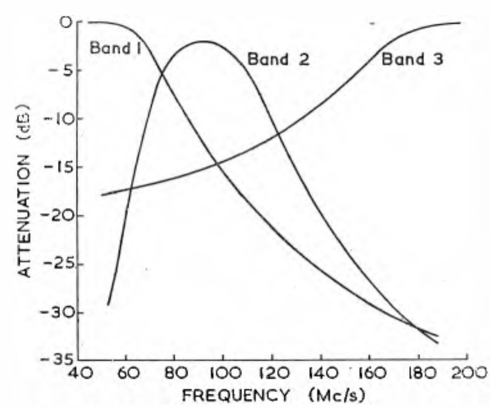

Fig. 9. Frequency/attenuation characteristics of network

Use of Printed Circuit Techniques for Production of Unit

The network was first made up with lumped components, it was then produced by printed circuit techniques which necessitated slight component value changes due to differences in self-capacitance of the coils and stray couplings.

Acknowledgments

The author is pleased to acknowledge the critical comments of Mr. R. A. Spears in preparing the final draft of the paper, and to Mr. J. T. Sparkes for his co-operation in executing the printed circuit techniques.

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Short News Items

Premiums for Technical Writing. Articles on electronics published in 1958 which are eligible for premiums awarded annually by the Radio Industry Council will be considered by the panel of judges early in the New Year. Six premiums of 25 guineas each are offered to any writer not earning his living wholly or mainly by writing. The judges, headed by Professor H. E. M. Barlow, Professor of Electrical Engineering, University College, London, are given the greatest possible freedom in deciding which articles merit awards, but they principally take into consideration the value of the article in making known British achievement in radio and electronics, originality of subject, technical interest, presentation and clarity. Writers are reminded that articles for consideration should be in the hands of the Secretary of the Radio Industry Council, 59 Russell Square, London, W.C.1, before the end of the year. Six copies of the journal or relevant pages, proofs or reprints with a signed declaration of the eligibility of the writer or writers should be submitted. This latter ruling does not apply to writers of articles published in the following journals as the members of the panel of judges automatically receive copies of these: *Wireless World; Electronic and Radio Engineer; Electronic Engineering; British Communications and Electronics; The Post Office Electrical Engineers' Journal; Radio Constructor.*

The Council of the British Institution of Radio Engineers, and the American Institute of Electrical Engineers have both advised members in Great Britain that under Section 16 of the Finance Act, 1958, the whole of the annual subscription paid by a member who qualifies for relief under that Section will be allowable as a reduction from his emoluments assessable to income tax under Schedule E. Particular attention is drawn to the fact that applications will only be accepted provided that the subscription is defrayed out of the emoluments of the office or employment. It is a further condition that membership should be relevant to the office or employment of the claimant.

IBM United Kingdom Ltd announces that the Government has authorized the Royal Army Pay Corps to install an IBM 705 computer, which will be supplied by IBM United Kingdom Ltd, and delivered in January 1961. It will be used to keep the central pay records for all soldiers of the Regular and Reserve Army. Among other jobs it will produce, for every man, the quarterly statement of his account, and the notification to the soldier of any adjustments to his pay, as they occur. Existing paying methods in the unit will be retained, but the speed of processing information at the central accounting unit will be increased. More accurate and efficient accounts will result in less correspondence, and, at the same time, the financial savings made will pay for the system in the first few years. The record of every man will be kept on magnetic tape which can be scanned, read or written at a speed of 900 000 characters a minute. Each tape reel holds over 5 million characters of information, and up to 100 of these can be used at any one time with such a computer.

Irwin Technical Ltd of Ormond House, 9 Ormond Close, London, W.C.1, give a comprehensive service for the production of instructional material such as manuals and wall charts and a full range of sales material from simple data sheets and leaflets to brochures and catalogues. The staff includes electronic and mechanical authors, technical illustrators, circuit draughtsmen and designers. Prirnted publications can be made to conform to any official specification or established style. Irwin Technical Ltd are contractors to a number of Government Departments and industrial companies.

The Plessey Co Ltd has made an award of 100 Guineas to an employee, Mr. F. J. Hancock. He has evolved a new method of packaging television and radio sets by which many of the expensive pieces of corrugated paper previously used can be eliminated. The new method has passed stringent tests by the company and has been approved by the Packaging and Allied Trades Research Association. It is the largest award so far made by Plessey for a suggestion put forward by an employee for consideration.

A British telecommunication firm, Standard Telephones and Cables Ltd, has been awarded four prizes for exhibits on its stand in the British Industries Pavilion at the Brussels Universal and International Exhibition. The Grand Prix was awarded in the "Transmission by Wire" class, in which the principal STC exhibit was a submerged two-way telephone repeater similar to those supplied by the company for the Newfoundland-Nova Scotia section of the first Transatlantic submarine telephone cable. More than 100 of these repeaters will be used in a new Transatlantic submarine telephone cable system which is scheduled for completion in 1961. This United Kingdom-Canada cable will in turn provide the first link of the new round-the-world Commonwealth cable project announced at the recent Commonwealth Trade and Economic Conference in Montreal.

The Hollerith Engineering Training School at Letchworth was officially opened recently by the Hon. David Bowes Lyon, Lord Lieutenant of Hertfordshire. The school is intended to serve all the engineering activities of the company, and includes in its courses all categories of engineers in the company's employment. It does not compete with technical colleges, but works in close collaboration with them. The opening was attended by about 150 guests, including representatives of many other local industries, heads of educational establishments in the area and the Chairman and representatives of Local Government authorities and Educational Committees.

Members of the British Plastics Federation were reminded by the Chairman at their recent Conference at Torquay of one of the most valuable activities of the Federation—its service of abstracts from journals, covering practically the whole world, and patents, of interest to all concerned in the manufacture or use of plastics. These deal with materials, processing methods, machinery and applications. The most important journals from which the abstracts are taken are available in the Federation library, 47-48 Piccadilly, London, W.1, to all subscribers and copies of articles can be supplied intact in the original journal, or in photostat or microfilm form.

The Independent Television Authority has placed a substantial contract with E.M.I. Electronics Ltd. for what, when completed, it is stated will be the tallest television aerial mast in the Commonwealth. The 1 000ft mast will be situated at Mendlesham, Suffolk, and will carry ITV's East Anglia programme from the studios of Anglia Television in Norwich. It will serve nearly two million people when the new station opens towards the end of next year. The order includes the supply and erection of the mast, special directional aerial system and feeders from the transmitter.

The office of the German company of the Solartron Electronic Group Ltd, Thames Ditton, Surrey—Solartron, G.m.b.H., is now at Bayerstrasse 13, Munchen, Germany.

Erratum. In the letter from R. S. Taunton on page 668 of the November issue, second column, line 18, should read ". . . . . any further increase in e, R_l is effectively . . . ."

LETTERS TO THE EDITOR

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

A Crystal Controlled Induction Heater

Dear Sir,—Mr. Cohen, in his article in the April issue, states on page 177 that the requirements of the Atlantic City Charter stipulate three permitted working frequencies of 13·66, 27·32 and 40·98Mc/s respectively, all with allowable bandwidths of 0·05 per cent.

My copy of the International Radio Regulations, Atlantic City, 1947, gives these frequencies as 13·56Mc/s, ± 0·05 per cent, 27·12Mc/s ± 0·6 per cent and 40·68Mc/s ± 0·05 per cent.

Being engaged on a somewhat similar project to the one described by Mr. Cohen, I am interested to know if there has been any recent change in these frequencies and bandwidths. I am certainly not aware of any.

Yours faithfully,

J. L. Ashford
Orpington, Kent.

The Author replies:

Dear Sir,—I have referred back to my references and find that the three frequency bands I have quoted were obtained from an F.C.C. Allocation Plan published in *Electronics,* March 1954. The frequency bands quoted by Mr. Ashford are the revised frequencies Atlantic City, 1947. This is borne out by a paper by Gieringer in *Electronics,* December 1948, entitled "The Frequency Stability of Diathermy Units". However, Hertwig, "Automatic Tuning Control of H.F. Generators with Varying Load", *Electron Appl. Bull.* 13 (1952), states that the lower frequency band of the Atlantic City agreement is 13·64Mc/s ± 0·05 per cent instead of Mr. Ashford's 13·56Mc/s ± 0·05 per cent.

I am most obliged to Mr. Ashford for drawing my attention to the matter.

Yours faithfully,

E. Cohen
Research Department,
Metropolitan-Vickers Electrical Co. Ltd.

Amendment No. 1, Published 20 December 1957, to Code of Practise CO1002 (1947) confirms the frequency allocations and bandwidths mentioned in Mr. Ashford's letter excepting that, in the U.K., "40·68Mc/s has been held unsuitable for general I.S.M. use because of its close proximity to Television Band 1 (41-68Mc/s)."—Editor.

Testing the Linearity of Modulators and Demodulators in Multi-channel F.M. Transmitters and Receivers

Dear Sir,—In connexion with the recently published article by G. C. Davey (August 1958), I would like to point out that the application of swept signal additionally modulated in frequency at a higher rate to the exploration of non-linearities in f.m. links is by no means restricted to the application mentioned by Mr. Davey in his article.

In this connexion, I would like to make reference to a paper by A. M. Reslock and the writer[1] where we pointed out that f.m. distortion caused by phase curve non-linearity can be quantitatively investigated in this manner in band-pass amplifiers.

Similarly, it can be proved that by addition of a second harmonic selector to the block diagram of Fig. 6 of the discussed article, direct display of the discriminator's curvature (or, in another scale, second harmonic distortion) can be obtained directly for numerical evaluation[2].

