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Commentary

THE Colour Television Sub-Group of Study Group XI of the International Radio Consultative Committee (C.C.I.R.) is to meet in Vienna towards the end of this month in an attempt to reach agreement on a common transmission system for colour television in Europe.

At an earlier meeting held some eighteen months ago no agreement was reached—in short it was decided not to decide and the only outcome of the meeting was that a long period should be set aside for closer study of the three systems under consideration.

It was felt at the time that although the American NTSC system had been in operation for some ten years there were a number of limitations and that it would be of advantage to look into the recently introduced French SECAM and the German PAL colour television systems which had been designed to overcome the several weaknesses of the NTSC system.

In the meantime large scale demonstrations of the three systems have been taking place throughout Europe and it was hoped that by this time, just prior to the final meeting in Vienna, the various Governments, broadcasting and other authorities concerned would be moving towards an agreed solution. But it seems, in fact, that the very opposite is happening, for at a meeting of delegates from East and West Europe in Paris last January agreement was as far away as ever.

The situation at the moment is that France is solidly behind SECAM, West Germany is equally behind PAL and this country is quite convinced that the NTSC system is the most suitable for our needs, and further, will adopt it even if the rest of Europe decides to adopt another system. Other countries in West Europe are divided in their opinions but East Europe and the Soviet Union have so far made no announcement. Much will depend on the Russian point of view as the Soviet countries between them hold some 50 per cent of the voting power.

It is the intention of Russia to make an early decision for she plans to have colour television installed as part of the programme for the fiftieth anniversary celebrations in 1967 of the founding of the Soviet Union.

Further confusion arises from the fact that while West Germany is in favour of PAL it will hesitate to install the system there if East Germany and other East European countries prefer a different one.

It is not easy at this distance to come down wholly in favour of one particular system since there are not only technical but political and economic considerations to be taken into account. As far as NTSC is concerned, the fact that it has been in operation for some twelve years is a point in its favour but this should not be stressed too much and this is certainly no reason why other new and possibly not yet fully developed systems should be rejected.

The BBC has been experimenting with the NTSC system almost from its introduction in America and over the years has acquired a good deal of operational experience so, understandably, this is why it favours this system. Its view is now supported by the Government and the Post Office, and within the last week or so BREMA—the British Radio Equipment Manufacturers' Association—has issued a report which states that the industry is also in favour of the system.

According to this report the advantages are that the NTSC system gives better picture quality, better compatibility and controllability than its rival systems. The cost of a colour receiver based on NTSC is also claimed to be cheaper.

But the NTSC system has its limitations as well, and possibly the principal one is that it is highly susceptible to phase distortions resulting from long distance transmission of the colour signal although this can be adjusted to some extent by the use of a hue control on the receiver itself. This phase distortion in long distance transmission links did not arise in America as the links across the country were operated by a single authority and were maintained at a high standard. It was only when the European links with their varying standards in different countries were studied that this problem arose.

In Britain the transmission links operated and maintained by the Post Office are of sufficient standard to eliminate this defect but even so the Post Office and the BBC have recently combined to produce a 'multi-burst' corrector device for insertion in the link to correct this distortion. The NTSC system, using this corrector device operated by BBC engineers, was demonstrated last month in Moscow using some 6 000km of transmission links in the Soviet Union and apparently was so successful that the Russians were in doubt that the signals were actually being sent over the links and had to re-assure themselves.

It was of course this phase distortion defect which led originally to the development of the SECAM, and later the PAL, systems and both systems seem to have overcome this trouble but at an increased cost of the colour television receiver.

In the initial stages this increase in cost has been put at about 6 per cent above that of the NTSC equivalent but it is expected that when production is fully under way the increase will be no more than one or two per cent, so this does not seem to be a serious disadvantage as far as the PAL and SECAM systems are concerned.

Another drawback was that the NTSC system did not lend itself too well to video tape recording and this was undoubtedly true in the early days, but here again its promoters claim that new tape recorders are being developed and will be available shortly, which will put the NTSC system on a level with its rivals.

From the technical point of view it seems that there is not a lot to choose between the three systems but the arguments for and against are likely to continue and it is almost certain that the March meeting will end with no decision again, as did the earlier meeting.

LOPAD
A Logarithmic On-Line Processing System for Analogue Data

By T. K. Cowell*, B.Sc.Eng., and M. Gordon*, B.A., M.Sc.

A computing system is described which carries out a wide range of algebraic operations directly on variables represented by electrical signals.

The principle used is to translate the signals into logarithmic form so that multiplication or division may be carried out by simple addition or subtraction operations.

Logarithmic conversion is obtained by utilizing the forward characteristic of a commercial semi-conductor diode. A technique has been devised which yields a computer performance independent of ambient temperature, despite the marked effect of the latter on the diode characteristics.

Operating directly on transducer signals the system can save much time in processing experimental results, and is currently finding a number of applications in the bio-medical field.

(Voir page 208 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 215)

THE past few years have seen considerable advances in the application of electronics to the bio-medical field, especially with regard to instrumentation. Nowadays, the continuous recording of blood pressure or heart rate for example, during a surgical operation or other treatment is commonplace, and in both experimental work and routine clinical investigation many other physiological quantities may be dealt with. This has resulted in a substantial increase in the amount of 'raw' biological data available, and has led to a situation where the efficient automatic recording of variables such as temperature, pressure or flow must often be followed by tedious point-by-point computation with a slide rule or desk calculator in order to obtain some physiologically significant function of the recorded quantities.

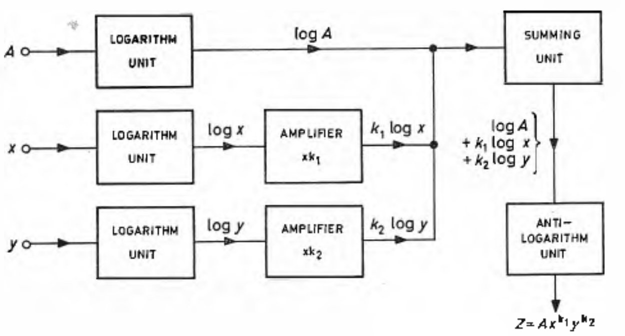

Fig. 1. Arrangement of logarithmic computer

Quite often the necessary computations involve repeated calculation of the products or quotients of variables raised to various powers and the computing system described here was developed with a view to reducing the time spent in this way. It operates directly on electrical signals derived from transducers, and permits on-line observation or recording of the required function of the variables.

Principle of Operation

The basic principle of the system is to translate the signals into logarithmic form so that the product or quotient operations may be effected by simple addition or subtraction.

For example, a computation such as:

$$z = Ax^{k_1}y^{k_2} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

may be written as:

$$z = \text{antilog}\,\{\log A + k_1 \log x + k_2 \log y\}$$

from which the computation can be seen to require the following operations:

(a) Obtain logs of A, x and y.

(b) Amplify $\log x$ and $\log y$ by factors k_1, k_2 respectively.

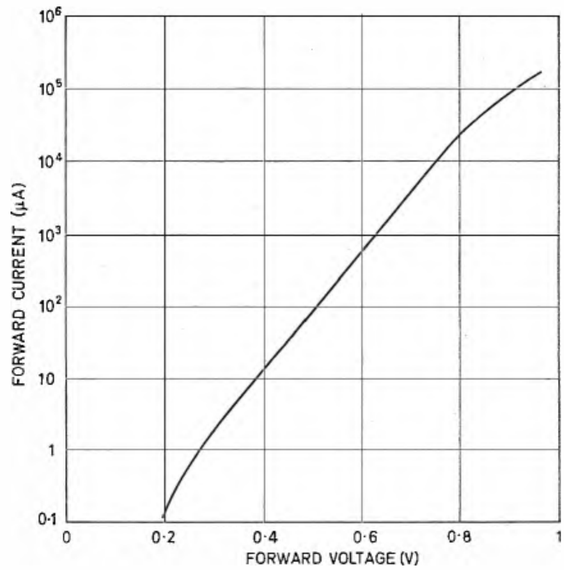

Fig. 2. Typical forward characteristic for SXL63 diode at a temperature of 25°C

(c) Sum quantities $k_1 \log x$, etc.

(d) Obtain antilog of sum.

A functional block diagram of a system carrying out these operations is shown in Fig. 1. Although operations (b) and (c) may be carried out using conventional summing techniques, the satisfactory performance of operations (a) and (d) requires circuits or circuit elements with an accurately logarithmic characteristic†. This can be achieved by using a conventional diode function generator, but a somewhat simpler solution is possible using a single semiconductor diode which has recently become available.

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† *Since the present work was undertaken, details have been published of a simple nine-decade logarithmic circuit*[3].

The Logarithmic Diode

Fig. 2 shows a typical forward characteristic for a Mullard SXL63 diode. Over the current range 3μA to 30mA the characteristic is linear to within 2 per cent when plotted on log-linear paper[2]. Over the range 10μA to 10mA or over any two decades of the full current range, the logarithmic slope is linear to within 1 per cent.

Analytically, the diode characteristic can be written[2] as:

$$V_f = V_o + T\{-S_o + S_1 \ln (I_f/I_o)\} \quad \ldots\ldots\ldots (2)$$

where V_f = forward voltage

I_f = forward current

T = junction temperature, (°K)

and V_o, S_o, S_1, I_o, are constants.

logarithmic diodes, it is necessary to feed the diodes from current sources proportional to the variables x and y. The logarithmic voltages across these diodes must then be summed by circuits which have a negligible effect on the diode currents, and the resultant voltage must be imposed on a final diode, whose forward current will then be a measure of the antilogarithm of the voltage. Again, this current must be measured by a device which causes only a negligible voltage drop in series with the antilogarithmic diode.

The way in which these requirements have been met is by using the diodes in operational amplifier configurations of the types shown in Figs. 3(a) and (b).

In the ideal case of infinite amplifier voltage gain, the amplifier input terminals become virtual earths, remaining

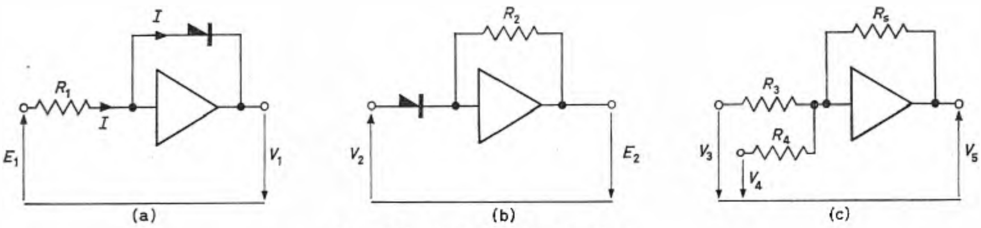

Fig. 3. Practical operational units
(a) Logarithm unit (b) Antilogarithm unit (c) Summing unit

This empirical expression approximates the diode characteristic (within the working current range), and is an adaptation to the SXL63 diode of the fundamental theoretical formula:

$$I_f = I_s\{e^{qV/kT} - 1\}$$

where I_s is the diode saturation current and the forward current is large enough for the unity term to be ignored.

In equation (2), the constants V_o and S_1 have approximately the values 1 200mV and 0·17mV/°K respectively. (This value for S_1 means that at room temperature V_f changes by about 115mV for each one-decade change in I_f.)

The constants I_o and S_o are interdependent, however, and one of them can be given an arbitrary value. The purpose of I_o is to make the quantity I_f/I_o, of which the logarithm is required, dimensionless. If I_o is chosen to be 1mA, S_o takes approximately the value 1·95mV/°K.

From Fig. 2 and equation (2) it can be seen that within the limits of current and accuracy stated, the forward voltage of the SXL63 diode is linearly related to the logarithm of the forward current. It follows that the forward current constitutes a similar measure of the antilogarithm of the forward voltage. Unfortunately, the relationship involves additional terms independent of these variables which are, moreover, temperature-dependent. Furthermore, because of production spreads, the constants of equation (2) will vary between one diode and another. (However, the interrelation of S_o and I_o means that the latter can be chosen to be the same for all diodes under consideration, the spreads affecting S_o only.) All of the above factors are important in the design of a practical computer, and in what follows it will be seen how they are accounted for. Before this, the way in which the diodes are used in practical logarithm and antilogarithm units will be explained.

Use of Operational Amplifiers

In order to carry out the operations of Fig. 1 using

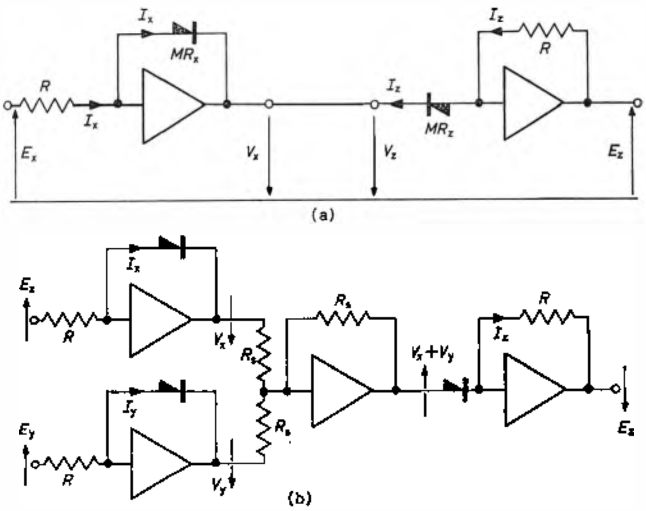

Fig. 4 (a). Simplest combination of operational units
(b). Basic arrangement for multiplication

at zero potential and drawing no current. In Fig. 3(a), the diode is then fed by an effective current source $I = E_1/R_1$. The output voltage of the amplifier is equal to the forward voltage of the diode at this current, and so gives an accurate measure of $\ln (E_1/R_1)$. Similarly, in Fig. 3(b), the diode is fed by an effective voltage source V_2, and the corresponding forward current passes through R_2, giving rise to an output voltage E_2. The forward current is thus E_2/R_2, which gives an accurate measure of antilog (V_2).

In Fig. 3(b), E_2 (or the current through R_3) is linearly related to the antilogarithm of V_2, but the circuit presents a non-linear load to the signal source. Thus, the latter must appear as a voltage generator (i.e. have low output impedance) if true antilogarithmic operation is to be obtained.

The circuit of Fig. 3(a) presents a linear load to the signal source so that logarithmic operation is not dependent upon source resistance. Note that in 3(a) either voltage or current can be considered as the input variable, while in 3(b) voltage or (feedback) current may be taken as the output variable. A more detailed analysis of these circuits is given in Appendix (2).

The necessary summing of logarithmic voltages can be achieved by the conventional circuit of Fig. 3(c). Providing the amplifier voltage gain is effectively infinite, the output voltage of the amplifier is given by:

$$V_5 = (R_5/R_3)V_3 + (R_5/R_4)V_4$$

so that any required multiplying factors for the logarithmic voltages can be obtained by adjustment of R_5/R_3 and R_5/R_4. The arrangement can of course be extended to sum a number of inputs, and since its output impedance is inherently low, it can be designed to be suitable for direct connexion to the antilogarithmic unit of Fig. 3(b).

identical characteristics (i.e. $V_{ox} = V_{oz}$ etc.) the current independent terms cancel, leaving

$$\ln (I_x/I_o) = \ln (I_z/I_o)$$

or:

$$I_x = I_z$$

whence $E_x = E_z$

If now a product computation of the type $z = xy$ is considered, where the variables x, y and z are represented by voltages E_x, E_y, and E_z respectively, a second logarithmic voltage V_y corresponding to E_y must be added to V_x before the antilogarithm operation is carried out (see Fig. 4(b)).

In this case, repeating the previous argument, the result will be:

$$(V_o - TS_o + TS_1 \ln (I_x/I_o)) + (V_o - TS_o + TS_1 \ln (I_y/I_o)) = V_o - TS_o + TS_1 \ln (I_z/I_o)$$

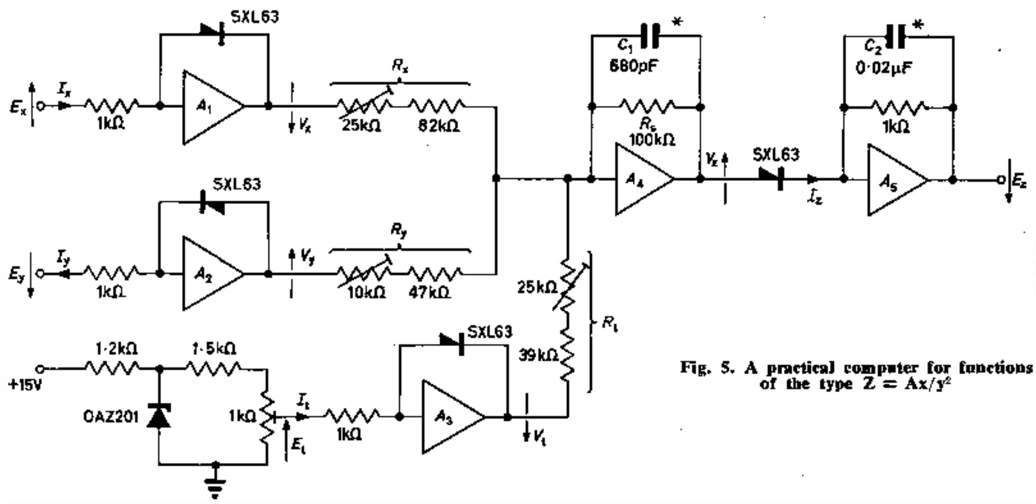

Fig. 5. A practical computer for functions of the type $Z = Ax/y^2$

For computations involving quotients, for example $z = (x/y) = xy^{-1}$, the required negative sign for log y can be obtained by adding a unit gain sign-reversing amplifier to the logarithmic unit. Alternatively, if a y signal of opposite polarity to the x signal is available, then it is only necessary to reverse the sense of the y diode.

The circuits of Fig. 3 constitute the basic units necessary for the construction of a practical computer, but before assembling them the ways in which some of the factors outlined in the previous section are accounted for will be explained.

Effects of V_0, S_0 and T

The circuit of Fig. 4(a) shows the simplest possible way of using the operational units, i.e., in converting a signal to logarithmic form and back again. Essentially E_x causes a current $I_x = E_x/R$ in diode MR_x so that the output voltage of the logarithm unit is, by equation (2):

$$V_x = V_{fx} = V_{ox} - T_x S_{ox} + T_x S_{1x} \ln (I_x/I_o)$$

This voltage is applied directly to the antilogarithm unit diode MR_z the forward current of which again satisfies the voltage/current relation of equation (2), i.e.

$$V_x = V_z = V_{fz} = V_{oz} - T_z S_{oz} + T_z S_{1z} \ln (I_z/I_o)$$

Therefore, provided that the diodes operate at the same temperature T,

$$V_{ox} - TS_{ox} + TS_{1x} \ln (I_x/I_o) = V_{oz} - TS_{oz} + TS_{1z} \ln (I_z/I_o)$$

Assuming for the time being that the diodes used have

which reduces only to:

$$V_o - TS_o + TS_1 \ln ((I_x/I_o) . (I_y/I_o)) = TS_1 \ln (I_z/I_o)$$

This relationship is dependent not only upon the diode constants V_o etc., but also upon temperature. However, if the forward voltage V_t of an additional diode carrying a steady current I_t is subtracted from the left-hand side of this equation, the relationship becomes

$$V_o - TS_o + TS_1 \ln ((I_x/I_o) . (I_y/I_o)) - (V_o - TS_o + TS_1 \ln (I_t/I_o)) = TS_1 \ln (I_z/I_o)$$

which reduces to

$$\frac{I_x \, . \, I_y}{I_t} = I_z$$

The use of this additional diode thus makes the computation independent of both temperature and the diode constants V_o, S_o, S_1 and I_o, although the output current is now a function of the current I_t. However, writing the relationship as

$$I_z = (1/I_t) \, I_x . I_y$$

it can be seen that $(1/I_t)$ constitutes an adjustable constant, which can be made to correspond to the constant A of equation (1). For example, if the currents are measured in milliamperes and I_t is set to 1mA the numerical relation

$$[I_z] = [I_x] \, [I_y]$$

is obtained, where the brackets [ ] denote dimensionless numbers (of milliamperes) and are used only to avoid the dimensional inconsistency which would arise in equating mA to mA × mA. If $R = 1000\Omega$ the corresponding voltage expression is

$$[E_z] = [E_x]\,[E_y]$$

where the brackets denote numbers of volts. Alternatively, for $I_t = 0{\cdot}1$mA, the relation becomes

$$[I_z] = 10\,[I_x]\,[I_y] \quad \text{(mA)}$$
$$\text{or} \quad [E_z] = 10\,[E_x]\,[E_y] \quad \text{(volts)}$$

The validity of the above method for obtaining operation independent of diode constants and temperature has rested on the assumption of identical diodes, and was for simplicity explained in terms of a simple computation $z = xy$. However, it can be shown theoretically (see Appendix 1) that essentially the same results may be obtained for any computation of the form $z = Ax_1^{k_1}x_2^{k_2} \ldots x_n^{k_n}$, by adding an appropriate multiple of V_t to the logarithmic sum. Furthermore, the technique is equally applicable where (as is usual) the values of V_o, S_o, etc., differ from one diode to another.

To show how the method works in practice, the design of a computer for a specific task will be explained.

A Practical Computer for $z = Ax/y^2$

Fig. 5 shows the circuit arrangement for a system computing functions of the form $z = Ax/y^2$ familiar to hydraulic engineers in the evaluation of nozzle coefficients, and of medical significance in the study of bladder function.

In the experimental model, the four logarithmic diodes were mounted in a small electrically heated stirred oil bath. This ensured that they all operated at sensibly the same temperature, at the same time enabling the actual operating temperature to be changed. During experiments, two different types of operational amplifier were tried—the Philbrick P65A and Fenlow A2. Both gave satisfactory logarithmic conversion, but the output capability of the former limited its associated diode current to about 2mA. The Zener diode used for stabilizing the E_t signal was chosen to have a low voltage/temperature coefficient.

Analysis of Operation

In analysing the system it is convenient to consider the currents I_x, I_y and I_z as representing the variables x, y and z, although in practice the signal inputs and output are voltages. Any derived current relationship may be directly converted to one involving voltages since $E_x = I_xR$, $E_z = I_zR$, etc.

Referring to Fig. 5, the summing unit output voltage can be identified as before with the antilogarithm unit input voltage, so that

$$V_z = M_xV_x - M_yV_y + M_tV_t$$

where

$$M_x = R_s/R_x,\ M_y = R_s/R_y,\ M_t = R_s/R_t$$

On substituting for V_x, V_z, etc., using equation (2) with appropriate suffices, this becomes

$$V_{oz} + T\left(-S_{oz} + S_{1z}\ln(I_z/I_o)\right) = \left\{\begin{array}{l} M_x[V_{ox} + T(-S_{ox} + S_{1x}\ln(I_x/I_o))] \\ -M_y[V_{oy} + T(-S_{oy} + S_{1y}\ln(I_y/I_o))] \\ +M_t[V_{ot} + T(-S_{ot} + S_{1t}\ln(I_t/I_o))] \end{array}\right\}$$

By transposing the current-independent terms on the left-hand side over to the right, dividing through by TS_{1z} and taking antilogarithms, this becomes:

$$I_z/I_o = \frac{(I_x/I_o)^{M_x(S_{1x}/S_{1z})}}{(I_y/I_o)^{M_y(S_{1y}/S_{1z})}}\,(I_t/I_o)^{M_t(S_{1t}/S_{1z})}\,e^{(\alpha+\beta)} \quad \ldots\ldots (3)$$

where:

$$\alpha = (1/TS_{1z})\,(M_xV_{ox} - M_yV_{oy} + M_tV_{ot} - V_{oz}) \quad \ldots\ldots (4)$$

and:

$$\beta = (1/S_{1z})\,(-M_xS_{ox} + M_yS_{oy} - M_tS_{ot} + S_{oz}) \quad \ldots\ldots (5)$$

The rather formidable-looking equation (3) may be simplified to yield the required relation between I_x, I_y and I_z by first establishing the x/z and y/z power laws, and then adjusting the independent variable M_t to eliminate the only temperature-dependent quantity α.

The directly proportional relationship required between I_x and I_z is obtained by adjustment of R_x, so that

$$M_x\,(S_{1x}/S_{1z}) = 1$$

Similarly for the inverse-square law relation between

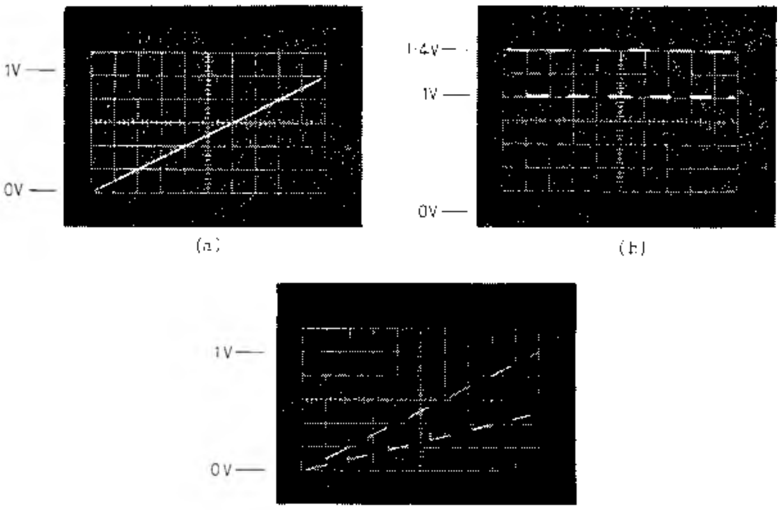

Fig. 6. x/y^2 computer signal waveforms

(a) x-input (b) y-input (c) z-input

I_y and I_z for which, by adjustment of R_y

$$M_y\,(S_{1y}/S_{1z}) = 2$$

Because S_{1x}, S_{1y} and S_{1z} are not much different from each other, the required values for M_x and M_y are approximately 1 and 2 respectively.

Since V_{ox}, V_{oy}, etc., are again of similar value, the value of M_t required to make the temperature-dependent quantity α disappear will be approximately two (see equation (4)). Also, since S_{ox}, S_{oy} are roughly the same, this value for M_t results in a nearly-zero value for β as well (see equation (5)).

If now the exponents of I_o are collected together, then with the above values for M_x, M_y and M_t, equation (3) becomes:

$$I_z = (I_x/I_y^2)\,.\,I_t^p\,I_o^q\,e^{\beta} \quad \ldots\ldots (3a)$$

where

$$p = M_t\,(S_{1t}/S_{1z}) \simeq 2$$
$$q = (2 - 1 + 1 - p) \simeq 0, \text{ so that } [I_o]^q \simeq 1$$

and $\beta \simeq 0$, so that $e^{\beta} \simeq 1$.

Since I_t (via E_t), may be adjusted freely, it can be set to give unity (or any other) constant of proportionality between the left- and right-hand sides of equation (3a), and in general

$$I_z = A\,.\,(I_x/I_y^2) \quad \ldots\ldots (6)$$

or in terms of voltages,

$$E_z = A'\,(E_x/E_y^2) \quad \ldots\ldots (7)$$

where $A' \simeq E_t^2$

Provided that all of the voltages, including E_t, are measured in the same units, equation (7) is always valid. Alternatively, E_t may be interpreted as defining the 'computing unit' for the system: this approach is discussed in Appendix (1).

The operations for setting-up the computer are not difficult in practice and may be summarized as follows:

(a) Initially set $R_x = R_s$, $R_y = \frac{1}{2}R_s$, and $E_t = 1V$

(b) Keeping the diodes at constant temperature and with E_y constant, trim R_x to obtain a directly-proportional relation between E_x and E_z.

(c) With the diodes at constant temperature and E_x constant, trim R_y to obtain an inverse-square relation between E_y and E_z.

(d) Trim R_t so that when the diodes are subjected to a temperature change, no alteration of output voltage E_z occurs.

(e) Adjust E_t for any required constant of proportionality between inputs and output.

Results

Fig. 6 shows some actual test signal waveforms for the circuit of Fig. 5 with E_t adjusted so that 1V at both inputs results in 1V output. The x input is a linear ramp of amplitude 1V, period approximately 50msec. The y signal has a castellated waveform, switching from 1V to $\sqrt{2}$V with a period of approximately 10msec. The output voltage $[E_x]/[E_y]^2$ therefore switches between two ramps, of amplitudes 1V and approximately $\frac{1}{2}$V: Fig. 6(c) gives an indication of the accuracy with which the computation is carried out.

The computer switching time (to 1 per cent of final value) is about 0·5msec, and is determined mainly by the capacitors C_1 and C_2 in Fig. 5. These capacitors were inserted to overcome high frequency instability in the computer, most probably due to the combined effects of phase shift in the operational amplifiers and the diode self-capacitances. The question of stability is being investigated further, and it is hoped that a more sophisticated solution will yield computing times considerably shorter than at present obtained. The limit is most probably set by the time-constant of the logarithmic diode itself, which is a rather complicated function of the signal amplitude, since both diode (chord) resistance and junction capacitance vary with forward voltage.

Conclusions

Results so far obtained with the experimental model are extremely encouraging, and a multi-purpose two-variable computer with switched function selection is under construction. Computing accuracy is dependent upon the type of calculation and the range of the variables. For computations where no variable (including the output) changes by more than two decades, the accuracy has been found to be better than 2 per cent.

A limitation of the simple arrangement described lies in its inability to deal with an input signal whose sign changes during computation, thereby causing reversed polarity of the associated logarithmic diode. In applications where this sign change could not be avoided, one solution might be to rectify the signal, and arrange for automatic sign-changing of the output.

Although developed primarily for simple data processing, the system could be useful in the construction of electronic analogue models. For example, a simple xy product circuit might with advantage replace the conventional quarter-squares or servo-multiplier in some applications. A further possibility is the computation of the $z=x^y$ type of function, by the use of two logarithm units in cascade, followed by two anti-logarithm units.

The authors believe that the basic LOPAD system could be developed to provide a general purpose computer costing not much more than a good oscilloscope, which would constitute a valuable laboratory tool. (In this connexion, a simple transistorized analogue store circuit has been developed which would extend the usefulness of the system in such applications.) It might in many instances provide an alternative to processing by digital computer, avoiding the necessity for programme-writing, analogue-to-digital conversion, etc.

Appendix

(1) Analysis of a Generalized Computer System

Consider input currents $I_1, I_2 \ldots I_i \ldots I_t$, logarithmically combined via summing unit voltage gains $M_1, M_2 \ldots M_i \ldots M_t$. I_t is held constant, for scaling purposes. The antilogarithm current I_a will then be given by:

$$V_{oa} - TS_{oa} + TS_{1a} \ln (I_a/I_o) = \sum_{i=1}^{t} M_i V_{oi} - \sum_{i=1}^{t} M_i TS_{oi} + \sum_{i=1}^{t} M_i TS_{1i} \ln (I_i/I_o) \quad \ldots\ldots \quad (8)$$

Rearranging:

$$I_a = I_o^{\left(1 - \sum_{i=1}^{t} (M_i S_{1i}/S_{1a})\right)} \left(\frac{\sum_{i=1}^{t} M_i V_{oi} - V_{oa}}{TS_{1a}} + \frac{S_{oa} - \sum_{i=1}^{t} M_i S_{oi}}{S_{1a}} \right) I_t^{M_t(S_{1t}/S_{1a})} \prod_{i=1}^{t-1} (I_i^{M_i(S_{1i}/S_{1a})}) \quad \ldots\ldots\ldots \quad (9)$$

where Π represents a continued product.

If the required computation is of the form:

$$I_1^{k_1} I_2^{k_2} \ldots I_i^{k_i} \ldots I_t^{k_{t-1}} = \prod_{i=1}^{t-1} I_i^{k_i}$$

then in equation (9), the M_i must be adjusted so that:

$$M_i (S_{1i}/S_{1a}) = k_i \ (i < t)$$

or:

$$M_i = k_i (S_{1a}/S_{1i}) \quad (i < t) \quad \ldots\ldots\ldots\ldots \quad (10)$$

$$\simeq k_i \text{ for normal production spreads.}$$

In equation (9), if all currents are measured in, say, milliamperes, the term in I_o can be made unity by choosing $I_o = 1$mA. The term in e becomes independent of temperature if:

$$\sum_{i=1}^{t} M_i V_{oi} = V_{oa} \quad \ldots\ldots\ldots\ldots \quad (11)$$

or:

$$\sum_{i=1}^{t} M_i \simeq 1 \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (12)$$

This requires that:

$$M_t = \frac{V_{oa} - \sum_{i=1}^{t-1} M_i V_{oi}}{V_{ot}} \quad \ldots\ldots\ldots\ldots \quad (13)$$

$$\simeq 1 - \sum_{i=1}^{t-1} M_i \simeq 1 - \sum_{i=1}^{t-1} k_i \quad \ldots\ldots\ldots \quad (14)$$

Under conditions (10) and (13), with suitable choice of I_o, equation (9) reduces to:

$$I_a = (Q\, I_t^{M_t(S_{1t}/S_{1a})}) \prod_{i=1}^{t-1} I_i^{k_i} \quad \ldots\ldots\ldots\ldots \quad (15)$$

where:

$$Q = e^{\frac{S_{oa} - \sum_{i=1}^{t} M_i S_{oi}}{S_{1a}}} \simeq 1 \quad \ldots\ldots\ldots\ldots\ldots \quad (16)$$

since, by equation (12):

$$S_{oa} \simeq \sum_{i=1}^{t} M_i S_{oi}$$

The bracketed factor is analogous to A of equation (1), and may be varied by adjustment of I_t. I_t can also be considered to define the 'computing unit' of the system, since equation (15) is approximately:

$$I_a \simeq I_t^{(1-\sum_{i=1}^{t-1} M_i)} \prod_{i=1}^{t-1} I_i^{k_i}$$

or:

$$I_a/I_t \simeq \prod_{i=1}^{t-1} (I_i/I_t)^{k_i}$$

Thus the system performs purely numerical computations on the I_i $(i < t)$, providing the I_i are measured in units of approximately I_t. The 'computing unit' can be set exactly to any required value by adjusting I_t, e.g. the application of $100\mu A$ at all $(t-1)$ computing inputs can be made to give $I_a = 100\mu A$ exactly, when $I_t \simeq 100\mu A$.

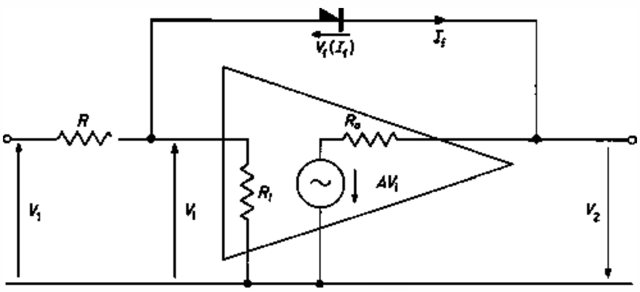

Fig. 7. Logarithm unit

so that the multiplier can now be controlled by the ratio of I_t and I_{t+1}.

(2) Analysis of Logarithm and Antilogarithm Units

Consider a logarithm unit with amplifier parameters as shown in Fig. 7; the diode voltage is written as $V_f(I_f)$.

In the equivalent circuit of Fig. 8, a fictitious diode of voltage $V_f(I_f)/(1 + A)$ precedes an ideal amplifier. It can be shown that I_f and V_2 are identical with those of the original circuit. Fig. 9 shows the circuit further transformed by Thèvenin's theorem.

For accurate logarithmic operation, the diode voltage must be negligible compared to resistive voltage drops, so that I_f is proportional to V_1, i.e.:

$$\frac{V_f(I_f)}{1+A} \ll I_f \left[R\|R_i + (R_o/A) - \frac{R_o}{A(1+A)}\right]$$

or:

$$Z_c \ll (1 + A)(R\|R_i) + R_o \quad \ldots\ldots\ldots\ (19)$$

where $\|$ represents parallel combination, and:

$$Z_c = V_f(I_f)/I_f = \text{diode chord impedance.}$$

Also, the voltage drop across the series negative resistance must be negligible compared to the diode voltage, so that V_2 is proportional to V_f, i.e.

$$\frac{V_f(I_f)}{1+A} \gg \frac{I_f R_o}{A(1+A)}$$

or:

$$Z_c \gg (R_o/A) \quad \ldots\ldots\ldots\ (20)$$

Finally, the circuit can be seen to operate independently of amplifier parameters if, additionally:

$$(1 + A)(R\|R_i) \gg R_o \quad \ldots\ldots\ldots\ (21)$$

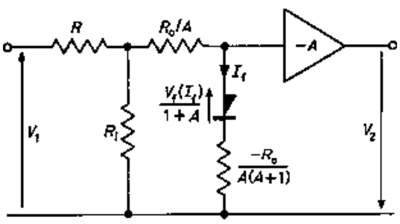

Fig. 8. Equivalent circuit of Fig. 7

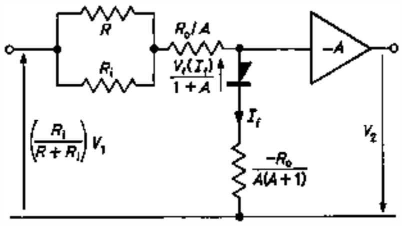

Fig. 9. Circuit transformed by Thevenin's theorem

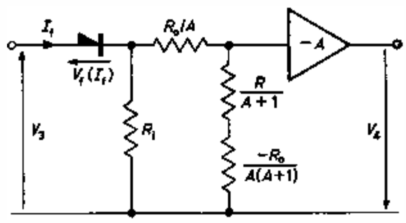

Fig. 10. Equivalent circuit of an antilogarithm unit

A special case occurs for computations with $\sum_{i=1}^{t-1} k_i \simeq 1$. Then, from equation (14), $M_t \simeq 0$ and in equation (15), I_t has only very weak control of the multiplier constant. In this case, two compensating diodes are necessary, carrying constant currents I_t, I_{t+1}. In equation (8), the summations now extend from $i = 1$ *to* $i = (t + 1)$, and the condition for vanishing temperature dependence becomes:

$$M_t V_{ot} + M_{t+1} V_{o,t+1} = V_{oa} - \sum_{i=1}^{t-1} M_i V_{oi} \quad \ldots\ldots\ (17)$$

or:

$$M_{t+1} \simeq -M_t$$

Providing equation (17) is satisfied, M_t, M_{t+1} may have arbitrary values. I_a is then given by:

$$I_a = Q' I_t^{M_t(S_{1t}/S_{1a})} I_{t+1}^{M_{t+1}(S_{1,t+1}/S_{1a})} \prod_{i=1}^{t-1} I_i^{k_i} \quad \ldots\ (18)$$

where $Q' = e^{\frac{S_{oa} - \sum_{i=1}^{t+1} M_i S_{oi}}{S_{1a}}} \simeq 1$.

From equations (17) and (18):

$$I_a \simeq Q'(I_t/I_{t+1})^{M_t} \prod_{i=1}^{t-1} I_i^{k_i}$$

and:

$$A \gg 1 \quad \ldots\ldots\ldots\ (22)$$

Conditions (19) and (20) are very convenient for worst-case design, since extreme values of Z_c for any particular application may be found from the V_f/I_f characteristic.

Fig. 10 shows the equivalent circuit of an antilogarithm unit, obtained from Figs. 7 and 8 by interchanging resistor R and the diode. In this case, resistive voltage drops must be negligible compared to V_f, so that:

$$V_f(I_f) \gg I_f \left\{ R_i \,\|\, \left[\frac{R}{1+A} + (R_o/A) - \frac{R_o}{A(1+A)}\right]\right\}$$

or:

$$Z_c \gg R_i \,\|\, \left(\frac{R + R_o}{1 + A}\right) \quad \ldots\ldots\ldots\ (23)$$

Finally, circuit operation is independent of amplifier parameters, if:

$$(1 + A) R_i \gg R + R_o$$
$$R \gg R_o/A$$
$$A \gg 1$$

The circuit of Fig. 10 must of course be fed from a source of impedance $\ll Z_c$.

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A Crystal Controlled Stimulator

By J. A. Reynolds*, A.M.I.E.R.E.

A double-pulse stimulator has been developed in which the timing between the output pulses may be varied in increments of 0·1msec over a total range of one second. Timing accuracy is ensured by the use of a crystal oscillator and a number of commercially available Dekatron units.

(Voir page 208 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 215)

MANY electrophysiological experiments require the use of a double pulse stimulator, and in some applications the delay time between the two pulses must be determined with a high degree of precision.

The use of crystal oscillators coupled with suitable scaling units is a well known technique in obtaining precise time measurements and allows a wide range of time intervals to be selected.

Using such a system an instrument employing a number of commercially available sub-units (Bendix Ericsson Ltd) enabled an accurately timed pulse system to be constructed, reducing the development time considerably.

General Requirements

The stimulator, which was intended for pulmonary studies was designed to generate a pair of voltage stimulus pulses whose interval between the leading edges could be varied in increments of 0·1msec over a total range of 1sec.

Ten pre-set pulse widths ranging from 20μsec to 50msec selected by a rotary switch for each pulse generator were desired, and to be combined in a suitable output amplifier. Wherever possible, all variable parameters, such as pulse width, interval time etc., were to be selected by switches reducing the amount of time required in setting up the instrument and allowing the maximum concentration to be focused on the experiment.

Coarse and fine control of the output voltage and provision for isolated stimuli by means of a radio-frequency probe, were to be included.

The usual facilities to be found in general purpose stimulators were also required, such as continuous running or triggered operation, initiated manually or from external sources, where synchronism with other apparatus is necessary.

Principles of Operation

Fig. 1 shows the block outline of the complete instrument.

The crystal oscillator performs the basic timing and is fed via a switch into the general purpose input unit, whose output is controlled by a gate unit, and thence to the Dekatron divider chain consisting of four decade dividers. The output pulse on cathode nine of the final Dekatron occurs 0·9999sec after the commencement of opening the gate unit.

Operation of the gate unit is by positive going pulses, these being start, stop and reset respectively. The number of pulses fed into the Dekatron chain is determined by the time interval between the start and stop pulse.

In the continuous running condition as a double pulse generator, the gate opens the input unit and pulses at the oscillator frequency are fed into the Dekatron divider chain. Four of the Dekatron cathodes are selected by switches and combined in a coincidence circuit such that, when a glow discharge rests simultaneously on these selected cathodes, the coincidence circuit initiates a trigger which operates the second pulse generator. The output pulse of this generator may be varied in width by switching various timing capacitors in its circuit, and its output is fed to a mixing stage in the amplifier together with the output from the first pulse generator. Both pulse generators are identical in design and various combinations of pulse widths, time delay, and amplitudes can be combined in the mixing stage.

The second pulse is also taken to a differentiating network where after suitable shaping and amplification, the leading and trailing edges are used to provide the stop and reset pulses to the gate unit. A separate reset pulse is also generated to set the Dekatrons to zero. The gate unit reset pulse also triggers a delay unit of 10μsec from which the start pulse and trigger to the first pulse generator are derived, initiating the next cycle of operations. The leading edge of the first pulse generator coincides with the first of the oscillator pulses now being fed into the divider chain. The process will now repeat itself at some specific frequency.

The repetition frequency is determined by the reciprocal of the delay time and second generator pulse width. This ensures that, as the pulse width is altered, the repetition frequency adjusts itself to accommodate the full width of the second pulse and eliminates the need for locked p.r.f./pulse-width controls necessary in the usual *RC* timed stimulators.

A further advantage of this system is a direct indication of delay time under darkened conditions without reference to the switch settings. For example, if the delay time has been set to 0·3250sec, Dekatrons one and two will show full sweeps over all their cathodes, while the third Dekatron will be seen to count to five, pause for the time duration of the second pulse width, and then fly back to zero. The fourth Dekatron in this case will always rest at zero. During the time determined by the second pulse generator pulse-width, the selected cathodes are at rest and appear as a relative brightening with respect to the others, and hence the delay time is simply read in the dark by these brighter cathodes without reference to the switch settings.

For single-pulse operation the trigger to the first pulse generator is switched out of circuit, otherwise the functions are identical with those for double-pulse conditions.

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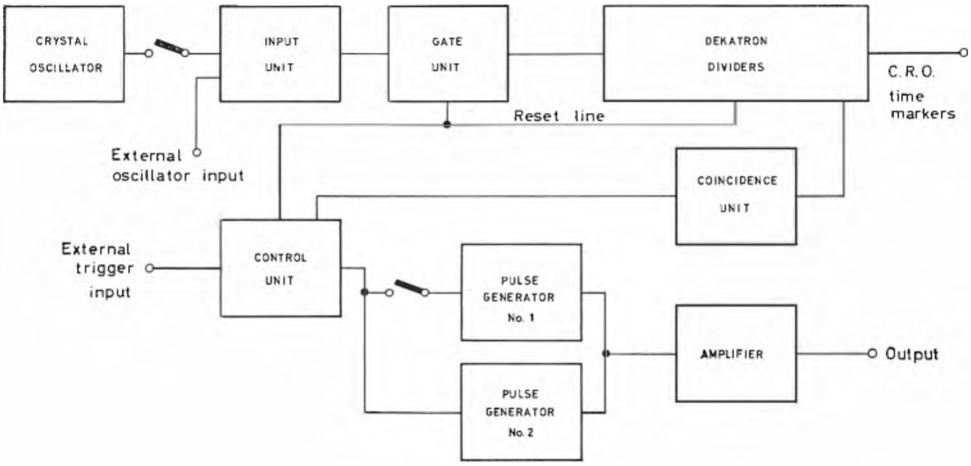


Fig. 1. General arrangement of the instrument

Single-shot conditions for double or single-pulses are obtained by deriving a suitable trigger from a resistance chain across the h.t. line and manually injecting this into the delay unit to produce the start pulse. An input socket to this delay unit also allows external triggers to be applied for synchronous operation with other apparatus.

The general purpose input unit may also be used independently from the gate unit and allows other modes of operation to be utilized. A fuller description of this unit and its use is described in the manufacturers' literature.