The technique described by Mr. Davey forms a logical experimental tool in exploring properties of networks, the data of which can be computationally obtained by the method of adjustment of transfer function derivatives[3,4], essentially by the application of Taylor series principle to the exploration of non-linearities of a small order of magnitude[5].

A similar technique can also be used for the quantitative evaluation of small non-linearities of characteristics of thermionic valves and transistors—for example, for the purpose of determination of intermodulation terms which are likely to arise[6].

It has always been my opinion that circuit design and tests by evaluation of derivatives of gain and phase transfer functions offers considerable and not yet fully realized possibilities and I am pleased to see yet another embodiment of this powerful method.

Yours faithfully,

Julius J. Hupert
Associate Professor of Physics,
De Paul University,
Chicago.

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The Author replies:

Dear Sir,—I would like to thank Dr. Hupert for his observations about my article. His references give a very comprehensive discussion of the possibilities of the technique involved, whereas my discussion was of a specific application now available as a commercially manufactured instrument.

Yours faithfully,

G. C. Davey
Marconi Instruments Ltd.
St. Albans.

A Multivibrator-Controlled Sinusoidal Oscillator

Dear Sir,—In this letter a circuit is described in which a multivibrator is made to control a sinusoidal oscillator. The frequency is dependent only on the frequency control element, and is virtually independent of temperature and voltage. Frequency control may be

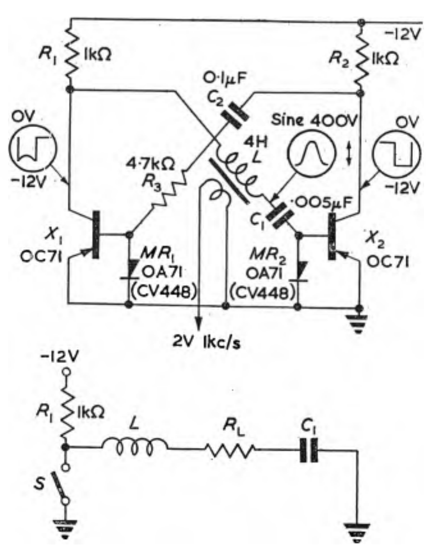


Fig. 1. Simplified equivalent circuit

derived from a series-mode operated crystal, or a series LC circuit. In the latter case the resonant coil can develop pure sine wave oscillations with a peak-to-peak amplitude in the order of 500V, using a 12V power supply. Power consumption is 12mA at 12V. The circuit works satisfactorily down to about 1V, the frequency remaining constant, amplitude being proportional to supply voltage.

In normal electronic techniques the requirement for a sinusoidal oscillator is met by providing an amplifier with a gain of $|\mu|$ and phase-angle Θ, using feedback with an attenuation factor $|\beta|$ and phase-angle ϕ. At the required frequency of oscillation it is arranged that $|\mu| = |\beta|$ and $\Theta = -\phi$. Variations in $|\mu|$ and Θ result in changes in the oscillator frequency. In the case of transistor amplifiers, special circuit precautions are necessary to off-set the effects of changes in temperature and voltage.

In the oscillator circuit to be described, the transistors are operated in an on/off condition. The oscillator frequency is thus largely independent of temperature and voltage. At 1kc/s the frequency decreased by 0·1 per cent/volt in the range 1 to 12V. A temperature change from 15°C to about 60°C produced a decrease in frequency $\sim$ 0·5 per cent.

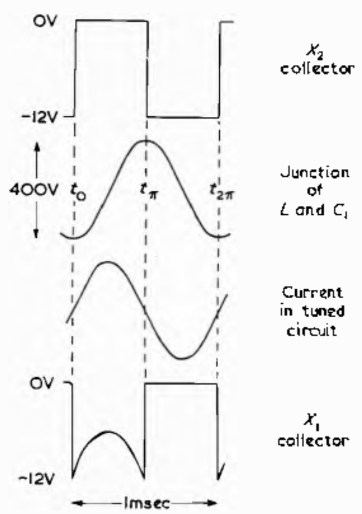


Fig. 2. Waveforms for circuit of Fig. 1

The oscillator circuit is shown in Fig. 1. The frequency controlling element LC is connected as a normal multivibrator cross coupling. The coil L is a Ferroxcube core type L.A.7 with 2500 turns of 42 s.w.g. enamelled wire, inductance $\simeq$ 4H. The secondary is 50 turns, giving an output of about 2V at an output impedance of 250Ω, at 1kc/s and with 12V supply.

When switched on the circuit quickly settles down to multivibrator operation, with waveforms as shown in Fig. 2. It will be seen from the phasing of the collector square waveform and the tuned circuit sine wave between L and C_1, that the transistor X_2 is bottomed when the current in the tuned circuit is negative-going. When the direction of the current reverses at the peak of the sine wave, X_2 is shut off, the negative square wave keeping X_1 bottomed for the period of the negative-going half cycle. At t_0 the tuned circuit current is positive-going into the base of X_2, which is cut-off, causing the collector to rise to -12V. The positive current flows into the diode MR_3 so that the C_1 end of the tuned circuit never moves far from earth. At t_π the current in the tuned circuit reverses, switching X_2 on and therefore X_1 off.

The circuit starts very readily and shows no signs of voltage transients at the transistor collectors. The current consumption is $E/2R_1 + E/2R_2$ where E is the applied voltage and R_1 and R_2 the collector loads: this simplifies to E/R_1 when $R_1 = R_2$. The collector waveforms are symmetrical square waves, the edges of the X_2 waveform being improved by the resistor R_3.

It is possible to use a series-mode crystal as the frequency controlling element, and in this case resistor R_1 was replaced by a choke of about 5H, and R_3 shorted out. Crystals tried in the experimental circuit were

5" XY Flexural at 1·6 and 2·5kc/s (Marconi)

Type D.J.C. at 2·5kc/s (G.E.C.)

A simplified equivalent circuit of the oscillator is shown in Fig. 1(b). It is shown as a simple series-tuned coil and capacitor, with series resistance R_L. The transistor X_1 operates as switch S, which connects the 1kΩ resistance to ground once per half cycle of the resonance frequency. The total resistance R in the tuned circuit is the sum of all the series resistances, $R_1 + R_L$.

The equation for the response of a series-tuned circuit to a square wave of voltage E, with zero initial conditions is:—

$$i = (E/nL)e^{-ut} \sin nt \quad \ldots\ldots \quad (1)$$

where $n = (1/LC - R^2/4L^2)^{\frac{1}{2}}$ ...... (2)

and $u = R/2L$ .... (3)

For a 4H coil, and $R = 2$kΩ, at 1kc/s equation (1) suggests that long term stabilities of the order of 0·3 per cent might be expected, and were in fact indicated by experimental results.

Although a complete analytical treatment of the multivibrator controlled oscillator has yet to be done, preliminary work has shown it to be an extremely stable and versatile oscillator of rugged construction and good frequency stability.

The author is indebted to U.K.A.E.A., Weapons Group for permission to publish this letter.

Yours faithfully,

D. D. McLeod
A.W.R.E.,
Aldermaston.

A Simple Pressure Operated Controller

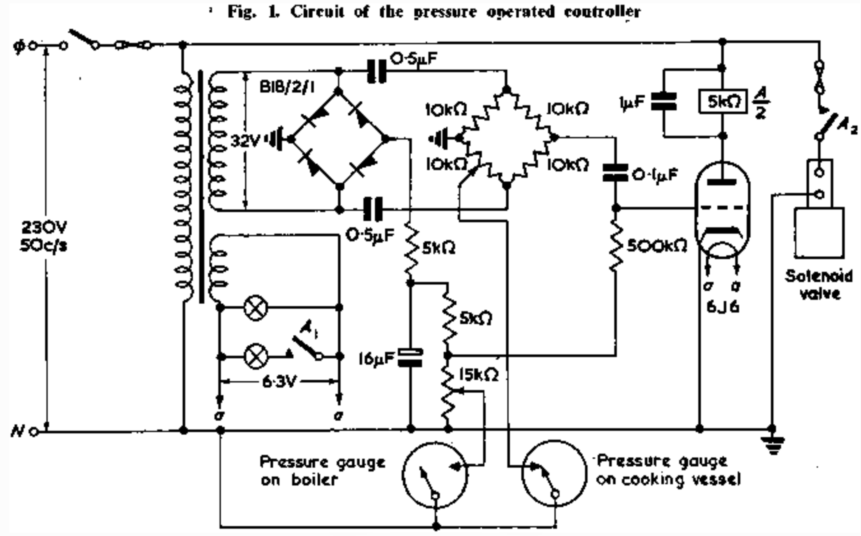
Dear Sir,—Fig. 1 shows a simple circuit which I have just developed. The problem was to start the cycle of a fully automatic cooking application. Before the cycle can start, steam pressure in a small high pressure boiler must be brought up to the desired value and the pressure within the cooking vessel must be down to zero. The obvious solution would have been to use two pressure switches, but these were not available in this country for the high pressure involved.

The solution was to use two pressure gauges which are cheap and readily available. One is fitted with an adjustable insulated wiping contact which is contacted by the pointer when the steam pressure is up to the required value. The other is fitted with a contact at the zero point on the gauge. The pointers on each pressure gauge are of course at earth potential.

As will be seen from the Fig. 1, when contact is made on one pressure gauge, the negative bias on the 6J5 is reduced and contact on the other gauge throws the a.c. bridge out of balance and applies a signal to the grid of the 6J5 approximately in phase with the anode supply. Under these conditions the anode current is sufficient to operate the pattern 3000 relay. With either pressure gauge contact broken anode current falls to below that at which the relay drops out, so the requirements of the design are fully met.

H. Crownshaw
M.I.Plant E.,
Dominion Breweries Ltd.,
Otahuhu,
New Zealand.

Fig. 1. Circuit of the pressure operated controller



BOOK REVIEWS

Foundations of Information Theory

By Amiel Feinstein. 137 pp. Demy 8vo. McGraw Hill Publishing Co. Ltd. 1958. Price 50s. 6d.

IT is an historical accident that Information Theory (or what many of us in Britain, with Shannon, call 'Communication Theory') was originally conceived within the fields of crytography and telecommunication. Much of the later work has been confined within the same disciplines—though more in the U.S.A. than in Britain and Europe. Here we have been much concerned with its relevance to physics, biology, linguistics, observation theory, in a wider form, as we know it under the title 'Information Theory'; most generally as a part of scientific method.

There is certainly a need at the present time for such a volume as this, a strictly mathematical text, concerned with mathematical definitions, with theorems and their rigorous proofs, and lemmas. Such a complete abstraction from all fields of interpretation is very laudable, but might have gone the whole way by avoiding the jargon of a particular science, telecommunication with its *messages, codes, signals, channels.* The book is for mathematicians, and a considerable mathematical knowledge is assumed, including that of measure theory.

The material is well ordered. After introducing the principal concepts the author studies the properties of H, the "entropic" measure of information-rate, in its various conditional forms. He then considers the simplest case of communicating with a finite set of symbols, which become converted to another finite set in the receiver, being perturbed by the noise according to some probability distribution. The decision processes, whereby perturbed signals are interpreted as unique symbols (with consequent loss of information) are considered inherent in such simple channels, but are, later, themselves studied. Again, the uses of stores ("memories"), question of coding, proof of the Capacity Theorem in various forms and other matters are taken up in very logical order and examined in rigorous, formal, mathematical terms.

There are two special points of criticism. First, the Index lists no more than 20 items. Second, the Bibliography is wholly inadequate; no mention will be found of the name of Hartley, nor of Gabor, Woodward, MacKay, Fisher, Barnard, and many others on this side of the Atlantic. The book is certainly not for beginners (its title does not suggest this) but rather for those already familiar with the main background and philosophy, who want the formal mathematics collected within a single volume.

COLIN CHERRY

Transistor Technology

Vol. I. Edited by H. E. Bridgers, J. H. Scaff and J. N. Shrive. 661 pp. Royal 8vo. D. Van Nostrand Company Ltd., New York, Toronto, London. 1958. Price 131s. 6d.

THIS book is the first of a series of three volumes which deal with the very extensive subject of transistor technology. The basic framework of the book is based on the record of a Symposium held by the Bell Laboratories in 1952 but brought up to date by the addition of new knowledge.

Parts I and II, which occupy 164 pages, deal with the germanium technology including purification, reclamation of scrap and the growing of single crystals, and is an excellent treatise on the subject.

Part III, occupying 279 pages, deals with the principles of device fabrication. A good descriptive summary of transistor design theory is given and the techniques used in the fabrication of point contact and grown junction transistors are described in considerable detail. It is a pity, however, that alloy techniques are not described since the majority of germanium junction transistors manufactured today are made by the latter technique.

Part IV occupying 153 pages deals exhaustively with the transistor characteristics, small signal, large signal and pulse parameters and the methods of measuring these parameters.

Part V of 13 pages deals briefly with transistor reliability.

Some of the subject matter of the book, in particular those parts concerned with point contact transistors, is mainly of historical interest, since very few such transistors are now being manufactured. However, the bulk of the book is concerned with basic principles and scientists and engineers in industry, as well as advanced students in Universities, who are concerned with semiconductors and semiconductor devices will find the book most valuable.

E. G. JAMES

Basic Electricity for Communications

Second edition. By William H. Timble. Revised by Francis J. Ricker. 538 pp. 80 figs. Demy 8vo. John Wiley & Sons, Inc., New York. Chapman & Hall Ltd., London. Price 50s.

THIS book can be simply summarized by stressing that the book is American, with American terminology and units which in many cases need translating by the English student, although this is less of a drawback to the experienced engineer who appreciates the difference between British and American practice. Symbols and abbreviations are not differentiated, although of course in British practice the two are distinct and each has its use. The references to the American wire gauge is to be expected, but the statement that the corresponding one here is the Birmingham wire gauge is untrue in electrical practice. Neither does an ammeter 'register' a current in our terminology. 'Capacitance' and 'capacity' are used indiscriminately and a student must be confused by this and naturally want to know what is the difference.

After all these points of difference the book can be considered in detail. The sixteen chapters deal with fundamentals; Ohm's law; power and energy; conductors; measurement of resistance; Ohm's law extended to further theorems; batteries; magnets and magnetic circuits; generators and motors; inductance; capacitance; a.c.; a.c. circuit applications; a.c. and Thevinin's theorem; vacuum tubes; gaseous conduction and semiconductor devices.

The subject matter is generally well treated although somewhat on the colloquial side as compared with good British practice. For example many common and excellent generator and motor characteristic curves relating voltage, current, speed, torque, etc., are absent, although some are given. A.C. theory includes vector treatment but the j-operator is not seen until it is later found in an appendix. Neither m.k.s. nor c.g.s. units appear to be mentioned anywhere. Much other bread-and-butter material is not included, and possibly this illustrates the difference between British and U.S. teaching practice.

The two chapters on electronics cover diodes and multi-grid valves, valve amplifiers, a simple valve-voltmeter, gaseous conduction, semiconductor devices, grid-controlled gasfilled valves, gaseous discharge tubes and voltage regulator tubes, metal rectifiers and transistors.

Each chapter sequence is—text; summary; problems (without answers). This is easy to follow and the book is very readable. There are several appendices, mainly tables of constants, etc. As will be expected, the book is likely to be more widely read in America than here.

E. H. W. BANNER

The Structure of British Industry

Edited by Duncan Burn. 499 pp. Medium 8vo. The National Institute of Economic & Social Research. Vol. 2. Cambridge University Press. 1958. Price 50s.

THIS is a survey for all who are interested in an up-to-date outline of the organization and functioning of the principal British industries. Each chapter treats one industry. The authors give an idea of the number, size, scope and interrelations of the firms or units in the industry; relations with overseas industries and the home government; the effect of the structure of the industry on its economic performance; its adaptability to markets and technical change; the degree of monopoly, and so on. Among the chapters of this volume are those on the aircraft industry and the electronics industry.

The Junction Transistor and its Application

Edited by E. Wolfendale. 394 pp. 397 figs. Demy 8vo. Heywood & Co. Ltd. 1958. Price 84s.

THIS book is the collected papers of nine physicists and engineers who have considerable experience in the field of semiconductors. Thus it is divided into ten chapters, each of which is written by a specialist in the work covered. The chapters have been chosen to give a fairly comprehensive coverage of the subject.

The book begins with an introduction to semiconductor physics followed by the theory of the pn junction and the transistor. The operation of the transistor is developed mathematically by means of the transmission line approach. The practical aspects of the current and voltage limitations are also briefly mentioned. In the second chapter, the more practical aspects of transistor characteristics are discussed. That is the different circuit configurations and the concept of small signal operation. In this the transmission line circuit is simplified to the T-network. Carrier storage and the effects of the bias conditions are also considered. Chapter three deals solely with the a.c. small signal representation of a transistor. Initially the "black box" four terminal network systems are explained, then the standard T and π networks are developed from the transmission line concept. The repetition in these three chapters of three slightly different aspects of equivalent networks illustrates a slight weakness of a textbook, in which the chapters are written by separate authors.

These fundamental aspects of transistor physics are followed by an introduction to the design of transistor circuits, starting with d.c. biasing arrangements and proceeding to the design of small signal and power amplifiers. The remaining sections of the book are more specialized and deal with high frequency amplifiers, oscillators, non-linear circuits, d.c. convertors, etc. The latter specialized chapters give the basic circuit arrangements in a qualitative manner together with some of the practical refinements. Some aspects, such as the design of narrow-band high-frequency amplifiers are dealt with in considerable detail, but others such as the use of catching diodes in bistable circuits are only briefly mentioned. Similarly some of the authors have tended to over-emphasize particular subjects. For example three quarters of the chapter on d.c. convertors is devoted to the ringing choke type at the expense of the equally, if not more, important transformer coupled arrangements. As the title states, the book is concerned solely with junction transistors and, in particular, germanium pnp alloy types. This is not sufficiently clearly stated in the text.

With such a large number of co-authors it is difficult to summarize briefly the manner of presentation. However, in general, the book is written in a clear concise manner with the use of short sentences and adequate sub-division of the chapters. The mathematical text and the figures are well laid out. Since the figures are numbered separately for each chapter, it is not readily possible for the authors to refer to other chapters; this has led to the repetition of some figures. As such this book is suitable as a textbook for electrical engineering students. Similarly it is suitable for engineers or physicists who are being introduced to transistors for the first time. Covering, as it does, a wide range of transistor applications, the book is not intended as a specialists book.

R. A. HILBOURNE

Conference on Extremely High Temperatures

Edited by H. Fischer and L. C. Mansur. 258 pp. 80 figs. Medium 4to. John Wiley & Sons Inc., New York. Chapman & Hall Ltd, London. 1958. Price 78s.

THIS book is based on the Conference on Extremely High Temperatures held in Boston in March 1958. It is a summary of the underlying basis for commercial or usable thermo-nuclear reactions. Emphasis is on high temperatures as they relate to the propulsion fields, with emphasis on the physics of the topics covered.