Circuit Description

The additional circuits to provide the various control functions and to generate suitable output pulses are shown in the following diagrams. Circuits of the units commercially available are not shown as these may be obtained from the manufacturers.

To ensure a high degree of accuracy and stability a 10kc/s crystal oscillator was built as a plug-in unit. The crystal for this oscillator is a Standard Telephone Co. Ltd. type 4414-NT, a suitable drive circuit being recommended by the makers.

The output of the oscillator is connected via a switch to the general purpose input unit which produces suitable pulses for driving the Dekatron divider chain. Switching out the oscillator enables external signals to be fed directly into the input unit. An amplifier is incorporated in the input unit and allows a range of external signals to be fed in independently of the oscillator over a range from 0·1 to 100V.

Each Dekatron unit is supplied as a complete plug-in decade counter. Resistors in the cathodes of the Dekatron tube are selected by a ten-way switch and brought out to a terminal block at the rear of the unit. The selector switch outputs are then combined in the coincidence unit.

Fig. 2 shows the coincidence circuit. V_1 is normally in the cut-off condition set by resistors R_{64} and R_{65}. The grid network comprises resistors R_{59}, R_{60} and R_{61}. The junction of resistors R_{60} and R_{61} is connected via four silicon diodes MR_7 to MR_{10} through the selector switches to the cathodes of the Dekatron tubes. R_{61} is effectively in parallel with these four selected resistors holding the grid of V_1 well below the cathode. When a glow discharge occurs at the cathode of a Dekatron, the potential across its associated resistor rises by some 45V; thus when all four selected cathodes coincide, valve V_1 is brought into conduction. The negative going pulse at the anode of V_1 is used to trigger the second pulse generator.

Fig. 3 is the circuit diagram of the control unit whose function is to provide the various synchronizing pulses and reset line voltages.

A positive pulse developed across R_{58} in the screen of the second pulse generator is inverted by the buffer amplifier V_2 and the negative going pulse at its anode is differentiated by C_2 and R_4. The leading edge of this differentiated pulse is transferred by MR_1 to the grid of V_3 whose positive going pulse at the anode is used as the stop pulse to the gate unit. MR_2 transfers the trailing edge to the flip-flop comprising V_4 and V_5. The pulse width at the output of V_5 is determined by C_6 and R_{10} and is set at 1msec to provide a suitable time interval for resetting the Dekatrons to zero. The output pulse at V_5 is coupled to the grids of V_5 and V_7.

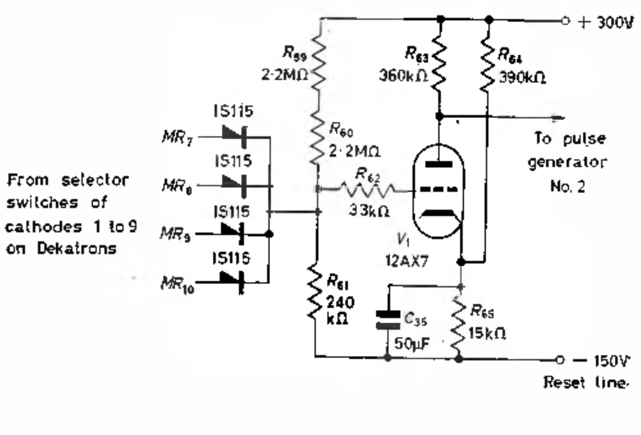
The resetting of the gate unit occurs at the trailing edge of the Dekatron reset pulse to ensure that all the Dekatrons are returned to zero before any further action of the circuit is initiated. V_7 and V_8, coupled as a trigger circuit, operate at the termination of the Dekatron reset pulse, the output pulse of V_8 being fed to the buffer valve V_9 providing the reset pulse for the gate unit.

The double triode V_{10} and V_{11} provides a delay of some 10μsec between its input trigger and the input to the cathode coupled phantastron V_{12}. The phantastron generates the start pulse and the trigger to the first pulse generator at the cathode and screen resistors respectively. A choice of three modes of triggering the start circuit is provided by switch S_2.

In the continuous running condition V_{10} grid is connected via C_{14} to the anode of V_9, thus the reset pulse initiates the start of a cycle of operations.

A single pulse or a double pulse is obtained by connecting the grid circuit of V_{10} to a positive potential formed by R_{32} and R_{33} across the h.t. line. In this mode of operation the selector switch S_2 has been spring loaded and will return to the external trigger position upon being released. A flexible lead and remote switch may be plugged in to

Fig. 2. Coincidence circuit



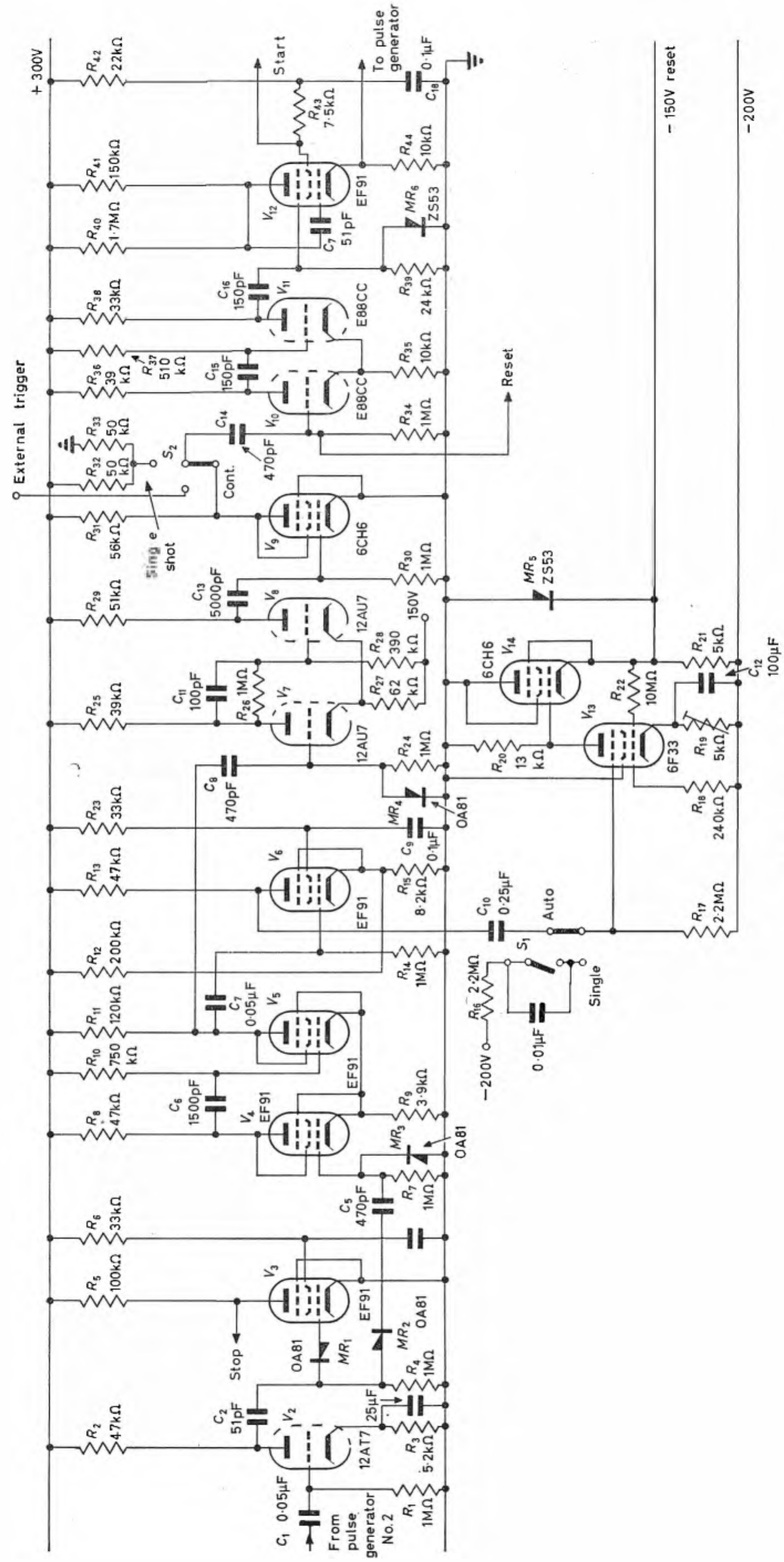


Fig. 3. The control unit

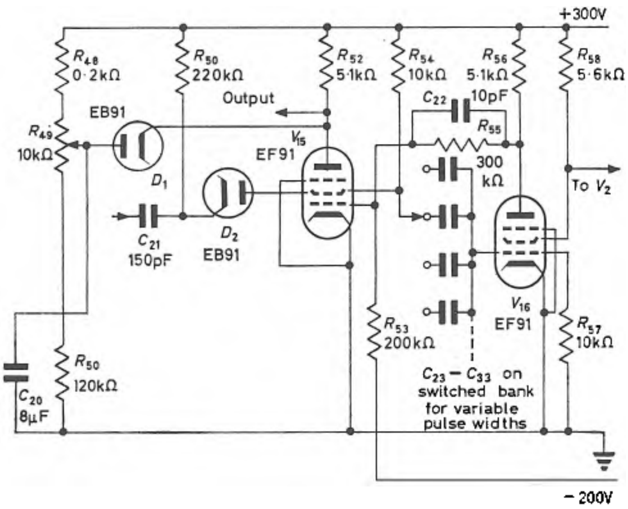


Fig. 4. The pulse generator

operate single-shot conditions some distance from the equipment.

Should the instrument be required to run in synchronism with an external source, a suitable pulse must be fed into the external trigger position. For example, the time-base of a c.r.o. can be used as a trigger to synchronize the display of stimulus and response.

V_5, V_{13} and V_{14} perform the dual purpose of providing a −150V negative line for the Dekatron dividers and also of switching this line to ground potential during the reset condition. To reset the Dekatrons to zero, the potential of the cathodes 1 to 9 is raised while those of 0 are left at −150V. Cathodes 1 to 9 are taken to the cathode of V_{14} while the 0 cathodes are taken to a fixed −150V line. V_{13} is a short suppressor grid pentode whose anode is directly coupled to the grid of V_{14}. Adjustment of R_{19} in the cathode of V_{13} enables the output of V_{14} to be set at −150V. When a negative pulse from V_6 anode is applied to the suppressor of V_{13} its anode rises to ground potential. V_{14} grid follows this rise and the reset line is taken positive, resetting the Dekatron tubes to zero. Manual resetting may also be accomplished by S_1.

Fig. 4 shows the circuit diagram of one of the pulse generators. Both pulse generators are identical in construction and function as one-shot multivibrators. Triggering is accomplished by means of a negative pulse applied to the screen resistor via diode D_2. Various pulse widths are selected by switching in different values of capacitance to cover the range required. Selection of the capacitors was accomplished by substituting a precision decade capacitor and measuring the pulse-width with an oscillograph and then selecting the appropriate value by use of a bridge. By this means, all pulse widths were held better than 1 per cent, this being adequate for the requirements specified. Should a fine control of pulse width be desired, it was found that by returning R_{57} to a variable positive potential by means of a potentiometer, continuous coverage over the whole pulse width range was possible.

The amplitude of the output pulse is varied by the slicing diode D_1 and R_{49}. Experience with the equipment has shown that a continuously variable control is not the best method of pulse amplitude control, and has now been replaced by a selector switch to give a fixed range of amplitudes.

A conventional d.c. coupled output amplifier is comprised of a mixer stage and cathode-follower. An attenuator in the output of the cathode-follower enables coarse and fine adjustment of the output voltage. To obtain an isolated stimulus, a conventional radio frequency probe is fed with the output pulses and then rectified to obtain the d.c. pulse. An output amplifier, with switched transformer loads for narrow and wide pulse widths, is under development for physiological investigations requiring this type of output.

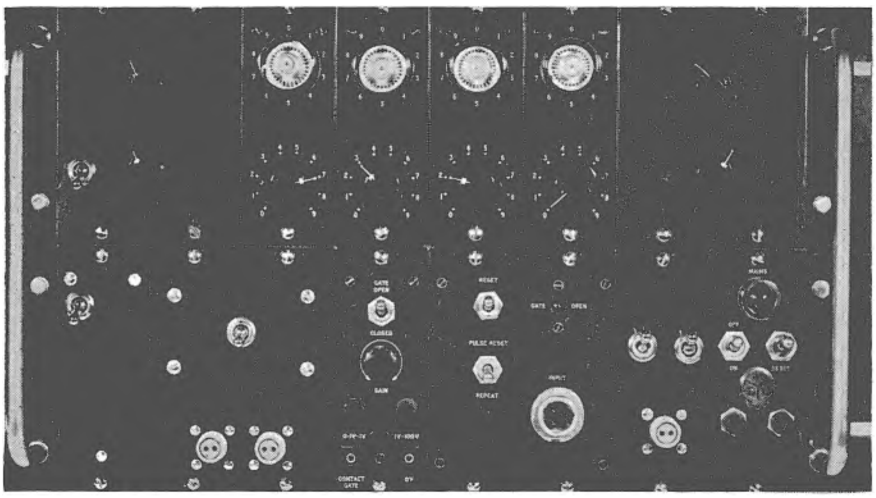
Where sub-units were commercially unobtainable, blank matching panels were obtained for the construction of these units.

The control unit (Fig. 3) is constructed as one complete chassis and is mounted lengthwise behind the Dekatron counters. A plug and socket with flying leads allows it to be easily removed in the event of servicing. The complete instrument is mounted in a double tier framework, a series of pre-pierced socket holes being provided for easy assembly and wiring. Fig. 5 shows a photograph of the front panel and controls.

Besides being used as a stimulator the instrument may be used as a counter with a maximum of ten thousand counts. Any count within this range may be pre-set by switches and an event marker in the form of a voltage pulse is obtained when the count has reached the pre-set number.

An ancillary unit enables the instrument to be used as a very accurate timer over the range of 0·1msec to 1sec. The timing range may be expanded by switching out the internal oscillator and using an accurate external source of lower frequency.

Fig. 5. The complete instrument



An electrical ruler for c.r.o. time calibration is available from the Dekatron cathodes.

The instrument has been found to be trouble free in operation for a number of years, and it provides a flexible instrument for use in the laboratory.

Acknowledgments

The author wishes to thank Dr. J. C. Gilson, Director of the Pneumoconiosis Research Unit, and the Medical Research Council for permission to publish this article.

The Metal-Oxide-Semiconductor Transistor

By R. A. Hilbourne* and J. F. Miles†

The metal-oxide-semiconductor transistor is a new semiconductor device, working on the field effect principle, but offering a much higher, temperature-stable, input resistance. The development has now reached the point where devices with stable characteristics are commercially available, and this article is intended to serve as an introduction to the device.

The construction of the m.o.s. transistor is described and the theory of its operation and its characteristics are given. This is followed by a description of several applications of the device both in linear circuits and as a switch.

(Voir page 208 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 215)

THE metal-oxide-semiconductor transistor is a new semiconductor device, working on the field effect principle, but offering a much higher, temperature-stable, input resistance. The development has now reached the point where devices with stable characteristics are commercially available, and this article is intended to serve as an introduction to the device.

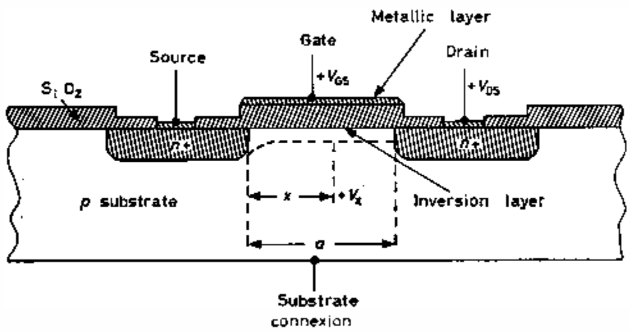


Fig. 1. Basic construction of metal-oxide-semiconductor transistor

The basic construction of a metal-oxide-semiconductor transistor is given in Fig. 1. These devices can be made on either p or n-type silicon giving complementary types of transistor. However, for simplicity, only those constructed on p-type silicon will be considered. As shown in the cross-sectional drawing, the transistor consists of a substrate of p-type silicon into which two n-type areas are diffused by conventional planar techniques. The separation between these two diffused areas, *a*, is small, but the length of the gap *b* is relatively large. Over the gap between the n+ areas, a thin silica layer is formed and on this is evaporated an aluminium layer.

Contacts are made to the substrate, the two n+ diffused regions and the isolated aluminium layer or gate. The operation of the device depends on the fact that the aluminium layer, the silica layer and the silicon form a parallel plate capacitor. Hence if a positive voltage is applied to the aluminium gate relative to the substrate then a positive charge will form on the aluminium and a negative charge in the surface of the silicon. The negative charge in the silicon is obtained by repelling any holes in the surface of the silicon and attracting electrons either from the n+ regions or those formed by thermal generation in the p-type silicon. The negative charge induced in the silicon will fill any positive states at the silica/silicon interface, and neutralize any positive impurity centres in the top surface of the silicon. Any excess electrons will form free carrier centres. Thus if a field is applied to the electrons along the surface a conduction path will occur. The transverse field can be obtained by applying a voltage between the two n+ regions. Thus if the right-hand region (or drain) is made positive with respect to the left-hand region (or source) there will be an electron flow from the source to the drain. Such a transistor in which conduction is by electrons is known as an n-m.o.s.t. The complementary type in which conduction is by positive holes is known as a p-m.o.s.t.

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Theory of Operation

To derive a simplified theory of the operation of this device cerain assumptions can be made. These are:

(1) All the electrons induced in the silicon are free carriers.

(2) The induced charge is determined only by the capacitance of the silica layer.

(3) The mobility of the electrons is constant.

In this case when a positive voltage V_{GS} is applied to the gate then the sheet resistivity of the induced charge layer, ρ_s is given by:

$$\rho_s = 1/Q\mu_n = \frac{1}{\mu_n C_o V_{GS}}$$

where μ_n is the mobility of the electrons and C_o is the capacitance per unit area of the silica layer.

Consider now the case where the source is shorted to the substrate and a positive voltage V_{DS} is applied to the drain, the voltage V_{DS} being less than V_{GS}. In this case a

SYMBOLS

a, b	=	width and length of the conduction channel.
ρ_s	=	resistivity of the conduction layer.
Q	=	the charge on the gate-substrate capacitor.
μ_n	=	electron mobility.
C_o	=	capacitance of unit area of the gate-substrate capacitor.
V_{GS}, V_{DS}	=	the gate and drain voltage with respect to the source.
β	=	the gain factor.
V_{GST}	=	the minimum voltage on the gate to cause conduction (also known as $V_{(P)GS}$).
I_{DSS}	=	drain current with gate shorted to source and substrate.
I_{DSX}	=	drain leakage current with gate biased more negatively than V_{GST}.
g_m	=	mutual conductance.

conducting skin will be induced in the surface of the silicon and current will flow between the source and drain. However, the current flow through the induced layer will result in a voltage drop along its length and the voltage across the dielectric layer at a point x from the source (see Fig. 1) will be $V_{GS} - V_x$. Hence the sheet resistivity of the induced layer at the point x will be:

$$\rho_s(x) = \frac{1}{\mu_n C_0 (V_{GS} - V_x)}$$

and $V_x = (I_{DS}/b) \int_0^x \rho_s(x)\, dx$

Solving these equations gives:

$$I_{DS} = \beta(V_{DS}V_{GS} - (V_{DS}^2/2)) \quad \ldots\ldots\ldots \quad (1)$$

where $\beta' = \mu_n C_0 (b/a)$

β is known as the gain factor.

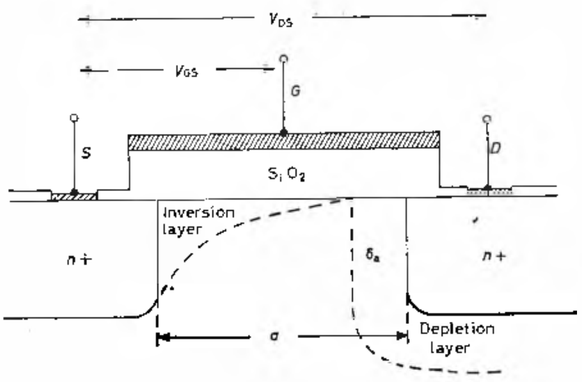


Fig. 2. Basic function of metal-oxide-semiconductor transistor

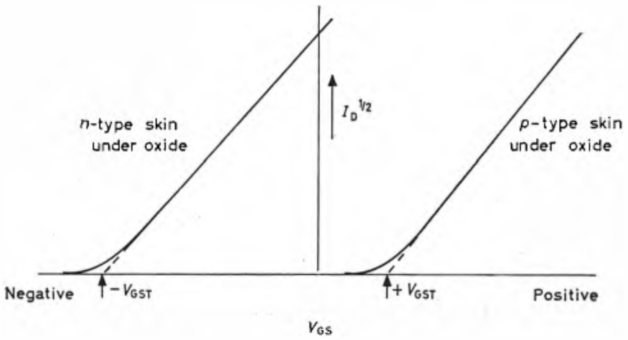


Fig. 3. Effect of n-type and p-type skin under oxide

Consider now the condition in which V_{GS} is zero and a voltage V_{DS} is applied to the drain (i.e. $V_{DS} > V_{GS}$). In this case no surface layer will be induced and the drain will simply be a reversed biased p-n junction, with a depletion layer of width δa around it. If now a voltage V_{GS} is applied to the gate then an inversion layer is formed on the p-type silicon until it reaches the depletion layer when $V_x = V_{GS}$ (see Fig. 2). However any electrons reaching the depletion region will be swept across to the drain. Under these conditions:

$$I_{DS} = (\beta'/2) \,.\, V_{GS}^2 \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

where $\beta' = \mu_n C_0 (b/a') = \mu_n C_0 \dfrac{b}{a - \delta a}$

where δa is the depth that the depletion layer spreads into the p-type silicon for a voltage $V_{DS} - V_{GS}$.

Expressions (1) and (2) are equal when $V_{GS} = V_{DS}$.

In a practical device it is possible to form an n-type or p-type skin in the top surface of the silicon under the oxide. With an n-m.o.s.t. if an n-type skin is formed under the oxide then a conducting path exists between the drain and source with zero gate voltage. This conducting path can be removed by applying a negative voltage to the gate. Conversely a p-type skin under the oxide will prevent conduction between the drain and source until a finite positive voltage has been applied to the gate. This effect is shown in Fig. 3. Thus expressions (1) and (2) must be modified to include the threshold or starting voltage V_{GST}.

Hence:

$$I_{DS} = \beta\,[V_{DS}(V_{GS} - V_{GST}) - (V_{DS}^2/2)] \quad \ldots\ldots \quad (3)$$

or:

$$I_{DS} = (\beta'/2)(V_{GS} - V_{GST})^2 \quad \ldots\ldots\ldots\ldots \quad (4)$$

Device Characteristics

The output characteristics of the Mullard m.o.s. transistor development type 95BFY are shown in Fig. 4. These are in close agreement with the expressions (3) and (4). This device has the conducting path at zero gate volts and can

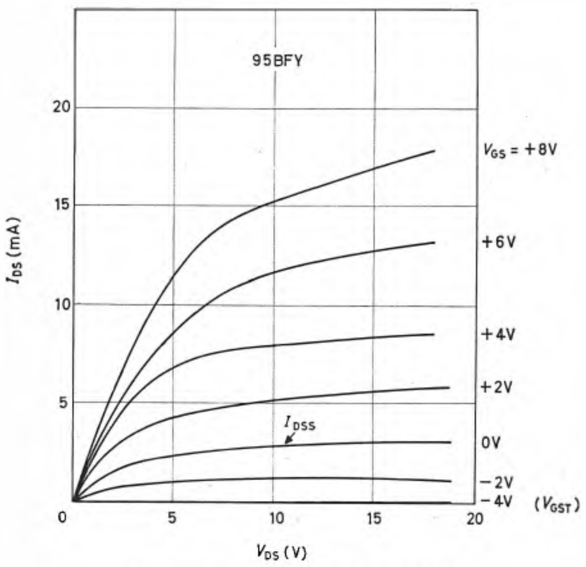


Fig. 4. Output characteristic of 95BFY

be cut-off by a negative voltage of about −4V. As expected the knee voltage is approximately $V_{GS} - V_{GST}$.

The device parameters can be summarized as follows:

Input Resistance:

This is the resistance between the insulated aluminium layer and the substrate in parallel with the surface leakage across the glass of the header. In practice the header leakage limits this to about $10^{14}\Omega$.

Output resistance:

The output resistance at voltages above the knee voltage can be obtained from expression (4). Thus:

$$r_{out} = 1 \Big/ \frac{1/\partial I_{DS}}{\partial V_{DS}} = a \Big/ \left(I_{DS} \,.\, \frac{\partial a}{\partial V_{DS}} \right)$$

Hence the output resistance decreases as the current increases. With a 95BFY, r_{out} is about 100kΩ at I_{DS} = 3mA.

At low values of V_{DS} where $V_{DS} < V_{GS} - V_{GST}$ expression (3) applies and:

$$R_{out} \simeq \frac{1}{\beta(V_{GS} - V_{GST})}$$

Mutual conductance

At voltages below the knee voltage there is a linear relationship between I_{DS} and V_{GS} (see expression (3)) and:

$$g_m = \beta V_{DS}$$

At higher drain voltages a square law characteristic is obtained and:

$$g_m = \beta'(V_{GS} - V_{GST})$$

or:

$$g_m = \sqrt{(2 I_{DS} \beta')}$$

Hence if the g_m is 2mA/V at $I_{DS} = 3$mA, $\beta' = 0{\cdot}66$mA/V^2.

Zero bias drain current:

Though not a basic parameter, the drain current for $V_{GS} = 0$ is a useful design parameter

$$I_{DSS} = (\beta/2)\, V_{GST}^2 \text{ for negative } V_{GST}.$$

For devices with positive V_{GST}, $I_{DSS} = I_{DSX}$, the drain to source leakage current.

Capacitances:

(a) C_{GS}: The gate to source capacitance is the parallel plate capacitance between the aluminium layer and the silicon. This has a value of about 3pF and is independent of bias conditions for normal operation.

(b) C_{DS}: The drain to source capacitance is mainly determined by the diffused p-n junction capacitance between the drain electrode and the substrate and as such has the normal voltage dependence; a typical value is 2·5pF at +20V. C_{DS} is independent of the drain current.

(c) C_{DG}: The feedback capacitance is due to the slight overlap of the aluminium gate layer over the drain electrode. This capacitance increases slightly with drain voltage since at the higher voltage the depletion layer spreads further under the gate.

Frequency limiting parameters:

The speed of operation of the device is ultimately limited by the rate at which the induced inversion layer can be established. From switching measurements using very low generator and load resistances it is found that the time to establish the conducting path is much less than 1nsec. In most practical applications the performance is limited solely by the capacitances given above.

Temperature dependence:

The mobility of charge carriers in the induced inversion layer decreases as the temperature increases. Thus in raising the temperature from 25°C to 100°C decreases the mobility to about 75 per cent of its original value. This variation in mobility produces a corresponding change in gain factor β. In addition the threshold voltage V_{GST} becomes approximately 1·5mV more negative for every 1°C rise in temperature.

The M.O.S. Transistor Amplifier

The theoretical analysis shows that the m.o.s. transistor differs from the bipolar transistor in being voltage driven, having a very high input resistance, both gate to source and gate to drain, giving it characteristics nearer to a valve than to a transistor. It is possible to use the device in most of the common circuit configurations replacing valves and transistors, and it has further possibilities due to the availability of the substrate connexion.

The simplest 'standard' circuit is the common cathode, or common emitter configuration. This becomes the common source circuit of Fig. 5(a). The device is biased to $I_D = I_{DSS}$. As the gate may be taken negative with respect to the source auto bias may be used (Fig. 5(b)), increasing the d.c. stability. For drain currents in excess of I_{DSS}, known as the enhancement mode, positive gate-source bias must be used, as in Fig. 5(c) or 5(d). In all cases the substrate must be connected to the source, to prevent forward biasing of the substrate junctions.

As the basic characteristic of the device is the square law $I_D = (\beta/2) \,.\, (V_{GS} - V_{GST})^2$, distortion would be expected. However, substitution of $V_{GS} = (V + v \sin \omega t)$ shows the second harmonic content to be $25v/(V - V_{GST})$ per cent. If $V = -2$ and $V_{GST} = -6$, then the harmonic content is $(25v/4) = 6v$ per cent, and v may be as high as 170mV before 1 per cent distortion is reached corresponding with possibly 6V output. The device may therefore be regarded as being linear except to very large signals.

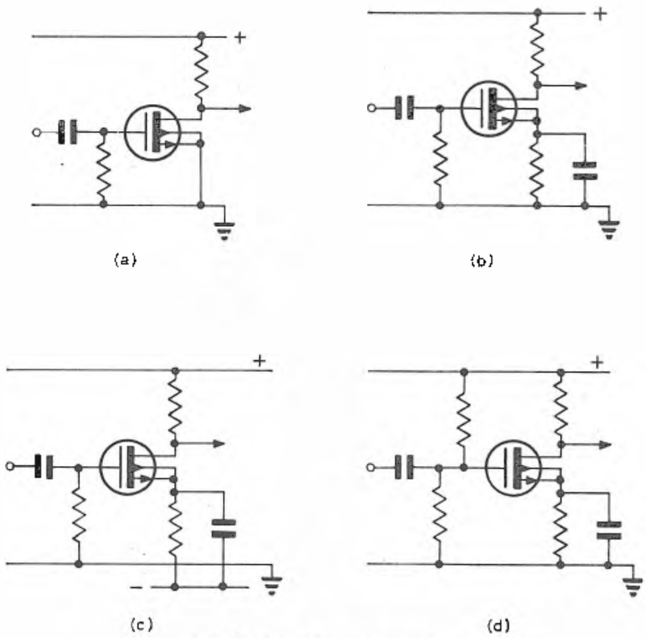


Fig. 5. Biasing arrangements

As a d.c. amplifier m.o.s. transistors may be used in long-tailed pair configurations. The temperature dependence of the characteristics is similar to that found in bipolar transistors and an improvement is found when pairs of devices are used. The degree to which pairs can be matched is being investigated.

The very high input resistance of the m.o.s. transistor suggests that its principal use will be as an impedance changer, the simplest of which is the cathode-follower type of circuit. Using the m.o.s. transistor this becomes the source follower of Fig. 6(a).

Comparing this with the equivalent valve circuit, the 95BFY has a g_m of greater than 1mA/V, which is of the same order; the maximum voltage, however, is lower, being limited to $V_{DS} = 35$V. This limits the current allowable with a given value of R_L, or vice-versa. Incremental gains of greater than 0·9 are obtained with $I_{DS} = 100\mu$A, but at $I_{DS} = 3$mA the gain falls to 0·85.

If higher current levels are desired the product $g_m R_L$ and hence the gain $A_v = g_m R_L/(1 + g_m R_L)$ may be increased by boosting the gain with a bipolar transistor as in Fig. 6(b). Using, for example, a BCY30 transistor with the 95BFY, gains of better than 0·9 are obtainable with a load current of 3mA. The impedance gain, if the biasing resistor is cancelled by Bootstrapping (Fig. 6(c)) can be $R_{in} = 10^9\Omega$ to $R_{out} = 200\Omega$. The limitation on the biasing resistor is that it must shunt the drain to gate current, which is less than 10^{-12}A.

Applications for this circuit are in charge amplifiers, for reading stored voltages in sample-and-hold or peak-reading circuits, for pH meters, etc. The limitation of its use with small signals is the generated noise, which is fairly high at low frequencies.

A Chopper Transistor

An interesting effect of the high resistance from the gate to both drain and source is that it makes the device extremely useful as a chopper transistor, as the current driven into the drain by the gate voltage when the device is switched off is of the order of $4V/10^{13}\Omega = 4 \times 10^{-13}$A, and is stable with varying temperatures compared with

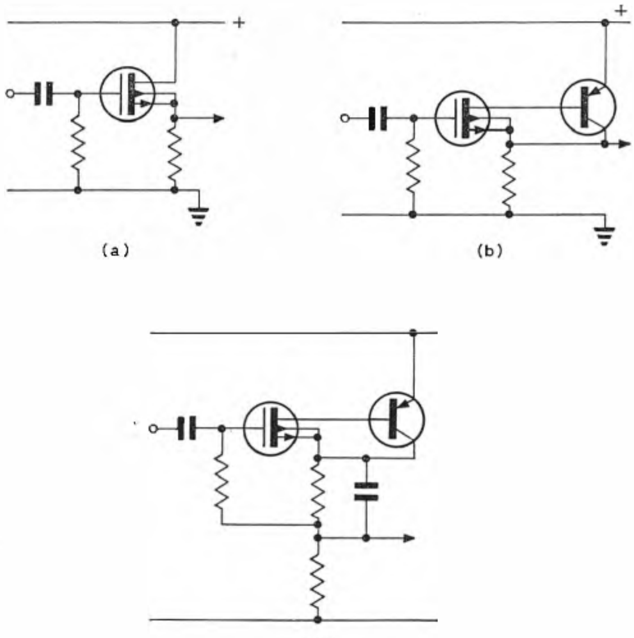


Fig. 6. Impedance changing circuits

the offset current in a bipolar transistor of the order of 2×10^{-9}A and exponential with change in temperature. A more important characteristic in the chopper configuration is that, as, in the 'on' state, the conduction is in an n-type layer between two n-type electrodes, there are no p-n junctions in the conduction path. While operating in the linear region of the characteristic the device has no depletion layer, and the conduction path is in a linear resistance, entirely eliminating the offset voltage found in all bipolar chopper transistors. The switching speed is limited by the transient ('spike') feedthrough of the gate waveform. This is due to the charging current of the drain to gate capacitance, and may be characterized by the capacitance charge $q = C_{dg} . V_{gs}$ where V_{gs} is the peak-to-peak gate voltage. If $V_{gs} = 6$V and $C_{dg} = 0{\cdot}5$pF, $q = 3$pC, compared with $2\text{V} \times 2\text{pF} = 4$pC for the best chopper transistors and $4\text{V} \times 8\text{pF} = 32$pC for the simple type of chopper transistor.

The ultimate sensitivity, at low chopping rates, using the m.o.s. transistor, is limited by the noise, which is lower in the linear part of the characteristic than in the 'square law' part, and 10μV d.c. signals have been amplified successfully.

The saturation resistance in the linear region of the characteristic may be seen from the operating equation:

$$I_{DS} = \beta\,[(V_{GS} - V_{GST})\;V_{DS} - (V_{DS}^2/2)]$$

to be

$$r_s = V_{DS}/I_{DS} = \frac{1}{\beta(V_{GS} - V_{GST})}$$

which is constant for changes in V_D, provided that V_{GS} is much greater than V_{DS}, which is the normal operating condition for a small signal chopper. The resulting value

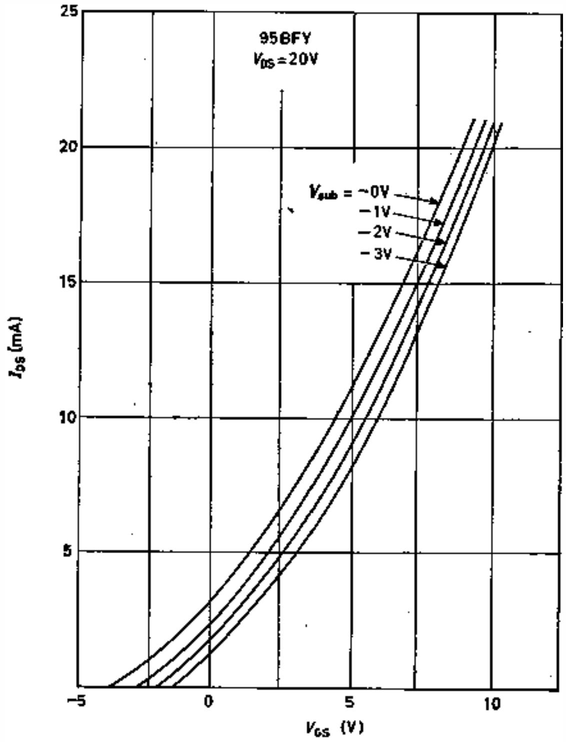


Fig. 7. Substrate control characteristic

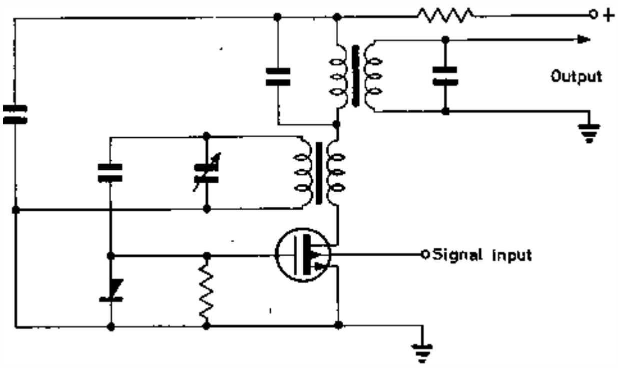


Fig. 8. Oscillator-mixer circuit

of 'on' resistance, around 1kΩ, is high. This affects only the gain, or chopping efficiency, of the device, and does not introduce an error signal. As the gain is normally stabilized by feedback, this is a small disadvantage.

As in all circuits using the m.o.s. transistor care must be taken not to forward bias the substrate junctions by more than ½V. The substrate should be taken to a bias equal to the maximum negative input signal.

Other Applications

The facility of connexion to the substrate has been mentioned, the drain substrate, and source forming a field effect transistor. Control of the voltage on this terminal controls the drain current as shown in Fig. 7, and the m.o.s. transistor therefore has two control terminals, the gate and the substrate, which independently control the output current. If both terminals are controlled, multiplication of the two signals occurs. Using this property an oscillator-mixer has been designed (Fig. 8), using a single device.

If one of the signals is a d.c. input the gain of the stage may be controlled either in open loop or as a.g.c., and the device has obvious possibilities as an analogue multiplier in the computer field and in power meters.

Further properties have to be investigated. When the chopper transistor was discussed it was mentioned that the device resistance varied as $r_{ds} = 1/\beta(V_{GS} - V_{GST})$, and its use as a voltage controlled resistor is possible. The high temperature coefficient of this resistance is a disadvantage in this application. It may also be possible to use the accurate square law under certain bias conditions, in the analogue computer field.

Switching Applications

The circuits discussed above have been linear applications but the device has great potential as a switch. As it is a majority carrier device there is no hole storage and a fast switching time is obtainable. A version of the 95BFY having a positive pinch off voltage is being produced which gives the advantage of threshold logic with no extra components. The high input resistance and low

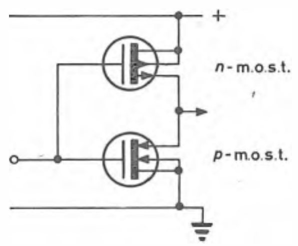


Fig. 9. Switch circuit

capacitance allow a high fan-out from the previous element to be used, and make the device ideal for resistive adding networks. A great potential is seen in micro-power and other applications with high component packing density, as the input power is negligible and the transient power small due to the short switching time. In the circuit of Fig. 9 an n-m.o.s.t. and a p-m.o.s.t. form a switch which may be operated from a common pulse, and consumes only a few nanowatts in each steady state.

An Airborne Data Acquisition System

British Eagle International Airlines has recently ordered the Sperry Airborne Data Acquisition System for installation in each of its 23 Britannia and Viscount aircraft and has taken an option on an additional quantity for further installation on BAC1-11 jet aircraft.

British Eagle thus becomes the first airline to adopt the new Sperry equipment.

The new Sperry system will provide British Eagle with comprehensive data relating to a wide variety of equipment on the aircraft. This will enable the airline to analyse equipment functions under differing flight conditions with a view to anticipating defects; advance information thus acquired will be used to maintain constant reliability of equipment.

By 1966/67, the installation of flight data acquisition equipment will become mandatory for all British aircraft. However the extent of forthcoming legal requirements are far exceeded by the information which will be supplied and recorded by SADAS. Minimum mandatory requirements will be six channels of information; the British Eagle equipment will provide initially 24 channels.

The system ordered by British Eagle International Airlines is a standard production version which has resulted from extensive research and development carried out by Sperry Gyroscope Co. Ltd at their Bracknell (Berkshire) Laboratories. The construction is completely solid-state and utilizes grown-circuit micrologic elements for its digital functions. The microcircuit design utilizes the advantages of digital techniques to the maximum possible extent.

The standard SADAS is a digital system which caters for the basic six parameters, viz. air speed, altitude, vertical acceleration, pitch attitude, aircraft heading and time, as required by the M.o.A. statutory requirements. In addition the standard system gives a 'plug-in' capability for the recording of a further 18 analogue and 30 on/off parameters at the discretion of the user. Plug-in modules have been developed for all the transducer outputs encountered on current aircraft systems. The transducer signals, irrespective of type, are converted at an early stage to a binary form. Subsequent handling of the transducer information is digital and will result in no loss of original accuracy.

Information is retrieved by a reverse process to that used for recording. As the system is digital in the first instance original accuracy and 'play-back' accuracy are the same.

SADAS will record parameters to 10 bits of digital information. The system is completely integrated and all system functions are synchronized to a crystal-controlled oscillator operating at 2·048Mc/s. Frame identification signals are also included to aid the recovery of the recorded information.

The digital information produced by the Sperry Airborne Data Acquisition System can be recorded either on wire or tape recorders whichever is appropriate to the particular application. For the British Eagle application (which is the accident/maintenance role) a Davall wire recorder has been selected. Recording time between cassette changes is approximately 300 hours.

Diesel Engine Test Plant

Automatic control equipment for the conversion to fully-automatic operation of the existing manually-operated engine-testing dynamometer equipments at the Dunstable commercial vehicle factory of Vauxhall Motors Ltd is being supplied by Bruce Peebles Co. Ltd, Edinburgh.

Each of the twenty-two new equipments will control one of the existing Bruce Peebles 40kW variable-speed d.c. dynamometers, and will enable the engine under test to be subjected to any one of a number of pre-determined load/speed test cycles. The required programme is encoded on a 24-track plastic card (itself colour-coded according to the test cycle it represents), and the passage of the programme card through a 'reader' provides the necessary control signals to the two independent closed-loop control systems. Any required combination of two hundred load settings and the same number of discrete speed steps can be selected, and the total test time can be pre-set to any chosen value up to a maximum of 75 minutes. A two-channel chart recorder provides a permanent record of each test run.

In a typical test cycle the diesel engine is first motored by the dynamometer at a low speed while the valves and the fuel injection system are checked and adjusted by hand. The motoring speed is then increased to 1 000rev/min, and the fuel supply is turned on; once the engine is running, the automatic control equipment takes charge, and the required test cycle is begun. The test can if necessary be interrupted by hand at any time.

The speed and the load are separately controlled. In the speed control circuit a speed-dependent voltage from a tachogenerator is compared with a voltage proportional to the required speed; the amplified 'error signal' difference between the two voltages supplies the control winding of a magnetic amplifier, whose output is fed to the field winding of the dynamometer to control the machine speed. In the load control circuit a similar comparison is made between a voltage proportional to the load-dependent armature current of the dynamometer and a pre-set voltage representing the required engine load; in this instance the amplified error signal controls a servo-motor which in turn operates the engine throttle. The speed response of the load control loop has been so designed as to prevent the possibility of such engine faults as uneven injection being masked by the automatic control.

Counters and Shift Registers Using Static Switching Units

By J. F. Young, C.G.I.A., A.M.I.E.E., A.M.I.E.R.E.

One advantage of counters incorporating universal high voltage transistor-diode switching units is that they can operate cold-cathode numerical indicators directly. A suitable binary-decimal counting code makes possible the use of direct coupled tens transfer as feedback while having sufficient redundancy to permit economy and a measure of loading equalization in the diode matrix used for decoding. A shift register arrangement using the the same static switching units is also given.

(Voir page 209 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 216)

TRANSISTOR-DIODE static switching units[1] using relatively high supply voltages are widely useful in industrial switching systems, one of the advantages of such devices being that they can switch neon indicator lamps. The introduction of cold-cathode numerical indicator tubes made possible a direct numerical display in instruments such as counters and voltmeters. However, because of the fairly high current required, it is not possible to operate numerical indicators directly from Dekatron counters. They can be operated from hot-cathode valve counters or cold-cathode counting chains. While the former only require eight triode sections, it is desirable in many cases to avoid the use of hot-cathode valves. The usual cold-cathode counter uses a chain of ten tubes, though presumably eight could be used in a binary arrangement[2]. However, any of these methods take up rather a lot of space for each decade. A Dekatron counter with ten transistor amplifiers could be used, but this is uneconomical. Consequently, it is worth considering the use of transistor binary counters to operate numerical indicators. This necessitates the use of fairly high voltage transistors because of the large difference between ignition and extinction voltages of numerical indicator tubes.

A simple four stage binary counter operates in a scale of sixteen. Where decimal counting is required, it is necessary to modify the counter by the addition of feedback. When this is done, six of the sixteen possible combinations of conditions in the counter are never used. Consequently there can be some redundancy in the counting code[3]. As an example, consider a binary-decimal counting code as follows:

DECIMAL	BINARY
1	1000
2	1001
3	1010
4	1011
5	1100
6	1101
7	1110
8	1111
9	0000
0	0001
1	1000

In this binary code, the left-hand digit is only zero in two places. In one of these the right-hand digit is zero and in the other it is one. Consequently, regardless of the middle two digits, the first and last digits alone convey sufficient information to identify a count of nine or zero. The middle two digits are therefore redundant in these two cases, and they can be given any convenient values without changing the definite nature of the code.