The conference was held under the sponsorship of the Air Force Cambridge Research Center, Electronics Research Directorate. This directorate is one of four Centers under the Air Research and Development Command concerned with electronics research and development for the United States Air Force.

Automatic Measurement of Quality in Process Plants

320 pp. 100 figs. Demy 8vo. Butterworths Scientific Publications. 1958. Price 50s.

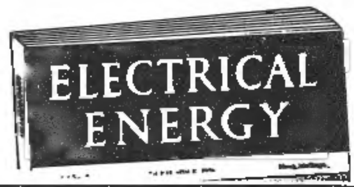
THIS book gives the edited proceedings of a conference sponsored by the Society of Instrument Technology, held at Swansea, in September 1957. The papers fall into two complementary groups; those covering a survey of the experience which has been gained with the wide variety of control instruments already in general use, while the second group explores the potential plant application of analytical techniques currently in use only in the laboratory.

Commercial Television

By Wallace S. Sharps. 496 pp. 90 figs. Demy 8vo. The Fountain Press. 1958. Price 63s.

THIS is a manual of advertising and production techniques and includes a dictionary of over 1 200 terms used in advertising film and television production.

The book covers all major techniques for live and film transmission and includes the latest available information on presentation from five-second slides to advertising magazines. The legal section contains a considerable amount of essential information for advertising practitioners.



THE COMPUTER EXHIBITION

A description, compiled from information supplied by the manufacturers, of a small selection of the equipment to be exhibited at the Computer Exhibition at Olympia, London, from 28 November to 4 December.

(The figures in parentheses indicate Stand Numbers)

AIR TRAINERS LINK LTD (1a)

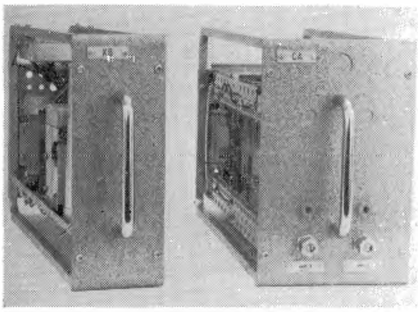
Bicester Road, Aylesbury, Buckinghamshire

A small mechanical computer will be shown as a working exhibit and also on the stand will be a scale model of the British European Airways training centre at Heston. Large electronic simulators have already been installed for Elizabethan and Viscount aircraft and work is proceeding on Comet 4 and Vanguard equipments. When complete this will probably be the largest group of d.c. analogue computers in the country. The individual components of these simulators are available for general use in industry and a few of these are described below.

STANDARD AMPLIFIERS

(Illustrated below)

A range of standard amplifiers has been designed for use in d.c. computing systems. Typical of these are the two illustrated: the type CA (right) and the XB (left).



The chassis of the type CA carries two entirely separate amplifiers which are identical in performance. The rear one, only, has connexions brought out from the anodes of the first stage. These connexions are used for the introduction of supplementary integration circuits when the amplifier is part of an electro-mechanical integrator. The supplementary circuits enable a very much wider range of rates to be integrated accurately.

The following description applies equally to either the front or the rear amplifier:—

The input stage employs a pair of high-gain, low drift pentodes connected in a differential long-tail circuit. Very high value summing resistors may be used as, for practical purposes, the input impedance is infinite and the grid current negligible.

Open loop gain is approximately $\times$ 60 000. Working gain may be set by the use of suitable feedback components to any value greater than $\times$ 0·01, subject to limitations imposed by the capacitive reactance of the load.

For load impedances of 1·5kΩ and over, the output can be driven into saturation without valve dissipation being exceeded. The rated output into various loads is $\pm$85V into 1·5kΩ, $\pm$100V into 2·5kΩ, $\pm$150V into 5·6kΩ. It is inadvisable to allow the output voltage to rise above 150 owing to the risk of damage to the valves.

A pre-set zeroing potentiometer and a test socket for checking output are mounted on the front panel. A changeover switch and resistance network must be provided externally to isolate the input from external signals while zero adjustment is in progress. The switch may be mounted on the front edge of the horizontal racking member.

Drift measured over a period of 120 hours is better than 20mV referred to the input. When used with a type XB chopper drift corrector at a loop gain of $\times$ 1 000, overall drift referred to the input is reduced to 100μV or less.

At unity gain the response is flat to 700c/s, falling to $-$40dB at 50kc/s. At open loop gain it is flat to 4c/s, falling to $-$17dB at 100c/s.

The type XB is a chopper type amplifier with a d.c. gain of about 150 times, and occupies the whole chassis. When connected to the grids of the differentially coupled long-tail pair input stage of amplifiers such as types CA, CD or CQ, a reduction of drift of the order of 150 times referred to the input is obtained. The chopper relay operates at a frequency of 50c/s. The standard d.c. amplifiers have excellent drift characteristics and in a typical computer drift correction is only justified in about 25 per cent of the amplifiers used. For this reason the XB drift reducing amplifier was introduced so that it could be used as an add-on refinement only where necessary, thus giving the maximum economy.

A low-pass filter with an attenuation of 24dB at 50c/s is used to remove unwanted ripple and ensure that the chopper samples principally the d.c. error voltage which can exist at the summing network junction if drift occurs in the main amplifier input. Input loading is negligible and in no way limits the use of very high value summing and feedback resistors. The d.c. gain is approximately $\times$ 150.

An external 10MΩ resistor connected to the second input of the main amplifier terminates the output of the drift corrector. Under these conditions the time-constant of the output smoothing filter in the drift corrector is of the order of 20sec, which is adequate to ensure loop stability and remove all traces of chopper ripple.

A limiter restricts the maximum output to about $\pm$0·8V. This adequately corrects all valve drift and avoids any serious blocking effects if the main amplifier is subjected to large temporary overload signals.

EE 4751 *for further details*

ELECTRONIC ENGINEERING will occupy Stand Number 21 and visitors will be welcome.

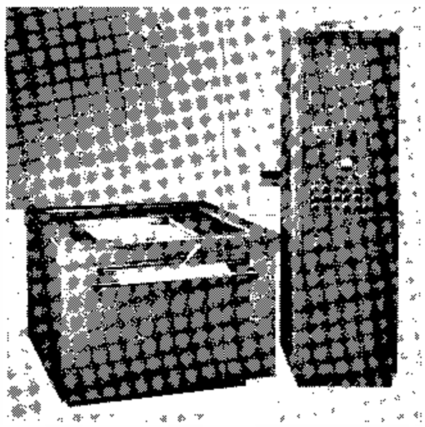
BENSON-LEHNER (G.B.) LTD. (12)

12, Bargate, Southampton, Hampshire

OSCILLOGRAPH TRACE READER

(Illustrated below)

The Oscar, model E, is a reading machine designed to expedite the analysis of various continuous trace records appearing either on film or paper. Records up to 12½ inches in width may be accommodated in the unit. While the actual alignment of the traces is performed manually, the machine automatically applies either linear or non-linear calibrations to the amplitude measurement and furnishes a resistance output proportional to the calibrated amplitude. Several types of Benson-Lehner plotters and automatic tabulators are available for



translating the resistance output of the Oscar into a graphic or printed record.

The machine has a back-lighted field of view 13in high and 26in wide. Variable control of light level is provided for convenience of reading traces of various densities. Film and paper records, either opaque or translucent, can be analyzed. The manual record transport will accommodate records up to 12½in in width and partly-filled spools can be interchanged without unspooling. Two alignment references are provided (one at each extreme of the horizontal field of view) for the alignment of the entire measuring system with the oscillograph reference trace, thereby aligning the machine to the paper instead of the paper to the machine. The record is positively secured by two rollaway magnetic clamps and a movable time reference can be positioned at any point over a distance of approximately 16in across the horizontal field of view.

A movable carriage incorporating two independent outputs for amplitude measuring (Y-axis) and two independent outputs for time interpolation (X-axis), provides for the mounting of either linear or non-linear calibration traces which are to be introduced in the trace reading. The origin of the calibration overlays can be positioned anywhere on the record and can be adjusted to compensate for channel zero drifts or scale changes. Provision is made for mounting several non-linear calibrations for the reading of adjacent multiple channels. Special overlays make possible slope reading, reading the sine, log, etc., of the amplitude, or for translating curvilinear to rectangular co-ordinates.

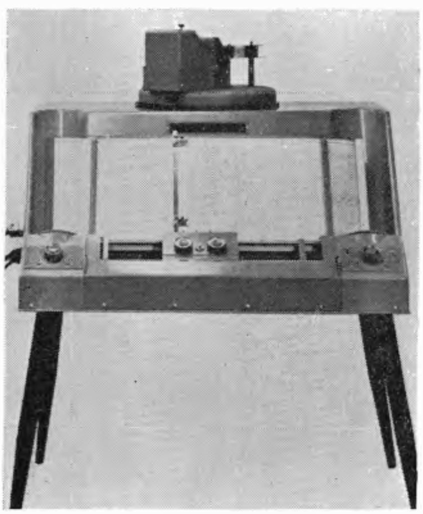
The accuracy of the measuring unit is $\pm 0{\cdot}1$ per cent of full-scale deflexion, although the actual system accuracy will always be a function of the quality of the record and the operator's ability.

EE 4752 *for further details*

ELECTRO-PLOTTER

(Illustrated below)

The electro-plotter model H is an automatic plotting machine for graphing business and scientific data. It will plot points and symbols and optionally draw straight lines between successive co-ordinates.



Available inputs include keyboard, punched cards, punched paper or magnetic tape and outputs from automatic data handling systems or record readers.