An output pulse should normally be obtained on a change from a count of nine to a count of zero for tens transfer to a following stage. With the code illustrated a transfer pulse can be obtained whenever the right-hand digit is one and the left-hand digit is zero, since the middle two digits are redundant. If the transfer pulse is also fed back to change the redundant two middle digits from 00 to 11 then the code for zero becomes 0111. The next input pulse to be counted will add to 0111 in normal binary fashion to give 1000, which is the code for the decimal one. Consequently this feedback of the tens transfer pulse in a counter to the middle two stages to change their state has the effect of changing a scale of sixteen counter to a scale of ten counter. The coding around the tens transfer point can be written as:

9	0000
0	0001
0	0111
1	1000

Several features will be noted here. The middle digits have been steady at 00 for a whole digit period before they are changed by feedback. The transfer and feedback pulse is obtained at a time when only the least significant digit is changing. Both of these features ensure that the counting speed is not limited by the feedback arrangements. The feedback pulse is obtained on a definite output pulse rather than on a change of an output, i.e. whenever the first digit is 0 and the last is 1, the middle two can be held at 11, so that direct coupling of the feedback is possible. The fact that the tens transfer pulse is used for feedback means that no special outputs need be obtained, the tens transfer output being available in any case. At the instant when the middle stages are changed they are not supplying output information.

A block diagram of a binary decimal transistor-diode counter using this feedback scheme is shown in Fig. 1. The waveforms at various points in the circuit are shown in Fig. 2. It will be noticed that the maximum counting speed required of the most significant stage is one half of that for the other stages.

In order to obtain ten outputs from the eight transistor collectors, a code converting diode matrix can be used. A complete matrix for a four digit code with ten outputs would require the use of 40 diodes, but it is possible to reduce this number by making use of redundancy in the counting code. While designing the matrix, it is desirable to minimize the number of diodes required while sharing the output load among the transistor stages as equally as possible. Each output circuit is in effect an

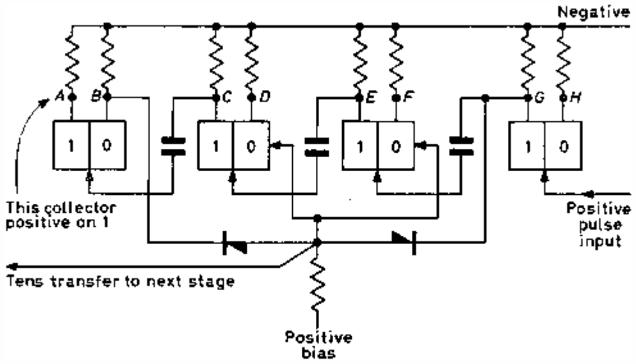


Fig. 1. Arrangement of the counter

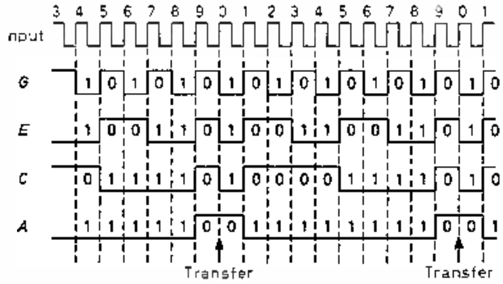


Fig. 2. Counter waveforms

AND gate, and at any instant it is necessary to supply current to nine of these gates to prevent an output voltage from being given from them, while the tenth gate is allowed to give an output. Since there are four stages, if it is assumed that each AND gate requires 12 units of current, the minimum possible loading on each stage at any instant is $(9 \times 12)/4 = 27$ units of current, if the load is shared equally.

A matrix which holds the maximum loading to 34 units while only requiring 32 diodes is shown in Fig. 3. The unit loads for this matrix are as follows:

	A	*B*	*C*	*D*	*E*	*F*	*G*	*H*
1	—	18	28	—	28	—	34	—
2	—	18	28	—	28	—	—	34
3	—	18	28	—	—	28	34	—
4	—	18	28	—	—	28	—	34
5	—	18	—	28	28	—	34	—
6	—	18	—	28	28	—	—	34
7	—	18	—	28	—	28	34	—
8	—	18	—	28	—	28	—	34
9	25	—	25	—	25	—	33	—
0	25	—	—	25	—	25	—	33

The addition of extra diodes would overload the *A* stage on counts of 9 and 0. The disadvantage of equalizing the loading in this way is the introduction of a slight chance of a short spurious output from 4 or 6 at the instant of tens transfer if one of the middle stages changes over before the other. Since this design was carried out empirically, it is not claimed that it is necessarily the best arrangement.

From an economic point of view, transistor static switching units should be usable in as many applications as possible. The use of high voltages gives a good tolerance on the pulse sizes which can be used. It is not easy to construct simple counters using low voltage static switching units because of poor pulse tolerance.

In order to count with bistable circuits such as transistor flip-flops, it is necessary to have a means of short term information storage in order to gate input pulses to the correct input to change the state of the device. Many ways of doing this are known. The simplest is the use of capacitors as shown in Fig. 4.

Here two transistor static switching units are coupled together in a memory or flip-flop arrangement. Extra input diodes are coupled to the input pulse line via capacitors C_1 and C_2. These capacitors can be charged by resistors R_1 and R_2 to a voltage depending on the collector voltage. With the supply voltages shown, if VT_1 is conducting then C_1 charges up to say 25V if the input to VT_1 is at 5V negative in this condition. Since VT_1 is conducting it holds VT_2 cut off. Consequently C_2 discharges via R_2 and the collector resistor of VT_2 to a zero voltage. Now if a pulse of +15V is applied to the pulse input line, the input to VT_1 is taken to +10V and VT_1 is cut off. The input to VT_2 is only taken to −15V, however, so that VT_2 cannot be affected by this input pulse. Once the collector voltage of VT_1 rises to more than −5V, VT_2 is allowed to conduct and it then holds VT_1 cut off even when the input pulse disappears. Provided that the input pulse is greater than 5V, it will cut off VT_1 and provided that it is less than 25V it will not affect VT_2, so there is a wide tolerance on the pulse size. The pulse should be limited to less than 15V with commonly available tran-

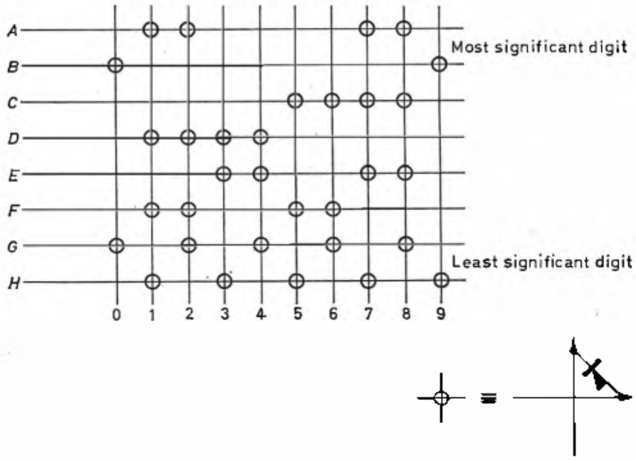
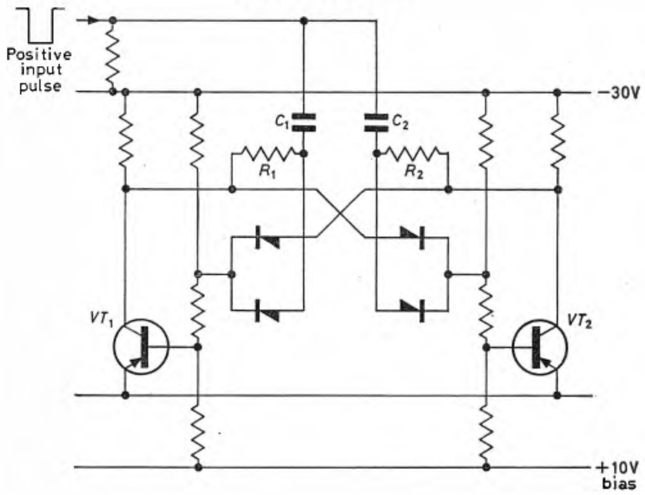


Fig. 3. Decoding matrix

Fig. 4. Binary stage



sistors in order to avoid exceeding the reverse base to emitter voltage rating. Thus a mean pulse amplitude of 10V should be suitable for use with the circuits given, permitting a maximum variation of pulse amplitude of ± 50 per cent. If the values of R_1 and R_3 are fairly high the slow negative feedback around a transistor via its input diode does not prevent the transistor from bottoming. The values of C_1 and C_2 are chosen so that they can charge to 25V via R_1 or discharge to at least 10V via

If it was not positive, do not apply a positive pulse to its base."

A counter is a device which summates input pulses, the sum being expressed in binary form. A shift register is a device which can store numbers in binary form and shift them any number of places, i.e. in effect multiply or divide them by 2^n. In a shift register, as in a counter, it is necessary to store information during the transfer interval and it is possible to use capacitors for this storage

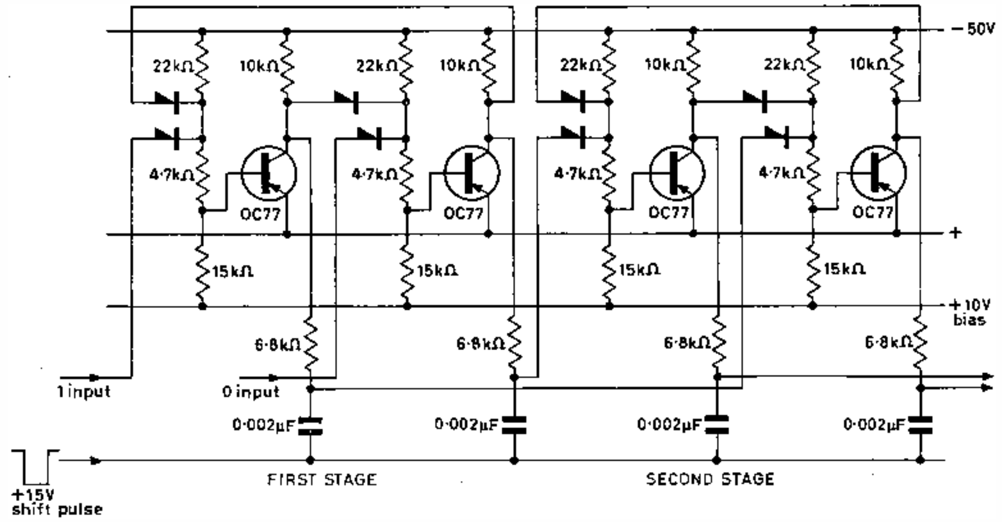


Fig. 5. Two stages of counter

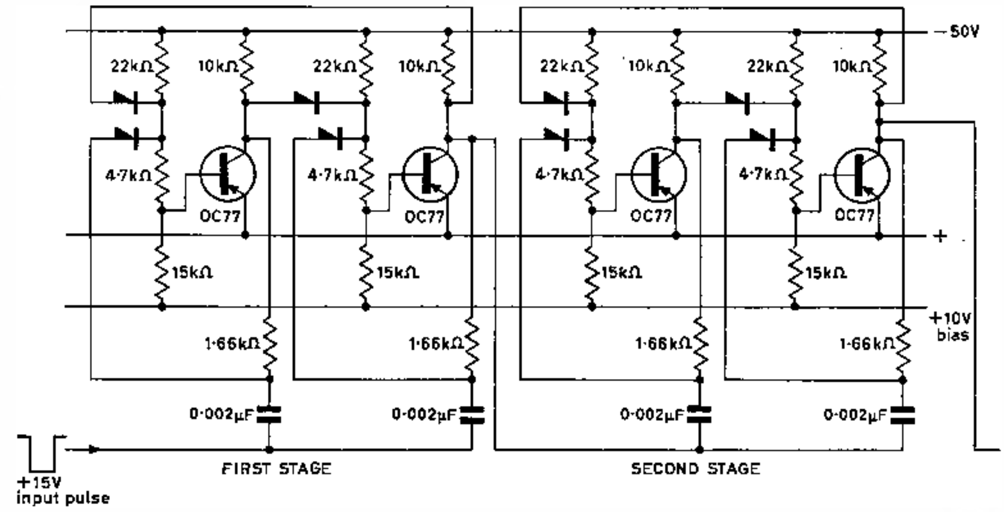


Fig. 6. Two stages of shift register

R_1 and the collector load resistor in the minimum anticipated interval between pulses. The capacitors should not be too small if the input pulses are not to be attenuated by stray capacitance. The input pulse must persist at least long enough for the flip-flop to change state. Too long an input pulse could allow the capacitor connected to a transistor which is cut off to discharge to zero and then charge in the opposite polarity. Care must be taken that this does not cause the transistor voltage rating to be exceeded when the pulse finally disappears.

A practical arrangement which has been used with pulse repetition frequencies up to 45kc/s is shown in Fig. 5. The logical principle of the counters described is: "if the collector was positive, apply a positive pulse to its base.

as in the counter. However, instead of the information about the next state of a stage being derived from its own previous state, it is now derived from the previous state of the previous stage in the register. In order to achieve this, the charging resistors for the capacitors must be taken to the collectors of preceding stages as shown in Fig. 6. Shift registers using this arrangement have been used up to pulse repetition frequencies of 25kc/s.

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The Design of H.F. Baluns for High Powers

By J. Wood*, M.I.E.E.E.

This article deals with the design of high frequency wideband high power transformers with particular emphasis on the use of unbalanced to balanced windings for the specific purpose of matching low impedance coaxial lines to a balanced feeder system or to high impedance aerials.

The use of nickel zinc ferrite makes a substantial contribution to the transformer design and performance which is characterized by a high efficiency (over 97 per cent) and very low v.s.w.r. contribution.

Design requirements are considered in some detail and proposals are given for a part theoretical, part empirical approach.

Experimental results relating to typical transformers are given with s.w.r. plots on Cartesian co-ordinates.

(Voir page 209 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 216)

WIDEBAND transformers[1] have hitherto been used mainly in receiver input circuits. It is sufficient in specifications for such transformers to quote the 'insertion loss' since the engineer is primarily concerned with producing a sufficient voltage at the input terminal to the receiver.

For the applications of wideband transformers to transmitters[2] the engineer requires to know very much more about the characteristics of such transformers, in particular:

(a) The v.s.w.r. introduced by the transformer when terminated in an ideal load.
(b) The efficiency of the transformer.
(c) Amount of distortion introduced by the transformer.
(d) Degree of unbalance in the balanced output.

Information relating to (a) is needed to assess the capability of the transmitter loading circuits to cope with the total mismatch of the transformer and the aerial.

Information relating to (b) and (c) has a significance which increases with the power rating of the transformer. In the case of a 5kW transformer these characteristics are extremely important.

Requirements for degree of balance (d) are different for different applications and perhaps the most stringent requirement is that where a wideband transformer is used to drive into a push-pull r.f. stage.

Equivalent Circuits

A transformer may be defined as a device for:

(a) Conversion from one impedance to another impedance.
(b) Converting circuits unsymmetrical to earth to circuits which are symmetrically disposed to earth.
(c) The isolation of different circuits.

For the purpose of this article the term wideband defines a transformer with a frequency coverage of at least 3 octaves.

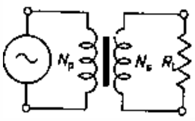
The ideal transformer is shown in Fig. 1 and its properties are defined by:

$$V_s/V_p = N_s/N_p;\quad I_s/I_p = N_p/N_s;\quad Z_s/Z_p = (N_s/N_p)^2$$

The ideal transformer, however, does not exist, and Fig. 1 can be redrawn to that shown in Fig. 2 to include the features found in a practical transformer.

The practical transformer differs from the ideal transformer by having:

(a) Leakage inductance (L_l) due to flux from primary current which fails to link with the secondary and represents the effect of magnetic energy stored between the windings.



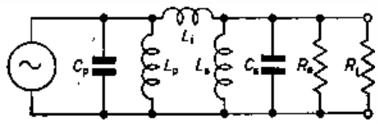


Fig. 1 (left). Equivalent circuit of ideal transformer
Fig. 2 (right). Equivalent circuit of practical transformer

(b) Distributed capacitances between turns and capacitance to earth, on the primary and secondary windings respectively. These distributed capacitances may be represented as lumped capacitances C_p and C_s at most frequencies of interest.
(c) Shunt resistance (R_e) representing the effect of eddy current losses and hysteresis losses.

Fig. 3 shows Fig. 2 redrawn to refer everything to the generator.

The greatest problems will be those presented by the presence of C_p, C_s and L_l at the highest operating frequency (defined as f_2) which is of the order of 30Mc/s.

SYMBOLS

L_p	=	Primary inductance
L_s	=	Secondary inductance
N_s	=	Secondary turns
N_p	=	Primary turns
V_p	=	Primary voltage
V_s	=	Secondary voltage
I_p	=	Primary load current
I_s	=	Secondary load current
I_o	=	Magnetizing current
L_l	=	Leakage inductance
R_p	=	Reflected primary resistance
R_L	=	Transformer secondary load
Z_p	=	Reflected primary impedance
Z_s	=	Reflected secondary impedance
R_e	=	Effective resistance of total core losses
C_p	=	Primary capacitance
C_s	=	Secondary capacitance
Y	=	Admittance
B	=	Susceptance
G	=	Conductance
f_1	=	Lowest operating frequency
f_2	=	Highest operating frequency
η	=	Efficiency
Z_o	=	Characteristic impedance of unbalanced line

* *Racal Communications Ltd.*

At this frequency it is possible that:

$$1/\omega C_p < R_p$$

and similarly:

$$1/\omega C_s < R_L$$

also L_l may be expected to have a significant effect upon the characteristics of the transformer.

At the lowest frequency (defined as f_1) which is of the order of 2·5Mc/s it may be assumed that:

$$1/\omega C_p \gg R_p$$

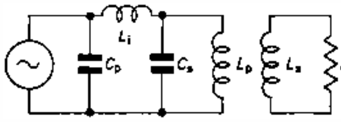
and similarly:

$$1/\omega C_s \gg R_L$$

Therefore it is reasonable to disregard completely the effects of C_p and C_s so far as f_1 is concerned.

The shunt inductive component L_p is, however, significant and it is this component which will determine the characteristics of the wideband transformer at f_1.

The individual problems associated with f_1 and f_2 become increasingly significant as the relationship between f_1 and f_2 approaches or exceeds four octaves. It is convenient to consider the design of a wideband transformer at these two frequencies.



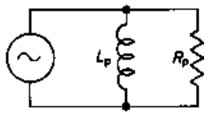


Fig. 3 (left). Equivalent circuit with reference to generator

Fig. 4 (right). L.F. equivalent circuit

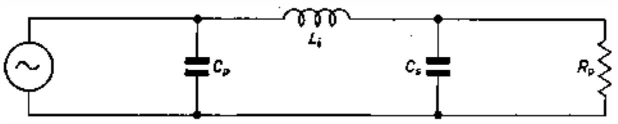


Fig. 5. H.F. equivalent circuit

Low Frequency Parameters

The equivalent circuit at f_1 may be reduced to that shown in Fig. 4 where it is assumed that the secondary is loaded by a resistive load R_L of the correct value and under these conditions the shunt resistive component of the total load seen by the generator is:

$$R_p = R_L\,(N_p/N_s)^2$$

which should be equal to the Z_o of the unbalanced line.

The successful design of a wideband transformer depends to a large extent on the use of high permeability core materials with the object of reducing the number of turns required to provide adequate primary and secondary inductance.

High Frequency Parameters

At high frequencies near f_2 only the shunt capacitances and the leakage inductance are significant and Fig. 4 can be redrawn to that shown in Fig. 5 for an equivalent circuit.

The load seen by the generator in Fig. 5 is no longer that which was given in equation (1) and the circuit should now be considered as a low-pass filter.

The load Z_p seen by the generator will probably be of incorrect magnitude and may possess a significant phase angle.

This equivalent circuit may be adapted to perform the function of a π-matching network by deliberately adding lumped reactances and these may be determined from the expressions for this network. It is not always convenient to change the value of the inductive reactance since this is determined by the practical design. One can, however, increase C_p or C_s so as to:

(a) Enable the circuit to be resistive.

(b) Provide the correct step down ratio as provided by the turns ratio at the low frequencies.

It is essential that the transformer possesses a low reactive to resistive power ratio, i.e. the reactive power handled by the transformer should not be sufficiently high to cause overheating.

Losses in Transformers

Definition of Losses

In any type of transformer design there are three distinctly separate circuits to be considered, and these are:

(1) The electric circuit.

(2) The magnetic circuit.

(3) The dielectric circuit.

The losses associated with these three branches are briefly summarized here.

I^2R Loss Due to Load Current

Since the currents are predetermined by the load one can only reduce the R value to the lowest possible minimum. This implies that heavy conductors are required. Since the inductance of a conductor decreases with increasing cross-section it follows that I^2R losses can be reduced at the expense of a slight increase in the magnetizing current.

I^2R Loss Due to Magnetizing Current

This loss is quite small in normal l.f. transformer design, where the magnetizing current is of the order of 3 to 10 per cent of the full load current. For wideband transformers there will be some significant increase at the lowest operating frequency.

I^2R Loss Due to Current Supplying Other Losses

In any transformer the input power will exceed the output power by an amount equal to the losses. Thus additional I^2R loss arises due to the increased primary current necessary. It is normally negligible.

Eddy Current Loss in Conductors Due to Leakage Fields

This loss is caused by the eddy currents in individual conductors which are set up by the stray magnetic fields. This loss is almost impossible to calculate, but is normally negligible in most transformer designs.

Hysteresis Loss in Core

This loss is dependent on the grade and size of the core material and can be calculated to a reasonable degree of accuracy from curves supplied. This loss may be of a significant order.

Eddy Current Loss in Core

This loss can be calculated from curves of the core material. It may be jointly considered with the hysteresis loss. Most of the considerations which apply to hysteresis loss apply equally well to eddy current loss.

Comparisons with Losses in L.F. Transformers

There are certain differences between low frequency power transformers and high frequency transformers. Low frequency power transformers are usually fed from a very low impedance source (e.g. the 50c/s grid system) and the transformer is therefore operated with a constant voltage source. Hence the magnetizing current and core losses are always present and are independent of the transformer load. The transformer efficiency so far as iron loss is concerned will improve with increasing load. Design considerations of the iron losses are therefore of paramount importance.

For transmitter applications the definition of no load to full load is different. Here, the load on the transformer secondary is usually maintained at a constant value and the primary voltage on the transformer is varied as the square of the transmitter output power. Thus the copper losses and iron losses both vary directly with the power load with which the transformer is fed.

It is usual in l.f. transformer design to have a copper loss factor which is equal to the iron losses and sometimes up to three times the iron loss. Copper losses are high because a large number of turns must generally be used to obtain an adequate primary inductance.

Such is not the case in h.f. transformer design. The amount of copper wire used in a modern ferrite cored 5kW transformer may perhaps not exceed 4ft and allowing for skin effect at the high frequencies a total copper loss of 0·3 per cent is not impossible. As a basis for comparison the electrical losses in a wideband transformer winding may be roughly equalled to those which would occur in about 8ft of coaxial cable. The core losses may perhaps range from 1 per cent to 2·5 per cent at the highest frequency. It should be possible to restrict ohmic losses in the dielectric materials to 0·2 per cent. The total efficiency of the transformer may be very high, and an efficiency of 99 per cent is not impossible. An average figure may be 98 per cent. The efficiency may be better at the low frequency end due to decreased skin resistance and probably a decrease in core losses. It is possible, however, that the increased magnetizing current at the lowest frequency with consequent I^2R losses may tend to have a levelling effect on efficiency over the band.

Standing Wave Ratio

Ideally a wideband transformer design should be such that when the secondary winding is terminated in an ideal load the reflected primary impedance Z_p should be resistive in nature and equal to the Z_0 of the coaxial line over the entire frequency range. Under these conditions the reflected wave will be zero and the s.w.r. will be 1 to 1.

In a wideband transformer design to cover a frequency range of more than 3 octaves it is inevitable that a significant v.s.w.r. will be introduced at some frequencies. It is desirable to restrict this figure to less than 1·5 : —1 at the worst point.

Reasons for this restriction are twofold:

(1) Some wideband aerials possess s.w.r. characteristics of 2 : 1. In order that the transmitter can cope with this it is essential that the transformer is as near ideal as possible.

(2) If the s.w.r. of the aerial sums with the transformer s.w.r. the effect may be one of excessive power loss due to the mismatch.

Core Design

Choice of Grades

For any one particular application it is inevitable that there will be more than one grade of ferrite which is applicable and that no one grade will be ideal in its behaviour. The final choice is determined after consideration is given to the characteristics of the ferrite with regard to:

(1) Permeability.

(2) Temperature effects—Curie point.

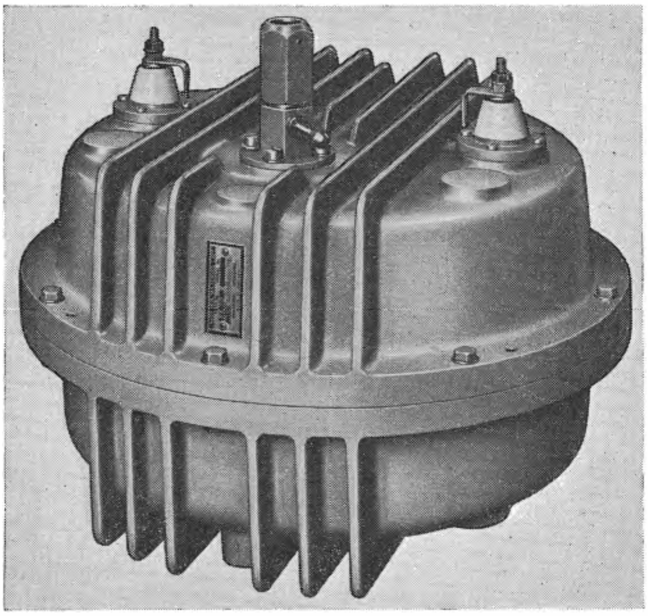
(3) Resistivity.

(4) Hysteresis losses.

From a study of ferrites the nickel zinc composition is the most suitable for high frequencies. In the nickel zinc composition there are a number of different grades available. For high power h.f. wideband transformer requirements, the Mullard grade B2 is ideal so far as hysteresis losses and distortion are concerned.

Permissible Core Losses

Maximum permissible core losses per cubic centimetre depend on the cross-sectional area, this figure of permissible loss decreasing with increasing core size due to the extremely poor thermal conductivity of ferrites, which are of a ceramic like nature.



Model suitable for pressurized cables

A figure of 0·1W/cm³ is accepted as being reasonable, and this corresponds to a temperature rise of the order of 40°C in free air. This holds good for cores up to 1 or 2cm radius. For larger cores, the figure must be reduced.

The use of heat sinks is desirable wherever possible, and with flat sided cores it is usually possible to use at least two of the four faces.

The use of oil cooling, although possible, is not always advantageous, for the following reasons:

(1) Increased capacitance effects.

(2) Increased complexity of design.

(3) Need for upright mounting.

The use of heat sinks will probably enable the figure of 0·1W/cm³ to be increased by some 30 per cent for the same temperature rise (40°C above ambient).

The maximum permissible figure for dissipation depends on the maximum permitted temperature, and for operation in temperate climates a figure of 65°C rise is reasonable. However, where the transformer is required to work in tropical conditions, with ambient temperatures of up to +55°C, then a figure of 40°C or 0·13W/cm³ with heat sinking should be adhered to.

Methods for Achieving Balance

In a transformer consisting of an unbalanced winding and a balanced winding, the physical design must be such as to satisfy the following requirements to enable the transformer to meet the balance requirements.

(a) Equal distributed capacitance to ground from both halves of the balanced windings.

(b) Equal degree of magnetic coupling from both halves of the balanced windings to the unbalanced winding.

A difference in voltage magnitudes can still occur even if requirements (a) and (b) are met as a result of the 180° phase difference between V_p and V_s if a voltage existing on the primary winding is coupled symmetrically to both halves of the balanced winding.

Various methods have been successfully employed to yield good balance requirements in unbalanced winding transformers. Tape windings are widely employed with the attendant advantage of increasing magnetic coupling and though the inter-turn capacitance may be increased it is usually acceptable.

As a further aid to balance an electrostatic screen may be interposed between the unbalanced and balanced windings. This is particularly necessary in the case where the number of turns on the unbalanced winding are so few in number that a significant voltage magnitude is developed across the earthed outer turn leading to the phase effects previously described.

Power Handling Capacity

For a given transformer design the maximum power that can be safely handled by the transformer is determined by one or more of the following:

(a) Sustained increase in core losses with consequent core fracturing (or total destruction in the case of a heavy overload) due to the ceramic like nature of ferrites.

(b) Effects of voltage flashovers leading to partial or total destruction of the electrical winding.

Limiting factors affecting (a) are:

(1) R.M.S. power input to transformer.

(2) Aerial mismatch.

Limiting factors affecting (b) are:

(1) Aerial mismatch.

(2) Type of service and peak power.

A mismatch of aerial impedance will, in certain circumstances, lead to increased voltage across the secondary of the transformer with increased magnetizing current and core loss.

Mechanical Requirements

CONSTRUCTION

The transformer is required to be portable, ruggedized, and suitable for operation in unattended aerial locations in all climatic conditions in ambient temperatures ranging from $-40°C$ to $+55°C$.

An ideal form of construction so far as outdoor applications are concerned seems to be a circular finned casting consisting of two identical halves with a sealing gasket and facilities for filling with dry air at a little above atmospheric pressure. This is to prevent breathing in of moisture and the possible growth of fungus within the transformer in certain tropical locations. The use of a finned casting will serve the dual purpose of providing a greater radiating surface for heat dissipation and yielding a stronger mechanical construction.

Regarding the construction of the transformer windings, the shape can be square, or round. Since most cores are of square cross-section, a square winding could be used for low power transformers. For higher powers, the circular winding offers less risk of voltage flashover due to the absence of sharp corners. A circular winding also results in the capacitance being evenly distributed round each turn and this is an advantage at high frequencies.

The temperature requirements are such that great care must be taken in the choice of insulating materials for radio frequency voltages. The list of available materials becomes narrowed where the voltage gradients are very high, as in the case of high power radio frequency transformers.

PROTECTIVE DEVICES

The essential differences between most transformer applications and the extraordinary conditions of use that are likely to be encountered with wideband aerial matching transformers need to be recognized. For instance, if the aerial load is suddenly removed from the transformer while the transmitter is delivering radio frequency power to the transformer primary, the transmitter tank circuit Q will rise to an unloaded figure of perhaps several hundred and in consequence the reflected voltage across the secondary winding of the transformer may rise to a dangerously high value. A figure of 10kV peak for a 5kW transmitter is a distinct possibility, and this would cause complete destruction of the transformer windings.

The possibility of dangerously high peak voltages being developed as a result of overmodulation exists, and also the transformer needs to be protected against lightning discharge.

The fitting of a horn-type spark gap across the secondary winding would give adequate protection against higher than normal peak voltages as a result of overmodulation. It would, however, give no protection whatsoever against lightning discharges. For the transformer to be protected against this phenomena a separate spark gap needs to be fitted from the output terminals to the casing of the transformer which is required to be earthed.

It should, however, be noted that the use of a safety spark gap can, under certain circumstances, render the transformer open to even greater hazards than those encountered if no spark gap is fitted. This is due to the fact that when the spark gap conducts the secondary load falls to a near zero value, and in consequence the load currents through the transformer windings rise to a very high value which may actually melt the windings due to the temperature attained.

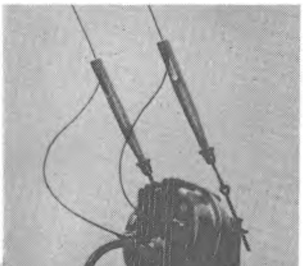
Design Approach

A complete theoretical design for a wideband radio frequency transformer is made extremely difficult, due mainly to the following reason:

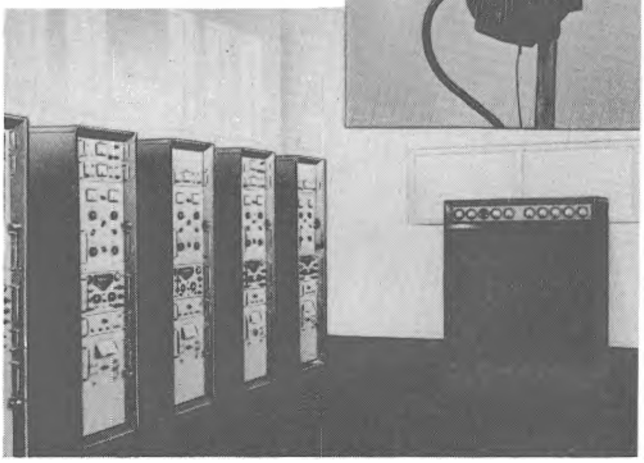
(1) The uncertainty with which the distributed capacitances to ground and the inter-turn capacitances can be calculated.

(2) Similar difficulties associated with the calculation of the leakage inductance to the degree of accuracy necessary.

A theoretical design is probably possible only in those cases where the octave range of the transformer specification does not exceed 3 to 1. Even in this case it is doubtful whether the results



A typical installation



will be within 10 per cent of the required performance.

The following method has been successfully employed:

(1) For a given maximum permissible value of s.w.r., the minimum value of primary winding reactance at f_1 may be determined. Thence the actual value of L_p and L_s can be determined. At this frequency the susceptance of C_p can be completely ignored.

(2) Select the appropriate core material and draw up a suitable design in terms of cross-section, having regard to the low flux density requirements.

(3) Calculate values of N_p and N_s or determine these experimentally.

(4) Calculate the voltage gradient per turn (this factor is very significant at the high powers where it may be as high as 1kV due to the small number of turns) and decide on a minimum turn spacing.

(5) Construct the experimental transformer and measure the characteristics L_p, L_l, C_p and C_s. At the highest operating frequency the reactance of the winding capacitances C_p and C_s should not be significantly less than R_p and R_L, otherwise there will be an appreciable circulating current.

(6) Admittance measurements with the transformer terminated in its ideal load should be carried out. The results will determine whether any high frequency equalizing circuits are necessary and what form they should take.

Electrical performance data and s.w.r. characteristic for a high power wideband balun are given in the Appendix.

Acknowledgments

The author is indebted to the Board of Directors, Racal Electronics Ltd., for permission to publish this technical note and acknowledgment is made to Mr. W. E. Robinson who contributed much to the early development and the making of the first experimental model in 1958.

Appendix

Experimental results relating to a high power balun transformer are given in Figs. 6, 7, 8 and 9. Fig. 8 reveals the transformer to possess an inductive characteristic over almost the entire frequency range with a phase angle increasing rapidly at the low frequency end due to a decrease in the reactance of L_p.

Fig. 6 shows the balance characteristic of the transformer and Fig. 9 shows the s.w.r. characteristics.

These experimental results were plotted on an experimental transformer, the ferrite core of which was constructed from a large number of small pieces of ferrite pieced together to form a larger core of a size 4in by 6in having a window area of 2in by 2in. This form of construction was necessary at the time because of the unavailability of large sections of ferrite.

Some loss of permeability was caused by this form of construction and this was subsequently proved when the production transformer constructed from a two piece core of a U and I limb was shown to have some significant improvement in electrical characteristics. The s.w.r. characteristics of this later model were of the order 1·1 to 1 from 30 to 4Mc/s and the lowest frequency was extended to below 2·0Mc/s for a s.w.r. of the order of 1·6 to 1.

Typical production models are shown in the photographs on pages 166 and 167.

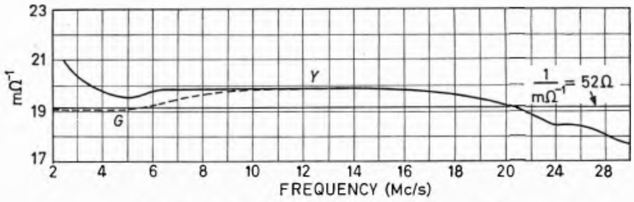


Fig. 7. Frequency-conductance characteristic

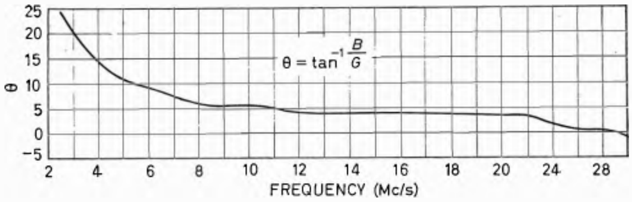


Fig. 8. Typical characteristic of 3kW model

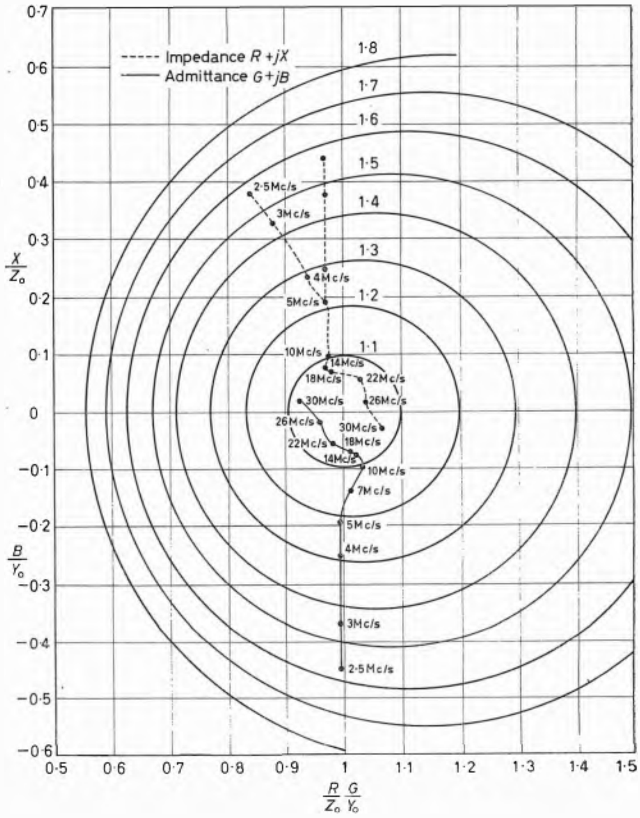


Fig. 9. S.W.R. characteristics

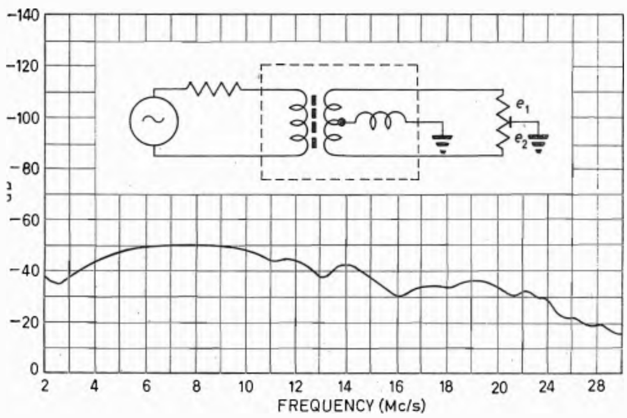


Fig. 6. Balance characteristic

e_1 = the larger voltage, e_1 and e_2 measured to earth from each side of a centre-tapped 600Ω resistor

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Solid State Devices for E.H.T. Rectifiers and Efficiency Diodes in Television Receivers

By F. D. Bate*, B.Sc., A.M.I.E.E.

An investigation of the problems involved in the use of solid state devices for e.h.t. rectifiers and efficiency diodes in television receivers. The effect of stored charge and forward resistance of solid state devices upon the regulation of e.h.t. diodes is considered. The heating which results from capacitive current flow when long chains of diodes are used is also investigated.

(Voir page 209 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 216)

In a television receiver two high voltage diodes are used; one, the e.h.t. rectifier, supplies the e.h.t. for the tube, and the other, the efficiency diode is used in the line scanning stage. In a black and white receiver a typical high voltage rectifier has an absolute maximum peak inverse voltage rating (p.i.v.) of 27kV, and a mean current of 200μA (U26) and a typical efficiency diode, the U193, has an absolute maximum peak inverse voltage of 6·6kV and 150mA mean current.

A colour receiver, on the other hand, will probably require a p.i.v. of 35kV and a mean current of at least 1mA for the e.h.t. rectifier and a p.i.v. of 7kV and 500mA mean current for the efficiency diode. Silicon diodes can be made where a single junction can withstand 2kV p.i.v. and if allowance is made for a 100 per cent overload safety factor about 35 junctions in series would be required for the 35kV p.i.v. e.h.t. rectifier and seven junctions for the efficiency diode. Selenium rectifiers connected in series can also be used for the e.h.t. rectifier and if a p.i.v. of 48V/junction is assumed then 730 junctions would be required[1].

E.H.T. Rectifier

In Fig. 1 regulation curves are plotted using a U26, a stack of selenium diodes, a chain of ordinary silicon diodes, BY105, and a chain of low stored charge silicon diodes. From the curves it can be seen that both the selenium and ordinary silicon devices show worse regulation than the U26 valve. The reason for the poor regulation in the case of selenium is the high forward resistance of the device whereas in the case of silicon it is the stored charge. The chain of silicon diodes having low stored charge gives a regulation curve very similar to the U26 valve diode. Fig. 2 shows typical conduction current waveforms for these various devices.

Effect of the Forward Resistance of a Diode

In Appendix (A) it is shown that the increase in the internal resistance of a rectifier, having a forward resistance R_D is given, for small values of R_D, by:

$$R_i = 4/3\ R_D (T_r/T_c) \quad \dots\dots\dots\dots\dots \quad (1)$$

where T_r is the repetition time and T_c the conduction time.

Equation (1), due to the simplification assumed in its derivation gives a value of R_i too high so that R_i, in practice, will always be less than this. Taking the U26, $R_D = 8k\Omega$ for a peak current of 15mA, $T_r = 64\mu$sec, $T_c = 1{\cdot}4\mu$sec for a mean current of 200μA, this gives $R_i \simeq 0{\cdot}5M\Omega$. Thus, roughly, half a megohm of the internal resistance of the supply is contributed by the resistance of the diode. Equation (1) also shows that when T_c, the conduction time, is small, then the effective resistance of the diode is large, this occurs for small values of mean current, moreover, the value of R_D also increases

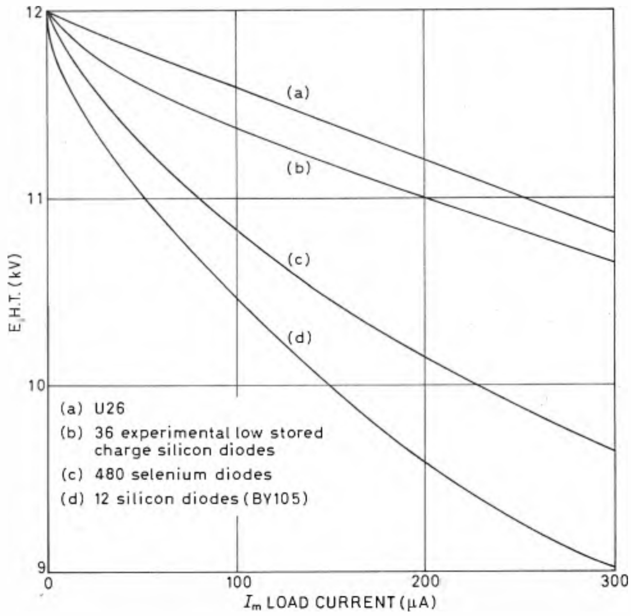
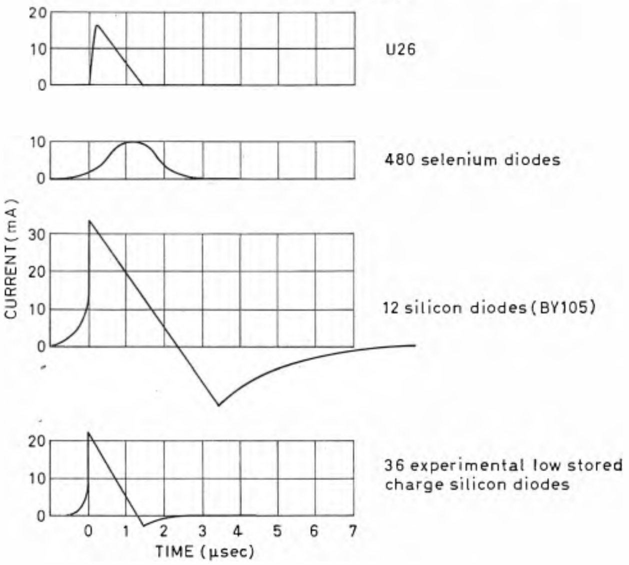


Fig. 1. Regulation curves of U26, selenium and silicon devices

Fig. 2. Shape of the conduction current of the rectifiers
Mean current = 200μA



* *Thorn-AEI Radio Valves and Tubes Ltd.*

with decreasing current. For instance if the peak current is about 5mA corresponding to a mean current of 20μA then $R_D = 14\text{k}\Omega$, $T_c = 0{\cdot}4\mu$sec and $R_i \simeq 3{\cdot}0\text{M}\Omega$. This explains why all regulation curves show a rather steep initial drop for the first 10 to 20μA of mean current.

In general the value of the internal resistance R_i is given by an equation of the type.

$$R_i \simeq k \,.\, R_D \;(T_r/T_c) \quad \ldots\ldots\ldots\ldots \quad (2)$$

where k is some constant that tends towards 4/3 for small values of R_D.