The plotting area is 30in by 30in and the data can be displayed as points, symbols or straight lines. It will plot points at from 30 to 60 points/minute with an accuracy of 0·05 per cent of full-scale.

Input registers are provided which present direct decimal display of the digits and signs for the X and Y values of the point to be plotted or the termination of the next line.

EE 4753 *for further details*

THE BRITISH TABULATING MACHINE CO. LTD. (5, 7)

17, Park Lane, London, W.1

GENERAL PURPOSE COMPUTER

(Illustrated above right)

One of the main exhibits on this stand will be the Hec 1201 general purpose computer. In this computer the input is by means of standard 80-column punched cards fed at a speed of 5 100 cards/hour by the input-output tabulator. One hundred and sixty binary digits may be read from up to 64 columns of each card at any time. Cards can be read up to three times at different sensing stations and information not computed can be stored in a 24-character storage unit or listed directly on the output printer.

The main internal storage of the machine is the magnetic drum. The periphery of the drum is divided into 64 sections or tracks, each track storing 16 words consisting of 40 binary digits; total capacity 1 024 words. Any track may be instantaneously referred to by means of a system of high speed electronic track switching. The drum revolves at 3 000rev/min, giving a basic word transfer time of 1·25msec.

The arithmetic unit consists of four valve-shift registers, an adder and a complementer. Additions or subtractions take place during the basic word time of 1·25msec and usually take 2·5msec. The average time for multiplication by a factor equivalent to about 8 decimal digits is approximately 20msec. When dividing, a 40 binary digit quotient can be found in a maximum of 50msec.

The output may be a printed document or a Hollerith 80-column punched card, or both. The 100 alphanumeric print wheels of the input-output tabulator, constitute a flexible output instrument

capable of producing a wide variety o documents. Time taken to print a line is 0·7sec; punching a card takes 0·6sec. Output can be in any notation.

EE 4754 *for further details*

THE BRITISH THOMSON-HOUSTON CO. LTD. (8)

Rugby, Warwickshire

CO-ORDINATE POSITION SETTER

(Illustrated above right)

An example of the Company's activities in the machine tool control field will be a working demonstration of a control desk for an electronic coordinate setter with an automatic card-reader in position. In this equipment, a mechanical card-reading device 'senses' the positions of holes punched in a card and enables the setting dials to be automatically positioned. Two sets of co-ordinates are fed in simultaneously, the operator merely dropping in a card and pressing a button. The set dimension is then displayed on the dials for checking purposes if required.

EE 4755 *for further details*

HELISYN POSITION CONTROL

A Helisyn position control will be demonstrated by a working model. The Helisyn is a device, at present undergoing development, for the purpose of precision linear mechanical positioning in response to a numerical input and with particular application in fields such as machine-tool

control and automatic inspection. It consists of a short cylindrical part which is movable along (but is not in contact with) a longer cylindrical part, each part being provided with conductive paths in the form of interlaced helices. One of the elements of the Helisyn is excited electrically by appropriate voltages set up in accordance with the input number, resulting in an electrostatically induced misalignment signal appearing on the other element.

The exhibit comprises laboratory apparatus arranged to accept decimal input defining a position to one ten-thousandth of an inch, and to show that the combination of digital/analogue convertor and Helisyn may be used to carry out mechanical positioning to an order of accuracy consistent with that of the input.

EE 4756 *for further details*

BURROUGHS ADDING MACHINE LTD. (28)

Avon House, 356, Oxford Street, London, W.1

DESK SIZE COMPUTER

(Illustrated below)

The Burrough's E101 computer is intended primarily for engineering and scientific applications, the magnetic drum storage capacity has been increased to 220 words. The computer is about the size of an office desk and meets the requirements of low capital cost.

Data is entered manually through the keyboard of an accounting machine, and programmed control is obtained by means of a plug board with pin insertion to accord with the instruction desired.

An alternative method of input has been designed to enter both data and instructions by means of punched paper tape or punched cards.

Output can be in the form of an accounting machine line printer or by punched paper tape. In the latter case this computer can be linked directly to an analogue computer or an XY plotter.

EE 4757 *for further details*

ELLIOTT BROTHERS (LONDON) LTD. (35, 43)

Century Works, Lewisham, London, S.E.13

Elliott Brother (London) Ltd., in association with the National Cash Register Co. Ltd. and Panellit Ltd. will be exhibiting a wide range of computing and data-processing equipment. Among the latest products of this group are the following.

High Speed Tape Reader

(Illustrated above)

This is designed to read standard 5, 7 and 8-hole punched paper tape at any speed up to 1 000 characters/second and can stop within a character. The characters are sensed by OCP71 phototransistors, which are illuminated via Perspex light guides. The transistors and guides can be withdrawn from the instrument in one unit, should a transistor require replacement. Transistor adjustment can be effected by removing an inspection plate.

The tape is drawn through by a pair of rollers, the upper roller being fixed on the end of a motor shaft. The position of the lower roller is controlled by an electromagnet, the amount of movement being sufficient to remove the drive from the tape, when desired. The tape also passes between two flat surfaces, the pressure between these being controlled by a second electromagnet. This provides a braking action on the tape, which always overrides the action of the driving rollers.

The instrument is complete with lamp, driving motor, electromagnet assemblies and transistor assembly, but does not include any electronic control circuits. The supplies required are 240V 50c/s mains, 12V 3A a.c. or d.c. for the lamp, and a low voltage supply not greater than about 20V for the photo-transistors. The electromagnet coils each have a total resistance of 65Ω, and will operate correctly on 100mA d.c. The overall dimensions are $10\frac{3}{4}$ by $6\frac{9}{16}$ by $8\frac{1}{4}$in high.

EE 4758 *for further details*

General Purpose Computer

(Illustrated below)

The National-Elliott 802 'Compudesk' is a fully automatic stored programme digital computer which can operate entirely without manual intervention, although this is possible and simple if required. The use of printed circuits, transistors and magnetic cores has made possible a compact construction which is little larger than a normal office desk and has eliminated the need for special cooling arrangements.

Orders are of the single address type and there are two orders, of 16 binary digits each, per word of 33 binary digits.

The extra digit is used for automatic modification of orders (B-modification). By means of this digit any location in the store may be used as a B-line. A B-lined order takes no longer to obey than the same order without a B-line. In previous computers facilities for doing this automatically have been limited to a few registers, often less than eight. The 1024 B-registers in the 802 make it possible to simplify the programming of extremely complex problems.

The functions are numbered in a logical way that is learnt very rapidly with no conscious effort on the part of the programmer. Each order refers implicitly to the single accumulator, and to one store address.

The large store has sufficient capacity for the computer to perform a wide range of complex problems, or to tackle large data processing tasks.

By using simplified operating methods the 802 will not require specialized programming staff and the ultimate user will be able to programme the 802 himself after only a short period of initial instruction.

The basic 802 'Compudesk' allows for punched paper tape input and output. Many applications require specialized input and output devices and provision has been made in the design to allow the 'Compudesk' to be attached simultaneously to several such devices. These include punched card input, manual keyboard input, analogue and digital recording mechanisms and others.

Three keys control operation and enable the user to start, stop and perform certain other operations. The simple control panel reduces the chance of error due to incorrect manual operation of the controls. A loudspeaker indicates the machine is calculating and produces a sequence of tones according to the instructions being obeyed.

A comprehensive library of subroutines is available, thus greatly reducing the amount of programming required.

EE 4759 *for further details*

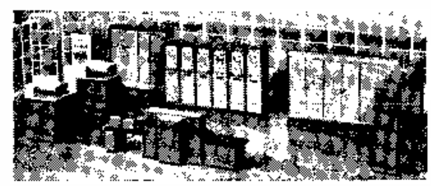
E.M.I. ELECTRONICS LTD. (26)

Hayes, Middlesex

High-Speed Data Processing System

(Illustrated below)

The EMIDEC 1100 is a high-speed digital electronic computer which, with its associated input and output equipment, and high-speed magnetic tape units, makes up a fully comprehensive data processing system.



Magnetic cores and transistors and unit construction are used throughout. Operating in the binary mode it has a fixed word length of 36 binary digits. Automatic conversion to and from the binary code is provided with all peripheral units.

There are three means of storing data:

(1) Immediate access store. A 1 024 'word' immediate access store of magnetic cores interlaced to form a matrix provides a range of accumulating registers and also holds the program currently in use.

(2) Magnetic drums. Drums of 8 192 or 16 384 words provide the main internal store. Any number of drums, within reasonable limits, may be used.

(3) Magnetic tape. High-speed magnetic tape decks with a tape storage capacity exceeding 13 million binary digits, or $2\frac{1}{2}$million alphanumeric characters and a very fast read/write rate of 20 000 characters per second play an integral part in the EMIDEC system.

The clock rate is about 112kc/s. Addition and subtraction takes 125μsec and multiplication and division 1 120μsec. (These times include any "B-line" modifications and three access times to the immediate access store.)

A variety of input and output equipment is available and the input and output may be in sterling, decimal or alphanumeric form.

Also on show will be several units from the EMIDEC 2400 which is a new high-speed data processing system with very comprehensive facilities.

EE 4760 *for further details*

THE ENGLISH ELECTRIC CO. LTD. (36)

Marconi House, Strand, London, W.C.2

Digital Computers

The 'DEUCE' digital computer is now available in two versions, 'DEUCE' Mk. I which is intended mainly for scientific work although it also has a wide range of uses as a general purpose computer, and 'DEUCE' Mk. II, which has been designed primarily to form the central unit of data processing systems, but is also a powerful unit for scientific as well as for commercial applications.