In the case of a chain of 480 selenium diodes

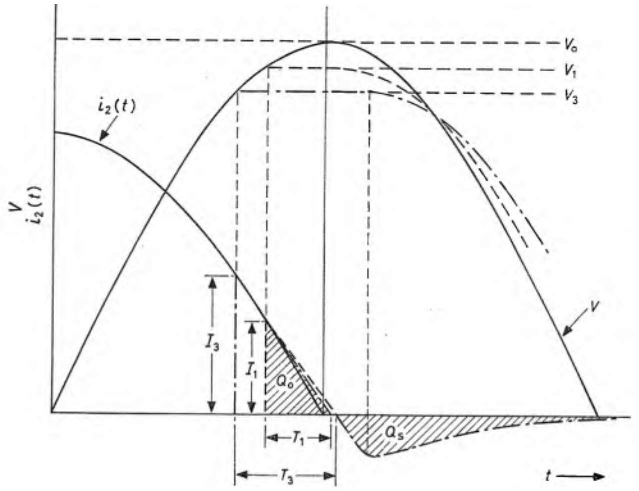


Fig. 3. Voltage and current waveforms associated with the e.h.t. diode

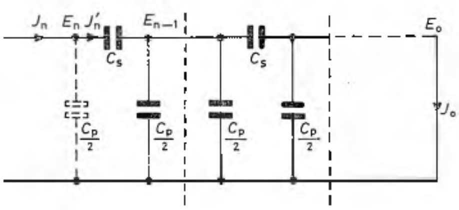


Fig. 4. Equivalent circuit of 'n' diodes with self-capacitance and capacitance to earth

$R_D \simeq 100\text{k}\Omega$, for a peak current of 10mA and mean current 200μA, $T_c \simeq 2{\cdot}5\mu$sec. Taking $k = 4/3$, this gives $R_i \simeq 3{\cdot}4\text{M}\Omega$ compared with 3MΩ measured. It would thus appear that k lies somewhere near a value of 1·2 to 1·3 for this value of R_D.

Effect of the Charge-Storage of a Solid State Diode

It can be shown from considerations of the waveform in Fig. 3 that the internal resistance R_2 of a device having stored charge and used as a line flyback e.h.t. diode is given approximately by:

$$R_2 \simeq R_0\big(1 + (Q_s/Q_0)\big) \quad \ldots\ldots\ldots\ldots \quad (3)$$

where R_0 is the internal resistance of the supply with a device having no stored charge, Q_s is the stored charge and Q_0 the charge taken out of the supply per cycle. Now the stored charge is not only a function of the current at any instant it is also a function of the current existing before that instant, so that, even when the forward current has collapsed to zero there is still some stored charge in the diode. This stored charge then flows out in the opposite direction to which it flowed in and produces a reverse current. The magnitude of the stored charge, for the same conduction angle, will increase with the repetition frequency, and is signficant at 15kc/s but not at 50c/s.

A further effect of using chains of solid state devices having stored charge occurs when a voltage transient is applied to the chain. In this case, ignoring capacitive current flow, which will be dealt with later, the diode with the lowest stored charge will recover first and hence have to stand the largest voltage, this gives an additional reason for using devices with low stored charge[2].

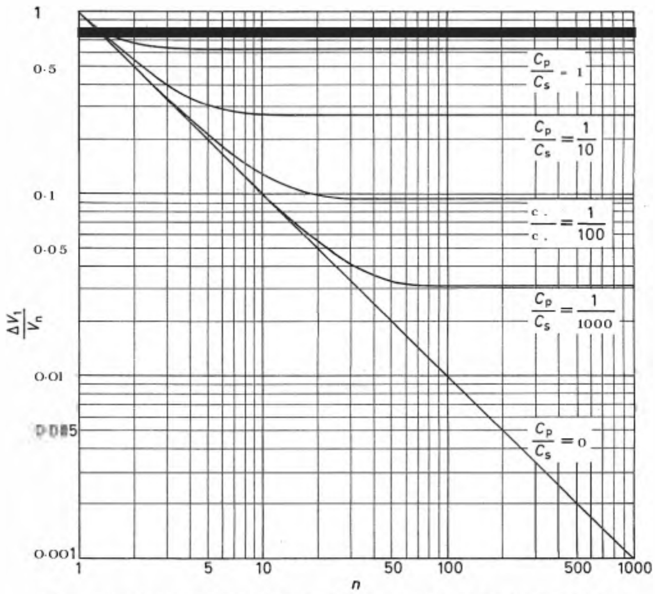


Fig. 5. Voltage across the first diode against the number of diodes

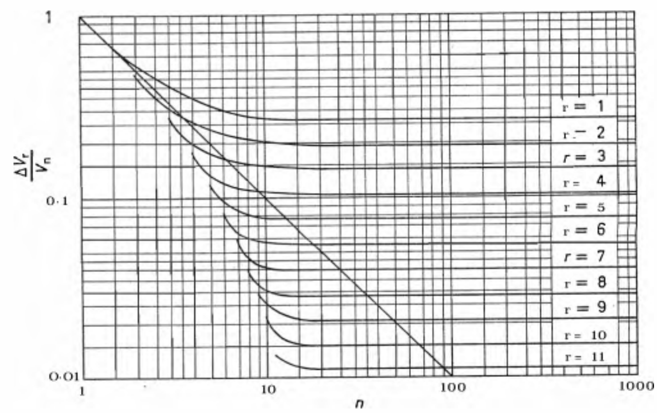


Fig. 6. Voltage across each of the first eleven diodes against the number of diodes
$C_p/C_s = 1/10$

Capacitive Current Heating of Chains of Rectifiers

So far only the effect of the forward resistance and stored charge upon the regulation of an e.h.t. supply have been considered. There is, however, the problem of the heating in the diodes caused by the flow of capacitive current. This heating is caused by the capacitive current producing a larger than average voltage across the first few diodes. This voltage reaches the breakdown voltage of the diode under consideration which is then clamped. The diode now has a constant voltage across it and a reverse current equal to the capacitive current of the rest of the diode chain. Heating takes place and the effect will be known as capacitive current heating.

A number of diodes connected in series, each diode

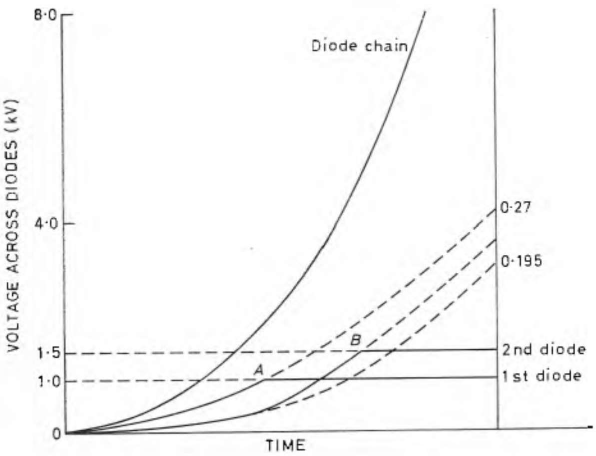


Fig. 7. Voltage appearing across the diode chain and the first and second diodes

having a capacitance across it, and to earth, forms, in effect, a ladder network as shown in Fig. 4. An analysis of this giving the voltage drop across each diode and the capacitive current flowing is given in Appendex (B). The graphs of Figs. 5 and 6 show the voltage drop across individual diodes in the chain as well as the capacitive current flowing in each individual diode. For instance, if $C_p/C_s = 1/10$, Fig. 6, the voltage drop across the first diode never falls below 27 per cent of the applied voltage no matter how many diodes are in the chain, and, provided more than a certain minimum number of diodes are present, this voltage drop falls off by about 73 per cent per diode as one moves down the chain. The capacitive current flowing through each diode falls off by the same amount. When $C_p/C_s = 1/100$ the voltage across the first diode never falls below 9·5 per cent of the applied voltage, each succeeding diode having 90 per cent of the previous diode voltage across it.

In the case of semiconductor devices the series capacitance is not a fixed value but is itself a function of the voltage across it, for instance, with a silicon diode, BY105, if the reverse voltage is about 100V, then the capacitance is about 2 to 3pF and at 1kV this falls to the order of 1pF. The capacitance to earth is more difficult to measure but is unlikely to be less than 1 per cent of this, that is, 0·01pF.

Consider now more fully what happens when the voltage across the diode chain is increasing. If the diodes could withstand a very large inverse voltage, then the voltages across individual diodes in the chain would be as already indicated, and no heating of the diodes would occur. However, taking the case where $C_p/C_s = 1/10$ the first diode in such a chain at 16kV would have to withstand over 4kV which is several times what present-day diodes can withstand. If breadkdown occurred at 1kV for instance, part *A*, Fig. 7, the diode would clamp at this voltage and the inverse current would be the actual capacitive current at the instant 1kV was reached across the diode. The voltage across the second diode would now rise more quickly and if this broke down at 1·50kV, part *B*, Fig. 7, would clamp at this voltage, but at a lower value of capacitive current. This process would pass down the chain until the inverse voltage was reached. In practice, this means that the first few diodes get quite hot while the rest of the chain remains cool. The first diode is not necessarily the hottest, because if it has a low breakdown voltage, then it clamps at this low voltage. The second diode may have a very high breakdown voltage and this high value of voltage may more than compensate for the lower value of capacitive current. This diode would consequently get hotter.

Avoidance of Capacitive Current Heating

In order to avoid capacitive current heating:

(a) The stray capacitance C_p should be made as small as possible so that the voltage distribution varies inversely as the capacitance across each diode.

(b) Assuming that C_p is finite, C_s could be made distributed, either by the first diodes having the largest self capacitance or by placing a spiral of wire over the first part of the chain, thus in effect increasing their capacitance[3].

(c) The first few diodes would have to have large inverse breakdown voltages so that the breakdown voltage was never reached.

(d) The capacitive current heating could be accepted but the first diodes should have a low breakdown voltage so that the consequent heating of individual diodes was reduced.

Efficiency Diode

Silicon diodes having low stored charge are the only diodes considered, the large forward resistance of the selenium diodes prohibit their use. A chain of 6 to 10 such diodes would probably be suitable for use as an efficiency diode. Experimentally chains of 12 diodes have been run successfully for 1 000 hours at a mean current of 150mA and peak voltage of 6kV, these had low inverse voltages per diode. The low forward resistance of these diodes, compared with a valve, results in a rather more efficient scanning system, saving about 5 per cent in power.

The voltages expected across each diode, assuming no breakdown of the diodes can be found from Figs. 5 and 6. Taking Fig. 5 where $C_p/C_s = 1/10$ the voltages to be expected across individual diodes can be quickly found. These are contained in Table 1 below for 6 , 8 and 10 diodes.

This table shows that for $C_p/C_s = 1/10$, and six diodes in the chain, then the first diode has to withstand nearly four times the inverse voltage of the last diode. When the number of diodes is increased to ten then the first diode has to stand over ten times the voltage of the last diode. It is also clear that increasing the number of diodes from six to ten only reduces the voltage across the first diode by about 10 per cent.

When $C_p/C_s = 1/100$ and there are ten diodes in the chain the first diode has 12·5 per cent of the voltage across it compared with 8·5 per cent for the last diode, compared with 10 per cent across each diode when $C_p/C_s = 0$. Thus even in the case of a very low capacitance

TABLE 1

Percentage Voltage Across Individual Diodes in the Chain ($C_p/C_s = 1/10$)

NUMBER OF DIODES	1st DIODE	2nd	3rd	4th	5th	6th	7th	8th	9th	10th	IF LINEAR $C_p/C_s = 0$
6	31	23	16·5	12	9·5	8					16·6
8	28	21	15·5	11·2	8·2	6	4·5	4			12·5
10	27	20·5	15	11	8	5·7	4·2	3	2·5	2·2	10

to earth the first diode has 47 per cent more inverse voltage than the last diode in the chain.

These results show that, from the capacitive point of view, it would be better to use a few diodes each having a large inverse voltage rating than a large number having a low inverse voltage.

Conclusion

It would appear that long chains of solid state diodes could be used both as an e.h.t. rectifier and efficiency diode in a television receiver. Such devices with low stored charge and low forward resistance would give a better regulation for the e.h.t. rectifier and increased efficiency for the efficiency diode. The heating in long chains of diodes at 15kc/s caused by the flow of capacitive currents and resulting inverse breakdown of the first few diodes is one of the main problems to be overcome.

Acknowledgments

The author would like to thank the management of the Thorn-A.E.I. Applications Laboratory for permission to publish this article, and also to Mr. W. J. Pitts for his help and advice in its preparation.

Appendix

(A) Effect of the Diode Resistance upon the Regulation of a Line Flyback E.H.T. Supply

The following theory is based upon two simplifications of the actual conditions:

(1) The forward resistance of the diode is linear.

(2) The shape of the diode conduction current is triangular, that is, its shape is unchanged by the value of the series resistance.

Applying the principle of the conservation of energy for the case of a resistance-less diode[4,5] one has that:

$$\tfrac{1}{2}C(V_o^2 - V_1^2) = I_M^2 R_L T_r \quad \ldots\ldots\ldots\ldots (4)$$

where V_o = output voltage on no load.

V_1 = output voltage with load R_L but $R_D = 0$, or load R_L' and $R_D \neq 0$.

I_M = mean output current.

R_L = load resistance.

C = capacitance in the oscillating circuit.

T_r = repetition time of each cycle.

For the circuit using a diode of resistance R_D and the same mean output current, one has that:

$$\tfrac{1}{2}C(V_o^2 - V_1^2) = (I_M^2 R_L' + I_{rms}^2 R_D)\, T_r \quad \ldots\ldots (5)$$

where R_L' is the new load resistance and R_D the diode resistance. Clearly from equations (4) and (5):

$$R_L = R_L' + (I_{rms}^2/I_M^2)\, R_D \quad \ldots\ldots\ldots\ldots (6)$$

Thus the diode behaves as an internal resistance of value $(I_{rms}/I_M)^2\, R_D$.

Now for a triangular conduction current of duration T_c:

$$(I_{rms}/I_M)^2 = (4/3)\,(T_r/T_c) \quad \ldots\ldots\ldots\ldots (7)$$

In practice, the presence of the diode resistance produces a 'rounding off' of the conduction current and an increase of T_c. The result is that the r.m.s. is always less than the value as estimated above, and hence the internal resistance R_i less than equation (7) suggests. A more general expression for the resistance would be:

$$R_i = k \,.\, R_D\, (T_r T_c) \quad \ldots\ldots\ldots\ldots\ldots\ldots (2)$$

where k is some number that tends to 4/3 for small values of R_D.

Hence the internal resistance of the line flyback e.h.t. supply becomes:

$$R_{in} \simeq T_r/C\ \left(1 + k\,(R_D C/T_c)\right) \quad \ldots\ldots\ldots (8)$$

where T_r/C is the internal resistance of such a supply assuming a diode of zero resistance.

(B) Voltage Drop Across Each Diode and Capacitive Current

The equivalent circuit for a chain of n diodes with self capacitance and capacitance to earth is shown in Fig. 4. E_n is the voltage to earth across the chain of n diodes, C_s is the series capacitance of each diode and C_p the capacitance to earth at each junction of two diodes. The output voltage E_o is zero so far as a.c. is concerned because normally a large capacitor C_o (Fig. 8) is present at this point. In terms of matrices this can be written:

$$\begin{bmatrix} E_n \\ J_n \end{bmatrix} = \begin{bmatrix} \cosh u & Z_o \sinh u \\ (1/Z_o)\sinh u & \cosh u \end{bmatrix}^n \begin{bmatrix} E_o \\ J_o \end{bmatrix} \quad \ldots\ldots\ldots (9)$$

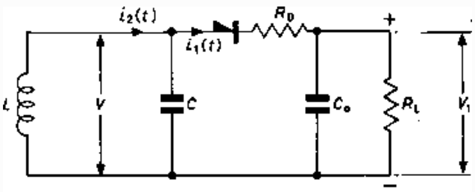


Fig. 8. Circuit having a solid state e.h.t. diode

where $\cosh u \equiv 1 + (C_p/2C_s)$

$$\sinh u \equiv \sqrt{\left[C_p/C_s\,(1 + (C_p/4C_s))\right]}$$

$$Z_o \equiv \frac{1}{p\sqrt{(C_p C_s + (C_p/4))}}$$

By simple Matrix Theory this becomes:

$$\begin{bmatrix} E_n \\ J_n \end{bmatrix} = \begin{bmatrix} \cosh nu & Z_o \sinh nu \\ (1/Z_o)\sinh nu & \cosh nu \end{bmatrix} \begin{bmatrix} E_o \\ J_o \end{bmatrix} \quad \ldots\ldots\ldots\ldots (10)$$

Writing this in ordinary equation form and putting $E_o = 0$ gives:

$$E_n = J_o Z_o \sinh nu$$
$$J_n = J_o \cosh nu \quad \ldots\ldots\ldots\ldots\ldots\ldots (11)$$

(a) *Voltage Across Each Diode*

Writing $(n-1)$ for n in equation (11) then:

$$E_{n-1} = J_o Z_o \sinh (n-1)\,u \quad \ldots\ldots\ldots\ldots (12)$$

$$\therefore \quad \frac{E_n - E_{n-1}}{E_n} = \Delta E_n/E_n = 1 - \frac{\sinh (n-1)\,u}{\sinh nu} \quad \ldots (13)$$

This equation gives the fraction of the total voltage applied to the circuit, which occurs across the first diode, expressed as a function of the number of diodes in the chain.

Equation (13) may also be written in the form:

$$\Delta E_n/E_n = 1 - \cosh u + \sinh u \coth nu \quad \ldots\ldots (14)$$

clearly when nu becomes very large, $\coth nu \rightarrow 1$.

so that:

$$\mathop{\mathrm{Lt}}_{n\to\infty} \quad \Delta E_n/E_n = 1 - \cosh u + \sinh u = (1 - e^{-u}) \quad \ldots\ldots (15)$$

using the values for $\cosh u$ and $\sinh u$ this becomes:

$$\mathop{\mathrm{Lt}}_{n\to\infty} \ \Delta E_n/E_n = \sqrt{\left[(C_p/C_s) + (C_p/2C_s)^2\right]} - (C_p/2C_s) \quad \ldots\ldots (16)$$

This equation gives the limiting value of the voltage that appears across the first diode, measured from the input voltage, when the number of diodes in the chain becomes large.

A graph of equation (14) for different values of C_p/C_s is shown in Fig. 5. When $C_p/C_s = 1$, this limiting value of the voltage is about 62 per cent of the input voltage and is practically reached when three diodes are in series; any increase in the number of diodes above this, does not affect the fractional voltage drop across the first diode.

When $C_p/C_s = 1/100$ the limiting value of the voltage is 9·5 per cent and is practically reached with thirty diodes in series.

In general, the voltage drop across the r^{th} diode is given by:

$$\Delta E_r/E_n = \frac{\sinh(n-r+1)u - \sinh(n-r)u}{\sinh nu} \quad \ldots\ldots (17)$$

and:

$$\operatorname*{Lt}_{n\to n} (\Delta E_r/E_n) = e^{-u(r-1)}(1 - e^{-u}) \quad \ldots\ldots\ldots\ldots (18)$$

Fig. 6 shows a graph of equation (17) when $C_p/C_s = 1/10$ up to the eleventh diode.

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Remote Control and Indication Equipment

Electronic remote control and indication equipment to enable up to 220 bits of digital information to be transmitted at speeds of 50, 240 or 750 bauds in time division multiplex over a single voice frequency telegraph channel to CCITT standards, has been introduced by Supervisory Systems Division of G.E.C. (Electronics) Ltd. Known at DT3, the system—which uses solid state components throughout—provides circuits over which equipment at remote locations can be operated or monitored from a central control point.

Designed to operate over commercial speech circuits whether on open wires, underground cable, carrier transmission system or radio links, systems are extensible in multiples of 10 circuits to a maximum of 220. Greater circuit capacity can be achieved by combining up to 24 basic systems using voice frequency telegraph equipment to channel information over the same speech circuit.

A system comprises a transmitting terminal and remote receiving terminal, linked by a single telegraph channel to constitute a unidirectional link. Multiple information to the link, supplied via the contacts of external electro-mechanical relays or switches, is interrogated in sequence by a continuous scanning system. These states are then transmitted in time division multiplex, and reconstituted into individual outputs at the receiving station. The outputs are then used to light external lamp displays or to operate telephone type relays associated with control circuits of the equipment to be operated. Information bits are checked by a parity system before display.

Use of modular construction in accordance with the Post Office type 56 practice where circuits are built up from a number of common, self-contained units, enables systems to be easily tailored to customers' requirements. Each unit comprises up to 14 circuit boards, interconnected by wrapped joints. Twelve different types of circuit board are used in the equipment, most of which contain two logic circuit elements.

The system is designed to operate from d.c. power supplies of +10V and −10V, but where these are not available, suitable rack-mounted mains units can be supplied to operate from 100-130V or 200-250V 40-60c/s. Ambient operating temperature range is −10°C to +40°C.

The principle of operation is briefly as follows. Information is presented to the transmitting terminal in parallel and may consist of up to 220 independent binary conditions. These conditions are normally obtained from the contacts of external electro-mechanical relays or switches, each contact presenting either an open or closed circuit. Each input is represented by a binary digit, an open-circuit equivalent to a binary '0' and a closed-circuit to a binary '1'.

At the transmit terminal a scanner, essentially an electronic distributor, is stepped at a speed equal to the transmission rate used by the system, e.g. if the system works over a 50 baud telegraph channel, the distributor steps once every digit period of 20msec. During each step of the distributor an electronic gate associated with a particular input will open and admit a binary '1'. If the input is '0', the gate does not open.

As the distributor steps, the conditions of the input gates, sampled in sequence, are used to modulate a voice frequency telegraph transmitter. A frequency shift system is used so that during each digit period the tone transmitted to line is one of two frequencies depending on the state of the gate being inspected.

At the receiving station, a v.f. telegraph receiver accepts the succession of tones from the line, and demodulates them to obtain an equivalent train of binary '0''s or '1''s. Each terminal contains a distributor, the stepping of which is synchronized to the transmitter. The receive distributor is arranged to open electronic gates which direct each group into a 12 bit temporary store. When each group has been entered and checked for the presence of errors, the information bits are transferred in parallel to a set of permanent stores appropriate to that group.

Each of the permanent stores may be used either to light an external lamp display, or to operate telephone-type relays associated with the control circuits of the equipment being controlled.

This equipment forms the nucleus of the largest resignalling project to be carried out by British Railways. This has just been commissioned as part of the electrification programme for the Euston/Crewe/Liverpool main line on the London Midland region. The new system enables a staff of three—two signalmen and a regulator in a central signalbox at Rugby—to control 59 route miles of track (159 single track miles). Previously, 22 mechanical signal boxes were required.

The G.E.C. Electronics' equipment, supplied under subcontract to S.G.E. Signals Limited (an associate of the G.E.C.), is used in conjunction with a comprehensive 16ft route-setting control panel and a 34ft track diagram which enable the operators to supervise train movements, and to control points and signals throughout the entire area.

The signalmen set up train routes by operating buttons laid out on the control panel in relation to the geographic position of the actual track signals etc. Buttons are pressed at the entry point, along the selected route, and at the desired exit point on the route diagram. As the train negotiates the predetermined route, the signals are automatically placed to danger behind it. Electrical interlocking equipment ensures complete safety along the selected route.

The 34ft track diagram provides coloured-light indication of the position of every train within the area controlled by the Rugby box, the routes set up, and the conditions ('on' or 'off') of the signals. In addition, the type of train on the track is shown in standard 4-digit code.

G.E.C. Electronics' Teledata v.f. tone system is used for the train describer circuits over which the identity of the trains entering the Rugby area are transmitted. This identity originates from a signal box in the area which a train is leaving. Similarly, as the trains leave the Rugby area their identity is transmitted automatically to the next receiving signal box over Teledata circuits.

D.T.3 equipment is used to supervise the control of colour light signals and points in seven satellite interlocking areas of the Rugby signalbox area. By means of D.T.3, digital information is transmitted cyclically by time division multiplex at 750 bauds over a single telegraph channel.

A Demonstration Time Division Multiplex System Using Pulse Amplitude Modulation Techniques

By R. Barrett*, B.Sc., A.M.I.E.E.

This is a description of a simple laboratory system built to demonstrate to communications students the basic principles of time division multiplexing techniques, the sharing in time of a common communication medium between several separate information channels.

(Voir page 209 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 216)

UNTIL recent years, when the advent of the transistor with its excellent switching properties, small size and low heat dissipation made systems as above practical, multi-channel communication over a common medium was usually effected by employing frequency division multiplexing techniques. In these systems each channel is allocated a particular slot in the frequency spectrum of the medium and the channels are stacked as shown typically in Fig. 1, combining and separating of channels being effected by frequency selective filter networks. The original information is usually translated to its allocated frequency band by single sideband amplitude modulation

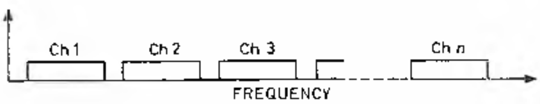
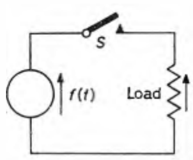


Fig. 1. Stacking of f.d.m. channels



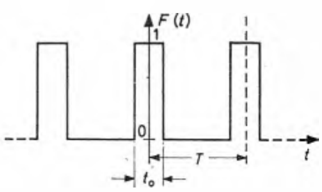


Fig. 2 (left). An elementary sampling switch

Fig. 3 (right). The switching function

techniques. Since each channel occupies a different band of frequencies, then ideally all could be using the common medium at the same time without mutual interference. Modern systems can combine more than 1 000 separate information channels on one pair of wires in this way but the filtering is complex and expensive and interference between channels occurs due to system non-linearities.

Time division multiplexed systems are basically much simpler, combination and separation of channels being effected by timing circuits rather than filters and inter-channel interference is less dependent on system non-linearities, due to the fact that only one channel is using the common communication medium at any instant in time. It will be seen that the number of channels that can be multiplexed is considerably less than by f.d.m. methods. Practical systems have been built with a 100 channel capacity.

Basic Principles of Sampling

Time division multiplexing is made possible by the fact that it is not necessary to continuously transmit a source of information in order that it shall be unambiguously interpreted by a receiver. Thus the information only needs to be 'sampled' at specific instants of time, as is shown by the following discussion.

Consider the simple switching circuit in Fig. 2, where the source of information, a time function $f(t)$, is connected to the load by the closing of the switch. Thus when the switch is closed the signal in the load $f_m(t)$ is equal to $f(t)$ and when it is open the signal in the load is zero. Analytically the action of the switch may be represented by a switching function $F(t)$ which is 0 when the switch is open and 1 when it is closed, as shown in Fig. 3.

Using the complex Fourier notation:

$$\left.\begin{aligned} F(t) &= (1/T)\sum_{n=-\infty}^{n=\infty} C_n \,.\, \exp(jn\omega_c t) \\ \text{where } C_n &= |C_n| \,.\, \angle\phi = \int_{-(T/2)}^{T/2} F(t) \,.\, \exp(-jn\omega_c t) \,.\, dt \end{aligned}\right\} \qquad \ldots\ldots\ldots\ldots (1)$$

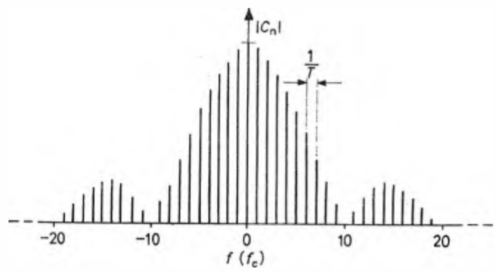


Fig. 4. Amplitude spectrum of the train of pulses when $k_o = 0{\cdot}1T$ Lines represent discrete frequency components

and $\omega_c = 2\pi f_c = 2\pi/T$

Thus from Fig. 3:

$$C_n = \int_{-(t_o/2)}^{t_o/2} \exp(-jn\omega_c t) \,.\, dt$$

$$\text{i.e. } C_n = kT \,.\, \frac{\sin(n\pi k)}{n\pi k} \qquad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (2)$$

$$\text{where } k = t_o/T = \frac{\text{pulse duration}}{\text{pulse period}}$$

Now $|C_n|$ gives an indication of the amplitude/frequency spectrum of the train of pulses, a typical plot being shown in Fig. 4. Hence, if k is made smaller the amplitudes of all frequency components are reduced and, if the criterion of bandwidth associated with effective transmission is taken as the first null, then the bandwidth is increased, since the frequency of the first null is f_c/k. These are two important factors to which reference will be made again later.

The result of switching produces the modulated function $f_m(t)$ according to the relationship:

$$\begin{aligned} f_m(t) &= F(t) \,.\, f(t) \\ &= (1/T)\sum_{n=-\infty}^{n=\infty} C_n \,.\, \exp(jn\omega_c t) \,.\, f(t) \qquad \ldots\ldots\ldots (3) \end{aligned}$$

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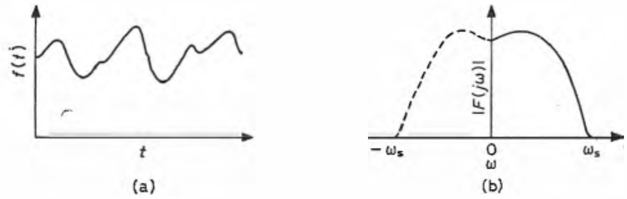


Fig. 5. The signal source $f(t)$ and its Fourier transform $F(j\omega)$
(a) Time spectrum (b) Frequency spectrum

If the Fourier transform of the signal source $f(t)$, shown in Fig. 5, is:

$$F(\mathrm{j}\omega) = \int_{-\infty}^{\infty} f(t) \,.\, \exp(-\mathrm{j}\omega t) \,.\, dt$$

being zero above ω_s, then for any particular component of the switching function, $n\omega_c$ say, the Fourier transform of the modulated signal $F_m'(\mathrm{j}\omega)$ is:

$$F_m'(\mathrm{j}\omega) = (1/T) \,.\, C_n \int_{-\infty}^{\infty} f(t) \exp(-\mathrm{j}\omega t) \,.\, \exp(\mathrm{j}n\omega_c t) \,.\, dt$$

$$= (1/T) \,.\, C_n \,.\, F(\mathrm{j}\omega - \mathrm{j}n\omega_c)$$

This is, apart from the multiplying factor, the spectrum of the original function shifted from the origin to $n\omega_c$. Therefore, the complete spectrum of the modulated signal $F_m(\mathrm{j}\omega)$ is shown in Fig. 6 (not to scale) and is given by:

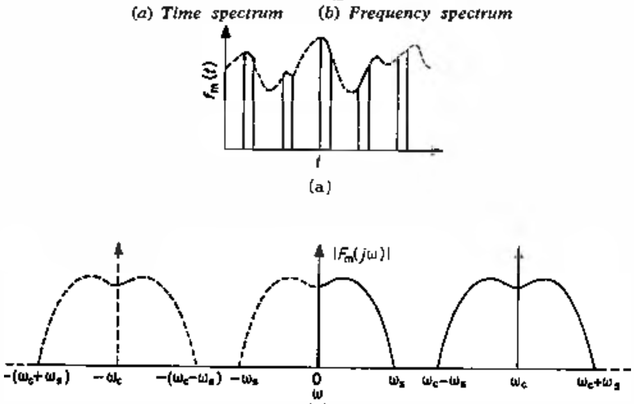
$$F_m(\mathrm{j}\omega) = (1/T) \sum_{n=-\infty}^{n=\infty} C_n \,.\, F(\mathrm{j}\omega - \mathrm{j}n\omega_c) \quad \ldots\ldots (4)$$

The use of negative frequencies of course is only a mathematical convenience and the actual spectrum will be that shown in full in Fig. 6 with all amplitudes multiplied by a factor of 2.

The most important conclusion to be drawn from this analysis is that, although the signal source is not continuously connected to the load, but only for a discrete interval of time during a switching period, the actual length of time depending on the value of k, the modulated function contains the original signal information $f(t)$. From this fact stems the possibility of time sharing of a communication medium between several channels, since during the period when a particular channel is disconnected other channels may be connected to the medium in turn.

From Fig. 6(b) it is seen that provided the lower sideband associated with the fundamental frequency of the sampling pulse train does not overlap the spectrum of the signal, then the signal is uniquely specified by the frequency components in the band up to f_s. The limiting case is when the edges of the two spectra coincide, i.e. when $f_c = 2f_s$. Thus for unique reconstruction of the signal from

Fig. 6. The sampled waveform $f_m(t)$ and its Fourier transform
(a) Time spectrum (b) Frequency spectrum



the samples there is the requirement that f_c needs to be just greater than $2f_s$. A much more stringent requirement is imposed, however, when practical means are considered of recovering the signal information. A study of Fig. 6(b) shows that it is only necessary to pass the modulated pulse train through a low-pass filter which passes f_s and sufficiently attenuates $f_c - f_s$ and all frequencies above. In order to obtain a realistic filter design requirement, sufficient guard space must, therefore, be allowed and so f_c must be appreciably greater than $2f_s$. Practical systems passing speech up to 3·4kc/s use sampling frequencies of 8 or 10kc/s. Ideally the filter should have a linear phase/frequency characteristic in the pass-band, but this is unimportant in speech systems.

Time Division Multiplexing

The process of sampling has produced a pulse amplitude modulated wave which may be transmitted over a communication medium and demodulated by means of a low-pass filter. In a single channel system there is obviously no point in this process but if account is taken of the fact that, in Fig. 6(a), $f_m(t)$ is zero for an appreciable amount of time, then samples of other channels may be interlaced with this set. Thus the medium is shared in time between the various channels and the combination process only requires the generation of sampling pulse trains in the correct time sequence.

De-multiplexing, or separation of the channels at the receiving terminal of the system, may be effected by 'coincidence' gating circuits. If, for instance, the channel 1 gate is opened by means of a properly timed gating pulse only during the intervals of time when a channel 1 pulse arrives at the receiving terminal, then the channel 1 low-pass filter receives the set of pulses containing all of the channel 1 information. A train of properly timed channel gating pulses is, therefore, required at the receiving terminal which is accurately synchronized to the original channel sampling pulses. The simplest method of achieving this in a pulse amplitude modulation system is the use of the sampling pulses of one channel as synchronizing pulses. If they are made appreciably larger in amplitude than the largest allowable sampled amplitude then it is a simple matter at the receiving terminal to detect the correct sequence and time the receiver channel gating pulses accordingly.

Channel Capacity

For a given information signal bandwidth and demodulating low-pass filter the spacing between sampling pulses is fixed. Therefore, the number of channels that may be multiplexed is determined by the duration of the sampling pulses. In practice there will be a small guard space between separate channel pulses to reduce crosstalk effects, but in the limit the number of channels will be $1/k$, where k is $\dfrac{\text{pulse duration}}{\text{pulse period}}$ as before. Thus to increase the channel capacity of the system it is necessary to decrease the pulse duration. The first effect of this, as deduced from equations (2) and (4), is that the amplitudes of the signal components in the signal waveform are reduced proportionately, requiring increased amplification after the low-pass filter and resulting in a degradation in signal-to-noise ratio.

A method of increasing the energy in the demultiplexed pulses prior to demodulation involves 'pulse stretching', i.e. increasing the duration of individual pulses without affecting their amplitudes. This process results in a series of flat-topped pulses and introduces a form of distortion, as is discussed below.

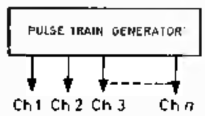
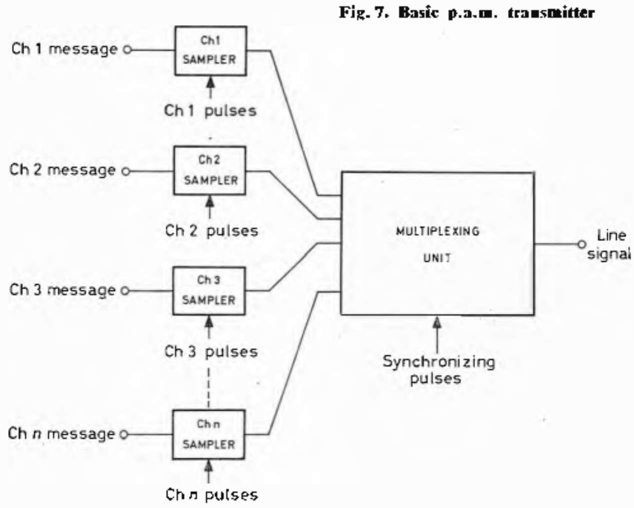


Fig. 7. Basic p.a.m. transmitter



Consider first the original modulated signal, described by Fig. 6(a) and equation (3). If the sampling pulses are made very narrow, i.e. $t_0 \to d\tau$, then $C_n \to d\tau$, since $\frac{\sin x}{x} \to 1$ as $x \to 0$ and thus:

$$f_m(t) = (d\tau/T) \sum_{n=-\infty}^{n=\infty} \exp(jn\omega_c t)\,.\,f(t) \quad \dots\dots\dots\dots (5)$$

The effect of pulse stretching can be represented by a summation of a large number of such trains, each successive train delayed by $d\tau$, say between the limits $\pm(t_0/2)$.

Hence:

$$F_m(j\omega) = (1/T) \sum_{n=-\infty}^{n=\infty} \int_{-(t_0/2)}^{t_0/2} \int_{-\infty}^{\infty} f(t-\tau)\,.\,\exp(-j\omega(t-\tau))\,.\,\exp(jn\omega_c(t-\tau))\,.\,dt\,.\,d\tau$$

$$= (t_0/T) \sum_{n=-\infty}^{n=\infty} F(j\omega - jn\omega_c)\,.\,\frac{\sin[(\omega - n\omega_c)\,t_0/2]}{(\omega - n\omega_c)\,t_0/2} \quad \dots\dots\dots\dots (6)$$

When this waveform is demodulated, the case $n = 0$ is of interest, i.e.

$$F(j\omega)_{\substack{\text{demod}\\(n=0)}} = k\,.\,F(j\omega)\,.\,\frac{\sin(\omega t_0/2)}{\omega t_0/2} \quad \dots\dots\dots\dots (7)$$

Attenuation distortion is thus produced in the signal band, the amount depending on the amount of stretch. In the limit, if the pulse is stretched for the full period,

$$\omega_s t_0/2 = \frac{\omega_s \pi T t_0}{2\pi T} = \omega_s k\pi/\omega_c$$

and $\omega_c = 2\omega_s$, $k = 1$

Therefore the multiplying term $\frac{\sin(\omega t_0/2)}{\omega t_0/2}$ varies from 1 at d.c. to $2/\pi$ at ω_s, resulting in a 3·9dB spread and necessitating equalization.

A second effect of increasing the number of channels

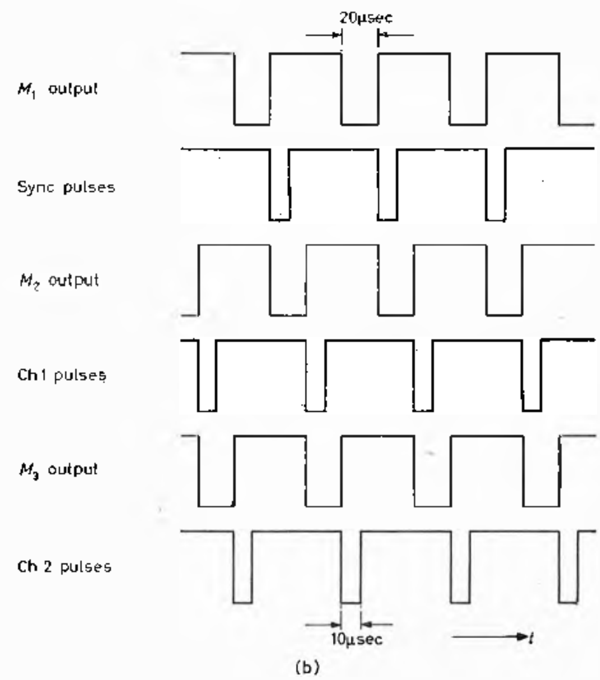
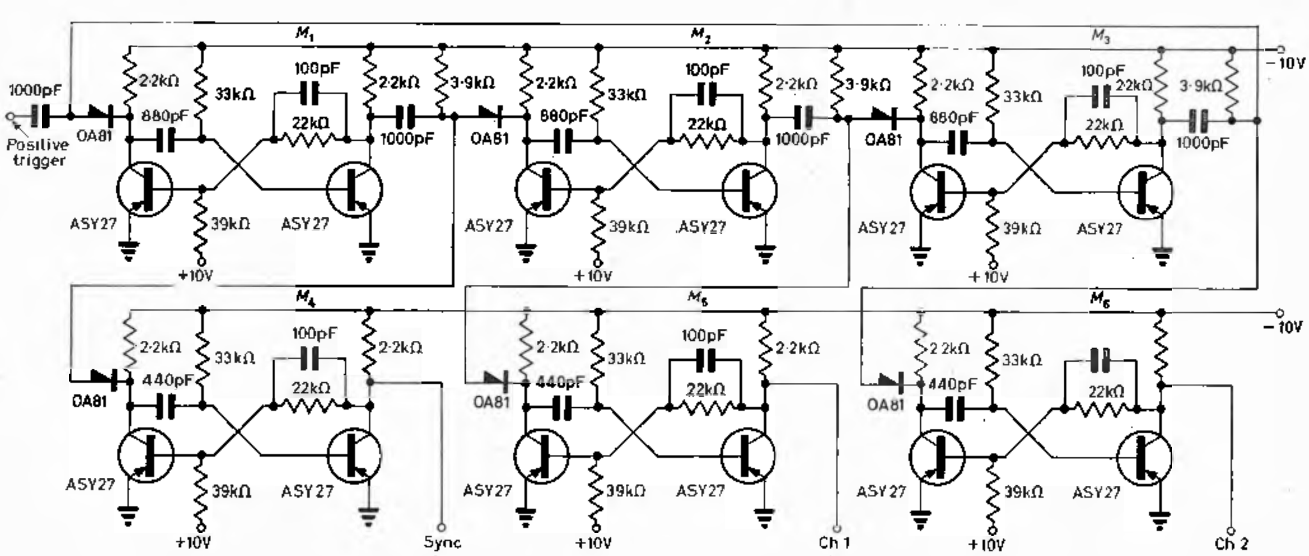


Fig. 8. Pulse train generator

(a) (below) The circuit (b) (above) The relevant waveforms



(a)

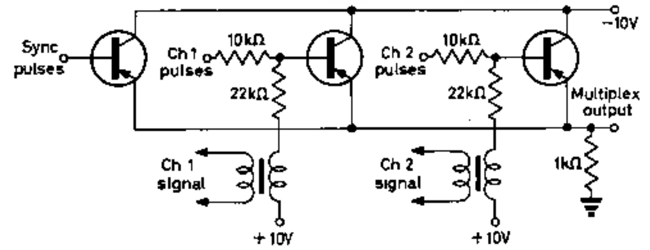


Fig. 9. Sampling and multiplexing unit

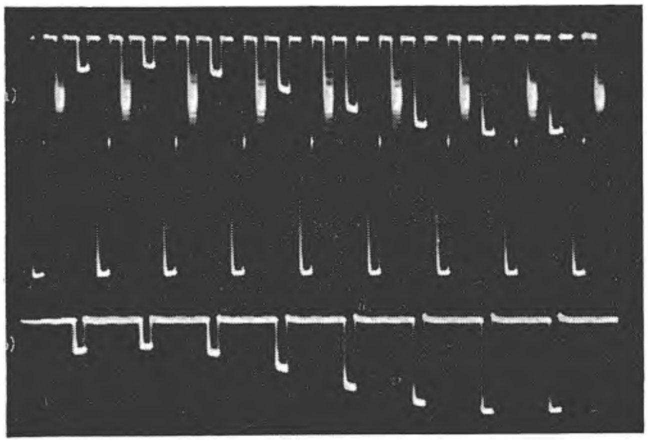


Fig. 10. System waveforms
(a) Line signal (b) De-multiplexed Ch. 2 signal

is an increase in bandwidth requirements of the transmission medium. It has been seen that in the limit the sampling frequency f_c must be twice the highest significant signal frequency f_s. If n channels are multiplexed then n pulses are transmitted to line during the sampling period and the line signal thus consists of a train of pulses occurring at a rate nf_c. Now these pulses completely specify a continuous signal with a highest frequency component of $nf_c/2$, i.e. nf_s. Thus the absolute minimum transmission bandwidth required for transmission of the multiplexed information is nf_s. The received line signal in such a case will be a continuous signal, but with its amplitude only having meaning at the sampling instants. In a practical system, of course, where it is necessary

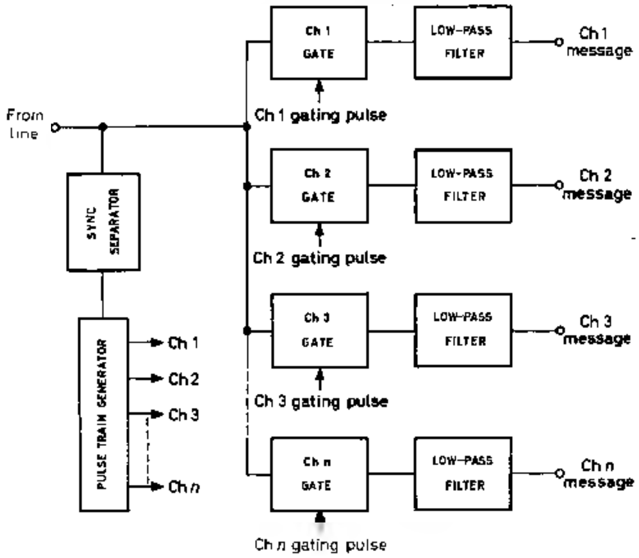


Fig. 11. Basic p.a.m. receiver

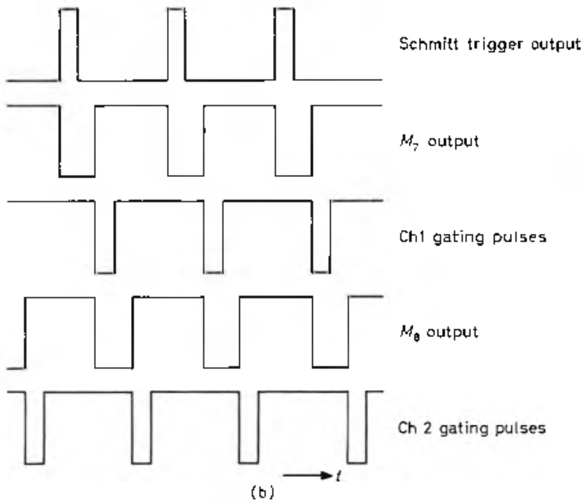
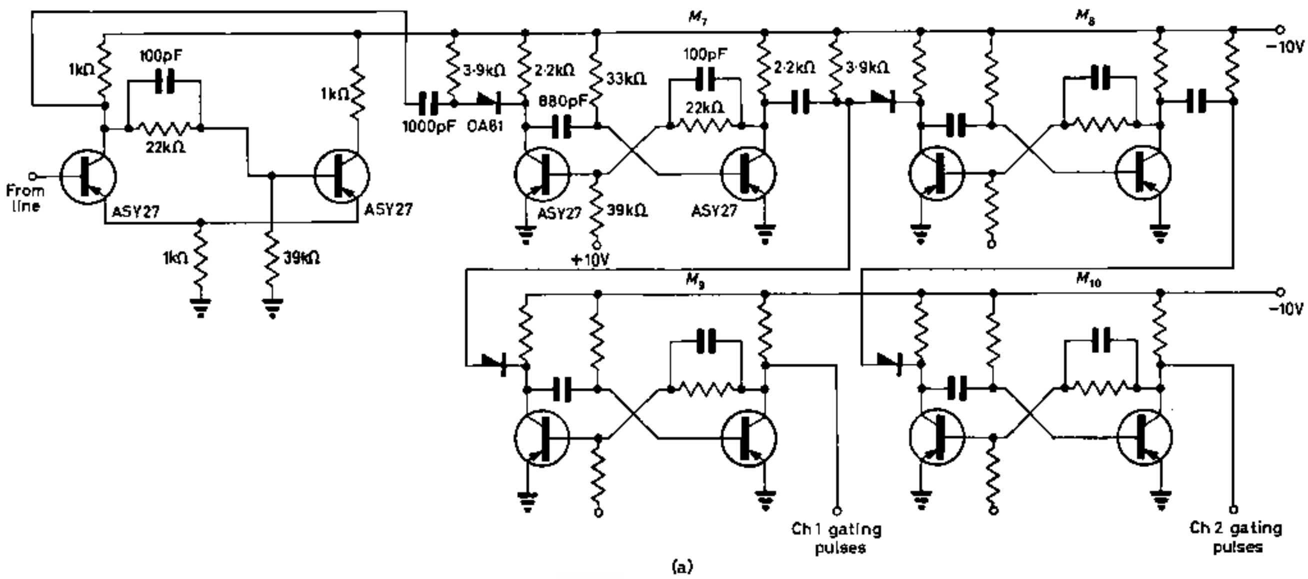


Fig. 12. Sync separator and gating pulse generator
(a) (below) The circuit (b) (above) The relevant waveforms



to extract synchronizing information from the line signal, some semblance of pulse shape must be retained, hence requiring a bandwidth greater than the theoretical minimum. If the criterion of bandwidth is taken as the frequency of the first null of the amplitude spectrum as before then the minimum bandwidth is nf_c.