'DEUCE' Mk. I uses 64-column punched card input and output, and can be fitted with magnetic tape auxiliary storage if required. The input and output system for 'DEUCE' Mk. II is combined in a single unit using all 80 columns of the standard punched card equipment, and normally this computer would be equipped with magnetic tape auxiliary storage.

'English Electric' high-speed paper tape input and paper tape output can be supplied for use with both the Mk. I and Mk. II versions of 'DEUCE'. Thus either computer can be integrated with any existing system which already uses paper tape.

One of the outstanding features of the 'DEUCE' system is the extensive library of programmes and subroutines which has been built up over the years since 'DEUCE' was first developed from the 'ACE' Pilot machine shortly after the last war. This library is available to all users of 'DEUCE', who are invited to add to it from their own operational experience of the computer.

EE 4761 *for further details*

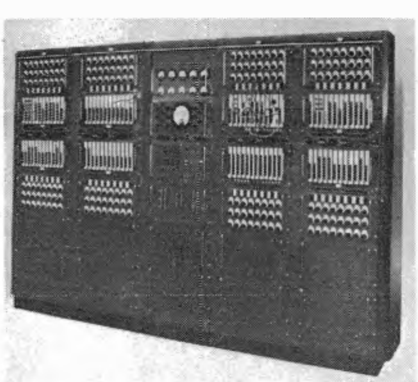
THE FAIREY AVIATION CO. LTD. (43a)

24, Bruton Street, London, W.1

Analogue Computer

(Illustrated below)

The Fairey electronic analogue computer is designed on a unitary basis to fit standard 19in by 6ft racks. A comprehensive range of elements is available enabling a computer of any desired size or complexity to be built up. From the operational standpoint, the machine has been so designed that an operator may set up a programme without having an intimate knowledge of the internal electronic circuits. The design also ensures a minimum loss of computer time in servicing. The basic arrangement comprises a control rack, housing control and distribution equipment for up to 96 operational d.c. amplifiers. Facilities for amplifier zero checking in the standby condition and overload indication during problem run, are also included.



The basis of the computer is the computing panel which houses 12 standard d.c. operational amplifiers. These amplifiers, which are in the form of a thin slab, plug into the back of the panel. The front portion of the panel contains a number of compartments into which function units, containing feedback components may be plugged.

These units are available for a wide range of linear and non-linear mathematical functions; depending on complexity they embrace one or more operational amplifiers.

A servo function unit containing an electromechanical function unit, a number of buffer amplifiers, and a servo amplifier, is also available. A range of interchangeable electromechanical function units is available having a number of applications which include multiplication, division, angle resolution, etc.

Two methods of problem patching are provided, one of which consists of directly interconnecting the function units by plugging in short lengths of cable, the second method being a central patching system where problem boards are interchangeable and removable for programming remote from the computer.

Solutions may be displayed or recorded on multi-channel cathode-ray oscillographs, quick response direct-writing recorders, or on X-Y plotters according to customers' requirements.

EE 4762 *for further details*

FERRANTI LTD. (10, 14)

Hollinwood, Lancashire

A large and comprehensive range of computing, data-processing and machine control equipment will be shown by Ferranti Ltd, including the following.

Data Processing System

(Illustrated above)

This large-scale commercial system, exhibited for the first time, uses magnetic tape, punched cards and punched paper tape ; it will be shown in model form carrying out the basic office procedure of up-dating a file of information. This routine task occurs throughout business and commerce, e.g., in wage accounting, stock control, consumer billing, sales statistics.

'Perseus' was specially designed for doing commercial work and is probably the largest data-processing system yet made in Western Europe. The system incorporates many novel features, particularly in data conversion and checking.

Among the facilities provided by this system are the following.

Up to 2048 words of 12 alphanumeric characters.

Mixed Radix Arithmetic (direct handling of sterling and decimal).

Field selection facility, for operating on parts of a 12-character word.

16 magnetic tape units controlled through four buffer stores, for simultaneous processing.

Built-in checks of all arithmetical and magnetic tape operations.

Two-card inputs, each can be of any punched card type.

Paper tape input and monitoring teleprinter.

Off-line magnetic tape operated Samastronic printer.

EE 4763 *for further details*

Digital Plotting Table

This digital plotting table unit has been designed to plot at high speed, two dimensional digital information, in conjunction with a digital analyser attachment or to operate from a suitable input as an automatic drafting machine. In this latter application, use can be made of the Ferranti computer service.

The M type transmission drive motors are transistor controlled and fed with pulsed information. Each pulse (positive or negative) causes a pen displacement in the appropriate sense of one digital unit.

The machine is designed on kinematic principles to reduce drive friction, and the carriage weight is controlled to give the best possible dynamic performance. The maximum traverse velocity is 60in/min with a digit size of 0·005in. Overall accuracy is better than 1 in 1 000.

Some flexibility is permissible both on machine size and on digit size to suit specific requirements.

EE 4764 *for further details*

IBM UNITED KINGDOM LTD. (34)

101, Wigmore Street, London, W.1

Data-Processing System

(Illustrated below)

The IBM305 Romac system will be shown in this country for the first time.

Transactions are processed individually as they occur (in-line processing), resulting in up-to-date records for close control over business data and control panel flexibility is combined with stored program instructions for internal processing of

information. The output is controlled independently of input and automatic data transfer of from 1 to 100 characters, with built-in validity check is possible.

There is a large-capacity magnetic disk storage and alphanumerical character representation is used throughout with direct interrogation of the disk file.

EE 4765 *for further details*

METROPOLITAN-VICKERS ELECTRICAL CO. LTD. (22)

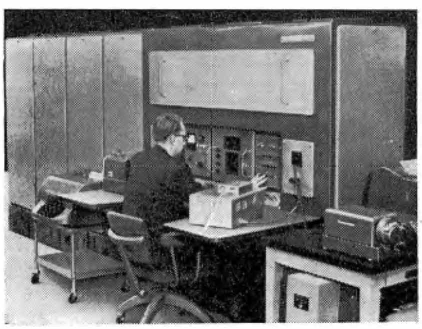
Trafford Park Manchester, 17

General Purpose Computer

(Illustrated below)

The Metrovick 950 general-purpose computer has been specifically designed to meet the requirements for a reliable medium-speed computer, at a price which is attractive to small firms, universities and technical colleges. It has many advantageous features, not usual in a computer of this size, which enable a wide range of mathematical, scientific and engineering problems to be solved.

It consists of five main parts; input equipment, store, arithmetical circuits, output equipment and control unit. Information is fed into the machine by means of a photo-electric punched-paper-tape reader, the paper tape having a 5-hole



code; the instructions and data are stored on an aluminium drum coated with a magnetic oxide, on which information is recorded in the form of magnetized spots on circumferential tracks. The arithmetical and control circuits incorporate junction transistors, and this means that size and power consumption are appreciably reduced. Output is by means of a 5-hole punched paper tape and printed page, the latter being very convenient during development of a computer programme. The control unit co-ordinates the operation of the whole machine.

The 32 instructions obeyed by the '950' have been chosen explicitly to make programming easy. The machine is a 1+1 address system, that is to say, in each instruction two addresses in the drum store are mentioned, the operand and the next instruction.

With the optimum timing the machine will obey between 150 to 200 instructions/sec. The optimum timing is often held to be a difficulty in programming, but the method used with the '950' of coding and optimizing simultaneously is quite simple.

EE 4766 *for further details*

Data Processing Computer

Also on show will be a scale model of the 1010 computer which has been designed for data processing and real-time process control, and is capable of solving complex scientific and engineering problems. It is a fast transistorized computer with parallel arithmetic circuits, magnetic-core working store and a variety of ancillary apparatus.

The fast arithmetical and control unit operates in the parallel mode, and carries out typical data-processing programmes at a speed of some 50 000 simple address instructions per second. Word length is 44 digits and contains two 22-digit instructions. A 'B' line has 13 digits and is capable of performing limited operations. The flexible instruction code provides a wide range of instructions; both fixed-point and floating-point calculations can be performed.

The immediate-access store holds either 2 048 or 4 096 words, each of 44 binary digits in a three-dimensional array of ferrite cores. Each word may contain two computer instructions, a 13-digit decimal number, or a group of coded alphabetical characters. Auxiliary storage is provided by two types of magnetic drum store and by a magnetic tape store.

A wide range of other ancillary equipment is available for use with the Metrovick 1010, including high-speed input and output apparatus. An installation may be enlarged by the addition of ancillary equipment up to a maximum of 32 units. Each unit is allowed to operate at its optimum speed by using a small buffer store to isolate it from the computer.

A novel feature of the Metrovick 1010 is the data scanner, which transfers information between the magnetic-core store and all ancillary equipment. The data scanner interlaces the transfers automatically word by word, with the execution of computer instructions, and it can maintain simultaneous transfers of blocks of data to or from all ancillary apparatus. This system allows relatively slow peripheral equipment to be connected 'on-line' to a very fast computer, without delaying computation by printing out, etc.

The instruction code provides facilities for the operation in parallel of a number of programs having compatible storage requirements. Programs may be interrupted in order to execute another program of high priority, as for instance, to carry out emergency operations when alarm conditions occur in a computer-monitored process plant.

EE 4767 *for further details*

MULLARD LTD. (2)

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

High Frequency Power Transistors

The OC23 and OC24 are two new alloy junction high frequency power transistors developed specifically for digital computing applications.