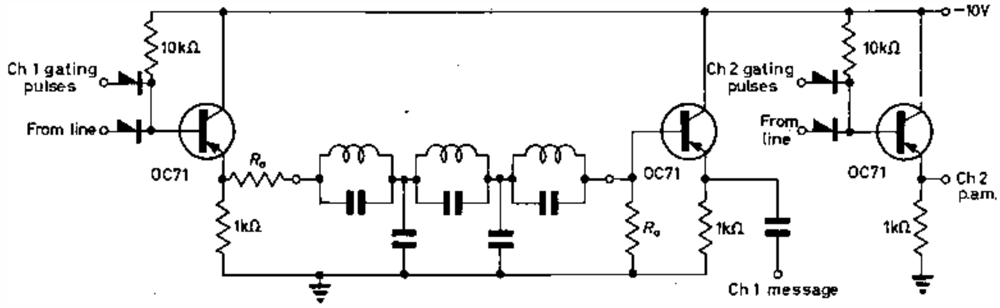


Fig. 13. De-multiplexing and demodulating unit

An Elementary System

The Transmitter

The basic units required to produce a multiplexed pulse amplitude modulated signal are shown in the block schematic of Fig. 7. The pulse train generator must produce n separate sets of sampling pulses, phased so that when they are combined in the multiplexing unit a correctly interlaced line signal is produced. In a commercial high capacity system, where time slots are very narrow and timing very precise, a shift register or ring counter, controlled by accurately timed shift pulses, would be used. In the elementary system used for demonstration purposes three channels are sufficient, two used as information channels and the third for synchronizing purposes. Timing is, therefore, not critical and a very satisfactory arrangement is obtained using a ring of monostable multivibrators driving subsidiary monostable circuits which act as individual sampling pulse generators. The circuit used, with associated waveforms is shown in Fig. 8. The channel bandwidth used extends to 3·4kc/s and the minimum sampling frequency must, therefore, exceed 6·8kc/s. Since only three channels are multiplexed, the actual sampling frequency can considerably exceed this figure and in order to simplify the design of the demodulating low-pass filter a frequency of 16·7kc/s was chosen. This results in a line signal of 10μsec pulses with 10μsec guard spaces. The circuit is not automatically self-starting but, once triggered, is self-sustaining.

The sampling and multiplexing circuits consist of resistor AND gates feeding a transistor OR gate as in Fig. 9. The synchronizing channel pulses are fed directly to line and are of greater amplitude than the maximum modulated amplitude of the information channels. The resulting line signal, with a 3·4kc/s bandwidth message on channel 1 and a 1·5kc/s sinusoidal tone on channel 2, is shown in Fig. 10(a).

The Receiver

The elementary block schematic of a receiver is shown in Fig. 11. A Schmitt trigger circuit is used to detect the synchronizing pulse train and drive the local monostable gating pulse generator as in Fig. 12. De-multiplexing is effected by gating these correctly phased pulses with the incoming line signal, Fig. 13, those pulses of the multiplexed signal coinciding with channel 1 gating pulses appearing at the output of channel 1 AND gate and similarly for channel 2. The channel 2 waveform after de-multiplexing is shown in Fig. 10(b).

The demodulating low-pass filter used in channel 1 has a relatively easy specification. The pass-band extends to 3·4kc/s and the stop band is from $16{\cdot}7 - 3{\cdot}4 = 13{\cdot}3$kc/s upwards, with a minimum discrimination of 40dB. A two-section m-derived image filter is satisfactory, with $f_c = 6{\cdot}8$kc/s, $m_1 = 0{\cdot}7$, $m_2 = 0{\cdot}85$.

A Simple Automatic Extremum Seeker

By P. Zorkoczy*, M.Sc.

A method of finding and tracking the optimum value of the output of a system is described. Information is extracted from the difference between the output $y(t)$ and the delayed output $y(t-\tau)$, and employed in a combination of simple analogue and switching elements to control the values of the input so as to attain the maximum output.

(Voir page 209 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 216)

A FREQUENT problem in control applications is to find and maintain that value of a controllable input variable $x(t)$ of a system which corresponds to the maximum (or minimum) value of the output $y(t)$. In the present article this general problem is restricted by assuming that (a) only one input to the system can be controlled, (b) at any instant there is only one maximum value of y for all permissible values of x and (c) the co-ordinates of this maximum are not constant but subject to random disturbances (Fig. 1).

The method involves obtaining information about the state of the system by observing the difference between the current value of $y(t)$ and its value τ seconds ago, $y(t-\tau)$ where τ is a suitably short time delay. A positive difference $D(t) = y(t) - y(t-\tau)$ will indicate that the peak is being approached, a negative $D(t)$ would show that the value of x is moving away from the optimum, while an approximately zero difference will imply that operation is in the region of the peak.

The block diagram of the extremum seeker is shown in Fig. 2. The drive for x is provided by an integrator whose

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input may be one of $+E$, $-E$ or 0 volts, the actual value depending on the state of the switching network. The operation is initiated by applying a small voltage pulse to the integrator or by utilizing the random disturbances present (N_x and N_y).

The function of the logic switching network (Fig. 3) is to apply the correct input to the integrator in the face of all possible disturbances affecting the x-y characteristic of the system. The operation of this network is based on the following rules:

(1) The input to the integrator is to be zero when the absolute value of $D(t)$ is less than a preset value A, and $+E$ or $-E$ otherwise. The choice of A determines the sensitivity in the y-direction.

(2) The polarity of the input to the integrator is to be reversed when $D(y)$ has been negative but greater in absolute value than A for more than T seconds. This safeguards against possible instabilities resulting from external disturbances.

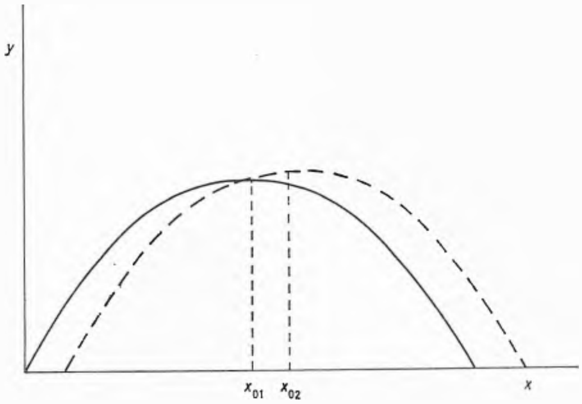


Fig. 1. System characteristic

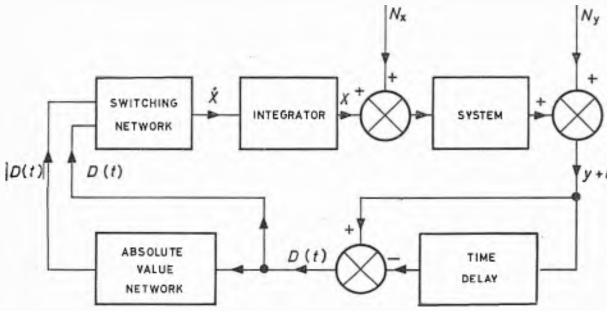


Fig. 2. Arrangement of the extremum seeker

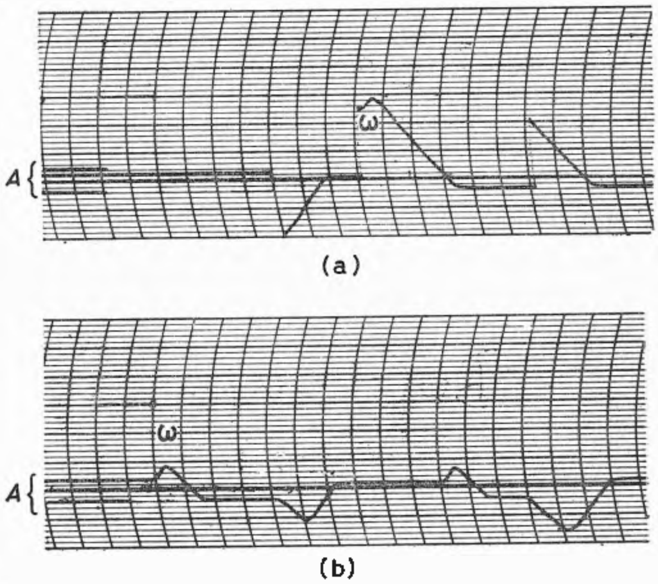


Fig. 4. Time variation of $(x - x_0)$ for step disturbances applied to (a) input and (b) output of the system

Horizontal scale 5sec; vertical scale 10V

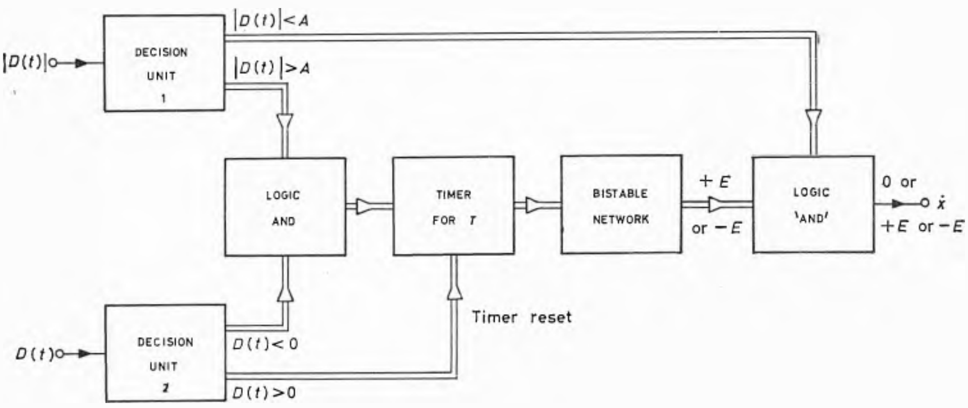
In an analogue computer realization of the extremum seeker the decision and switching operations can be carried out using fast comparator relay units, and the time delay may be produced by an operational amplifier network[1] or by tape recording techniques.

In selecting the parameters for a particular application, the time delay τ is made small but finite and E is then adjusted to a value just smaller than that resulting in hunting around the peak. This value of E will clearly depend on the shape of the x-y characteristic and on the choice of A. If now T is made slightly greater than τ it becomes possible to produce the correct response to disturbances which necessitate a change in the direction of sweep of x and to those requiring no such change.

Figs. 4(a) and 4(b) present the response of an experimental extremum seeker, programmed on a general-purpose analogue computer, to step disturbances applied to x and y respectively while in equilibrium at the peak. The x-y characteristic, simulated by a diode function generator, was of the form shown in Fig. 1, and the size of the steps was 20 per cent of the full range of each variable. The values of the parameters were as follows: Integrator time-constant = 1sec, τ = 1·2sec, T = 2sec, E = 2V, x-range = 70V, y-range = 30V. No dynamic lags were assumed.

Fig. 3. The switching network

Double lines denote binary variables



Acknowledgment

The numerous discussions with Dr. V. Zakian on the topic of optimization and on the presentation of this article are gratefully acknowledged. Thanks are also due to the staff of the Analogue Computing Laboratory.

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The Channel Electron Multiplier

By J. Adams* and B. W. Manley*

The channel electron multiplier consists of a continuous resistive dynode formed on the inner surface of a tube having a ratio of length to diameter of about 50. When a potential of about 2·5kV is applied between the ends of the tube, an electron entering the low potential end will initiate a cascade of secondary electrons from the wall. The number of resulting electrons emerging from the high potential end of the tube may exceed 10^6.

Such devices may be small and rugged, and are potentially useful in space applications. Various channel geometries have been investigated at the Mullard Research Laboratories, and it has been found that by making the channel multiplier of spiral form, pressure effects are avoided completely.

(Voir page 209 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 216)

UNTIL a few years ago direct study of the earth's outer atmosphere was impossible; information could only be gathered by indirect means. The development of the high powered rocket has changed this picture completely and it is now possible to send complete laboratories to study, for example, the various belts of energetic particles that surround the earth.

The payloads of such rocket laboratories are necessarily limited and so a demand has arisen for miniature, rugged versions of instruments for which size has never before been a limitation. Such instruments include particle detectors, multiplier tubes, mass spectrometers and so on. A particular requirement is for a small, robust, windowless electron multiplier.

The channel electron multiplier seems to be ideally suited to this need. It is so called because the multiplying process takes place within a tube or channel.

It was in 1930 that Farnsworth[1] first proposed an electron multiplier consisting simply of a resistive tube. By establishing a potential between the ends of the tube, the inner surface becomes a continuous dynode, replacing the separate dynodes used in a conventional electron multiplier. Secondary electrons generated at the low potential end of the channel by the incident radiation are accelerated down the tube until they strike the wall again. These electrons in turn generate more secondaries, and so a cascading process ensues which results in a large number of electrons emerging from the end of the tube. The process is shown schematically in Fig. 1. Many electron paths through the channel multiplier are possible, and the number of stages is thus indeterminate as it is in a multiplier with separate dynodes.

There is no evidence in fact that Farnsworth succeeded in making such a multiplier. He presumably failed because he was unable to obtain the required resistivity for the channel walls. Indeed many years were to elapse before the techniques of preparing suitable resistive coatings for the channel walls were successfully established. In 1960, Oschepkov[2] published details of a multiplier tube made of a special ceramic, and in 1962 Goodrich[3] described a multiplier of glass with a resistive coating on the inner surface of the tube.

Performance of Channel Multipliers

Typically such tubes are about 1mm diameter, and 5cm in length. Electron gain in excess of 10^5 is normally obtained with a potential difference of 2·5kV between the ends of the tube. The maximum continuous output current which can be drawn from a channel multiplier is usually a few microamperes. Geometrical scaling appears to be possible over a very wide range of dimensions. Providing the ratio of length to diameter of the tube exceeds about 30, the performance is largely independent of the actual size.

Whatever output current is drawn from the multiplier must be supplied by the wall, which is analogous to the resistor chain of the conventional electron multiplier. In practice it is found when using thin conductive coatings

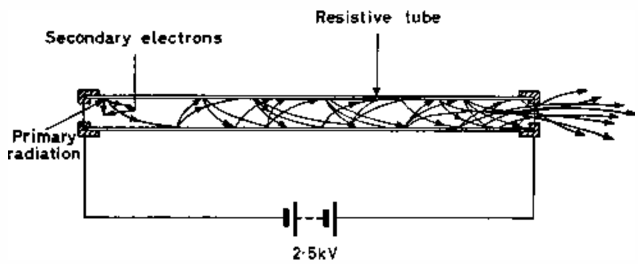
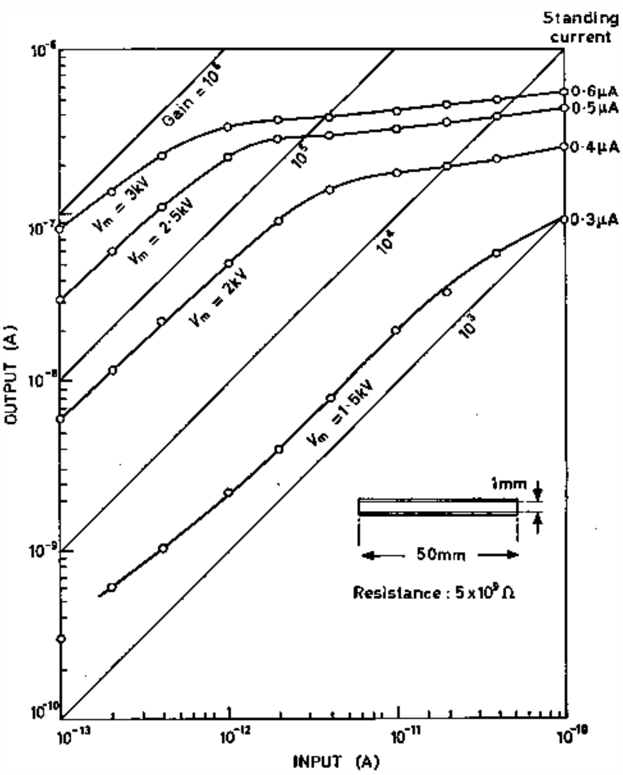


Fig. 1. Channel electron multiplier

Fig. 2. Characteristics of a channel electron multiplier



* *Mullard Ltd.*

that saturation effects are evident when the output current reaches about one half the standing current in the tube wall. This effect is apparent in the gain characteristics of a 1mm diameter channel multiplier shown in Fig. 2.

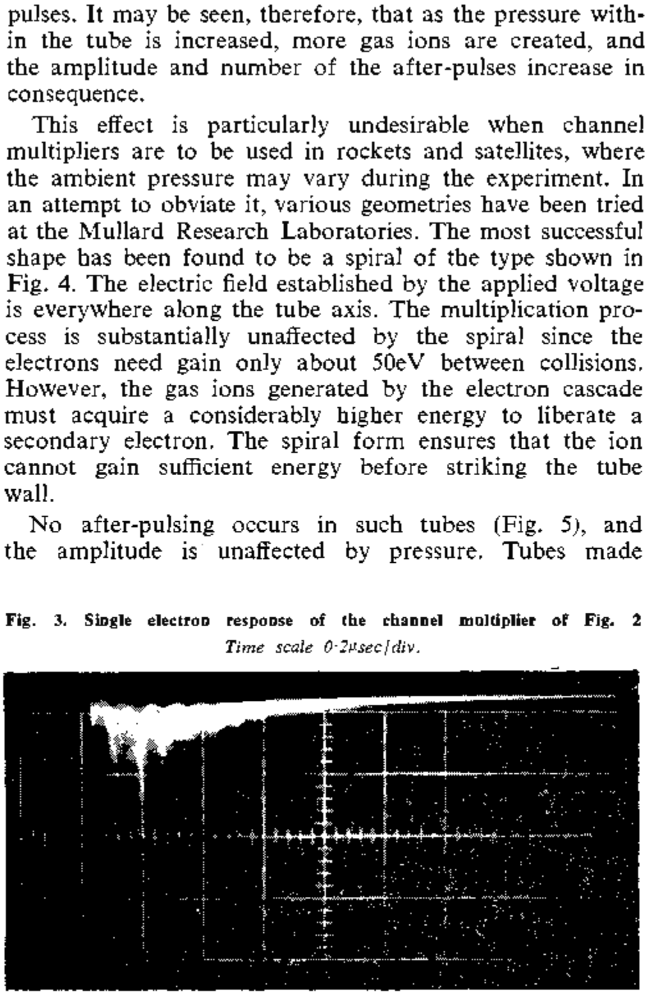
The channel multiplier is particularly well suited to detecting single electrons. The gain can exceed 10^8 for counting rates less than 10^3 electrons per second. Also, providing that the multiplier is carefully matched to its application the output pulses are nearly all equal in size; the pulse height distribution is extremely narrow. This is apparently due to a limitation of the maximum charge which can be drawn from the channel wall during a pulse. It arises from the potential distortions along the channel wall caused by the pulse itself. It is an extremely useful property when a multiplier is being used for pulse counting.

A less desirable feature of the channel multiplier in this form is that the charge multiplication factor is pressure dependent. A change in pressure of a factor two at 10^{-5}mm Hg can result in change in gain by a similar factor. A closer examination of the output pulse reveals that it is really a series of pulses which lasts for about 1μsec (Fig. 3). The pulse train is caused by ionic feedback. When the first cloud of electrons arrives near the output end of the channel it is sufficiently dense to ionize appreciable numbers of residual gas atoms. The positive ions so created drift back to the low voltage end of the tube where they may possibly collide with the wall. These collisions have a small but not negligible chance of producing secondary electrons, which then continue the multiplication process and give rise to further output pulses. It may be seen, therefore, that as the pressure within the tube is increased, more gas ions are created, and the amplitude and number of the after-pulses increase in consequence.

This effect is particularly undesirable when channel multipliers are to be used in rockets and satellites, where the ambient pressure may vary during the experiment. In an attempt to obviate it, various geometries have been tried at the Mullard Research Laboratories. The most successful shape has been found to be a spiral of the type shown in Fig. 4. The electric field established by the applied voltage is everywhere along the tube axis. The multiplication process is substantially unaffected by the spiral since the electrons need gain only about 50eV between collisions. However, the gas ions generated by the electron cascade must acquire a considerably higher energy to liberate a secondary electron. The spiral form ensures that the ion cannot gain sufficient energy before striking the tube wall.

No after-pulsing occurs in such tubes (Fig. 5), and the amplitude is unaffected by pressure. Tubes made at present have a resistance of about $10^{10}\Omega$, and the pulse height is maintained up to about 10^4 pulses per second.

Fig. 3. Single electron response of the channel multiplier of Fig. 2
Time scale 0·2μsec/div.



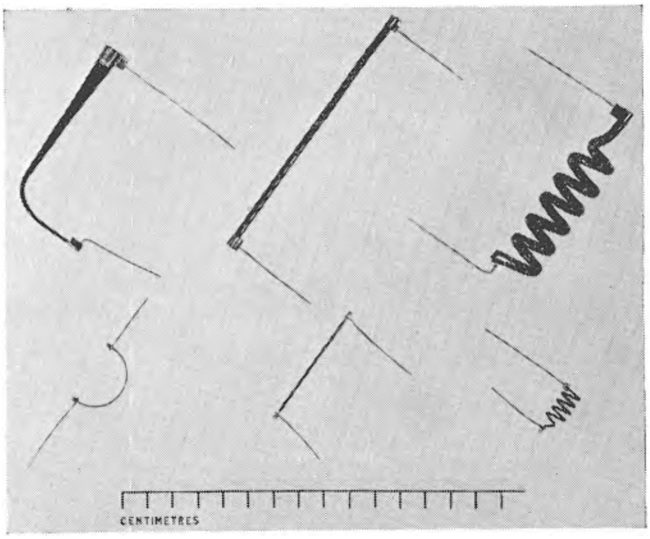


Fig. 4. Varied forms of channel multiplier. Spiral forms which obviate after-pulsing are shown on the right

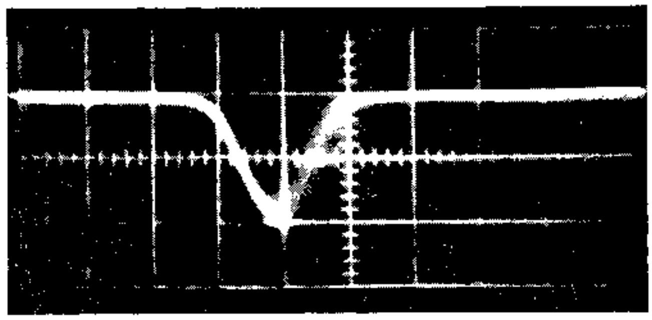


Fig. 5. Single electron response of spiral channel multiplier
Time scale 20nsec/div.

Applications

So far, the channel electron multipliers have been used only in space applications[4,5]. Although no absolute measurements of detection efficiency have been made, such multipliers are found to respond to β particles, gas ions, far ultra-violet, X-rays, and of course electrons over at least the energy range 50eV to 20keV. If a gold layer is evaporated at the mouth of the tube then sensitivity to ultra-violet is extended to about 0·25μ. In space applications the multiplier is open to the ambient atmosphere. A particular advantage of the channel multiplier in this respect, is that it is completely unaffected by exposure to normal atmospheres. Thus, unlike conventional multipliers, no precautions are necessary during the pre-launch period.

Multipliers of this type are not yet widely available. When they are, their particular features are certain to be of interest in the fields of mass spectrometry and ultra-violet spectrometry. They may also find applications as multiplying structures in conventional photomultipliers.

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Frequency Control of Astable Transistor Multivibrator Circuits

By S. L. Hurst*, B.Sc. (Eng.), A.M.I.E.E., A.M.I.E.R.E., and C. L. Seal†, Dip.Tech. (Eng.)

The frequency of usual collector-coupled astable transistor multivibrator circuits is normally sensibly independent of voltage. By minor circuit modifications this is no longer true, and the resultant frequency of oscillation may now be substantially linear with voltage. Possible uses for such modified circuits are suggested.

(Voir page 209 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 216)

THE circuit of the conventional symmetrical collector-coupled astable transistor multivibrator circuit using pnp transistors, is shown in Fig. 1 below. Analysis of this simple circuit gives the frequency of oscillation as

$$f = \left(\frac{1}{(C_1R_2+C_2R_3)\ln 2}\right)\text{c/s}, = \frac{1}{1{\cdot}38\,C_CR_B}\ \text{c/s}.$$

when $C_1 = C_2 = C_C$ and $R_2 = R_3 = R_B$. This analysis assumes that the collector supply voltage V_{CC} is much greater than the saturated collector-to-emitter and base-to-emitter transistor voltages $V_{CE(sat)}$ and V_{BE}, an assumption that is usually true.

The frequency of oscillation of this circuit is sensibly independent of the collector supply voltage V_{CC}, a feature which is of great value in many circumstances, and in general if frequency adjustment is required it is performed by mechanical means, varying C_1 and C_2, and/or R_2 and R_3.

Frequency adjustment of the free-running circuit may also be made by the well known technique of injecting pulses into the base of the transistors, such that conduction of each transistor during its 'off' half-cycle is caused to occur before the natural half-cycle time delay of the circuit. This technique, however, only allows the time per half-cycle to be shortened, i.e., the frequency to be increased above the natural frequency of the circuit, and does not permit plus-and-minus adjustment of the circuit frequency.

It may also be argued that as the frequency of the circuit is synchronized to the frequency of these injected pulses, then there is little point in using a free-running astable circuit, and a bistable circuit triggerd by the injected pulses may be preferable unless the free-running characteristic of the astable circuit is desirable in the absence of the injected or 'synchronizing' pulses.

Frequency adjustment of the astable circuit by means of a variable d.c. voltage instead of either component value variation or 'synchronizing' pulses is however quite feasible with little circuit modification. The following sections detail two simple ways of achieving this adjustment, one being a fairly wide-range adjustment with a moderately low input impedance presented to the adjusting d.c. voltage, the other being a more precise small-range 'trimming' adjustment, with a possibly high input impedance.

Wide Range Frequency Adjustment by Means of a Variable D.C. Voltage

The fundamental reason why the collector d.c. supply voltage V_{CC} does not materially influence the frequency of the circuit of Fig. 1 is because the base resistors R_2 and R_3 are themselves returned to this same d.c. supply. If R_2 and R_3 are taken to an alternative negative d.c. supply $-V_X$, see Fig. 2, then the resultant frequency of oscillation is dependent upon the value of both V_{CC} and V_X as well as the time-constants C_1R_2 and C_2R_3.

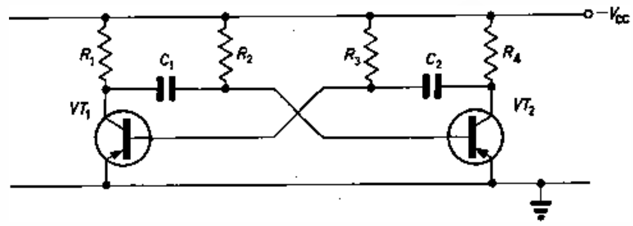

Fig. 1. The conventional symmetrical astable circuit

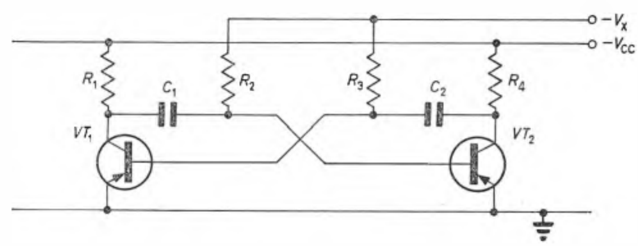

Fig. 2. Symmetrical astable circuit with separate d.c. base supply

The frequency of oscillation of this circuit may be shown to be approximately

$$f = \left[\frac{1}{2C_CR_B\ln(1+|V_{CC}/V_X|)}\right]\text{c/s},$$

(see Appendix A).

where $C_C = C_1 = C_2$, and $R_B = R_2 = R_3$.

The usual restrictions on the permissible values of R_B and V_X apply to this circuit, in order to ensure that each transistor remains fully 'on' during its appropriate half cycle. It may be shown (Appendix B) that the equation

$$(h_{FE} - 1) \geqslant V_{CC}/V_X\,((R_B/R_C) + 1)$$

should be satisfied, where $R_C = R_1 = R_4$, $R_B = R_2 = R_3$. In the majority of cases this equation may be further approximately simplified to $R_B \leqslant h_{FE}\,.\,R_C\,(V_X/V_{CC})$. Failure to maintain saturated conditions will cause the frequency stability of the circuit to be impaired, and in the extreme, the frequency of the circuit will increase with decreasing value of V_X due to the 'on'/'off' steps of collector voltage becoming smaller.

As is shown in Fig. 3, the variation of frequency with

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base supply voltage V_X is practically linear once voltage V_X exceeds about $\frac{1}{2}V_{CC}$, and of course provided fully saturated conditions on both transistors are being maintained. The graph for $R_B = 50\text{k}\Omega$ indicates that fully saturated conditions are not being maintained in the particular test circuit when V_X is less than about -3V; with $R_B = 100\text{k}\Omega$ a correspondingly higher minimum satisfactory limit for V_X is experienced. It may also be noted that there is no immediate upper voltage limit to which V_X may be increased, as there is with V_{CC}; the principal limit to V_X is one of transistor base-emitter *current* rating, V_X/R_B being the 'on' base current which must be held within acceptable limits.

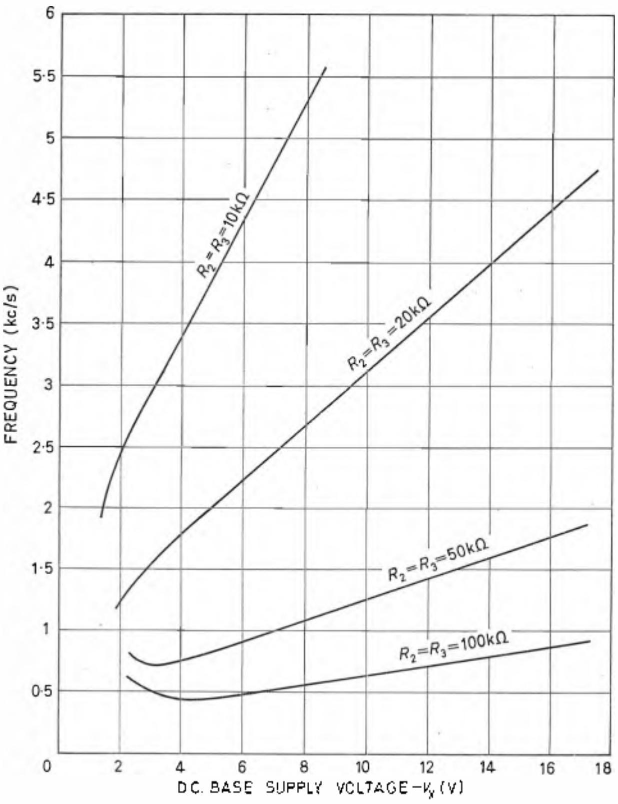

Fig. 3. 'Wide-range' circuit; variation of frequency with base supply voltage V_X, V_{CC} constant

($V_{CC} = 12\cdot0V$, $C_1 = C_2 = 0\cdot01\mu F$, $R_1 = R_4 = 2\cdot2\text{k}\Omega$)

The linear relationships between V_X and frequency shown in Fig. 3 agree very closely with theory, and give values for $\Delta f/\Delta V_X$ for the circuit as shown in Table 1. V_{CC} being held constant at $-12\cdot0$V, with $R_1 = R_4 = 2\cdot2\text{k}\Omega$, $C_1 = C_2 = 0\cdot01\mu$F.

The value of $\Delta f/\Delta V_X$ as well as varying with R_2 and R_3 also varies with C_1 and C_2, being inversely proportional to the value of C_1 and C_2, and also with the value of V_{CC}.

The input impedance of the circuit presented to the variable d.c. voltage supply V_X is approximately $(R_2 . R_3)/(R_2 + R_3), = \frac{1}{2}R_B$ when $R_2 = R_3 = R_B$. This is not in general a very high impedance, only say tens of thousands of ohms, and thus this circuit may be inconvenient to use if supply V_X must not be loaded appreciably.

Small Range 'Trimming' Frequency Adjustment by Means of a Variable D.C. Voltage

In the circuit shown in Fig. 4, the negative collector supply voltage V_{CC} provides the majority of the required base current drive for VT_1 and VT_2, via resistors R_2 and R_3 as in the normal astable circuit. The presence of resistors R_5 and R_6 however modifies the discharge impedance seen by the capacitors C_1 and C_2 during each respective 'off' half-cycle, and also the effective d.c. terminal voltage towards which C_1 and C_2 commence their exponential decay.

The frequency of oscillation of this circuit may be shown to be approximately

$$f = \left[\frac{1}{2C_C R_B' \ln(1 + |V_{CC}/V'|)}\right] \text{c/s},$$

where $C_C = C_1 = C_2$ as before,

$$R_B' = \frac{R_2R_5}{R_2 + R_5} = \frac{R_3R_6}{R_3 + R_6}$$

and

$$V' = \left[\frac{V_{CC} + V_X . (R_2/R_5)}{1 + (R_2/R_5)}\right] = \left[\frac{V_{CC} + V_X . (R_3/R_6)}{1 + (R_3/R_6)}\right]$$

(see Appendix C).

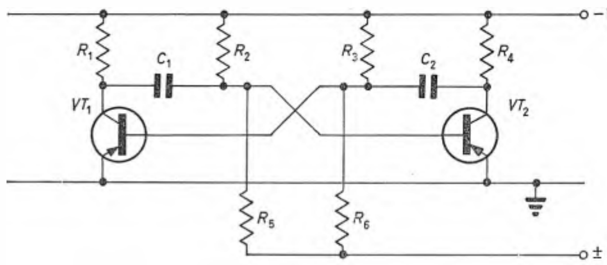

Fig. 4. Symmetrical astable circuit with 'trimming' frequency adjustment voltage V_X

The sign of V' must be negative in this circuit or else each transistor will remain steadily and permanently biased 'off'. Voltage V_{CC} is inherently a negative supply voltage but V_X may be positive or negative as desired. The appropriate signs for V_{CC} and V_X must therefore be inserted in the above expression for V' in order to obtain the correct numerical value for this negative voltage V'.

Now if $R_5 = R_6 \gg R_2 = R_3$, thus giving the relatively high input impedance of approximately $R_5R_6/(R_5 + R_6) = \frac{1}{2}R_5 = \frac{1}{2}R_6$, the effect of V_X on the frequency of the circuit will be such as to give a relatively low value for $\Delta f/\Delta V_X$, i.e. V_X acts as a sensitive 'trimming' control on the frequency of the astable circuit.

Fig. 5 indicates that for various values of resistor $R_5 = R_6$, a substantially linear relationship exists between frequency and trimming voltage V_X. The measured and theoretical values of $\Delta f/\Delta V_X$ are in close agreement, typical values are shown in Table 2 for a circuit with $V_{CC} = -12$V, $R_1 = R_4 = 2\cdot2\text{k}\Omega$, $R_2 = R_3 = 50\text{k}\Omega$.

The reduction in frequency variation with increase of R_5, R_6, can be partly offset by working the circuit at a

TABLE 1

$R_2=R_3$ (kΩ)	$\frac{\Delta f}{\Delta V_X}$ (c/s/V)
10	480
20	220
50	88
100	42

TABLE 2

$R_5=R_6$ (kΩ)	$\frac{\Delta f}{\Delta V_X}$ (c/s/V)
50	95
100	47
250	15·4
500	8·4

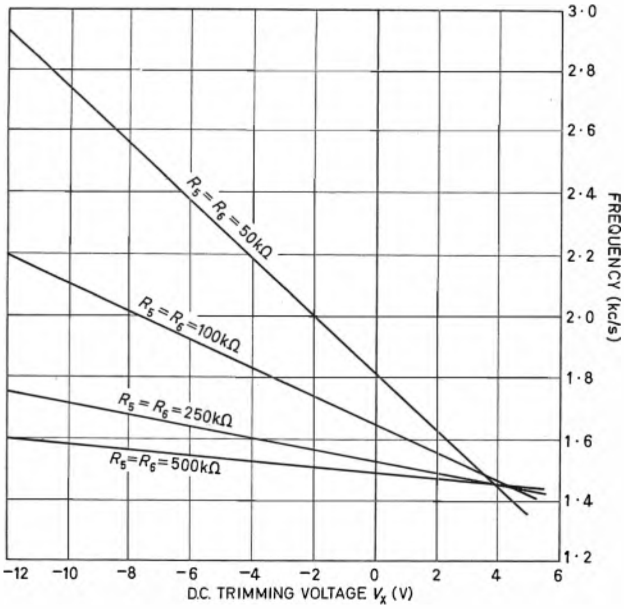

Fig. 5. 'Small-range' circuit; variation of frequency with 'trimming' voltage V_X, V_{CC} constant

($V_{CC} = 12{\cdot}0V$, $C_1 = C_2 = 0{\cdot}01\mu F$, $R_1 = R_4 = 2{\cdot}2k\Omega$, $R_2 = R_3 = 50k\Omega$)

lower value of collector supply voltage V_{CC}. The penalty paid for lowering V_{CC} is that variations of leakage current and base-emitter saturation voltage with temperature now cause the frequency of the circuit to be more temperature dependent.

Possible Applications

The substantially linear relationship between Δf and ΔV_X for both the above adjustable frequency circuits suggests that possible use may be made of these circuits in circumstances where a frequency source varying with a d.c. voltage level is required. It was not the object of the project from which the above results are taken to pursue this applications side to any degree, and the remarks following are thus introductory rather than comprehensive.

The possible use of one or other of the above circuits used in conjunction with a digital frequency meter to give a digital presentation of the unknown voltage V_X was briefly considered. With the first 'wide range' circuit, linearity only holds for V_X greater than about $-6{\cdot}0$V and thus using this circuit puts an immediate limitation on the lower voltage that may successfully be allowed.

Consider, however, this 'wide range' circuit. If circuit values can be adjusted such that $\Delta f/\Delta V_X$ = say 100c/s/V with V_{CC} constant, and the frequency when $V_X = -6{\cdot}0V = 600$c/s, then presenting this frequency on a digital frequency meter and 'inserting' a decimal point between the hundreds and the tens digit will give a reading equal to the value of voltage V_X. Unfortunately as both $\Delta f/\Delta V_X$ and the frequency are dependent upon V_{CC}, V_X, $R_2 = R_3$ and $C_1 = C_2$, the solution to the resultant equations requires V_{CC} to be a very low voltage, (approximately $-0{\cdot}5$V), to give the required values for $\Delta f/\Delta V_X$ and frequency, when $V_X = -6{\cdot}0$V. The suggested circuit arrangement is thus impractical.

Turning, however, to the second 'small range' circuit, one immediate advantage is that the circuit is approximately linear from V_X = zero. If the circuit is thus preset by suitable adjustment of V_{CC}, $R_2 = R_3$, $R_5 = R_6$, and $C_1 = C_2$ so that with V_X = zero, the frequency f_o is, say, 1 000c/s, and $\Delta f/\Delta V_X$ is, say, 10 or 100c/s/V, then a digital presenation of V_X from zero upwards becomes a possibility. Many variations of the above parameters are possible in this circuit to give the desired value for $\Delta f/\Delta V_X$ and f_o; for example $V_{CC} = -12$V, $C_1 = C_2 = 0{\cdot}015\mu$F, $R_2 = R_3 = 79{\cdot}1k\Omega$, and $R_5 = R_6 = 31{\cdot}2k\Omega$ is one solution which gives $f_o = 1\,000$c/s and $\Delta f/\Delta V_X$ = 100c/s/V. However, although this simple circuit arrangement gives a useful presentation of voltages, from zero up to 9·99V in the above case, its stability and linearity does not offer any major attraction. Possible refinements of this basic circuit may improve the performance to more acceptable standards.

The most practical use of the above circuits, and in particular the circuit of Fig. 2, is possibly as frequency sources in control systems, where V_X is a derived voltage dependent upon the difference between the frequency of the astable circuit and some other reference signal. If the higher impedance circuit of Fig. 4 is employed, and if V_X is smoothed by some reasonable sized capacitor, then momentary loss or mutilation of the reference signal may be successfully accommodated without any material change in the automatically trimmed frequency of oscillation of the astable circuit. The circuit thus acts as a form of 'flywheel' circuit, a possible advantage in some control systems that are subject to occasional noise or other interference.

Acknowledgments

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Appendix

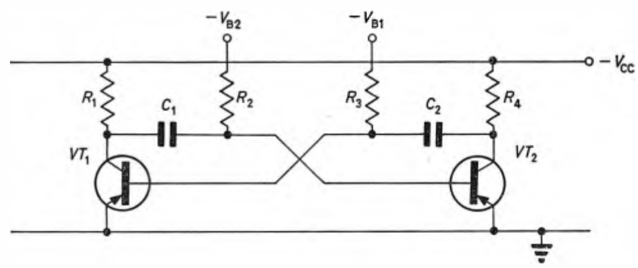
(A) Frequency of Oscillation

Consider the circuit of Fig. 6.

A full analysis of this circuit gives expressions for t_1 = the time transistor VT_1 is 'off', and t_2 = the time transistor VT_2 is 'off', as follows, provided the collector circuit time-constants C_1R_1 and C_2R_4 are both small in comparison with the base circuit time-constants C_1R_2 and C_2R_3 (i.e., provided the collector voltages recover substantially to the steady-state 'off' voltage level during each respective half-cycle:

$$t_1 = T_1 \ln \left[\frac{-V_{B1} - R_3 \, . \, I_{BEX} - \dfrac{C_2}{C_2 + C_{BE}}(V_{CC} - V_{CE(sat)})}{-V_{B1} - R_3 \, . \, I_{BEX} + V_{CE(sat)}} \right]$$

Fig. 6. Standard astable circuit, with separate base supply voltages



and:

$$t_2 = T_2 \ln \left[\frac{-V_{B2} - R_2 \, . \, I_{BEX} - \dfrac{C_1}{C_1 + C_{BE}}(V_{CC} - V_{CE(sat)})}{-V_{B2} - R_2 \, . \, I_{BEX} + V_{CE(sat)}} \right]$$

where T_1 = time-constant = $(C_2 + C_{BE}) \, . \, R_3$,

T_2 = time-constant = $(C_1 + C_{BE}) \, . \, R_2$,

C_{BE} = transistor base-emitter capacitance when reverse-biased,

I_{BEX} = transistor base-emitter leakage current when reverse-biased,

$V_{CE(sat)}$ = transistor saturated collector-emitter voltage.

Now for most circuits, C_1 and $C_2 \gg C_{BE}$, and $V_{CE(sat)} \ll V_{CC}$. Also as I_{BEX} is small, $R_2 \, . \, I_{BEX}$ and $R_3 \, . \, I_{BEX}$ may be neglected. Thus the above times t_1 and t_2 simplify to:

$$t_1 = C_2 R_3 \ \ln \left[\frac{-V_{B1} - V_{CC}}{-V_{B1}} \right] \text{sec}$$

and:

$$t_2 = C_1 R_2 \ln \left[\frac{-V_{B2} - V_{CC}}{-V_{B2}} \right] \text{sec}$$

Thus the frequency of oscillation of the circuit, $= 1/(t_1 + t_2)$ c/s, may be obtained.

When the base resistors are both taken to the same base supply voltage, i.e., $V_{B1} = V_{B2} = V_X$, and when $C_1 = C_2 = C_C$, $R_2 = R_3 = R_B$, then the frequency of the circuit becomes:

$$f = \left[\frac{1}{2 C_C R_B \ln \left((-V_X - V_{CC}) / -V_X \right)} \right] \text{c/s}$$

$$= \left[\frac{1}{2 C_C R_B \ln \left(1 + |V_{CC}/V_X| \right)} \right] \text{c/s.}$$

When $V_X = V_{CC}$, this final equation simplifies to the commonly quoted expression for a conventional astable transistor circuit, namely:

$$f = \frac{1}{2 C_C R_B \ln 2} = \frac{1}{1{\cdot}38 \, C_C R_B} \text{ c/s.}$$

In spite of the appreciable number of simplifications made above in deriving these final formulae, the latter nevertheless remain accurate to within, say, 3 per cent for circuits where C_1 and $C_2 > 0{\cdot}0001 \mu$F, and V_X and $V_{CC} > -6$V.

(B) 'On' Time

Consider the conditions surrounding transistor VT_1 (Fig. 7) just after VT_1 has been switched 'on'.

The collector and base currents after switching 'on' at time $t = 0$ are as shown in Fig. 8.

At time $t = 0$, both the collector and the base currents of VT_1 are greater in magnitude than the final steady-state values. The worst state of affairs for holding VT_1 saturated is if the above base circuit time-constant is smaller than the collector circuit time-constant.

Taking the extreme case where $C_3 R_4 \ll C_1 R_2$, then the base current has fallen to its steady-state value of $i_B = -((V_X - V_{BE})/R_3)$, while the collector current is still approximately its peak value at $t = 0$. This peak value of collector current, neglecting the effects of collector-base capacitance and noting that the voltage existing across capacitor C_1 immediately prior to time $t = 0$ is $-(V_{CC} - V_{BE})$, is thus approximately

$$i_{C(max)} = -\left[\frac{V_{CC} - V_{CE(sat)}}{R_1} + \frac{V_X + (V_{CC} - V_{BE}) - V_{CE(sat)}}{R_2} \right]$$

Now $h_{FE} \geqslant i_{C(max)}/i_B$

$\therefore$ Neglecting the small V_{BE} and $V_{CE(sat)}$ terms in the above equations:

$$h_{FE} \geqslant \left[\frac{V_{CC}/R_1 + V_X/R_2 + V_{CC}/R_2}{V_X/R_2} \right]$$

Therefore:

$$h_{FE} \geqslant [V_{CC}/V_X \left((R_2/R_1) + 1 \right) + 1]$$

or $(h_{FE} - 1) \geqslant V_{CC}/V_X \left((R_2/R_1) + 1 \right)$.

(C) 'Off' Conditions

Considering the circuit of Fig. 4 with VT_2 'off'.

With VT_2 'off', its presence in the circuit may be neglected, and thus the circuit of R_2 and R_5 to the right

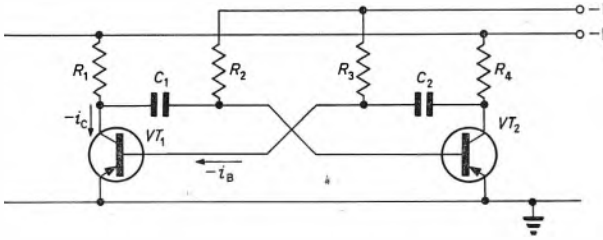


Fig. 7. Transistor VT_1 conducting, V_2 non-conducting; dominant currents = collector and base currents of VT_1

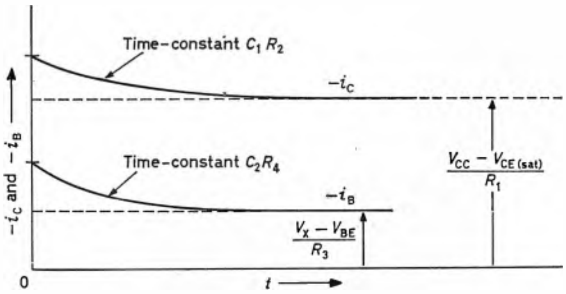


Fig. 8. VT_1 collector and base currents, following switch-over to VT_1 conducting/VT_2 non-conducting

of capacitor C_1 may be replaced by a single equivalent resistor of value $(R_2 \, . \, R_5/R_2 + R_5)\Omega$, connected to a d.c. source of voltage of value $V_{CC} - (V_{CC} - V_X)\left(R_2/(R_2 + R_5)\right)$ volts.

Similarly when VT_1 is 'off', the circuit of R_3 and R_6 to the left of C_3 may be replaced by an equivalent resistor of $R_3 R_6/(R_3 + R_6)\Omega$ and voltage $V_{CC} - (V_{CC} - V_X)\left(R_3/(R_3 + R_6)\right)$ volts.

These equivalent values may now be placed in the general frequency equation of Appendix A, thus giving the result:

$$f = \left[\frac{1}{2 C_C R_B' \ln \left(1 + |V_{CC}/V'| \right)} \right]$$

where $R_B' = \dfrac{R_2 \, . \, R_5}{R_2 + R_5} = \dfrac{R_3 \, . \, R_6}{R_3 + R_6}$

and $V' = V_{CC} - (V_{CC} - V_X)\left(\dfrac{R_2}{R_2 + R_5} \right)$

$$= V_{CC} - (V_{CC} - V_X)\left(\frac{R_3}{R_3 + R_6} \right)$$

$$= \left[\frac{V_{CC} + V_X (R_2/R_5)}{1 + (R_2/R_5)} \right] = \left[\frac{V_{CC} + V_X (R_3/R_6)}{1 + (R_3/R_6)} \right]$$

Single and Matched Pair Transistor Choppers

By A. Lydén*

The refined theory using the modified Ebers's and Moll's relations for the transistor chopper is reviewed. Equations are given for both the steady state and transient behaviour. Matched pair switches have smaller offset voltage, smaller spikes and lower drift than the single transistor switch. The two-transistor-in-parallel configuration has inherently the lowest drift.

(Voir page 209 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 216)

BRIGHT[1,3] and Kruper[2,3] showed the suitability of the transistor as a low level switch, and found that the offset voltage V_{EC} in the 'on' state is smallest in the inverted or common collector connexion shown in Fig. 1.

The author[4] has given a physical explanation of why V_{EC} is so small in the inverted connexion and presented a more complete theory for the low level transistor switch. Firstly this theory will be shortly reviewed.

Switch Theory

Modification of Ebers's and Moll's Relations

To determine the characteristics of the transistor as a switch equations are needed describing the d.c. signal

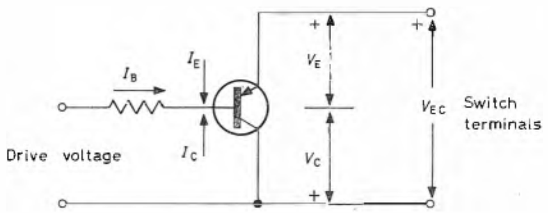


Fig. 1. The common collector transistor switch

behaviour of the transistor. Ebers's and Moll's[5] relations contain four constants which vary with the working point, namely the collector cut-off current I_{CBO}, the emitter cut-off current I_{EBO}, the common base normal direct current gain A_N and the common base inverted direct current gain A_I.

However the relations can be modified to include only two variable constants, namely A_N and A_I. These modified relations are:

$$I_E/I_o = -(1/A_N)\,[\exp(\gamma V_E)-1]+[\exp(\gamma V_C)-1] \quad \ldots\ldots\ldots\ldots (1)$$

$$I_C/I_o = \exp(\gamma V_E) - 1 - (1/A_I)\,[\exp(\gamma V_C) - 1] \quad \ldots\ldots\ldots\ldots (2)$$

where $\gamma = q/kT$

q = magnitude of electron charge

k = Boltzmann constant

T = absolute temperature

I_o = Sah's cut-off current

Sah[6] has shown that I_o does not vary with I_C and I_E. I_{EBO} and I_{CBO} are given by equations (3) and (4).

$$I_{EBO} = \frac{1 - A_N A_I}{A_N}\, I_o \quad \ldots\ldots\ldots\ldots (3)$$

$$I_{CBO} = \frac{1 - A_N A_I}{A_I}\, I_o \quad \ldots\ldots\ldots\ldots (4)$$

A_N is dependent on V_E, but rather little on V_C and A_I is dependent on V_E, but rather little on V_E. When $V_E > (1/\gamma)$ and $V_C \leqslant 0$ then for fairly small values of I_C (< 1mA):

$$B_N = \frac{A_N}{1 - A_N} \simeq \text{constant} \,.\, I_C^{a} \quad 0 \leqslant a \leqslant 0{\cdot}5$$

where B_N = common emitter d.c. gain.

Alloy and Mesa silicon transistors normally have $a = 0{\cdot}4$ to $0{\cdot}5$. That means $B_N < 0{\cdot}01$ when $V_E < (1/\gamma)$ Planar silicon transistors normally have $a = 0{\cdot}2$ to $0{\cdot}3$ when $V_E < (1/\gamma)$, hence $B_N < 1$.

The ON State

The offset voltage V_{EC} is (pnp):

$$V_{EC} = (1/\gamma) \ln A_N \quad \ldots\ldots\ldots\ldots\ldots (5)$$

$$V_{EC} \simeq -(1/\gamma B_N) \text{ if } B_N \gg 1 \quad \ldots\ldots\ldots\ldots (6)$$

The correct value of B_N for equation (6) is that measured at $V_C = 0$ and at emitter current I_E':

$$I_E' = -\frac{A_I}{1 - A_I}\, I_B = -B_I I_B \quad \ldots\ldots\ldots\ldots (7)$$

where I_B is the base current in the 'on' state. The correct value of B_I in equation (7) is that measured at $V_E = 0$ and at collector current I_C':

$$I_C' = -(1 + B_I)\, I_B \quad \ldots\ldots\ldots\ldots\ldots (8)$$

Since equation (8) contains the quantity asked for a trial calculation is necessary. (In the derivation of equations (7) and (8) it has been assumed $B_N \gg 1$.)

The 'on' resistance R_S is:

$$R_S = \frac{1}{\gamma\,|I_B|}\left(1/B_I + \frac{1}{1 + B_N}\right) \ldots\ldots\ldots\ldots (9)$$

where correct values of B_N and B_I are those mentioned above.

At larger I_B ($>$ 1mA) equation (5) does not hold. The measured value of V_{EC} becomes greater than the calculated value. This is partly due to the collector bulk resistance r_C which gives a contribution to V_{EC} which $= r_C I_B$. However, the main reason is that the linear diffusion equation is not valid at high excess hole densities in the base, which means equations (1) and (2) are not valid.

In a chopper amplifier the drift is mainly due to the variation of V_{EC}. The temperature variation can be determined by differentiation of equation (6).

$$dV_{EC}/dT = V_{EC}\,[(1/T) - (1/B_N)\,.\,(dB_N/dT)] \quad \ldots . (10)$$

Sometimes it is possible to find an I_B which gives $dV_{EC}/dT = 0$.

The variation in V_{EC} due to variation in B_N is at constant temperature:

$$dV_{EC}/V_{EC} = -(dB_N/B_N) \quad \ldots\ldots\ldots\ldots (11)$$

The OFF State

Assume that the junctions are reversed biased with voltages much larger than $1/\gamma$. Equation (1) gives the offset current I.

$$I = -I_E = -(I_o/B_N) \quad \ldots\ldots\ldots\ldots (12)$$

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If the reverse voltages are less than $1/\gamma$ expansion of equation (1) gives:

$$I = \gamma V_C (I_o/B_N) \text{ if } V_C \approx V_E \text{ .... } (13)$$

The offset current is hence reduced by the factor $|\gamma V_C|$. The 'off' resistance R_o when the reverse voltages are zero is:

$$R_o = \frac{A_N}{\gamma\,|I_o|} \text{ .... } (14)$$

If:

$$\frac{1}{R_B \gamma\,|I_o|} + (1/B_I) \gg (1/B_N)$$

which is mostly the case.

The equivalent circuit of the switch is shown in Fig. 2.

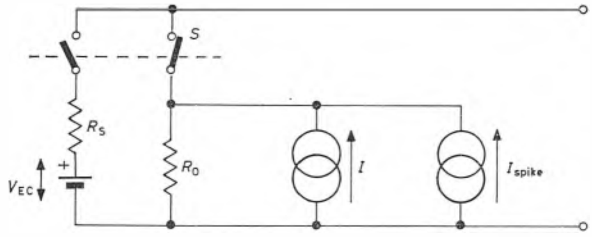
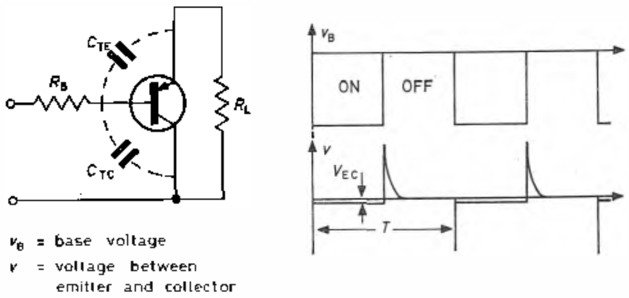
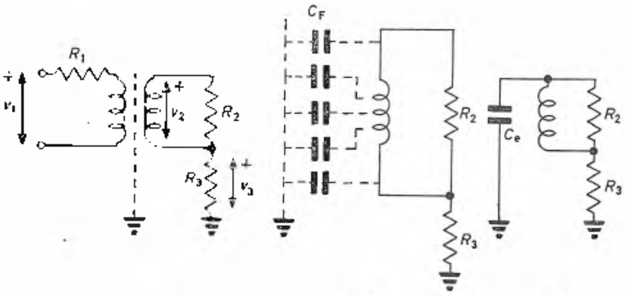


Fig. 2. The equivalent circuit of the low level switch



(a) (b)

Fig. 3. Spikes appear over the switch at switching on and off



(a) (b) (c)

Fig. 4. (a) Drive circuit with a transformer
(b) There is a distributed capacitance C_F between the screen and the secondary winding
(c) C_F replaced by an equivalent capacitance C_e

The Behaviour at Switching On and Off

Fig. 3 shows the circuit assumed and the voltage response v over R_L. The spikes are mainly due to the base-emitter depletion-layer capacitance C_{TE}. (Silicon alloy transistors have high V_{EC} at low I_B. This gives a contribution to the spikes. However, these spikes are very short if the drive voltage has a good square form.) An analysis of the shape of the spikes is as a rule of minor interest. In most applications it is, instead, the area of the spikes which is of importance.

Switching Off

The area of the spike at switching off $\phi_{sp(off)}$ is:

$$\phi_{sp(off)} = R_L \,.\, C_{TE} \,.\, V_E' \text{ .... } (15)$$

where:

$$V_E' = V_E - (1/\gamma) \ln \frac{R_B + R_L}{R_S} \text{.......... } (16)$$

V_E is the emitter voltage immediately before switching off. Equation (13) states that $\phi_{sp(off)}$ is independent of the cut-off frequency f_α. This is true if $f_\alpha >$ a few Mc/s. In a chopper amplifier $\phi_{sp(off)}$ will cause a faulty voltage $V_{sp(off)}$ in the same way as V_{EC} does:

$$V_{sp(off)} = 2f_o \,.\, \phi_{sp(off)} \text{ .... } (17)$$

Mostly $V_E' \simeq 0{\cdot}4$V. Since V_E' varies little with R_L, $V_{sp(off)}$ is roughly directly proportional to R_L. In the equivalent circuit it is therefore convenient to use a current I_{spike} defined by:

$$I_{spike} = 2f_o \,.\, C_{TE} \,.\, V_E' \text{ .... } (18)$$

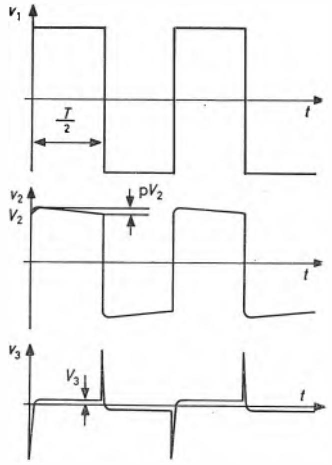


Fig. 5. Waveforms in the drive circuit Fig. 4(a)

Switching On

The voltage spike at switching on has an area $\phi_{sp(on)}$ which as a rule is much smaller than $\phi_{sp(off)}$. $\phi_{sp(on)}$ increases with C_{TE} and decreases with f_α, but is almost independent of R_L. It is convenient to define a voltage $V_{sp(on)}$.

$$V_{sp(on)} = 2f_o \,.\, \phi_{sp(on)} \text{ .... } (19)$$

Mostly $V_{sp(on)} \ll V_{EC}$, and can be neglected.

Chopper Transistor Properties

A chopper transistor should have the following properties:

(1) Stable and high B_N (also at low I_E)

(2) B_I ought to be > 1

(3) Small r_C

(4) Small I_o (silicon transistors preferable)

(5) Small C_{TE}

(6) $f_\alpha >$ a few Mc/s.

Matched Pair Switches

The imperfections of the low level switch can partly be eliminated in different ways. It is desirable to decrease especially V_{EC} and the spikes. Switches have earlier been presented which cancel V_{EC} by creating a voltage $-V_{EC}$ in series with the switch in the 'on' state. This does not decrease the drift (dV_{EC}), but is only a zero displacement and this can be done in a simpler way. Here two types of matched pair switches which have smaller drift than the variation in V_{EC} will be described and analysed.

THE INFLUENCE OF THE CAPACITANCES OF THE TRANSFORMER

Matched pair switches sometimes contain a transformer. This gives rise to a certain problem, which will be treated separately here. The transformer is in general included in a drive circuit as shown in Fig. 4(a). If the transformer had no capacitances the voltage v_3 would be zero. Now this is not the case and therefore $v_3 \neq 0$. Capacitive transmission directly between the windings can mainly be eliminated by a screen between the windings as shown in Fig. 4(a). However, this is not sufficient to get $v_3 = 0$. The secondary winding has a certain distributed capacitance to earth. This is illustrated by C_t in Fig. 4(b). C_t can be replaced by an equivalent capacitor $C_e < C_t$ connected between earth and one end of the secondary winding as shown in Fig. 4(c).

Assume now that the drive voltage is a square wave such as v_1 in Fig. 5. The secondary voltage v_2 has a certain tilt of the top*. Between the steps v_3 is a constant voltage $= v_3$.

$$V_3 = C_e R_3 (du_2/dt) \approx C_e R_3 \frac{s \,.\, V_2}{T} \,.\, 2 \quad \ldots\ldots\ (20)$$

where s is the percentage tilt of the top of v_2.

Suppose that $C_e = 20\text{pF}$, $T = 1\text{msec}$, $R_3 = 100\text{k}\Omega$, $p = 0{\cdot}05$ and $V_2 = 5\text{V}$. This gives $V_3 = 1\text{mV}$. R_3 is often equal to R_L. V_3 will therefore appear across the switch in the 'off' state and 1mV will be much too great to be tolerated. The spikes in v_3 might also be intolerably large.

Two Transistors in Series[2]

Fig. 6 shows some examples of how this switch can be realized. (All quantities will in the sequel be indexed 1 or 2 showing that they belong to transistor No. 1 or No. 2.) If $B_{N1} \simeq B_{N2}$ and they change, a certain dV_s will arise, which is:

$$dV_s = -V_{EC}\left((dB_{N1}/B_{N1}) - (dB_{N2}/B_{N2})\right) \quad \ldots.\ (21)$$

Equation (21) should be compared with equation (11).

$$dV_s = 0 \text{ if } (dB_{N1}/B_{N1}) = (dB_{N2}/B_{N2})$$

Circuit (a) is the one usually described. If $R_{B1} = R_{B2} = 0$ the bases are directly connected. This gives $V_{C1} = V_{C2}$, but not necessarily $I_{B1} = I_{B2}$. If $I_{01} = I_{02}$ the excess hole densities in the bases are equal. This gives the best chances of B_{N1} being equal to B_{N2}, and hence the best matching. Circuit (b) can be realized very conveniently with a newly presented transistor which has two emitters, as shown. In circuit (c) there is no transformer, but the switch must be followed by an a.c. differential amplifier. The switch drives a current V_{EC1}/R_{L1} through the source in the 'on' state. This has in general no significance. The spikes are subtracted in the differential amplifier $C_{TE1} = C_{TE2}$ and $R_{L1} = R_{L2}$.

If terminal No. 2 in circuits (a) or (b) is earthed, a certain charge current to C_t goes through R_{L2}. Furthermore the spikes will not eliminate each other owing to the capacitances of the transformer.

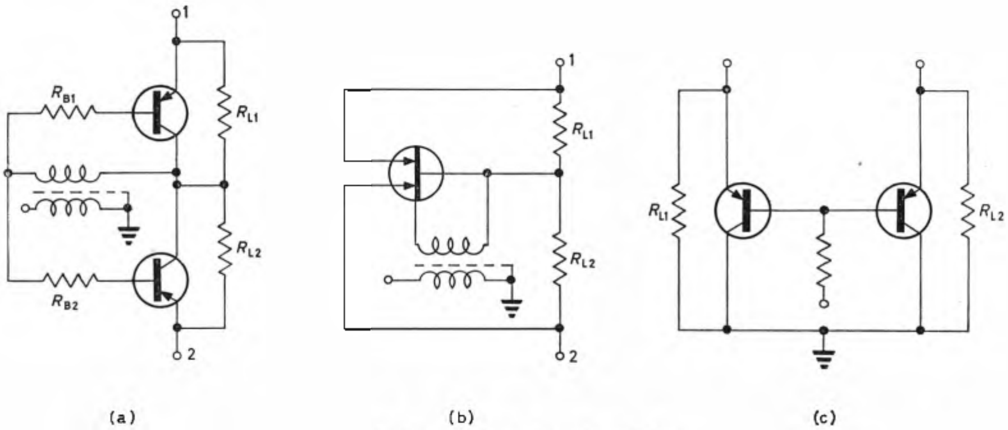


Fig. 6. Two transistors in series switches

Two Transistors in Parallel[8]

This switch can be realized in two ways as shown in Fig. 7. In this case the transistors in the 'on' state drive a current through each other, so $V_s = 0$. If one terminal of the switch in Fig. 7(a) is earthed, charge current to C_t will give a certain voltage across R_L. The circuit with complementary transistors in Fig. 7(b) contains no transformer, which is a great advantage. In this circuit the spikes, too, are eliminated if $C_{TE1} = C_{TE2}$.

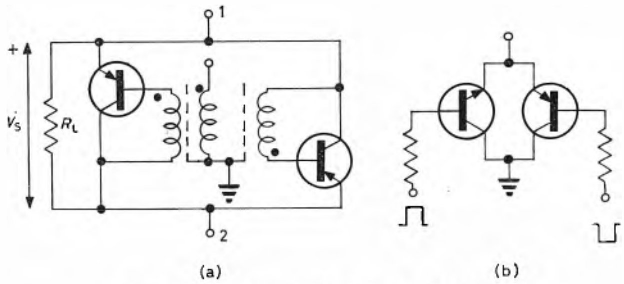


Fig. 7. Two transistors in parallel switches

It can be shown[7] that $V_s = 0$ when:

$$I_{B2}/I_{B1} = \frac{1 + (B_{N2}/B_{I2})}{1 + (B_{N1}/B_{I1})} \quad \ldots\ldots\ldots\ldots\ (22)$$

and that:

$$dV_s \simeq \frac{-V_{EC}}{2(1+(B_I/B_N))}\left((dB_{N1}/B_{N1}) - (dB_{I1}/B_{I1}) + (dB_{I2}/B_{I2}) - (dB_{N2}/B_{N2})\right) \quad \ldots.\ (23)$$

if $B_{N1} = B_{N2}$, $B_{I1} = B_{I2}$ and $I_{B2}/I_{B1} = 1$.

$dV_s = 0$ if $dB_{N1}/B_{N1} = dB_{N2}/B_{N2}$, which also was the case in the circuit in Fig. 6, but $dV_s = 0$ also if $dB_N/B_N = dB_I/B_I$, which often happens in symmetrical transistors. The factor $2\,(1 + (B_I/B_N)) \simeq 2$ for unsymmetrical transistors and $\simeq 4$ for symmetrical. Consequently this circuit ought to have, at the most, only half as great a drift as the circuit in Fig. 6 (compare with equation (21)). To test this the drift of five pairs of OC44 (unsymmetrical

* C_e and R_3 will function as a differentiator, and give a voltage v_3 across R_3 as shown in Fig. 5.

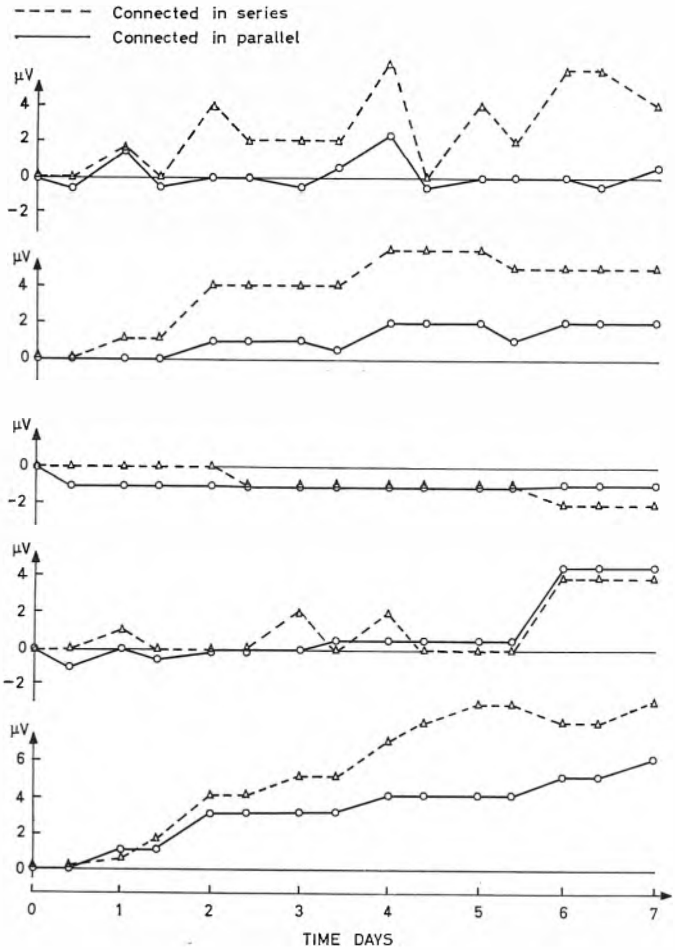


Fig. 8. The drift of 5 pairs of OC44 during 7 days

alloy germanium) during 7 days was measured. The transistors in each pair were in the 'on' state and connected in series or parallel depending on the position of a switch. (Between the measurements the transistors were always connected in series.) $dV_s = f(\text{time})$ is presented in Fig. 8, which shows that dV_s for all pairs is smallest in the parallel connexion.

For chopper applications suitable complementary planar transistors are now available (for example 2N2466 and 2N2593). Perfect complementary transistors are in principle impossible to obtain. (Holes and electrons have different mobilities.) In this case, however, only B_N, B_I and C_{TE} need to be similar. Unfortunately there do not yet seem to be any suitable complementary symmetrical silicon transistors on the market. However, symmetrical epitaxial transistors will probably soon be available.

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Communications for North Sea Oil Rigs

During the intensive search for oil and natural gas under the North Sea, a G.P.O. radio telegraph system will provide communications between the many mobile drilling rigs and one of two shore stations at Humber, on the Lincolnshire coast, and Stonehaven, near Aberdeen. At the same time Cullercoats Radio will be equipped to provide two single sideband radiotelephone channels for the exclusive use of the oil drilling rigs and telephone subscribers on shore.

The shore stations at Humber and Stonehaven will be fitted entirely with Marconi multi-channel telegraph transmitting and receiving systems and will provide up to 30 individual communication channels.

This order was placed by the General Post Office and is worth over £150 000. An accompanying order from The Marconi International Marine Co. Ltd., calls for radio equipment which they will install on one of the drilling rigs. This equipment is housed in a transportable shelter and is similar to that sold in large quantities to the British Army.

The first terminal is due to be opened on 1 May and will be at the existing G.P.O. Humber station at Mablethorpe on the Lincolnshire coast. The second which is at Stonehaven, 10 miles south of Aberdeen will be opened a few months later.

These drilling rigs, the first of which is now in position off the Dogger Bank, are classified by the General Post Office as 'ships'. They are in fact, fairly mobile and it is imperative that at all times contact can be made with each one individually. During their searchings they may move out of range of one of the transmitters. Therefore, a very flexible system is necessary to cater for their continual movement.

Both shore terminals will use initially, a 24-channel radio telegraph system operating on the same transmitting and receiving frequencies. The 24 channels will be divided between the two terminals depending upon the distribution of drilling rigs between their respective areas. Each terminal will be capable of transmitting all 24 channels, although at any one time, it will be impossible for the same channel to be transmitted or received at both terminals. This channel-division system will ensure that each rig can use its one allocated channel at all times, irrespective of its location.

A further step towards improving the system reliability is the use of Marconi Autospec error correcting equipment on each channel. This equipment considerably reduces the error rate in telegraph transmission by using a special error correcting, self-checking code.

In each terminal, both transmitters (main and standby) will use extremely high stability crystal oscillators, specially developed by The Marconi Company. It is most unusual for such high accuracy oscillators to be used in low power systems, but they will be necessary to enable both terminals to operate on the same frequency.

The two shore stations at Humber and Stonehaven will each be fitted with the following equipment:

2 Low power (1kW) wideband transmitters, type HS113.
2 Single path i.s.b. receivers, type HR21.
28 Autospec error correcting multiplex terminals, type H5100.
2 24-channel voice frequency telegraph transmit terminals, type H5000.
2 24-channel voice frequency telegraph receive terminals, type H5001.
2 12-channel voice frequency telegraph transmit/receive terminals (standby), type H5000/5001.
2 i.s.b. drive units, type HD57A, controlled by external high-stability oscillators.

The transportable shelter supplied to The Marconi International Marine Company will contain the following equipment for the rig to shore radio telegraph circuit:

2 Low power (350W p.e.p.) h.f. transmitters, type HS27.
2 multi-service h.f. receivers, type HR28.
2 Autospec error correcting multiplex terminals, type H5100.
2 One-channel, voice frequency telegraph transmit/receive terminals, type H5000/5001.
1 Teleprinter.

The Marconi International Marine Company will install in this shelter additional transmitting and receiving equipment to cater for all other communications requirements.

These shelter wireless stations are approximately 17ft long, 7ft wide and 7ft high. They provide a fully equipped radio room containing everything necessary except the power supplies. Heating, lighting and ventilation facilities, fixed furniture and storage lockers are all built in and the shelter needs only bolting down to the rig deck. They can operate on a number of different services including c.w., f.s.k., telegraphy, s.s.b., d.s.b. and i.s.b. telephony.

Dynamic Calibration of Electronic Integrator Gain

By D. K. Taylor*

An accurate method for measuring and calibrating the gain of an electronic integrator is described in this article, and an experimental estimate of the accuracy is given.

(Voir page 209 pour le résumé en français:
Zusammenfassung in deutscher Sprache auf Seite 216)

BASICALLY, an electronic integrator is composed of a high gain amplifier, a resistor R, and a capacitor C, arranged as in Fig. 1.

If it is assumed that the gain of the integrator is infinite, the output voltage e_o and the input voltage e_i are related by the equation:

$$e_o = -G \int_0^t e_i dt \qquad (1)$$

where the integrator gain G is given ideally by:

$$G = 1/RC \qquad (2)$$

G is assigned a value by choosing R and C appropriately. However, the ideal situation never exists, and equation (2) is not strictly correct. Errors caused by such sources as the internal resistances and capacitances of the amplifier enter into the determination of the actual gain. Thus the calibration of the external resistor and capacitor of the integrator does not guarantee that the gain is calibrated.

The gain should be calibrated by measuring G and then adjusting the capacitance C until the correct value of G is measured. Alternately, one could adjust R, if trimming resistors are present, and in that case the following discussion will be equally correct if everywhere R and C are interchanged.

A simple method for measuring G is to insert a constant voltage e_{oi} as an input into the integrator and to measure the output voltage e_o at time t_1. Then the gain is calculated from:

$$G = e_{o1}/e_i t_1 \qquad (3)$$

where the output voltage and time are usually obtained from an oscillograph trace. However, this method has poor accuracy if standard instruments are used, for either e_i or t_1 is limited to a small value, because of intrinsic limitations of analogue computers, so that the percentage error is large.

A more accurate method for measuring G is to determine it from the period of a sine wave produced by a sine wave generator which has the integrator to be calibrated as a component, as shown in Fig. 2.

Time is a quantity which can be measured with extreme accuracy, and in this method the measurement of gain is accomplished by an accurate measurement of time. For this, one uses a common analogue computer laboratory instrument, the EPUT (events per unit time) meter. This method has the added advantage that the capacitor can be adjusted while observing the period continuously, which gives faster convergence to the correct gain. However, periodically the sine wave should be stopped and restarted to verify its value since a small error is introduced by adjusting during operation. It is assumed that the gain of the invertor has been previously calibrated by adjusting its external resistor until the ratio of the output and input voltages has the correct value. The voltage y, which is defined in Fig. 2, is governed by the differential equation:

$$y'' + \omega_o^2 y = 0 \qquad (4)$$

where $\omega_o^2 = g_i g_j g_o$. The solution is given by:

$$y = y_o \cos \omega_o t$$

which has the period:

$$T_o = 2\pi/(g_i g_j g_o)^{\frac{1}{2}} \qquad (5)$$

if all the gains are equal to the ideal values.

The gain of the test integrator t will be calibrated with the aid of two auxiliary integrators a_1 and a_2. First, with the circuit of Fig. 2 consisting of integrators a_1 and a_2

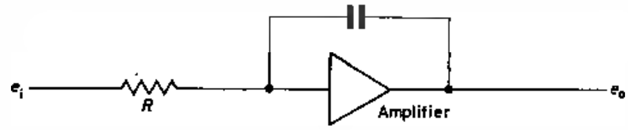

Fig. 1. Basic electronic integrator

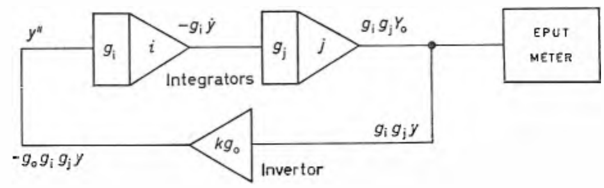

Fig. 2. The calibrator circuit

and invertor k ($i = a_1$, $j = a_2$) one measures the period T' of the circuit. Then with ($i = a_1$, $j = t_2$) adjust the capacitance of the test integrator t until the period of the circuit is again T'. Now the gain of integrators t and a_2 are equal:

$$g_t = g_{a2} = g' \qquad (6)$$

Next one sets ($i = a_2$, $j = t$) and measures the period T. The corresponding angular frequency is:

$$\omega^2 = g'^2 g_o^2 \qquad (7)$$

Finally, one adjusts the capacitor of integrator t and obtains a final angular frequency ω_F given by:

$$\omega_F^2 = gg'g_o = \omega_o \omega \qquad (8)$$

where the gain g was chosen so that:

$$\omega_o^2 = g^2 g_o \qquad (9)$$

The auxiliary integrator a_2 now has the gain g' and the test integrator t has the gain g, the correct value.

Any other integrator may now be calibrated very easily by placing it in the circuit with integrator t and adjusting its capacitor until $T = T_o$. It is seen that once an integrator is given the correct gain the calibration of any other one is trivial.

A experimental estimate of the accuracy of this method was obtained by comparing the actual periods obtained after 16 integrators were calibrated. The observed periods were within ·005 per cent of the theoretical ones in every case.

Acknowledgments

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* *University of Maryland, Department of Physics & U.S. Naval Ordnance Laboratory.*

BOOK REVIEWS

Introduction to Microwave Theory and Measurements

By A. L. Lance. 308 pp. Med. 8vo. McGraw-Hill Book Co. 1964. Price 66s.

ACCORDING to the preface this book is intended as "a reference for engineers and technicians involved in microwave measurements and microwave systems or components." It contains 15 chapters on "Field Theory, Transmission Lines, Waveguide and -Components, Measurement of Microwave Frequency, -Attenuation and -Power, Ferrite and Active Devices," etc.

This list of topics is certainly impressive but, alas, not so the contents. The author has attempted to present this highly technical subject, involving complex electronic phenomena and sophisticated devices with a minimum of theory and mathematics. Such an endeavour—very commendable by itself—would require great care in the selection of topics and in the manner of presentation: neither of these are evident in this book.

Loose statements abound throughout and there are some occasional errors. In the introductory chapter one reads (p. 5): "When lines of electric force are displaced laterally, lines of magnetic force are induced in the immediate adjacent space". In the first equation dealing with Coulomb's law of interaction between electric charges q, their dimensions are given as "coulombs per metre". The continual use of abbreviations like V.S.W.R., C.C.W., P.S.W.R. etc. is irritating, particularly as the reader has to thumb through the whole book in the hope of finding their meanings defined. On p. 168 the author talks of "ferrimagnetic resonant line width $\Delta\mathbf{H}$" (sic!). Bold symbols are used for some vectors, but not for others: thus equation (1.5) is written as $\mathbf{E} = F/q$. At the start of Chapter 4: "'The instantaneous reference voltage is represented by the complex number $E = (E/\sqrt{2})e^{j0}{}_{o}$, and so it goes on. Typical for the whole approach is perhaps the start of Chapter 4.3: "It is not necessary to understand the mathematical derivation in order to use the circle diagram" (p. 38).

Some of the illustrations have been taken direct from other publications or from trade literature, often without proper explanation. Several graphs are given without indication of the variables plotted, a particularly poor example being Fig. 7.7. Many of the sketches are very scanty and the reader is left to guess their significance (see Fig. 7.9). Chapters 10 and 11 on "Ferrite and Active Devices" are perhaps the worst offenders, purporting to explain the operation of a large number of highly complex devices, including paramagnetic amplifiers, masers and lasers. This is done in a very superficial manner, while using all the 'high-falutin' modern jargon, "stimulated emission," "population inversion," etc. etc. While a good case can be made for using such jargon in specialist scientific literature, it is patently wrong to use such terms in a book of this sort, particularly when no effort is made at explaining their meaning.

This reviewer certainly would not know to whom to recommend this book: hardly to the expert. And the semi-qualified technician would only be misled into believing that he could understand the basic principles of these devices and their modes of operation, when, in fact, he is only given their jargon. A statement e.g. (p. 206) that "the original ammonia maser had a deviation less than one part in 100 billion" might sound a lot, even in America, but hardly helps with the understanding of maser action.

E. BILLIG

Elements of Pulse Techniques

By O. H. Davie. 197 pp. Med. 8vo. Chapman & Hall Ltd. 1964. Price 35s.

THE 'Elements of Pulse Techniques' is an interesting book, being both clearly written and well produced. The preface states the book is intended to assist the student, technician and serviceman as well as the many people in other industries who have to use electronic equipment.

By keeping mathematics to a minimum the author admirably achieves the right level for his potential readers. But so thoroughly are subjects dealt with that few circuit designers who read it can fail to learn something.

There are six chapters. The first is concerned with defining the pulse and the effect of applying it to simple networks. The second and longest chapter deals with the generation, shaping and delaying of the pulse. Such generators as multivibrators, time-bases, blocking oscillators, sine-squared, nanosecond and high voltage impulse types are adequately treated. Then follows a short chapter containing an excellent treatment of coaxial, artificial and ultrasonic delay lines. Chapter four is concerned with pulse amplification by linear pulse, direct coupled and distributed amplifiers. It is unfortunate not more attention is paid here to transistor pulse amplifiers. Chapter five deals with measurement of the pulse and nearly half is taken up with a good and welcome description of the rationale behind modern precision measuring oscilloscope design. The final chapter is the one of most general interest as it considers such applications of the pulse as to the determination of transmission line faults, ultrasonic measurement, high voltage techniques, and in laser action, etc.

Appended to each chapter is a substantial list of references. Unfortunately some third of these are not mentioned in the text and some sections have no references where they would have been useful.

Perhaps the one serious fault of an otherwise excellent book, in which the author clearly knows and understands his subject, is that too much attention is paid to valve circuits and too little to their transistor equivalents. One would venture to suggest that in current pulse circuit practice valves are largely obsolescent having less application than the author would have one believe.

M. E. BRYAN

Stromversorgung Elektronischer Schaltungen und Geräte

By S. W. Wagner. 728 pp. Med. 8vo. R. V. Deckers Verlag. 1964. DM68

IN all branches of electronic engineering the problem of selecting and designing the most suitable power supply circuit is becoming increasingly more difficult and complicated. Much of the necessary information is available in various text books and technical articles, however, it is the aim of this book to collect and amplify this information in a single reference work. Of necessity the outcome is a lengthy book which required the collaboration of 21 contributors.

The subject is divided into 12 chapters. The first is introductory, and contains definitions of the characteristics of power supplies and describes their measurement. The second chapter deals with the various forms of two-terminal supply: primary cells, accumulators, fuel cells, nuclear batteries, solar cells, and thermo-electric generators. In the third chapter the basic elements of mains equipments: transformers, rectifiers, smoothing chokes and capacitors, are introduced, and filter circuits are covered in the next chapter. In the fifth chapter the basic single and three-phase rectifier circuits are described. In chapters 6 and 7 the basic elements of stablized supplies, for example, reference tubes and Zener diodes, and stabilizer circuits using valves, transistors, thyratrons, and thyristors are considered. In chapter 8 some account is given of magnetic stabilizers and the design of mechanical, transistor, thyratron, and thyristor invertors and convertors is presented in chapter 9. In chapter 10 the production and stabilization of high voltages is examined, and in chapter 11 power supplies containing energy storage elements are mentioned. The final chapter deals with miscellaneous topics such as special measuring techniques, fuses and circuit-breakers.

In all sections the text is clearly

written and points are illustrated by numerous diagrams and tables. Worked examples and practical circuits add greatly to the usefulness of the book. In a work covering so wide a field it is perhaps unfair to pick on topics, such as protection in semiconductor rectifier circuits, which appear to be treated inadequately, however, in general the book achieves its objective and should remain a useful reference work on power supplies for a number of years.

G. D. BERGMAN

Electronic Precision Measurement Techniques and Experiments

Edited by John E. Remich. Written by members of the staff of Philco Technological Centre. 366 pp. Med. 8vo. Prentice-Hall. 1964. Price 78s.

THIS manual is claimed to provide a broad source of information on precision measurement techniques, with primary emphasis on electrical and electronic equipment. It appears to be addressed to the practising user of field and laboratory test equipment to help him to improve the accuracy of his measurements. It is a great pity that the book itself is not as worthy as the objective. The information given is confused, incomplete, misleading and in some places erroneous. Only one or two examples can be given in a short review. In the chapter on errors, true and mean value are confused, subscripts and superscripts alternate in designating the same quantity and a graph has the wrong numerical values assigned to one axis. When dealing with calibration methods for thermometers the triple point cell is not mentioned, the hypsometer shown is unusable and possibly dangerous and no reference is made to pressure corrections. Even the areas of primary emphasis are not exempt from this sort of shortcoming. Standing and travelling waves are confused in the chapter on microwave equipment and no mention could be found of noise factor or noise generators in receiver testing.

There is a great need for a comprehensive manual on measurement techniques which will assist newcomers to this field or will assist those with a specific measurement problem. It is to be doubted however that this volume will supply that need.

I. J. RICHMOND

Physics of III-V Compounds

By O. Madelung; translated by D. Meyerhofer. 409 pp. Med. 8vo. John Wiley & Sons Ltd. 1964. Price 98s.

IN 1952 Welker's first paper on the semiconducting properties of III-V compounds appeared. Over the next few years there was an approximately exponential rise in the number of papers published on such compounds, but since 1961 the annual number of papers has remained almost constant. It is to be inferred that the first-order features and many of the second-order features are now known and understood. In consequence a specialized review can be expected to have a lasting value.

This book is an authoritative review of the physical properties of these compounds. Their technology and applications are not considered, although the physical properties of p-n junctions are discussed. The other detailed topics are: crystal structure, band structure, optical properties, transport properties, impurities and defects, mixed crystals and related ternary compounds, and certain miscellaneous properties including some information on thin films. There is a summarizing discussion in the last chapter. More than 1 000 references have been culled, and these are given in alphabetic-chronological order at the end. They are referred to where relevant throughout the book.

Between them the author and translator have produced an excellent text and this is matched by the production of the volume. The subject is a specialized one and is presented at a level which requires prior understanding of the basic theory of semiconductors and of their properties. To the reader who is so equipped it can be very warmly recommended. As a reference work it is invaluable.

F. J. HYDE

Circuits with Periodically Varying Parameters Including Modulators and Parametric Amplifiers

By D. G. Tucker. 175 pp. Demy 8vo. Macdonald & Co. Ltd. 1964. Price 40s.

FROM time to time new theories and concepts emerge to explain the functioning of physical phenomena. As the old theories are found to be incorrect or lacking in the ability to explain all the related facts, they are discarded, often peremptorily. It is frequently claimed that many times more theories find their ways into waste paper baskets than ever survive to become accepted. It is therefore unusual to find a new book devoted to a concept which was first introduced perhaps 60 years ago but which has only become prominent during the last decade.

The functions of frequency changing and frequency conversion have traditionally been described in terms of non-linear circuits but this concept causes difficulty over the process of modulation. The alternative approach is to consider linear circuits with parameters which vary as a function of time and it is this approach which forms the basis of Professor Tucker's book.

The first five chapters are largely theoretical. They introduce and contrast the theories associated with the two methods of approach to the subject. Later chapters deal with more specific applications of the concept of linear circuits with periodically varying parameters to modulators, parametric amplifiers, and harmonic generators, one chapter being devoted to lattice networks. Problems and their solutions are included not merely to test the understanding of the student or reader but to assist in extending the explanations in the various chapters.