The outstanding feature of these new transistors is their exceptionally good high frequency performance combined with large power handling capabilities: the average cut-off frequency is 2·5Mc/s and the collector dissipation 8W at an ambient temperature of 45°C (with infinite heat sink). These characteristics make them suitable for several other computing applications besides their primary purpose of driving rectangular-loop ferrite computing elements and memory matrices. They can, for example, also be used to provide drive pulses for magnetic drum and tape recording circuits; in the drive and output stages of clock-pulse generators; in servo-mechanisms associated with computers; and as fast-response series power regulators for transistor h.t. supplies dealing with large-current microsecond pulses.

For driving rectangular-loop ferrite devices they provide 700mA pulses of, for example, 2μsec duration and with a rise time of less than 0·5μsec. Used in sine wave applications two OC23s or OC24s in class-B push-pull deliver upwards of 3W at a frequency of 500kc/s.

The transistors have an extremely high slope—16A/V—and give average current gains of 200 at 100mA collector current and 150 at 1A. The bottoming voltage is only 0·4V for 1A collector current, with 30mA base drive.

Maximum emitter current is 1·2A peak or mean, the maximum continuous collector-emitter voltage is 24V (with a higher peak voltage) and the maximum thermal resistance 3°C/W.

EE 4768 *for further details*

MULLARD EQUIPMENT LTD. (11)

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

Magnetostrictive Delay Lines

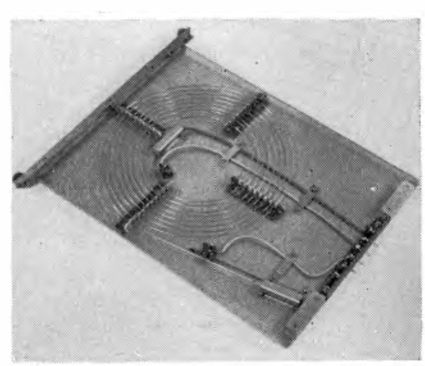
(Illustrated below)

An improved type of magnetostrictive wire delay line with facilities for continuous adjustment of the delay period over a wide range will be shown by Mullard Equipment Ltd.

Lines giving maximum delays of any duration up to 4·5msec can be supplied, built to users' specific requirements and incorporating, if required, input and output amplifiers and pulse re-forming circuits.

The maximum pulse repetition frequency that can be stored depends on the value of the delay required: in a 4·5msec line a p.r.f. of at least 250kc/s can be stored, while with a 1msec delay 1Mc/s is attainable. No temperature control is necessary since the lines have a temperature coefficient of delay as low as 5×10^{-6}/°C.

The lines use a newly-designed transducer that makes possible a continuous



variation in the delay period. Essentially the transducer consists of two coils connected back-to-back and mounted on a former which fits around the wire. Adjustment of the delay is obtained by movement of the transducer along the line. The range of adjustment given by a single transducer is 100μsec, determined chiefly by the distance between the brackets supporting the wire, but by using a number of transducers spaced along the line the range can be extended indefinitely and it is also possible to produce from one line a number of signals each delayed from the initiating pulse by pre-selected intervals.

To keep the dimensions of the line as small as possible the acoustic energy generated by the transducers is propagated along the wire in the form of torsional waves. These have the advantage over longitudinal waves of a lower velocity in the wire and thus require a physically shorter line to produce a given delay. They also give rise to less dispersion, enabling the wire to be wound compactly in spiral form. A line giving a maximum delay of 1msec requires 10ft of wire and can be mounted in a unit of only 10in by 7½in by ¾in overall dimensions.

EE 4769 *for further details*

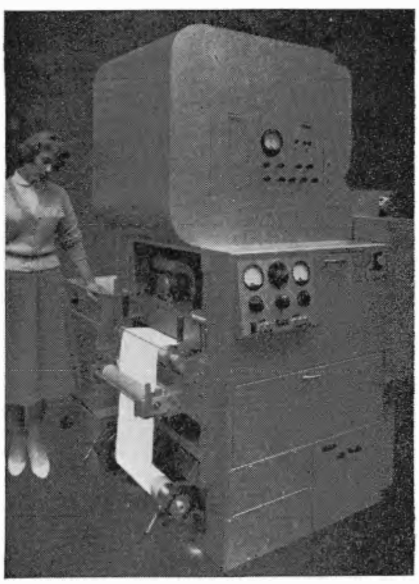
RANK PRECISION INDUSTRIES LTD. (48)

37-41, Mortimer Street, London, W.1

COMPUTER PRINTER

(Illustrated below)

The object of this high quality printer is to print the output from a computer at computer speed. The experimental model on display prints 1 500 lines/minute, but this will be increased to 3 000 lines/minute. At the exhibition, pre-printed forms will be used, but plans are already well ahead for it to print its own forms at the same time as it prints the computer's output, giving greater flexibility to the system and considerably reducing the running costs. By coding certain instructions or 'orders' in the input, the layout of data on the form may be arranged as desired.

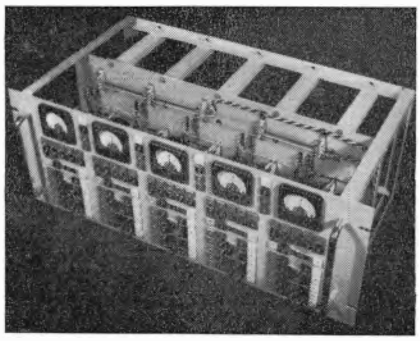


This printer works either from the computer's output recorded on magnetic tape or straight from the computer's buffer store. A beam of electrons writes the letters and figures on an ordinary cathode-ray tube, and they are instantly copied by xerography.

The xerographic machine used is the Rank-XeroX Copyflo Continuous Printer. This copies an original (whether it is on paper or on a cathode-ray screen) on to plain, unsensitized paper at the rate of 20ft/min. The process, which is completely dry and automatic, works by electrostatics.

For the printing, any styling (or typeface) can be used, with capitals and small letters, and italics if required and the number of characters it can print is not limited. The unit on display will have a repertoire of 50 characters, printing up to 128 characters across paper 13in wide and the input will be from magnetic tape. The final model will use paper 26in wide and will produce 2 copies of 128 characters each or 4 copies of 64 each per line.

EE 4770 *for further details*



SAUNDERS-ROE LTD. (19)

East Cowes, Isle of Wight

ANALOGUE COMPUTING EQUIPMENT

(Illustrated above)

The range of analogue computing equipment shown will include a complete computer, a Miniputer and also those individual units from the range of standard computing units which are not included in the computer. These separate units will include a diode function unit, a quarter squares multiplier, a limiter and a multiplier/divider Mk. 2. In addition an amplifier mounting panel and five amplifiers will be exhibited separately although several of these units are in fact included in the computer. This is because this particular unit demonstrates the new method of construction which has been adopted for the computers of 5 plug-in computing amplifiers fitting into the rear of a mounting panel, the front of which forms the patch panel. All the necessary control relays for the five amplifiers are also contained in this unit. Thus an amplifier mounting panel together with its five plug-in amplifiers only requires the addition of 300V power units and a source of 6·3V for the valve heaters to become a complete analogue computer.

The complete computer will be built in a three-bay forced ventilated rack and will comprise twenty computing amplifiers, forty coefficient potentiometers, a servo-multiplier and a force function generator.

EE 4771 *for further details*

SPERRY GYROSCOPE CO. LTD. (39)

Great West Road, Brentford, Middlesex

MAGNETIC STORAGE DRUM

(Illustrated below)

Originally designed for use in automatic telephone switching systems and developed in collaboration with the Automatic Telephone and Electric Co. Ltd, this equipment is capable of application to a much wider range of computer systems where a requirement exists for high-density storage of information.

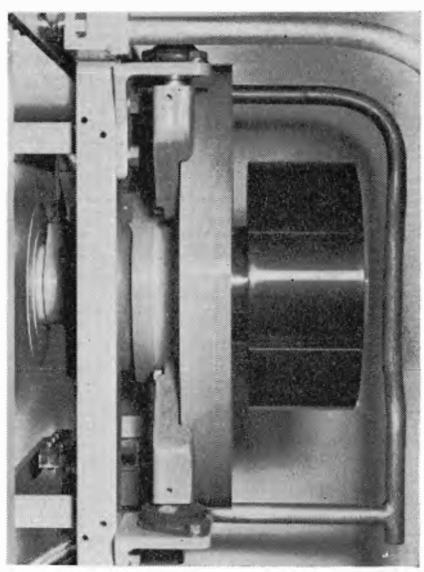
The design of this drum has been based on marine gryo compass techniques where an operating life of high-speed wheels of the order of 20 years is commonplace Thus, two of the most important features of the storage drum are its long running life and automatic lubrication system.

The recording surface of the drum is a nickel plating providing high-density storage using valve or transistorized circuits.

The reading and writing heads may be spaced axially along the mounting posts and also adjusted circumferentially to any position around the 360° of the drum periphery by means of the continuous 'T' slot in the mounting plate.

Two versions of the drum are available at present, both 9in in diameter, one with an effective surface area of 100in^2 and the other of 157in^2. The surface finish is better than 16μin and no surface defect exceeds an amount 8×10^{-9} of the total surface area. Maximum eccentricity or 'run-out' on the bearing spin axis and maximum non-linearity of the surface is less than ·0001in.

The illustration shows a side view of the drum mounted in its carrying frame. On the left-hand end of the shaft can be seen the eddy current brake which is used for speed regulation.



EE 4772 *for further details*

Meetings this Month

BUSINESS COMPUTER SYMPOSIUM

To be held at Olympia, London, W.14, during the Electronic Computer Exhibition.

Admission charges: £2 12s. 6d. per Session.
£15 15s. 0d. for all Sessions

Communications should be addressed to:
Mrs. S. S. Elliott, M.B.E.,
Electronic Computer Exhibition and Business Symposium,
11/13 Dowgate Hill,
London, E.C.4.