The book is designed for final-year undergraduate and immediate postgraduate levels in Electrical Engineering and Physics. It is based on a part-course of lectures given by Professor Tucker at the University of Birmingham. The success of the book will be measured by the number of future courses in university and technical colleges which will be founded on these concepts when advanced network theory is taught.

R. A. BONES

Mathematical Methods in Reliability Engineering

By N. H. Roberts. 300 pp. Med. 8vo. McGraw-Hill Book Co. 1964. Price £5

THIS book requires a reasonable knowledge of mathematics in order to obtain the useful information contained in it. There is nothing original in the book but it brings together most of the mathematics and statistics which are useful to engineers engaged in reliability studies.

The first few chapters are concerned with mathematical foundations, while the remaining sections consider the reliability of a system as it passes through normal development. The chapters which will be of interest to the reliability engineer are those on accelerated testing and on redundancy.

While a few examples are given, it would have been helpful if more details on practical problems were supplied.

There are a number of errors in the book, but on the whole the book should be useful to electronic equipment designers and in particular to those engaged in military projects.

G. W. A. DUMMER

Pulse Technology

By William A. Stanton. 255 pp. Med. 8vo. John Wiley & Sons. 1964. Price 53s.

IT is easier to assess a condidate's performance by his problem solving ability than to judge his understanding by means of a verbal exposition. This may be the reason why our technical education concentrates on work involving the derivation of expressions and the numerical evaluation of formulae. It is not uncommon to find graduates perfectly versed in the art of problem solving who nevertheless find it difficult to formulate the problem in the first place since they lack understanding of the physical phenomena involved. True, these men soon learn this aspect of their work from practical experience as they have "mathematics as a crutch to lean on" as Mr. Stanton puts it in his preface. He is, however, of the opinion that the technician should have an immediate physical understanding of electronic circuits and, therefore, in writing this book on electronic circuits he has deliberately jettisoned the mathematical crutches, a generous action performed for the good of his readers. There can be no doubt that it is infinitely more difficult to write on electronics with the aid of words only, that is without mathematics. In fact the explanations do not always come off. One has the feeling that, with the asides, the blackboard and audience participation the orally delivered lecture might have been crystal clear, but in cold print ambiguities often seem to appear and many sentences have to be read more than once. The purist reader will be appalled by such verbs as 'Théveninize', yet it is clear and expressive—but why a capital 'T'? The use of equivalent networks is rightly emphasized and the many good examples should leave the reader with a practical facility in the use of Thévenin's and Norton's great time savers. In treating transient phenomena the author speaks of "time of problem solution". Of course, in its context it is quite clear what is meant, but the impish suspicion that the answer to the problem might depend on whether it is tackled before or after tea cannot easily be suppressed.

In spite of these criticisms, to which many more examples could be added, one is left with the impression that this is an excellent book wholly written for the sake of its readers. Although primarily aimed at the technician, a text of this type should be compulsory reading in the early stages of any degree course in electronics and it may even bring into conscious focus some of the vagaries that lie dormant in even an experienced engineer's mind. The first five chapters on circuits making up over two-thirds of the book are the best and are well supported by examples, problems and answers. When it comes to 'Number Systems' and 'Logic Fundamentals' the approach appears inadequate and no more than the sketchiest of introductions is given in the 50 or so pages concerned.

The book is beautifully printed and produced and its price, on present-day standards, would not be much too high for only the glossary and extensive, well annotated list of bibliographical references. It is hoped that the book will have sufficient success to encourage the author and publishers to iron out the many irritating oddities for a second edition.

K. L. SELIG

Transistor Circuits

By K. W. Cattermole. 470 pp. Demy 8vo. 2nd Edition. Temple Press Books Ltd. 1964. Price 70s.

The text for this second edition has been thoroughly revised and brought up to date. The most important device development in recent years is the drift transistor. Its various forms of construction are described briefly, and its distinctive high-frequency and transient properties are analysed in some detail. Less applied, but of some interest, are the newer negative-resistance devices: tunnel pnpn, and similar structures are described, and some of the circuit theory originally inspired by the point-contact transistor has been re-written to accommodate them. At the same time applications of transistors have multiplied so extensively that no general survey could include all of them. In consequence the survey in its original form has been omitted and items of general interest have been transferred from it to other chapters. Subjects of which the treatment has been considerably expanded include logical combinatory circuits, audio and video amplifiers, current-switching binary circuits, sine-wave oscillators and counters to arbitrary roots.

There is a classified bibliography for the reader in search of further information.

Fundamentals and Components of Electronics Digital Computers

By G. Haas. Med. 8vo. 247 pp. Cleaver-Hume Press Ltd. 1963. Price 42s.

The first part of this book deals with mathematical fundamentals. After a short introduction which gives a synopsis of the classification, operation and fields of application of computers.

The second part examines the special requirements of individual components, who particular attention paid to static and dynamic properties, while the last section gives detailed examples of practical circuits and discusses methods of calculating and dimensioning.

Infra-red and its Thermal Applications

By M. La Toison. 147 pp. Med. 8vo. Philips' Technical Library. 1964. Price 42s.

This book deals with both the theory of infra-red radiation and the practical side (construction and installation of ovens).

The first four chapters are devoted to the theoretical concepts essential for proper understanding. The next chapter, on infra-red sources, is followed by a number dealing with the operation of infra-red lamps and their industrial applications.

Reliability Engineering for Electronic Systems

Edited by R. H. Myers, K. L. Wong and H. M. Gordy. 360 pp. Demy 4to. J. Wiley & Sons Ltd. 1964. Price 94s.

This book contains contributions from a number of authors dealing with reliability as required for present-day military and space systems.

Computer Applications 1962

Edited by M. W. Gutterman and R. S. Hollitch. 224 pp. Med. 8vo. Macmillan & Co. 1964. Price 56s.

Contains the papers together with discussions which were presented at The 1962 Computer Applications Symposium sponsored by the Illinois Institute of Technology Research Institute, formerly known as the Armour Research Foundation.

An Electronic Organ for the Home Constructor

By A. Douglas. 100 pp. Demy 8vo. Sir Isaac Pitman. 1964. Price 20s.

The second edition includes information on additional features such as a polyphonic upper manual and a new arrangement for continuous tuning.

Minor errors in the first edition, published in 1958, have been corrected.

Introduction to Fortran

By S. C. Plumb. 203 pp. Med. 8vo. McGraw-Hill Book Co. 1964. Price 44s.

This book is a self-instructional programme organized into progressive steps designed to enable the reader to proceed at his own speed and capability.

Examinations and exercises are provided with each chapter and a test group required 16 hours for the completion of the programme and the exercises.

Solid Circuits and Microminiaturization

By G. W. A. Dummer. 346 pp. Med. 8vo. Pergamon Press. 1964. Price 60s.

This book contains the proceedings of the Conference held at West Ham College of Technology, London, in June 1963 at which some 30 papers were presented.

Electrical Interference

By R. F. Ficchi. 262 pp. Demy 8vo. Iliffe Books Ltd. 1964. Price 50s.

This book was originally published in America and contains the author's experience in interference problems connected with the Ballistic Missile Early Warning Systems and subsequent Minuteman Systems.

Complete chapters are devoted to the principles and practice of shielding and filtering, interference reduction in cables and four chapters deal with various aspects of earthing.

Design of Low-Noise Transistor Input Circuits

By W. A. Rheinfelder. 160 pp. Demy 8vo. Iliffe Books Ltd. 1964. Price 30s.

The first two chapters, dealing with the noise figure concept and its measurement, provide the basic foundation for calculations where the noise is uniform. The following chapter is devoted to non-uniform noise, such as cross-modulation, noise modulation and intermodulation. Noise sources in valves and transistors are then discussed. The next three chapters cover in detail the design of practical circuits for low noise—Chapter 5, where uniform spectral density of both signal and noise is involved, and Chapter 6, where noise distribution is uneven, while Chapter 7 gives the essentials for low-noise audio design. The final chapter describes typical modern circuits developed by American and Continental manufacturers.

Sensitivity Analysis of Dynamic Systems

By R. Tomovic. 142 pp. Med. 8vo. McGraw-Hill Book Co. 1964. Price 76s.

Analytical, computer and simulation methods for defining the influence of parameter perturbations on the behaviour of dynamic systems are dealt with in this book.

In the first part of the book engineering methods are studied for defining parameter influence coefficients and the remaining chapter illustrate important applications in the synthesis and design of control systems.

Data Acquisition and Processing in Biology and Medicine—Volume 3

By K. Enslein. 344 pp. Med. 8vo. Pergamon Press. 1964. Price 100s.

This volume contains the proceedings of the Conference held in 1963 at Rochester University, N.Y., at which 22 papers were presented.

The BEAMA Directory, 1964-65

Demy 8vo. Pergamon Press. 1964. Price 60s.

This BEAMA Directory and Buyers' Guide is designed to provide, in convenient form and in five languages (English, German, Spanish, Portuguese and Russian), a ready reference to well over 1 000 different electrical and allied products, their manufacturers in Britain and their distribution all over the world.

The first edition, which replaced the BEAMA catalogue, was published in 1961-62.

Short News Items

"Electronics in Industry—the next five years," is the subject of a symposium to be organized jointly by the Institution of Electrical Engineers (South Midland Centre Electronics Section), and the Institution of Electronic and Radio Engineers (West Midland Section),[4] will be held at the University of Birmingham in the Department of Electronic and Electrical Engineering on Tuesday 6 April 1965.

A series of six papers will be presented, giving an authoritative review of the trends of development in industrial electronics that are likely to be seen in the next five years.

Further details and forms of application for registration may be obtained from: G. K. Steel, Electrical Engineering Department, College of Advanced Technology, Gosta Green, Birmingham 4.

The Electronic and Electrical Engineering Department of the University of Birmingham is offering a series of twelve month Postgraduate Courses based on a combination of lectures, tutorials and laboratory and project work. The Courses lead to the degree of M.Sc. for students already holding a Bachelor's degree or the equivalent, or to the Diploma in Graduate Studies for Non-Graduates. The following topics are available:

- Information Engineering (communications, computers and control, together with either solid state devices or microwave engineering).
- Air Traffic Engineering (radar and navigational aids, air traffic control, computers, communications and traffic studies).
- Solid State Electronic Engineering.
- Control Engineering.
- Machine Control Systems.

Application forms, further details and notes on financial assistance may be obtained from The Registrar, University of Birmingham, Edgbaston, Birmingham 15.

The Electronics Group of the Institute of Physics and The Physical Society is arranging a conference on "Non-metallic Thin Films" to be held at Chelsea College of Science and Technology, London, from 23 to 24 September 1965. The conference will be devoted to the discussions of the physical properties of continuous non-metallic films deposited on substrates of a different material. It is proposed to include the following main subjects: formation of thin films, electronic structure, electrical conduction, electron emission, electronic applications especially in active devices.

Short contributions of about 15 minutes' presentation time are welcomed and outlines of not more than 300 words should be sent to Dr. A. K. Jonscher, Physics Department, Chelsea College of Science and Technology, London, S.W.3, not later than 15 June 1965. A leaflet (ref. ACB78) giving guidance on the preparation and submission of outlines of talks is available from the Institute and Society.

Further details and application forms are available from the Meetings Office, The Institute of Physics and The Physical Society, 47 Belgrave Square, London, S.W.1.

An International Conference on Noise Abatement is to be arranged in Dresden by the East German Chamber of Technology from 23 to 26 November this year.

The following themes will form the basis of the conference:

- Physiological and psychological problems of noise with special emphasis on working conditions.
- Noise measurement and evaluation.
- Architectural acoustics and protection from noise in town planning.
- The development of noise in industrial and traffic centres and protective measures against it.

The conference will be held in German, Russian and English. Official invitations may be obtained from the Kammer der Technik, 108 Berlin, Clara-Zetkinstrasse 114-117, East Germany.

The British Scientific Instrument Research Association (SIRA) is to hold an Instrument Industry Conference at the Grand Hotel, Eastbourne, on 13 to 14 May this year to discuss recent developments and future trends in materials and manufacturing processes, and their relevance to the British instrument industry.

Registration forms and the final programme for the conference will be available shortly on application to Miss E. A. S. Mills, Head of Information Group, British Scientific Instrument Research Association, South Hill, Chislehurst, Kent (tel.: IMPerial 5555/9).

The British Society for Strain Measurement has recently been formed, the principal objects being as follows:

(1) To promote interchange of information on the engineering aspects of strain measurement.
(2) To promote technical discussions of practical aspects by means of public meetings, publication of papers and exhibitions.
(3) To foster new lines of development and discuss how they may best be pursued.
(4) The dissemination of ideas on new equipment and techniques.
(5) To maintain an interest in and support educational matters relating to strain measurement.

Details of membership and subscription rates are available from the Hon. Secretary, Mr. V. Lewis, Department of Mechanical Engineering, Brunel College, Acton, London, W.3.

The Paul Instrument Fund Committee, composed of representatives of the Royal Society, the Institute of Physics and the Physical Society and the Institution of Electrical Engineers, has recently made the following research grants:

£3 500 to Dr. J. E. Baldwin, assistant director of research in the Cavendish Laboratory, Cambridge, for the construction of a receiver for measuring power spectrum of a radio frequency noise source using auto-correlation techniques in a large number of parallel channels.

£7 800 to Dr. R. D. Davies, senior lecturer in radio astronomy, University of Manchester, for the construction of a spectrometer for radio astronomy employing digital auto-correlation techniques with a sampling switch rate of 4Mc/s which it is hoped can be increased to 10Mc/s and even to 100Mc/s.

£2 700 per annum for three years to Professor J. D. McGee, professor of applied physics, Imperial College of Science and Technology, London, for the development of a photo-electronic image device for time resolution of rapidly changing optical images.

The Scientific Instrument Manufacturers' Association (SIMA) has formed a Traffic Engineering Instruments Group with the object of achieving good liaison between the instrument industry and those able to determine or influence policy in this important field.

It is hoped to make the liaison a two-way process. In one direction to give the instrument industry early knowledge of possible requirements and trends so that U.K. products are available to meet the needs when they arise, and in the other to create a keen appreciation of the capabilities of the industry and of the possible applications of its products and research.

It is believed that such co-operation can lead to a more rapid application of scientific methods and instrumentation to traffic engineering problems and give the industry a better opportunity of building up a strong export market in this expanding and world-wide field.

SIMA has also set up an Economics and Statistics Committee to meet the needs of the Association and its members. This committee will co-operate closely with other bodies doing similar work, and will act as an advisory body to Council for the provision of economic and statistical information. They will determine the information likely to be of value to the industry, advise on its collection, analysis and dissemination and consider methods of improving it. Another aspect of the work of the committee will be the formulation of terms and definitions for use in the statistics work of the Association.

The Fifth International Congress on Acoustics is to be held at The Palais des Congres, Liége, Belgium, on 7 to 14 September this year and will deal with the following subjects:

- Speech analysis.
- Physiological and psychological acoustics.
- Molecular acoustics.
- Ultrasonics.
- Noise control.
- Room acoustics.
- Electroacoustics.
- Physical acoustics.
- Mechanical vibrations and musical acoustics.

Authors who are proposing to submit papers are asked to send a summary of not more than 50 words to the Congress Secretariat not later than 1 April and the complete paper not later than 1 June.

Further details are obtainable from the Congress Secretariat, 33 rue Saint-Gilles, Liége, Belgium.

Mullard Research Laboratories have established Research Fellowships to enable visiting scientists to spend a period at the Mullard Electronics Research Laboratories at Redhill, Surrey.

The fellowships will enable scientists engaged in academic, government or industrial laboratories to work for a period with well-established teams and will facilitate individual contributions with the minimum of non-productive preparatory work.

Fellows should normally have experience in a sphere in which the Mullard Laboratories are active, and will be appointed for about a year. They may be from the U.K. or overseas. Stipends appropriate to the status and seniority of the applicant will be paid.

To begin with, a limited number of these Fellowships will be awarded on a basis of personal invitation.

The General Post Office is to install a camera system for recording telephone exchange meter readings.

After extensive tests at Edinburgh—where it cut the reading time for 1 000 meters from four hours to ten minutes—the new photographic system is to be introduced nationally.

Earlier Post Office experiments with photographic meter reading over a number of years has been abandoned due to technical difficulties and prohibitive costs. The present system is due to two factors—improved photographic and reading equipment and the adoption of new mechanical and computer-aided methods of telephone billing.

The specially designed equipment which the Post Office is now using consists of a pre-focused camera mounted at the back of a 'hood' resembling a pyramid-shaped loudhailer. All the meter reader has to do is to press the mouth of the hood on to a selected square of an upright rack which houses the meters—and pull a trigger. This sets off a flash inside the hood, giving a shadow-free picture of 100 meters on under a square inch of film. 3 500 meters can be 'read' before the camera needs reloading.

Films taken from the cameras are wound on to automatic projectors which show up individual meter readings in numerals half an inch high. Each reading is recorded twice on a punched card by different operators. Both readings are checked against each other automatically before the final figure is passed to the telephone billing department.

The Telecommunications Division of Ultra Electronics Ltd has received a further order for 200 packsets for use by the British Police Forces. With 450 ordered by the Home Office for police and fire service use in August last, this brings the total to 650 during the current financial year.

The Ultra 3A4 packset, weighing less than 3 lb, forms part of the 'Vigilant' range of communications equipment. It operates in communications bands of up to 179Mc/s, with provision for a maximum of three channels. Battery life is up to eight hours between charges, and the operational range is up to five miles.

The Decca DASR-1 long range air surveillance radar purchased by the Ministry of Aviation and installed at Clee Hill in Shropshire as part of the radar network for the United Kingdom Airways System has been officially handed over to the Ministry.

The Clee Hill installation is capable of remote control so that unattended operation of the transmitters is possible. The radar information obtained by the DASR-1 will be relayed by microwave link equipment to the Southern Air Traffic Control Centre at Heathrow. Decca have designed, built and supplied the signal processing equipment necessary for use in conjunction with the link itself.

The radar coverage provided by the DASR-1 extends over part of the Amber 1 Airway to the east, a section of the Amber 25 Airway extending from Manchester Control Zone in the north to Brecon in the south, and over the adjoining airspaces.

Choice of the equipment was largely governed by the particularly difficult high terrain in the Clee Hill area, since 10cm radars of this type offer complete gap-free cover not materially affected by the contours of the ground in the vicinity of the aerial.

Supreme Headquarters Allied Powers Europe (SHAPE) has selected Marconi transportable microwave link equipment for an Allied Command Europe (ACE) Communications System requirement.

Marconi Italiana of Genoa, a subsidiary of The Marconi Company, was awarded this contract worth more than £100 000, for the supply of all the necessary equipment for a number of complete terminal and repeater stations providing either a single long-distance radio link or a number of independent medium-distance links.

The system consists of a similar number of terminal and repeater stations which are extremely mobile and can be rapidly deployed in the field along line-of-sight transmission paths. Each separate hop between terminal or repeater stations may therefore cover up to 50 miles depending upon topographical conditions.

In addition, any two of the terminal stations may be coupled as a back-to-back repeater station. Using this configuration, a single long-distance link can be made. However, with the terminal stations operating separately, independent links working over distances of about 100 miles are available.

Each terminal station is made up of MH150 super-high-frequency radio link transmitter/receivers, six-channel telephone terminals from the MX150 series and six-channel voice-frequency telegraph terminals from the MC124 series. In addition, aerial systems comprising parabolic dishes, passive reflectors and 40ft mast assemblies, together with a number of engine alternator power units will be provided.

The Danish P.T.T. has placed an order with Telefunken for 26 remote control systems for the control and monitoring of unattended repeater stations in its telephone system. The remote control network will be divided into several separate groups. In each group the stations are all similar so that at each station reports can be received and appropriate controls initiated by instructions.

G.E.C. (Telecommunications) Ltd, of Coventry, England, and Ateliers de Constructions Electriques de Charleroi (ACEC), of Belgium, have reached agreement to co-operate in the broadband microwave radio equipment field. Under this arrangement, ACEC will assemble, manufacture and market G.E.C. products, but the existing G.E.C. distributor arrangements remain unchanged.

ELECTRONIC EQUIPMENT

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

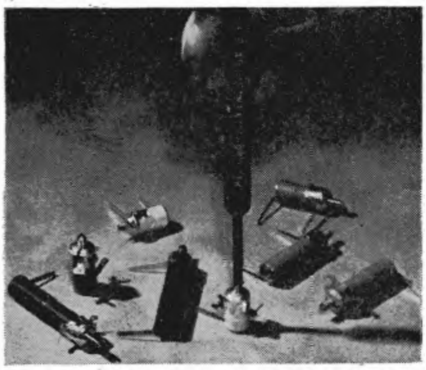
(Voir page 203 pour la traduction en français; Deutsche Übersetzung Seite 210)

PISTON TRIMMER CAPACITORS

Distributed by: Standard Telephones & Cables Ltd, Capacitor Division, Brixham Road, Paignton, Devonshire

(Illustrated below)

The Stangard, low cost, tubular glass piston trimmer range is now being marketed in the U.K. by STC Capacitor Division. The trimmers are suitable for use in such equipment as oscilloscopes, bridges, signal generators, frequency standards, calibrators, spectrum analysers, f.m. tuners and many applications previously limited to rotary ceramic disks, tubular ceramics and air variables.



Stangard trimmers are manufactured by JFD Electronics Corporation of New York. They are ruggedly built and have a vinyl encapsulation which protects the glass dielectric against shock. Low cost is achieved through the use of solid electrode bands and automated production lines.

These trimmers are available with two exterior designs—one for printed circuit board insertion and the other for panel or chassis mounting. D.C. working voltage is 1kV and the four trim ranges, in picofarads, are: 0·5 to 3·0, 0·8 to 5·0, 1·0 to 8·0 and 1·0 to 12·0.

Electrical characteristics include: Q better than 1000 at 1Mc/s, insulation resistance 10^6MΩ at 500V d.c. and a dielectric that will withstand in excess of 2kV. In addition, Stangard capacitors have a smooth adjustment torque and multiturn adjustment for sensitive tuning. These properties are fully retained through a temperature range of −55°C to +125°C.

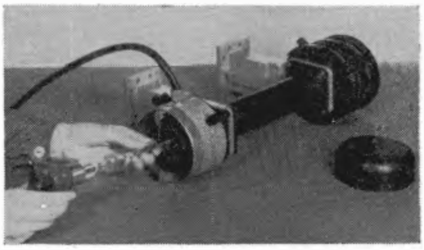
EE 79 751 *for further details*

HIGH POWER TRAVELLING WAVE TUBES

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

(Illustrated above right)

One of the most advanced travelling wave tubes yet offered in the telecommunications microwave field is the



new LB4-20 introduced by Mullard Ltd. It gives a saturated output power of 20W, an increase of 100 per cent over previous types designed specifically for telecommunications, combined with a low a.m./p.m. conversion rate of only 1·5° to 2°/dB.

With a gain of 40dB and a frequency range of 3·7 to 4·2Gc/s, either a wider bandspread, providing for more channels, or a longer link distance becomes available wih no increase in distortion.

The tube, which is contained in mount type P4L5, is of unpackaged construction; this means that a replacement tube can be fitted, using the existing mount, with only simple re-alignment of the mount being necessary. The focusing controls have been brought out to the front end of the mount to facilitate alignment. Positive location of the tube is by a special valve base which eliminates the need for a bulky cable harness.

Travelling wave tube LB6-20, which has been developed as a complement to the LB4-20, also gives a saturated output power of 20W, but operates over the frequency range 5·9 to 6·5Gc/s with a gain of 37dB. Despite this high power output the LB6-20 has an a.m./p.m. conversion rate of 1·5° to 2°/dB, thus its noise level is extremely low.

With both the LB4-20 and the LB6-20 natural cooling is sufficient with horizontally mounted tubes, but must be assisted by a convection duct or low velocity air flow when the tubes are vertically mounted. A conduction-cooled model is available for special applications.

EE 79 752 *for further details*

EDDY CURRENT TEST SET

C.N.S. Instruments Ltd, 61 Holmes Road, London, N.W.5

(Illustrated above right)

This equipment, the type 600, is intended for: inspection of bolt holes for fatigue or stress corrosive cracking; measurement, on a comparative basis, of thinning (for example, on aerofoil sections); sorting of mixed materials (provided a difference is resistivity exists). The test set, which is fully transistorized, provides the following facilities. A switched frequency range of 4kc/s, 10kc/s and 25kc/s (or up to 100kc/s for certain applications). Three modes of operation, either amplitude detection, phase acceptance, or phase rejection are selectable by a switch, enabling unwanted variables to be ignored by the operator; a six-inch meter scale for ease of reading; a wide range of probes, including surface probes, feed through coils and probes suitable for detection of cracks in bolt holes.

Special probes can be made to suit individual requirements and stocks of

printed circuit boards are held to facilitate servicing, should this be necessary.

EE 79 753 *for further details*

STRAIN MEASURING BRIDGE

The M.E.L. Equipment Co. Ltd, 207 Kings Cross Road, London, W.C.1

(Illustrated on page 197)

A new portable strain measuring bridge, Philips type PR9205, is now available from The M.E.L. Equipment Co. Ltd. Although weighing only a little over 10 lb, the bridge is rugged in construction and is suitable for field use in a variety of industries. It may be used with strain gauge transducers of 120Ω to 2kΩ resistance in half or full bridge configuration.

When setting the instrument up in a particular application, the zero point may be adjusted independently of the measuring range. With the instrument set up, the zero balance range is plus or minus 10 000 microstrain and the measuring balance range is plus or minus 8 250 microstrain, giving a total calibrated range of plus or minus 18 250 microstrain. Accuracy is better than 1 per cent ±6 microstrain.

Balance is indicated by a moving-coil meter and is effected by a single helical potentiometer control having a 3 600° scale calibrated in microstrain. Short term measurements can thus be made quickly and accurately, while the accuracy of long term measurements is ensured by high stability and accurate

zero re-adjustment facilities. Changes in ambient temperature have little affect on the instrument, drift being less than 1 microstrain/°C between −10 and +40°C.

The instrument is completely transistorized and operates from internal rechargable batteries. It can, however, and may equally well, be operated from five ordinary unit cells, changes in voltage between 5·5 and 7·5V having no affect on performance.

EE 79 754 *for further details*

MULTI-TRACK RECORDING HEAD

Gresham Lion Electronics Ltd, Lion Works, Hanworth Trading Estate, Feltham, Middlesex

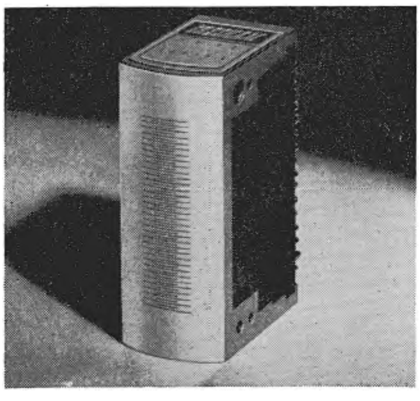
(Illustrated below)

Gresham Lion Electronics Ltd has developed a 33 track magnetic recording head for use on 1in tape. This new head gives the highest track packing density commercially available in the United Kingdom, and is designed primarily for digital applications where high track packing densities are required.

The 33 tracks are contained in a compact head 1¼in high, 1¼in deep and 0·6in wide. The track pitching is 0·030in and the track width is 0·010in.

The head is fitted with a gap length of 0·00025in, and the windings of each track have four terminals which allow the two balanced windings to be connected singly, in series or in parallel. It has an output of 1·5mV peak-to-peak at 300 bits/in at 15in/sec.

Crosstalk is better than −20dB on a square wave signal at or below 10kc/s at 15in/sec tape speed. (The crosstalk is



measured by recording simultaneously in phase signals on two alternate tracks and measuring the signal induced in the interposing tracks when reproducing).

EE 79 755 *for further details*

TIMER-COUNTER

Elesta Electronics Ltd, 34 Cambridge Park, London, E.11

(Illustrated below)

This portable transistorized instrument with neon indicator display can count to 9999 with a counting rate up to 3kc/s. Used as a timer with the internal oscillator switched at either 1 000 or 100c/s, the maximum ranges are 9·999sec or 99·99sec. An optional frequency of 10c/s is available, other frequencies can be provided on request. The input signal should be rectangular and not less than

7V peak-to-peak. An optional internal Schmitt trigger, increases sensitivity to 3V peak-to-peak and ensures positive operation with all input waveforms.

EE 79 756 *for further details*

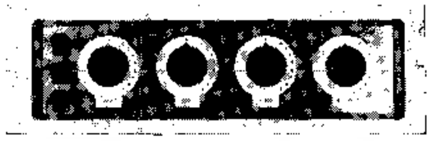
POTENTIAL DIVIDER

Cambridge Instrument Co. Ltd, 13 Grosvenor Place, London, S.W.1

(Illustrated above right)

This Kelvin Varley potential divider has been designed to provide multiplying ratios between zero and ×1 in steps of ×0·0001, while maintaining a constant input resistance. Three versions of the divider are available, having input resistances of 1kΩ, 10kΩ and 100kΩ, and maximum loadings of 75, 250 and 750V.

The divider carries four ratio dials having ratios of ×0·1, ×0·01, ×0·001 and ×0·0001. The upper three dials each has nine positions, the lowest dial having a tenth position that brings the maximum ratio to ×1. The switches have laminated brushes and silver-faced contacts, and are not subject to contact resistance variations greater than ±0·00002Ω/decade. All resistors are of carefully aged Manganin wire and—except for those of the lowest dial—are made up of Ayrton-Perry windings on mica cards in strain-free mountings. They are adjusted



to an accuracy of within ±0·05 per cent of their nominal value, ratio errors being less than ±0·1 per cent of the setting on the dials.

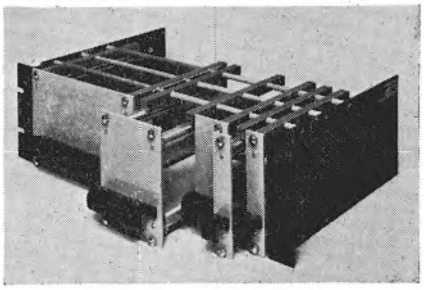
EE 79 757 *for further details*

MODULAR RACK UNITS

Vero Electronics Ltd, South Mill Road, Regents Park, Southampton

(Illustrated below)

With the introduction of a 5¼in modular rack unit Vero Electronics Ltd now cover the full range of popular standard sizes—5¼in, 7in and 8¾in unit heights, and module panel widths of 1in, 2in, 4in and 8in. The overall width of the unit is 19in with depths of 11in and 14in. This system permits a range of vibrations, using standard components, in excess of



400, only requiring a screwdriver to complete a rigid and permanent structure of any size or combination in the space of a few minutes.

EE 79 758 *for further details*

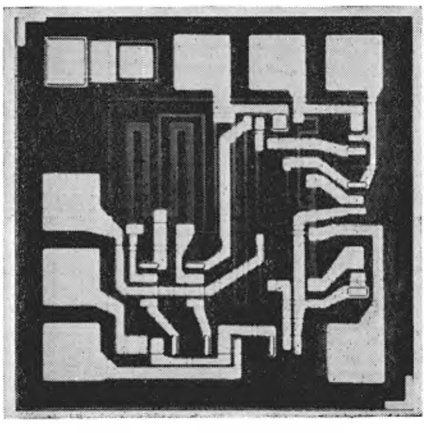
INTEGRATED D.C. AMPLIFIER

SGS-Fairchild Ltd, 23 Stonefield Way, Ruislip, Middlesex

(Illustrated below)

SGS-Fairchild Ltd has announced the availability in production quantities of the μA702, a complete d.c. amplifier constructed on a single chip of silicon using the planar epitaxial process.

The μA702 is a high gain operational type amplifier meant for use with external feedback elements to determine operating characteristics. Significant features are



low d.c. offset and drift, wide bandwidth, low power consumption, large output swing and operation over a wide range of supply voltages.

This amplifier can be used as a d.c. voltage amplifier in a variety of configurations, depending on how feedback is connected. Applications could include amplifying the output of transducers such as resistive bridges, Hall effect devices, photo-diodes, etc. With proper feedback elements the μA702 can perform the integrating, differentiating, summing and subtracting functions required in an analogue computer. In addition, the amplifier can be used with reactive feedback elements for precise frequency shaping of the gain characteristics or with a bridge-T filter to make a bandpass amplifier. Diodes can be incorporated in the feedback loop to form a low-threshold detector and instrument rectifier.

The μA702 has a peak output current of 10mA; maximum power dissipation of 250mW; typical open loop gain of 63dB; and typical 5μV/°C thermal drift. The amplifier is available in both μA702 military (−55°C to +125°C) and μA702C professional (−25°C to +100°C) versions.

Illustrated is a photomicrograph of the μA702 operational amplifier showing its geometry.

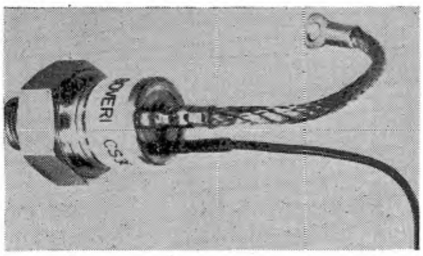
EE 79 759 *for further details*

THYRISTORS

British Brown-Boveri Ltd, Glen House, Stag Place, London, S.W.1

(Illustrated below)

British Brown Boveri Ltd has recently introduced a new thyristor into its manufacturing programme. This new device, designated CS.30, has a mean forward current of 70A at a stud temperature of 80°C and 180° conduction angle. It is



available for rated voltages in the range 100V to 600V at 100V intervals and minimum forward breakover will not occur below voltages which are 40 per cent above these rated values. The peak one-cycle (20msec) surge current of the CS.30 is 1 300A.

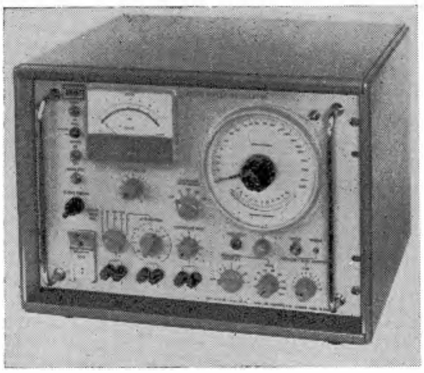
EE 79 760 *for further details*

AUDIO SWEEP OSCILLATOR

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustrated above right)

A new sweep oscillator covering the audio range from 20c/s to 20kc/s on a single logarithmic scale has been introduced by Dawe Instruments Ltd. Designated the type 443B, it provides a constant voltage output over the whole frequency range, even when connected to a non-linear or frequency-dependent network. The frequency scale is fully rotatable and provision has been made for coupling the frequency control to the Dawe type 1406 high-speed level recorder to record responses automatically. The sweep can be started by operating a front



panel switch or by remote control to synchronize with the type 1406. An adjustable frequency marker is provided.

The oscillator works on the heterodyne principle, the outputs from two high-frequency oscillators (one working at fixed frequency, the other at a frequency varying over a small range) being fed into a mixer. The difference frequency in the audio range is filtered and amplified, a transformer-coupled output stage giving output impedances of 6, 60, 600 or 6 000Ω. Output is variable in 10dB steps from 12·5V to 125μV, output voltage at the load terminals or to the input of the attenuator being continuously variable and displayed on a front panel meter.

Frequency is varied by a specially designed capacitor giving a truly logarithmic scale within ±1 per cent ±1c/s. Course and fine manual frequency drives are provided on the front panel or externally from the high-speed level recorder, and the frequency can be modulated internally or externally by up to 200c/s at any setting.

A voltage compression circuit controls the output of the oscillator to maintain constant current or voltage output. Distortion is less than 0·1 per cent from 200 to 2 000c/s rising to not more than 1 per cent at 20c/s and 20kc/s on attenuator output of 10V. The instrument is available for rack or bench-mounting and draws about 70W from a 50 or 60c/s mains supply. Weight is 55 lb with case.

EE 79 761 *for further details*

PULSE TIME GENERATOR

Telefon Fabrik Automatic A/S, 7 Amaliegrade, Copenhagen, K, Denmark

(Illustrated on right)

The type EN5-01 pulse time generator is designed primarily as a precision instrument for testing automatic telephone components and equipment, but may also be used for other applications where accurate and stable pulses with known duty cycle are required.

By means of two 'static contacts' (solid state) the generator provides normally open and normally closed contact functions. Further it provides a time-base signal and marker pulses for the leading and trailing edge of the contact function.

Although primarily designed as a precise timing device the generator can simultaneously be used as a precise amplitude and power unit. By connecting the static contacts to an external power source such as TFA type EC3-01, an output signal with an amplitude accuracy corresponding to that of the power source is obtained.

Due to the fact that the static contact has well defined opening and closing characteristics it is possible to control very accurate doses of electrical charges in small increments.

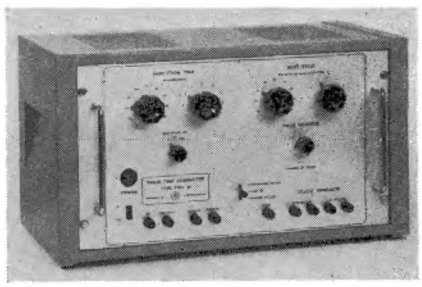
A crystal-controlled oscillator supplies pulses with 10μsec repetition time. In the succeeding frequency divider, or time-base multiplier 10μsec, 100μsec and 1msec repetition time-base may be selected.

The time-base multiplier is followed by a pulse selector which can be set to a desired number of pulses (*N*) between 1 and 99.

Whenever the selector has received *N* pulses it resets to zero and it supplies one pulse to a 100 step ring-counter.

The ring-counter gives one output pulse for each group of 100 received pulses, this output pulse is called the *O*-pulse and its repetition time is the desired repetition time for the final signal. The ring-counter also provides an extra pulse, which occurs *P* steps later than the *O*-pulse. This *P*-pulse can be delayed in steps of 1 per cent from 1 to 99 per cent of the repetition time.

Two static contacts are activated by the two output pulses from the ring-counter. The *O*-pulse opens one of the contacts (the normally closed contact) and closes



the other (the normally open contact). The *P*-pulse resets the contacts to their normal position.

As seen from the foregoing the delay of the *P*-pulse gives direct the duty cycle in percent of the repetition time.

The *O*-pulses pass a gate circuit between the ring-counter and the static contacts and the gate is controlled by a three-position lever key on the front panel. In the upper, locked position of the key all *O*-pulses pass the gate and the contacts operate continuously. In the lower, unlocked position of the key the gate is open only for a desired number of pulses preset on the front panel, and the contacts operate the desired number of times. In the middle, or stand-by position of the lever key, the gate is closed and no O-pulses reach the contacts which therefore remain in their normal position. However, in stand by position it is possible to control the gate by means of an external negative voltage and thereby generate a composite pulse signal, simulating for instance a multi digit dial signal from a telephone subscriber.

A special unit for programming a multi digit dial signal will soon be available.

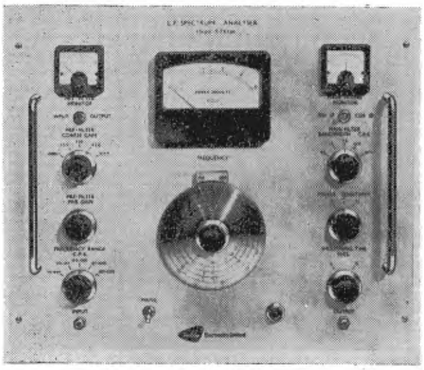
EE 79 762 *for further details*

SPECTRUM ANALYSER

Fenlow Electronic Ltd, Springfield Lane, Weybridge, Surrey

(Illustrated below)

The Fenlow spectrum analyser type SA3 is a new instrument designed to provide the spectral density of waveforms over the frequency range from 1·5c/s to 5kc/s. Fixed filter widths of 0·3, 1·5, 37·5



and 187c/s are available. The noise power being passed by the filters is measured by use of thermocouples, the output of which is indicated on a meter and is also available to drive the automatic plotter.

The analyser makes use of the heterodyne principle in which the incoming signal is multiplied by the oscillator signals in quadrature, the multiplier outputs are passed through low-pass filters, the outputs of which drive the thermocouples.

Six operational amplifiers are used to give a high degree of accuracy. Digital methods are employed to provide the internal oscillator signals for the two quadrature channels.

Facilities are provided for monitoring the signals at all key points in the instrument.

EE 79 763 *for further details*

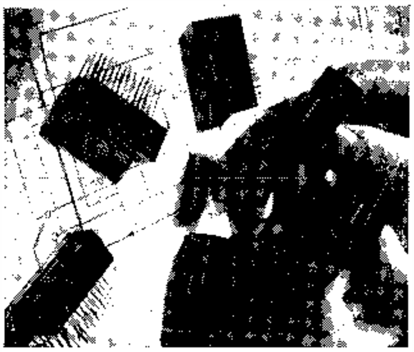
LOGICAL CIRCUIT ELEMENTS

Ferranti Ltd, Ferry Road, Edinburgh, 5, Scotland

(Illustrated above right)

Three new series of logical circuit elements are now available from Ferranti Ltd, extending the appeal of this type of circuit to a wider range of users. The new elements offer rectangular shapes, higher switching speeds and lower costs without the sacrifice of quality or reliability.

The 400 series is an alternative to the cylindrical 300 series but in the rectangular shape. This provides comparable



packaging density to the 300 series and in many instances the new shape allows superior printed circuit modular stacking. The modules contain silicon active elements, have switching speeds in excess of 3Mc/s, and operate over the temperature range -40 to $+100$°C.

The 500 series is a low cost range of rectangular modules with switching speeds in excess of 5Mc/s. They contain silicon transistors and gold-bonded germanium diodes of proved reliability, and have a temperature range of -20 to $+90$°C.

The 700 series modules (illustrated) are of simplified design to offer a cheap range of elements to the industrial designer. Switching speeds are in excess of 1Mc/s, and the operating temperature range is -20 to $+90$°C.

Great care has been taken to ensure that reliability remains at a high level, silicon planar transistors being used throughout. All units are fully encapsulated in epoxy resin, are of low weight ($<$10g) and are mechanically robust.

The modules are presented in rectangular form and measure 1·48in by 0·45in by 0·55in. Connector outlets are in the form of flying leads, and the maximum number of outlets is 14, in a single line on 0·1in pitch. All circuits operate on standard supply rails of $+12$V and -6V; the tolerance on all lines is nominally $\pm$0·5V. Logic levels for this series of modules are similar to that of the rest of the Ferranti range; i.e., state 1 = $+6$V (dependent on load) and state 0 = 0V (approx.).

Ferranti Ltd offer an application service and in many cases can provide modules mounted on printed circuits boards to fulfil a number of standard functions, such as reversible and non-reversible decade counters, shift registers, adders, decoders, etc.

EE 79 764 *for further details*

AUTOMATIC SAMPLING CHANGER

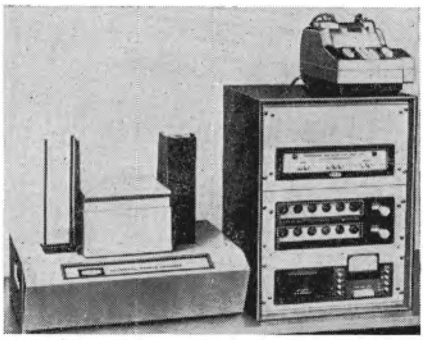
Isotope Developments Ltd, Bath Road, Beenham, Reading, Berkshire

(Illustrated above right)

A new range of automatic solid-state changing systems for use in medical, veterinary and other clinical and research applications which necessitate the automatic measurement of radiation energy, has been developed by Isotope Developments Ltd, a member of the Elliott-Automation Group. The range includes the I.D.L. 'Alphamat', 'Betamat' and 'Lobetamat' automatic counting systems for the measurement of alpha, beta and low-energy beta radiations respectively.

The new 'Lobetamat' automatic solid-state changing system, which is already being employed in medicine, automatically feeds up to 50 separate samples consecutively to a lead-shielded Geiger-Muller twin-tube assembly. This consists of a thin end window tube surrounded by a 'guard' counter which, in conjunction with an anti-coincidence circuit, provides additional shielding for eliminating the effects of background and cosmic radiation. A programme and read-out unit operates in conjunction with the scaler and a digital printer to provide a continuous record giving the particular patient's sample number, the count and time.

The 'Lobetamat' system can be pre-programmed in three modes: to switch off when all fifty samples have been



counted; to re-stack the samples in their original order before switching off; and to re-stack and re-count, repeating this cycle automatically as often as may be desired.

EE 79 765 *for further details*

EVAPORATION SOURCE

Thin Film Components Ltd, 26 Mead Lane, Chertsey, Surrey

This equipment is a versatile evaporation source for mounting within any high vacuum system for the deposition of metallic and non-metallic materials as thin films. It is suitable for use in micro-circuit, integrated circuit production and general research and development of micro-miniature components.

The system uses the electrons emitted from a hot filament to bombard the evaporant material which is placed in a water-cooled crucible located below the filament.

The evaporant head consists of four crucibles mounted on a water-cooled copper block, which is in turn attached to a rotary vacuum seal allowing consecutive depositions of four different materials without contamination. There is also provision for conventional resistance heating of the evaporant by the attachment of heavy current mounting bars.

Power for the operation of the source is derived from a console approximately 21in wide, 22in high, and 16in deep, suitable inter-connexions are also provided. A fully variable supply (low voltage) for the emitter filament and also a variable supply of 0 to 10kV d.c. is provided for controlling the heat input to the source. The maximum power is 1·5kVA.

The equipment is capable, by virtue of water-cooling the evaporant holder, of producing ultra-pure thin films with a minimum of effort and with maximum reliability. This, together with extreme versatility makes the equipment a suitable tool for research and production alike.