Date: 1 December. Time: 2.15 p.m.-4.30 p.m.
Chairman: The Earl of Courtown—I.C.I. Ltd.

Opening Address.

Computers: Retrospect and Prospect.
By: Earl of Halsbury, F.R.I.C., F.Inst.P., Managing Director, National Research Development Corporation.

SESSION 1:

Papers:
Payroll, Production and Sales Applications.
By: N. C. Pollock, M.B.E., Manager, O. & M. Dept., Stewarts and Lloyds Ltd.
The Computer as an aid to Production Management.
By: J. W. Grant, F.S.S., Manager, Production Control Section, The British Tabulating Machine Co. Ltd.
Computer Services in the United Kingdom.
By: D. Wragge-Morley, Scientific Correspondent, Financial Times.

SESSION 2:

Date: 1 December. Time: 6 p.m.-8.15 p.m.
Chairman: Mr. Lewis Wright, Chairman of the Scientific Advisory Committee Trades Union Congress.
Papers:
Large Scale File Maintenance.
By: D. West, Assistant to Financial Director, Radio Rentals Ltd.
Technical Planning of Steel Tube Manufacture.
By: R. G. Hitchcock, B.Sc., A.M.I.P.E., Executive Officer, Computer Steering Committee, Tube Investments Ltd.
Public Utility Accounting.
By: G. Sherlock, F.A.C.C.A., A.C.I.S., Deputy Accountant (Mechanisation), South Western Gas Board.
Electronic Data Processing in the Nationalized Industries.
By: Dudley W. Hooper, M.A., A.C.A., Chief Organising Accountant, National Coal Board.

SESSION 3:

Date: 2 December. Time: 10.15 a.m.-12.30 p.m.
Chairman: Sir Walter Puckey, British Institute of Management.
Papers:
Inventory Control, Accounting and Payroll.
By: A. Bradley, F.C.I.S., Head, Administration Department, Ford Motor Co. Ltd.
Production Control by Buying Computer Time.
By: R. B. Baggett, Joint Managing Director, Job White & Sons Ltd. and G. M. Davis, M.A., London Computing Services, English Electric Co.
Accounting for Farmer's Sugar Beet Production.
By: J. P. Lawler, C.A., Chief Accountant, Comhlucht Siuicre Eireann Teo. (Irish Sugar Co. Ltd.).
Electronic Data Processing.
By: A. J. Brookbank, Sales Office Manager, Glaxo Laboratories Ltd.

SESSION 4:

Date: 2 December. Time: 2.15 p.m.-4.30 p.m.
Chairman: President of the Office Management Association.
Papers:
A Practical Application in Mail Order Business.
By: P. Cotter, O. & M. Div. Littlewoods; H. E. C. Nash, Mail Order Stores Ltd., and P. Shackleton, Systems Advice, Elliott Brothers (London) Ltd.
The Application of Linear Programming to the Design of Animal Feeding Stuffs.
By: A. Muir, B.Sc., J. Bibby & Sons Ltd.
Integrating the Procedures of an Insurance Office.
By: K.-E. Schang, Managing Director, A.B. Datacentralen—(Trygg-Fylga Insurance Co.'s Group, Sweden).
The Approach to EDP of a Large User.
By: S. G. Furniss, A.C.I.S., Imperial Chemical Industries Ltd.

SESSION 5.

Date: 3 December. Time: 10.15 a.m.-12.30 p.m.
Chairman: Sir John Simpson, C.B., Controller, H.M.S.O.
Papers:
Merchandise Accounting.
By: D. S. Greensmith, M.B.E., A.C.I.S., General Office Manager, Boots Pure Drug Co. Ltd.
(a) Stores Control and Control by Exception.
By: K. F. Turner, B.Comm. Chief Computing Engineer, Rolls-Royce Ltd.
(b) Computers for Research and Design.
By: L. Griffiths, B.Sc., Secretary of Computer Investigation Committee, Rolls-Royce Ltd.
Technique for Statistical Analyses of Family Expenditure.
By: M. A. Wright, B.Sc(Eng.), A.C.G.I., Control Mechanisms and Electronics Div. National Physical Laboratories.
Government Experience.
By: J. Merriman, O. & M. Division, H.M. Treasury.

SESSION 6.

Date: 3 December. Time: 2.15 p.m.-4.30 p.m.
Chairman: President Designate, Institute of Cost and Works Accountants.
Papers:
Analysis of Sales Statistics.
By: C. A. Wilkes, F.C.I.S., A.C.W.A., Imperial Chemical Industries Ltd., Dyestuffs Division.
Wages Accounting.
By: W. H. Sargent, Central Wages Executive, British Railways, Western Region.
Industrial Mathematics.
By: D. Owen, Operations Research and Cybernetics Div., United Steel Co. Ltd.

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 17 December. Time: 6.30 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Lecture: A Vidicon Camera for Industrial Colour Television.
By: I. J. P. James.

South Wales Section

Date: 10 December. Time: 6.30 p.m.
Held at: Glamorgan College of Technology, Treforest, Glam.
Annual General Meeting followed by discussion: The Training of Electronics and Radio Engineers.

West Midlands Section

Date: 10 December. Time: 7.15 p.m.
Held at: Wolverhampton and Staffordshire College of Technology, Wulfruna Street, Wolverhampton.
Lecture: Electronic Instruments in Motor Vehicle Research.
By: J. C. Dixon.

North Western Section

Date: 4 December. Time: 6.30 p.m.
Held at: Reynolds Hall, College of Technology, Sackville Street, Manchester 1.
Lecture: High Input Impedance d.c. Amplifiers.
By: S. Stuart.

THE BRITISH SOUND RECORDING ASSOCIATION

Date: 19 December. Time: 7.15 p.m.
Held at: The Royal Society of Arts, John Adam Street, Adelphi, London, W.C.2.
Lecture: Microphone Balance Techniques.
By: J. Borwick.

THE INSTITUTION OF ELECTRICAL ENGINEERS

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

Ordinary Meeting

Date: 4 December.
Lecture: A Review of Work towards Nuclear Energy from Controlled Thermonuclear Reaction.
By: D. W. Fry.

Radio and Telecommunication Section

Date: 10 December.
Lecture: Bridging the Atlantic.
By: A. H. Mumford.
Date: 15 December.
Lecture: The Acoustic Design of Talks Studios and Listening Rooms.
By: C. L. S. Gilford.

East Midland Centre

Date: 4 December. Time: 7.15 p.m.
Held at: The De Montfort Hall, Leicester.
Faraday Lecture: Automation.
By: H. A. Thomas.
Date: 12 December. Time: 7.15 p.m.
Held at: The Technical College, The Newarkes, Leicester.
Lecture: The National-Elliott 405 Computer and its Application to Production Procedures.
By: H. A. Mathews.

Cambridge Radio and Telecommunication Group

Date: 9 December. Time: 8 p.m.
Held at: The Cavendish Laboratory, Free School Lane, Cambridge.
Lecture: B.B.C. Sound Broadcasting on v.h.f.
By: E. W. Hayes and H. Page.

North-Eastern Radio and Measurement Group

Date: 15 December.
Held at: King's College, Newcastle-upon-Tyne.
Lecture: Magnetic Tape for Data Recording.
By: C. D. Mee.

Sheffield Sub-Centre

Date: 17 December. Time: 6.30 p.m.
Held at: The Grand Hotel, Sheffield.
Lecture: A New Cathode-Ray Tube for Monochrome and Colour Television.
By: D. Gabor, P. R. Stuart and P. G. Kalman.

North-Western Centre

Date: 2 December. Time: 6.15 p.m.
Held at: The Engineers' Club, 17 Albert Square, Manchester.
Informal Lecture: Thermonuclear Fusion and the Development of Zeta.
By: E. R. Hartill.

North-Western Radio and Telecommunication Group

Date: 10 December. Time: 6.45 p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Lecture: Electronic Devices in Guided Missile Systems.
By: C. Heys.

South Midland Centre

Date: 2 December. Time: 7 p.m.
Held at: The City of Birmingham Town Hall.
Faraday Lecture: Automation.
By: H. A. Thomas.
Date: 8 December. Time: 7.30 p.m.
Held at: Malvern Winter Gardens.
Lecture: ZETA.
By: R. Carruthers.

Rugby Sub-Centre

Date: 10 December. Time: 6.30 p.m.
Lecture: Recent Uses of Ultrasonics in Investigating the Characteristics of Materials.
By: J. Lamb.

Southern Centre

Date: 3 December. Time: 6.30 p.m.
Held at: R. A. E. Technical College, Farnborough.
Lecture: Transistors in Communication and Control Equipment.
By: E. Wolfendale.

Western Centre

Date: 8 December. Time: 6 p.m.
Held at: The University Engineering Laboratories, Bristol.
Lecture: Operational Experience at Calder Hall.
By: K. L. Stretch.

THE INSTITUTION OF ELECTRONICS

Date: 5 December. Time: 7 p.m.
Held at: The Lecture Hall, Anglesea Road Annexe, College of Technology, Portsmouth.
Lecture: The Applications of Transistors in Communications and Control Equipment.
By: E. Wolfendale.

THE SOCIETY OF INSTRUMENT TECHNOLOGY

Date: 10 December. Time: 6 p.m.
Held at: Manson House, Portland Place, London, W.1.
Lecture: An Experimental Approach to Some Control Problems of Multi-Stand Rolling Mills.
By: H. Gill.

THE TELEVISION SOCIETY

Date: 4 December. Time: 7 p.m.
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.
Lecture: A New Development in Flying Spot Film Scanners.
By: E. H. Traub.