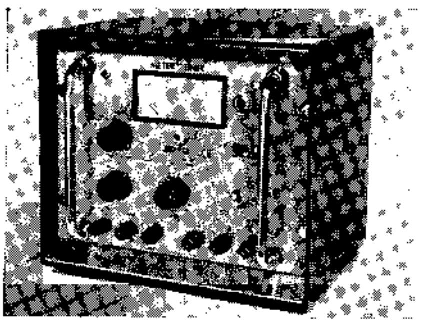
EE 79 766 *for further details*

METER UNIT

Brookdeal Electronics Ltd, Myron Place, Lewisham, London, S.E.13

(Illustrated below)

This is a compact, self-contained unit for use as an indicator in balanced systems having a source of impedance of 2kΩ, such as the Brookdeal phase sensitive detector, PD629. The basic accuracy of the instrument is determined by a



high-grade moving-coil movement having a scale length of 2·6in. A filter is incorporated to remove flutter at the sampling frequency, and this, together with the long scale ensures readings of high reliability and repeatability.

Attention has been paid to detail in this instrument: the backing-off supply is Zener stabilized to eliminate zero drift, and backing-off currents greater than full-scale deflexion can be obtained on all ranges without loss of accuracy. A low impedance output is provided for driving a pen recorder. A reverse-sense switch reverses both meter and pen recorder indications, but does not alter either backing-off or zero setting. The instrument provides a choice of three time-constants, and has a step attenuator. It is housed in a standard Brookdeal 'half-cabinet', or may be teamed with the Brookdeal phase shift network in a full 19in rack mounting.

EE 79 767 *for further details*

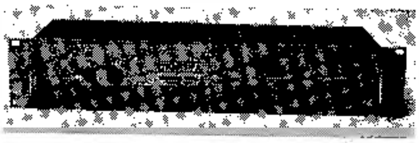
TELEPRINTER POWER SUPPLY

Industrial Instruments Ltd, Stanley Road, Bromley, Kent

(Illustrated above right)

The Transipack 309/B is a transistorized, stabilized power supply specially designed for use with teleprinters. It will deliver +80V and −80V at a load of 1A maximum.

The 309B performs a similar function to the rectifier set No. 66C and fits into a 19in G.P.O. rack. It is suitable for applications where space and weight are



at a premium as it is only 3½in high and 8in deep and it weighs 14 lb.

Overload protection is provided by a resettable high speed cut-out with an operating coil in each output line which trips at 3A; electronic protection is provided against short duration short-circuits.

EE 79 768 *for further details*

MEASURING INSTRUMENT SYSTEM

Dymar Electronics Ltd, Rembrandt House, Whippendell Road, Watford, Hertfordshire

(Illustrated below)

Dymar Electronics Ltd has designed a completely new range of laboratory and general purpose measuring instruments utilizing plug-in techniques.

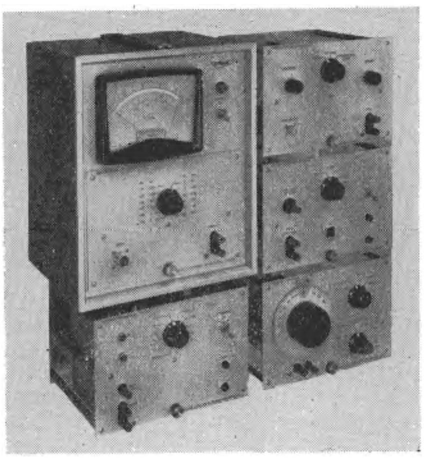
The basis of the system is the type 70 meter unit which will accept any one of the range of plug-in measuring instruments.

The meter unit, available in both bench top and 19in rack mounting versions, provides a precision 5in meter movement and comprehensive power supply facilities.

The meter is fitted with an anti-parallax mirror and is compensated against temperature variation.

Of the initial range of plug-in instruments the following have been released, type 701 wideband millivoltmeter; type 721 d.c. microvoltmeter; type 722 d.c. kilovoltmeter; type 741 a.f. signal generator; type 761 noise factor meter. Further units are currently under development and will be available shortly.

The wideband millivoltmeter type 701 is designed to measure a.c. voltages over a wide range of amplitude and frequency. It features a 10MΩ input impedance on all ranges, full-scale sensitivities from 1mV to 300V and a frequency range extending to 5Mc/s. It employs a four stage amplifier and rectifier circuit stabilized by approximately 50dB of overall feedback thereby ensuring excellent linearity and full accuracy with valve aging. Output terminals are provided per-



mitting the ins...ment to be used as a sensitive pre-amplifier for driving oscilloscopes or power amplifiers.

The d.c. microvoltmeter type 721 is intended for the measurement of small d.c. voltages with negligible loading on the circuit under test. A choice of two input terminals is provided giving either 100μV f.s.d. sensitivity at 1MΩ input impedance or 10mV f.s.d. sensitivity with an input impedance of 100MΩ. A low noise synchronous chopper is used at the input giving virtual freedom from drift and a phase-sensitive detector provides automatic polarity indication of the measured voltage. The calibrating voltage, derived from a Zener diode is accurate to ±1 per cent.

The d.c. kilovoltmeter type 722 will measure accurately voltages up to 30kV with negligible loading. Due to the very high input resistance of 3 000MΩ; the current drain on the voltage under test is a maximum of 10μA at 30kV and correspondingly less on other ranges. The use of stabilized h.t. and heater supplies in the d.c. amplifier ensures freedom from drift. A reference potential of 100V ±1 per cent is provided from a neon tube for calibrating purposes.

The type 741 general purpose signal generator features an extended frequency coverage from 30c/s to 300kc/s and an accurately monitored output level. It employs a Wien-Bridge oscillator with automatic amplitude stabilization by means of a thermistor. The maximum output level is 10V and can be attenuated by 80dB in 10dB steps by means of a 600Ω precision attenuator. A fine amplitude control permits interpolation of setting between steps so that the output voltage can be set precisely to the required level without the need for additional monitoring instruments.

The noise factor meter type 761 permits rapid evaluation of receiver noise performance in the band 1 to 220Mc/s. It incorporates a noise generator using a saturated diode and a noise detector which includes an accurate 3dB attenuator thus eliminating any possible errors due to non-linearity. Two versions of this instrument are available having either 50 or 75Ω output impedance.

EE 79 769 *for further details*

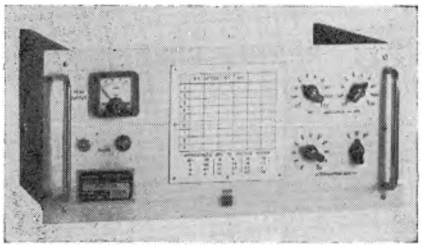
PULSE AMPLIFIER

Research Electronics Ltd, Bradford Road, Cleckheaton, Yorkshire

(Illustrated on page 201)

A new high grade a.c. coupled pulse amplifier designated the 'high accuracy wide band linear pulse amplifier model 9040A' is announced by Research Electronics Ltd.

The instrument incorporates many refinements particularly with regard to stability for the exacting requirements of gamma spectroscopy and pulse height studies. It has been designed to be used in conjunction with Research Electronics 'high accuracy single channel pulse height analyser model 9050' but may be used with any other high grade pulse height analyser.



Various head amplifiers enable the instrument to be used with fast ionization chambers, scintillation counters and proportional counters including boron trifluoride counters for neutron counting. An interesting feature is the provision of a meter which gives indication of the maximum instantaneous pulse height delivered at the output. This considerably simplifies the setting-up procedure and gives visual indication of overloading.

EE 79 770 *for further details*

HEAVY DUTY RELAYS

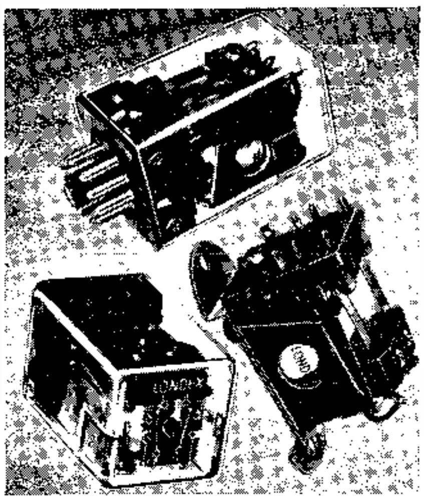
Londex Ltd, 207 Averley Street, London, S.E.20

(Illustrated below)

An entirely new range of a.c. or d.c. relays which have a minimum mechanical life of at least 20 million operations, has been introduced by Londex Ltd, a member of the Elliott-Automation Group. The average life of the silver contacts used in these relays is at least 5 million operations on non-inductive loads at typical ratings of 6A 240V/a.c. or 12V d.c.

These new relays which will be known as the Londex type TOP, are oustandingly rugged, reliable and quiet in operation. Two and three-pole change-over models are available in either enclosed or unmounted forms. Enclosed relays are provided with Makrolon dust covers and are fitted to plug-in bases—standard octal in the case of two-pole relays and standard 11-pin for three-pole types. A connexion diagram is imprinted on the top of the cover, which also carries lugs to locate a spring retaining clip. Unmounted relay contacts terminate in solder tags fitted in a contact panel.

Maximum contact ratings, non-inductive, are: a.c. 6A, 440V, 1500V; d.c. 6A, 250V, 75W up to 50V and 20W at 100V or over. Coils, which will operate the relay over the range of +6 per cent to −15 per cent of the nominal voltage, consume 4VA continuous, 7·5VA inrush, at 50c/s a.c. or 2 to 3W on d.c. Operate and release times are 3 to 10msec on a.c., 15 to 30msec on d.c.



EE 79 771 *for further details*

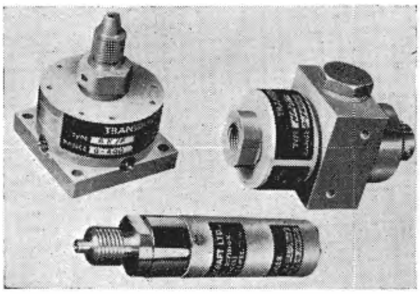
PRESSURE TRANSDUCERS

Westland Aircraft Ltd, Saunders-Roe Division, Osborne Works, East Cowes, Isle of Wight

(Illustrated below)

The Saunders-Roe Division of Westland Aircraft Ltd has introduced three new pressure transducers for liquids or gases, all embodying bonded foil strain gauges in a full bridge configuration.

These new transducers feature small physical size, low volumetric change with pressure, very high natural frequency and the ability to withstand high acceleration values without effect on the electrical output.



Three types cover the ranges from 0 to 30lb/in² up to 0 to 40 000lb/in² gauge, one model can be used as a differential transducer in the ranges 0 to 50 up to 0 to 1 500lb/in².

Maximum applied voltage is 10V into a 220Ω bridge, and a sensitivity in excess of 2mV/V is obtainable on most ranges.

EE 79 772 *for further details*

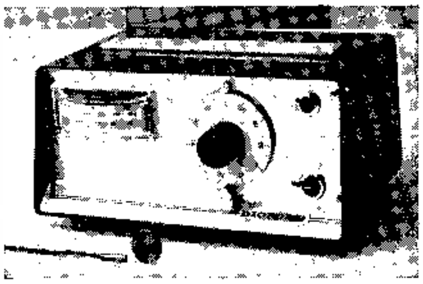
TEMPERATURE CONTROLLER AND INDICATOR

Fords (Finsbury) Ltd, Chantry Avenue, Kempston, Bedford

(Illustrated above right)

A new temperature controller, fitted with a visual temperature indicator, has been produced by the Electronics Division of Fords (Finsbury) Ltd.

A special feature of this instrument, marketed under the name of Dioford is the remote semiconductor probe. This is only ¼in long and 3/16in in diameter and may be used at distances up to 75ft from the instrument, using uncompensated wiring without loss of efficiency. Fitted with the indicator this instrument is suitable for the close control and indication of temperatures up to 150°C. Without the indicator it will control temperatures up to 260°C. When supplied without the indicator, which is independent of the control circuit, it is particularly suitable for use under conditions where considerable vibration is experienced. A plug-in relay, contained within the instrument, is adequate for switching powers of up to



1¼kW without the use of other contactors.

The instrument is of compact and robust construction being only 4½in high by 9in long by 6in deep. Both bench and panel mounted models are available. Semiconductors are used throughout and components are adequately rated to give a high order of reliability.

EE 79 773 *for further details*

DECADE MICA CAPACITOR

H. W. Sullivan Ltd, Murray Road, Orpington, Kent

(Illustrated below)

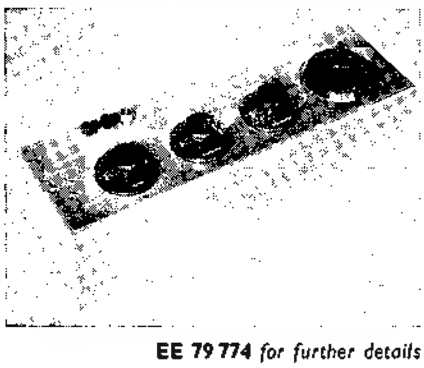
This is a completely screened three decade mica capacitor with an additional continuously variable air capacitor. The inclusion of a fourth (variable) dial is essential in order to overcome the inconvenience and inaccuracy which would otherwise be caused by the addition of an independent variable capacitor to a three dial instrument.

Temperature coefficient is of the order ±20 parts in 10⁶/°C and peak working voltages up to 500V may be used.

The capacitor is adjusted to be entirely direct reading throughout the range to an accuracy of 0·1 per cent. Moreover, it has been carefully designed to have a low series resistance and inductance and so its behaviour at the higher frequencies is excellent in both effective capacitance and power factor. The upper limits of frequency without loss of accuracy are given below:—

$C(\mu F)$	f(kc/s)
0·001	150
0·01	50
0·1	20
1·0	7

Up to these frequency limits the power factor is not augmented above the order 0·0005. If inaccuracies of the order 1 per cent are admissible the upper limits of frequency may be raised to 20kc/s to 500kc/s depending upon capacitance value.



EE 79 774 *for further details*

MEETINGS THIS MONTH

THE INSTITUTION OF ELECTRICAL ENGINEERS

Ordinary Meetings

All meetings will be held at Savoy Place, commencing at 5.30 p.m. unless otherwise stated.

Date: 4 March
Lecture: Research and Development in Control Engineering
By: J. H. Westcott

Date: 24 March Time: 6 p.m.
Held at: Institution of Mechanical Engineers, 1 Birdcage Walk, Westminster, London, S.W.1 (Joint meeting with the Institutions of Civil and of Mechanical Engineers)
Lecture: Incentives to Invention and Innovation
By: J. C. Duckworth

I.E.E.-R.Ae.S. London Joint Group on the Application of Electricity in Aircraft

Date: 23 March Time: 6 p.m.
Discussion: The Problem of Aerial Installation and Siting in Modern Aircraft
Opened By: R. A. Burberry and H. Hill

Electronics Division

Date: 8 March
Lecture: Screened Surface Waves and Some Possible Applications
By H. E. M. Barlow

Date: 10 March
Lecture: Electronic Circuits—Past, Present and Future
By: G. King

Date: 15 March
Discussion on: Irreversible Effects resulting from Breakdown in Transistors and Rectifiers
Opened By: C. F. Drake and A. F. Newall

Date: 16 March
Discussion on: Periodically Varying Elements
Opened By: D. G. Tucker and D. P. Howson

Date: 17 March Time: 6 p.m.
Discussion on: Implanted Stimulators for Bladder and Rectum
(Joint meeting with the I.E.R.E. Medical and Biological Group at the I.E.R.E., 9 Bedford Square, W.C.1)

Date: 22 March
Lecture: Experimental Instrumentation in Scientific Satellites
By: A. P. Willmore

Date: 29 March
Lecture: Vertical Radiation Patterns of h.f. Curtain Arrays of Plateau Sites
By: P. Knight, R. E. Davies and R. G. Manton

Power Division

Date: 1 March
Discussion on: Control of Room Storage Heaters
Opened by: R. B. Rowson

Date: 11 March
Discussion on: The Need or otherwise for the Provision of Stabilizing Windings on Star/Star Connected Transformers
Opened by: A. J. Wakeling

Date: 17 March
Lecture: The Utilization of Electricity in Medicine
By: D. G. Melrose

Science and General Division

Date: 9 March
Lecture: A Multijunction Thermal Convertor—an Accurate d.c./a.c. Transfer Instrument
By: F. J. Wilkins, T. A. Deacon and R. S. Becker

Date: 2 March
Lecture: Energy Storage
By: K. E. V. Willis

Date: 11 March
Discussion on: Two point Boundary Value Problems in Control
Opened by: C. Storey

Date: 25 March
Colloquium on: Critical Path Analysis

Date: 26 March
Discussion on: Insulating Thin Films

Date: 29 March
Colloquium on: Earth Loop Impedance Testing (at 10 a.m., 2.30 p.m., 5.30 p.m.)

THE INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Education Group

Date: 3 March Time: 6 p.m.
Held at: I.E.R.E. Lecture Room, 9 Bedford Square, London, W.C.1
Discussion: Teaching of Control Engineering (Tickets will be required)
By: J. C. West, H. A. Prime and H. Barlow

Electro-acoustics Group

Date: 10 March Time: 6 p.m.
Held at: London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1
Paper on: Problems in Listening
By: F. H. Brittain

Joint I.E.E.-I.E.R.E. Computer Groups

Date: 12 March
Discussion: The Training of Computer Engineers

Joint I.E.R.E.-I.E.E. Medical Electronics Group

Date: 17 March Time: 6 p.m.
Held at: I.E.R.E. Lecture Room, 9 Bedford Square, London, W.C.1
Paper on: Implanted Stimulators for Bladder and Rectum
(Tickets will be required)
By: P. G. Lale

Southern Section

Date: 18 March Time: 7 p.m.
Held at: Basingstoke Technical College
Paper on: Digital Computer Control of Gas Turbine Engines
By: D. Underwood and C. R. Kill

Date: 3 March Time: 6.30 p.m.
Held at: Highbury Technical College, Cosham, Portsmouth
Paper on: Digital Computers in Road Traffic Control
By: H. A. Codd

South Western Section

Date: 24 March Time: 7 p.m.
Held at: University of Bristol Engineering Laboratories
Paper on: Electronics and Process Control
(Joint meeting with the Bristol Branch of the British Computer Society)

Date: 25 March Time: 7 p.m.
Held at: Plymouth College of Technology
Paper on: Electronics and Business Management
(Joint meeting with the Western Centre of the I.E.E.)
By: W. C. Henshaw

South Wales Section

Date: 10 March Time: 6.30 p.m.
Held at: Welsh College of Advanced Technology, Cardiff
Paper on: Electronic Timing
By: A. McKenzie

Date: 29 March Time: 6 p.m.
Held at: Welsh College of Advanced Technology
Paper on: Computers in Control of Processes, the Coming Revolution in Industry
(Joint meeting with the Western Centre of the I.E.E.)
By: D. N. Truscott

West Midland Section

Date: 15 March Time: 7.15 p.m
Held at: Lanchester College of Technology, Coventry
Paper on: Synchronous Satellite Communication
By: F. J. D. Taylor

Date 10 March Time: 7.15 p.m.
Held at: Wolverhampton College of Technology
Paper on: Radio-astronomy
By: H. Gent

Scottish Section

Date: 17 March Time: 7 p.m.
Held at: Department of Natural Philosophy, University of Edinburgh
Paper on: Hybrid and Analogue Computers
By: A. J. Collins

Date: 18 March Time: 7 p.m.
Held at: Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow
Paper on: Hybrid and Analogue Computers
By: A. J. Collins

Yorkshire Section

Date: 3 March Time: 6.30 p.m.
Held at: University of Leeds Electrical Engineering Department
Paper on: Teaching Machines
By: Max Sime

East Midland Section

Date: 25 March Time: 6.30 p.m.
Held at: University of Leicester
Paper on: Electronics in Automobile Research
By: R. A. Evans

North Western Section

Date: 11 March Time: 7 p.m.
Held at: Manchester College of Science and Technology, Room F.1
Paper on: Recent Techniques in Magnetic Recording
By: W. Silvie

North Eastern Section

Date: 10 March Time: 6 p.m.
Held at: Institute of Mining and Mechanical Engineers, Westgate Road
Paper on: Some Applications for Microwave Techniques
By: J. Bilbrough

THE TELEVISION SOCIETY

Meetings are held at the Conference Suite I.T.A., 70 Brompton Road, London, S.W.3, and commence at 7 p.m.

Date: 5 March
Lecture: Non-entertainment Television
By: V. J. Cooper

Date: 18 March
Lecture: Video Storage for Standards Convertors
By: E. Rout

THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Date: 2 March Time: 5 p.m.
Held at: The Conference Room, Fleet Building, 70 Farringdon Street, E.C.4
Lecture: The Use of P.E.R.T. for Planning and Control
By: C. A. May

PUBLICATIONS RECEIVED

MULLARD INDUSTRIAL VALVE GUIDE, 1965 edition is now available. Like the 1964 edition the new guide has been produced in two sections. The equivalents guide contains an index of all Mullard valves and tubes for industry, communications and radar, together with a comprehensive guide to various valves and tubes for which Mullard types may be used as replacements. Abridged data on all current Mullard valves and tubes is published in a similar manner to previous issues. Requests for copies of these booklets should be addressed to Mullard Limited, Industrial Markets Division, Mullard House, Torrington Place, London, W.C.1.

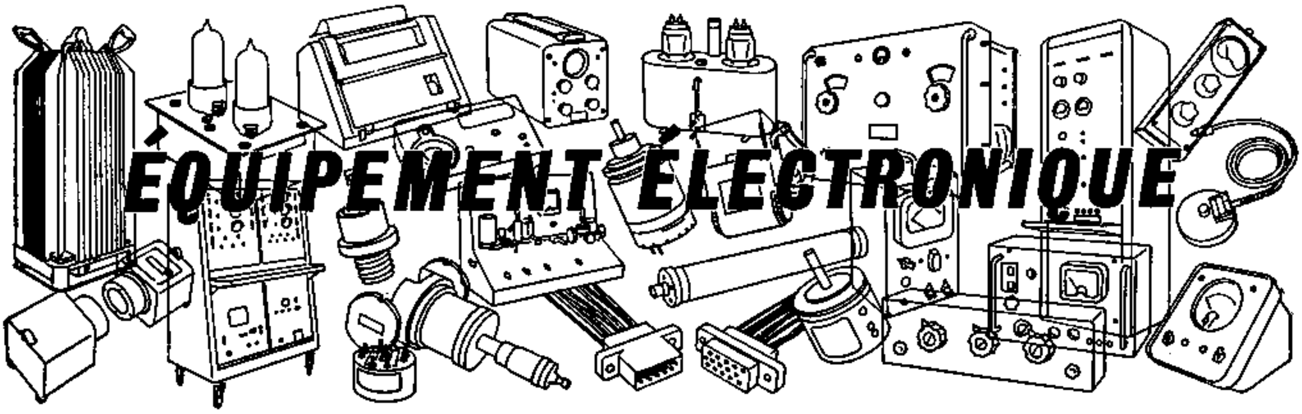
STYROFLEX CAPACITORS is a leaflet describing polystyrene capacitors for professional use manufactured by Siemens and Halske of Germany. These capacitors and copies of the leaflet are available from the U.K. distributors, R. H. Cole Electronics Ltd, 7-15 Lansdowne Road, Croydon, Surrey.

SILICON POWER TRANSISTORS are described in three new leaflets available from Westinghouse Brake and Signal Co. Ltd. These are technical publications 36-100, 36-101 and 36-104 which describe respectively 6, 7·5 and 20A transistors.

FAST SWITCHING (INVERTOR GRADE) THYRISTORS is the title of four new Technical Publications numbers 36-50, 36-51, 36-52 and 36-53 which give full details of fast switching invertor grade thyristors having forward currents of 3, 9·55, 35 and 70A respectively. Copies are available from Westinghouse Brake and Signal Co. Ltd, Rectifier Division, 82, York Way, King's Cross, London, N.1.

GREY INSTRUMENT KNOBS is a leaflet giving details and illustrations of a wide range of instrument knobs in a grey moulding to B.S. 381C : 1948 colour number 632. The illustrations are a close match as possible to the colour of the knobs. Copies of the leaflet are available from A. F. Bulgin & Co. Ltd., Bye Pass Road, Barking, Essex.

MULLARD SPECIAL QUALITY VALVES AND EQUIVALENTS is a reference chart in which Special Quality valves are listed together with their CV Services, American and Mullard Standard types number equivalents. Abridged data on Mullard Special Quality Valves is also included. Copies of the 1965 edition of the SQV reference chart are available from Mullard Ltd, Industrial Markets Division, Mullard House, Torrington Place, London, W.C.1.



ÉQUIPEMENT ÉLECTRONIQUE

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

Traduction des pages 196 à 201

CONDENSATEURS SHUNT D'ÉQUILIBRAGE À PISTON

Distributeurs: Standard Telephones & Cables Ltd, Capacitor Division, Brixham Road, Paignton, Devonshire

(*Illustration à la page 196*)

La gamme des condensateurs shunt d'équilibrage à piston Stangard à bas prix et dont le piston est formé de verre tubulaire, est vendue maintenant dans le Royaume-Uni par la division des condensateurs STC. Ces condensateurs peuvent être utilisés pour les oscilloscopes, les ponts, les générateurs de signaux, les étalons de fréquence, les calibreurs, les analyseurs de spectres, les blocs d'accord de modulation de fréquence ainsi que pour de nombreuses autres applications qui étaient limitées jusqu'ici aux disques en céramique rotatifs, aux pièces en céramique tubulaires et à d'autres variantes.

Les condensateurs Stangard sont fabriqués par la JFD Electronics Corporation de New York. Ils sont de construction robuste et sont enrobés de vinyle qui protège le diélectrique de verre contre les chocs. Leur bas prix est dû à l'emploi de bandes d'électrodes solides ainsi que de lignes de montage automatique.

Ces condensateurs sont livrables en deux modèles extérieurs: l'un pour l'insertion dans les plaquettes de circuit imprimé et l'autre pour montage sur panneau ou châssis. La tension de régime continu est de 1 kV et les 4 gammes d'équilibrage sont de 0,5 à 3,0, 0,8 à 5,0, 1,0 à 8,0 et 1,0 à 12,0 picofarads.

Les caractéristiques électriques comprennent: facteur Q supérieur à 1000 à 1 MHz, résistance à l'isolement de 10^6 MΩ à 500 V c.c. et un diélectrique dont la résistance dépasse 2 kV. En outre les condensateurs Stangard ont un couple d'ajustement souple ainsi qu'un réglage à plusieurs tours pour l'accord sensible. Ces propriétés sont pleinement maintenues dans la gamme de température de $-55°$ C à $+125°$ C.

EE 79 751 *pour plus amples renseignements*

TUBES DE GRANDE PUISSANCE POUR ONDES PROGRESSIVES

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

(*Illustration à la page 196*)

Le nouveau tube pour ondes progressives LB4-20 réalisé par la société Mullard Ltd, est un des plus perfectionnés présentés à ce jour dans le domaine des télécommunications par microondes. Il fournit une puissance de sortie saturée de 20 W, ce qui constitue une augmentation de 100% par rapport au type précédent conçu spécifiquement pour les télécommunications, ainsi qu'un taux de conversion m.a./m.p. de 1,5° à 2°/dB.

Grâce à un gain de 40dB et une gamme de fréquences de 3,7 à 4,2 GHz on peut obtenir soit un étalement de bandes plus étendu et, partant, un plus grand nombre de canaux, soit une plus grande distance de liaison sans accroissement de la distorsion.

Le tube est tenu par un support type P4L5 et n'est donc pas dans un boîtier. On peut ainsi fixer un tube de remplacement sur le support au moyen d'un simple réalignement de ce dernier. Les commandes de concentration se trouvent sur l'extrémité frontale du support afin de faciliter l'alignement. La fixation du tube se fait au moyen d'une base de lampe spéciale qui rend superflue l'emploi d'un harnais de câble encombrant.

Le tube pour ondes progressives LB6-20 qui a été réalisé pour servir d'appoint au tube LB4-20 donne également une puissance de sortie saturée de 20 W mais il fonctionne dans la gamme de fréquence de 5,9 à 6,5 GHz avec un gain de 37 dB. Malgré cette puissance de sortie élevée, le tube LB6-20 a un taux de conversion m.a./m.p. de 1,5° à 2°/dB, son niveau de bruit étant donc extrêmement bas.

Tant pour le tube LB4-20 que pour le tube LB6-20, le refroidissement naturel est suffisant pour des tubes montés horizontalement, mais ce refroidissement doit être aidé par un conduit de convection ou un flux d'air à faible vitesse lorsque les tubes sont montés verticalement. Il existe, en outre un modèle à refroidissement par conduction pour des utilisations spéciales.

EE 79 752 *pour plus amples renseignements*

CONTRÔLEUR DE COURANT DE FOUCAULT

C.N.S. Instruments Ltd, 61 Holmes Road, London, N.W.5

(*Illustration à la page 196*)

L'appareil type 600 a été conçu pour le contrôle de la fissuration corrosive due à la fatigue ou aux tensions dans les trous de boulons; pour la mesure, sur une base comparative, de l'amincissement (par exemple, des sections de plans aérodynamiques); pour le triage de matériaux mixtes (à condition qu'il existe une différence de résistivité). Cet appareil d'essai, entièrement transistorisé, comporte une gamme de fréquences à commutation de 4 kHz, 10 kHz et 25 kHz (ou jusqu'à 100 kHz pour certaines applications), trois modes de fonctionnement, à savoir: la détection d'amplitude, la réception de phase ou le rejet de phase. Ces trois modes peuvent être choisis, selon le cas, à l'aide d'un commutateur, permettant à l'opérateur de ne pas tenir compte de variables superflues. Il comprend en outre une échelle de mesure de 15 cm pour faciliter la lecture des indications, une gamme étendue de sondes, comprenant en outre des sondes de surface, ainsi que des sondes et des bobines d'injection pour la détection des fissures dans les trous de boulons.

Des sondes spéciales peuvent être fournies sur demande pour certaines applications particulières, de même que des plaquettes de circuits imprimés pour faciliter l'entretien au cas où il serait nécessaire.

EE 79 753 *pour plus amples renseignements*

PONT EXTENSOMÉTRIQUE

The M.E.L. Equipment Co. Ltd, 207 Kings Cross Road, London, W.C.1

(*Illustration à la page 197*)

Un nouveau pont extensométrique portatif, Philips type PR9205, est maintenant offert par la société The M.E.L. Equipment Co. Ltd. Bien que ne pesant qu'environ 4 kg, ce nouveau pont est de construction fort robuste et il peut être utilisé sur place dans diverses industries. En particulier, il peut être employé avec des transducteurs extensométriques de 120 Ω à 2 kΩ de résistance dans une configuration de pont complète ou mi-complète.

Lorsque l'instrument est mis au point pour une application particulière, le 0 peut être réglé indépendamment de la gamme de mesure. Une fois l'instrument mis au point, la gamme d'équilibre nul est de plus ou moins 10 000 microcontraintes et la gamme de mesure d'équilibre est de plus ou moins 8 250 microcontraintes, ce qui donne une gamme étalonnée totale de ±18 250 microcontraintes. La précision est supérieure à 1 % ±6 microcontraintes.

L'équilibre est indiqué par un instrument de mesure à cadre mobile et il est effectué par une commande potentiométrique hélicoïdale unique comprenant une échelle de 3600° étalonnée en microcontraintes. Des mesures à court terme peuvent ainsi être effectuées rapidement et avec précision, cependant que la précision des mesures à long terme est assurée par les possibilités de réajustement à haute stabilité et à zéro précis. Les changements de température ambiante ont peu d'effet sur l'instrument, la dérive étant inférieure à 1 microcontrainte/° C entre −10 et +40° C.

L'appareil est entièrement transistorisé et fonctionne sur piles intérieures rechargeables. Il peut cependant fonctionner tout aussi bien sur cinq cellules ordinaires, les changements de tension entre 5,5 et 7,5 V n'affectant nullement le rendement.

EE 79 754 *pour plus amples renseignements*

TÊTE D'ENREGISTREMENT MULTIPISTES

Gresham Lion Electronics Ltd, Lion Works, Hanworth Trading Estate, Feltham, Middlesex

(Illustration à la page 197)

La société Gresham Lions Electronics Ltd a mis au point une tête d'enregistrement magnétique à 33 pistes pour bande de 2,5 cm. Cette nouvelle tête assure la densité de tassement de pistes la plus élevée qu'on puisse obtenir dans le commerce au Royaume-Uni et elle a été conçue principalement pour les applications numériques exigeant des densités de tassement de pistes très élevées.

Les 33 pistes sont contenues dans une tête compacte mesurant 3,75 cm de haut, 3,17 cm de profondeur et 1,52 cm de largeur. L'espacement des piste est de 0,76 mm et la largeur de la piste est de 0,25 mm.

La tête comporte une largeur d'entrefer de 6.4 μm et les enroulement de chaque piste ont quatre bornes qui permettent de relier deux enroulements équilibrés de manière individuelle, en série ou en parallèle. Sa puissance de sortie de crête à crête est de 1,5 mV à 300 chiffres/pouce à 37,5 cm/sec.

La diaphonie est supérieure à −20dB sur signal d'onde carrée à ou au-dessous de 10 kHz à 37,5 cm/seconde de vitesse d'avance de la bande. (La diaphonie est mesurée par l'enregistrement simultané en phase de signaux sur deux pistes alternatives et par la mesure du signal induit dans les pistes interposées durant la reproduction.)

EE 79 755 *pour plus amples renseignements*

COMPTEUR-MINUTERIE

Elesta Electronics Ltd, 34 Cambridge Park, London, E.11

(Illustration à la page 197)

Cet instrument portatif transistorisé avec indicateur au néon peut compter jusqu'à 9999 à une vitesse de comptage maxima de 3 kHz. Utilisé comme minuterie avec l'oscillateur intérieur commuté à 1000 ou 100 Hz, les gammes de temps maxima sont de 9,999 sec ou 99,99 sec. Une fréquence facultative de 10 Hz est prévue, d'autres fréquences pouvant être fournies sur demande. Le signal d'entrée est rectangulaire et non inférieur à 7 V de crête à crête. Un déclencheur intérieur facultatif Schmitt, augmente la sensibilité à 3 V de crête à crête et assure le fonctionnement positif avec toutes les formes d'ondes d'entrée.

EE 79 756 *pour plus amples renseignements*

DIVISEUR DE POTENTIEL

Cambridge Instrument Co. Ltd, 13 Grosvenor Place, London, S.W.1

(Illustration à la page 197)

Ce diviseur de potentiel Kelvin Varley a été réalisé pour fournir des rapports de multiplication entre 0 et ×1 en plots de × 0,0001, tout en maintenant une résistance d'entrée constante. Il existe trois versions du diviseur dont les résistances d'entrée sont de 1 kΩ, 10 kΩ et 100 kΩ et les charges maxima de 75, 250 et 750 V.

Le diviseur comporte quatre cadrans de rapports ayant des rapports de × 0,1, × 0,01, × 0,001 et × 0.0001. Les trois cadrans supérieurs ont chacun neuf positions, le cadran le plus bas ayant une dixième position qui porte le rapport maximum à × 1. Les commutateurs sont munis de balais laminés et de contacts argentés et ils ne sont pas sujets à des variations de résistance de contact supérieures à ±0,00002 Ω/décade. Toutes les résistances sont en fil Manganin soigneusement vieilli et—à l'exception de celles du cadran le plus bas- sont constituées d'enroulements Ayrton-Perry sur cartes de mica fixées sur support anti-contraintes. Elles sont réglées à un degré de précision de ±0.05 % de leur valeur nominale, les erreurs de rapport étant inférieures à ±0,1 % du calage sur les cadrans.

EE 79 757 *pour plus amples renseignements*

ELÉMENTS MODULAIRES POUR BÂTIS

Vero Electronics Ltd, South Mill Road, Regents Park, Southampton

(Illustration à la page 197)

Avec l'introduction de l'élément modulaire pour bâti de 13,3 cm, la société Vero Electronics Ltd couvre maintenant l'ensemble de la gamme des dimensions populaires: 13,3 cm, 17,78 cm et 22,2 cm de hauteur d'élément ainsi que les largeurs de panneaux modulaires de 2,5 cm, 5 cm, 10 cm et 20 cm. La largeur hors-tout de l'élément est de 48 cm, la profondeur étant soit de 27,94 cm soit de 35,5 cm. Le système permet toute une gamme de vibrations n'utilisant que des composants courants dont le nombre dépasse 400. Il suffit d'un tournevis pour compléter une structure rigide et permanente de n'importe quelle grandeur ou dans n'importe quelle combinaison dans l'espace de quelques minutes.

EE 79 758 *pour plus amples renseignements*

AMPLIFICATEUR DE TENSION CONTINUE INTÉGRÉ

SGS-Fairchild Ltd, 23 Stonefield Way, Ruislip, Middlesex

(Illustration à la page 197)

La société SGS-Fairchild Ltd vient d'annoncer la production en série de l'amplificateur de tension continue, type μA702, construit d'une seule pièce de silicium au moyen du procédé épitaxial planaire.

L'Amplificateur μA702 est un appareil du type opérationnel à gain élevé destiné à l'emploi avec des éléments de réaction externe pour déterminer les caractéristiques de fonctionnement. Il se distingue en particulier par son décalage réduit et sa faible dérive de tension continue, par sa largeur de bande étendue, par sa consommation électrique réduite ainsi que par une puissance de sortie élevée et son fonctionnement dans une gamme étendue de tensions d'alimentation.

L'amplificateur μA702 peut être utilisé comme amplificateur de tension continue dans une grande variété de configurations, qui dépendent de la manière dont la réaction est connectée. Ces applications comprennent l'amplification de la sortie de transducteurs tels que les ponts résistifs, les dispositifs à effet Hall, les diodes photoélectriques, etc. Lorsqu'il est muni des éléments de réaction appropriés, le μA702 peut réaliser les opérations d'intégration, de différenciation, de sommation et de soustraction exigées d'une calculatrice analogique. En outre, il peut être utilisé avec des éléments de réaction réactifs pour la formation de fréquences précises des caractéristiques de gain ou en liaison avec un filtre à pont T pour constituer un amplificateur passebande. Des diodes peuvent être incorporées au circuit de réaction pour former un détecteur à seuil réduit et un redresseur d'instruments.

L'amplificateur μA702 a un courant de sortie de pointe de 10 mA, une dissipation de puissance maxima de 250 mW, un gain de circuit ouvert typique de 63 dB et une dérive thermique caractéristique de 5 μV/° C. Il est livrable en version pour usage militaire (−55° C à +125° C) sous la référence μA702 et pour usage professionnel (−25° C à +100° C) sous la référence μA702C.

On voit dans notre illustration une microphotographie de l'amplificateur opérationnel μA702 montrant sa géométrie.

EE 79 759 *pour plus amples renseignements*

THYRISTORS

British Brown-Boveri Ltd, Glen House, Stag Place, London, S.W.1

(Illustration à la page 198)

La société British Brown-Boveri Ltd. a récemment ajouté un nouveau thyristor à son programme de réalisations. Ce nouveau dispositif, type CS.30, a un courant direct moyen de 70 A à une température de 80° C et un angle de conduction de 180°. Il est prévu pour des tensions nominales de 100 V à 600 V à intervalles de 100 V. La surtension de pointe d'un cycle (20 msec) du CS.30 est de 1300 A.

EE 79 760 *pour plus amples renseignements*

OSCILLATEUR DE BALAYAGE ACOUSTIQUE

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustration à la page 198)

Un nouvel oscillateur de balayage couvrant la gamme acoustique de 20 Hz à 20 kHz sur une seule échelle logarithmique a été réalisé par la société Dawe Instruments Ltd. L'oscillateur type 443B fournit, en effet, une tension de sortie constante sur toute la gamme de fréquences, même lorsqu'il est branché sur un réseau non-linéaire ou dépendant de la fréquence. L'échelle de fréquence est entièrement rotative et on peut accoupler la commande de fréquence à l'enregistreur de niveau à vitesse élevée Dawe type 1406 pour l'enregistrement automatique des réponses. Le balayage peut être mis en oeuvre en actionnant un commutateur du panneau frontal ou par télécommande de synchronisation avec le type 1406. L'appareil comprend également un marqueur de fréquence réglable.

L'oscillateur fonctionne suivant le principe de l'hétérodyne, les sorties des deux oscillateurs à haute fréquence (l'un fonctionnant à une fréquence fixe et l'autre à une fréquence variant dans une gamme réduite) étant injectées à un mélangeur. La fréquence de différence dans la gamme acoustique est filtrée et amplifiée, un étage de sortie accouplé à un transformateur donnant des impédances de sortie de 6, 60, 600 ou 6 000 Ω. La sortie varie par plots de 10 dB de 12,5 V à 125 μV, la tension de sortie aux bornes de charge ou à l'entrée de l'atténuateur étant à variation continue et affichée sur un appareil de mesure du panneau frontal.

On fait varier le fréquence au moyen d'un condensateur de conception spéciale donnant une échelle logarithmique réelle de ±1 % à ±1 Hz. Des commandes de fréquence à réglage manuel précis ou approximatif sont prévues sur le panneau frontal ou à l'extérieur de l'enregistreur de niveau à vitesse élevée et la fréquence peut être modulée intérieurement ou extérieurement jusqu'à 200 Hz à n'importe quel calage.

Un circuit de compression de tension contrôle la sortie de l'oscillateur afin de maintenir constante la sortie de courant ou de tension. La distorsion est inférieure à 0,1 % de 200 à 2000 Hz, ne s'élevant pas à plus de 1 % à 20 Hz et 20 kHz sur une sortie d'atténuation de 10 V. L'instrument est livrable pour montage sur bâti ou sur banc d'essai et sa consommation électrique sur courant secteur de 50 ou 60 Hz est de 70 W. L'appareil et son coffret pèsent environ 25 kg.

EE 79 761 *pour plus amples renseignements*

GÉNÉRATEUR D'IMPULSIONS

Telefon Fabrik Automatic A/S, 7 Amaliegade, Copenhagen, K, Denmark

(Illustration à la page 198)

Le générateur d'impulsions EN5-01 a été conçu principalement comme instrument de précision pour le contrôle de composants et de matériel de téléphone automatique mais il peut également être employé pour d'autres applications exigeant des impulsions précises et stables avec cycle de régime connu.

Le générateur assure des fonctions de contact normalement ouvert et normalement fermé au moyen de deux "contacts statiques" (éléments solides). Il fournit en outre un signal de base de temps et des impulsions de marquage pour la durée d'établissement de la fonction de contact.

Bien qu'originellement conçu comme dispositif de minutage précis, le générateur peut être employé en même temps comme élément de puissance et d'amplitude précis. En reliant les contacts statiques à une source de puissance extérieure telle que le TFA type EC3-01, on obtient un signal de sortie d'une précision d'amplitude correspondant à celle de la source de puissance.

En raison du fait que le contact statique a des caractéristiques d'ouverture et de fermeture bien définies, il est possible de contrôler des doses très précises de charges électriques par petits incréments.

Un oscillateur piloté par quartz fournit des impulsions à taux de répétition de 10 μsec. On peut choisir dans le diviseur de fréquence ou le multiplicateur de base de temps des taux de répétition de base de temps de 10 μsec et 1 msec.

Le multiplicateur de base de temps est suivi d'un sélecteur d'impulsion qui peut être réglé au nombre voulu d'impulsions (N) entre 1 et 99.

Lorsque le sélecteur a reçu des impulsions N, il se réenclenche à zéro et il fournit une impulsion à un compteur annulaire à 100 plots.

Le compteur annulaire fournit une impulsion de sortie pour chaque groupe de 100 impulsions reçues, cette impulsion de sortie s'appelle l'impulsion 0 et son taux de répétition constitue le temps de répétition voulu pour le signal final. Le compteur annulaire fournit également une impulsion secondaire qui se produit à p plots après l'impulsion 0. Cette impulsion p peut être retardée par plots de 1 % de 1 à 99 % du taux de répétition.

Les deux contacts statiques sont amorcés par les deux impulsions de sortie du compteur annulaire. L'impulsion 0 ouvre l'un des contacts (le contact normalement fermé) et ferme l'autre, (le contact normalement ouvert). L'impulsion p réenclenche les contacts à leur position normale.

Comme on vient de le constater, le retard de l'impulsion p indique directement le cycle de régime en pourcentage du taux de répétition.

L'impulsion o passe à travers un circuit de porte entre le compteur annulaire et les contacts statiques et la porte est contrôlée par une clef à levier à trois positions sur le panneau frontal. A la position supérieure et verrouillée de la clef toutes les impulsions 0 passent par la porte et les contacts fonctionnent de manière continue. A la position inférieure et non verrouillée de la clef, la porte n'est ouverte que pour le nombre voulu d'impulsions présentes sur le panneau frontal et les contacts ne fonctionnent que le nombre de fois voulu. A la position médiane ou d'attente de la clef à levier, la porte est fermée et aucune impulsion 0 n'atteint les contacts qui demeurent donc à leur position normale. Toutefois, dans la position d'attente on peut contrôler la porte au moyen d'une tension négative extérieure et produire ainsi un signal d'impulsions composées simulant par exemple le signal d'un cadran à plusieurs chiffres d'un abonné au téléphone.

Un appareil spécial pour la programmation d'un signal de cadran à plusieurs chiffres pourra être obtenu bientôt.

EE 79 762 *pour plus amples renseignements*

ANALYSEUR DE SPECTRES

Fenlow Electronics Ltd, Springfield Lane, Weybridge, Surrey

(Illustration à la page 199)

L'analyseur de spectres Fenlow type SA3 est un nouvel instrument étudié pour indiquer la densité spectrale de formes d'ondes dans la gamme de fréquence de 1,5 Hz à 5 kHz. Des largeurs de filtre fixes de 0,3, 1,5, 37,5 et 187 Hz sont prévus. La puissance de bruit transmise par les filtres est mesurée au moyen de thermocouples dont la sortie est indiquée sur un instrument de mesure et sert également à actionner l'instrument de report automatique.

L'analyseur utilise le principe de l'hétérodyne suivant lequel le signal d'entrée est multiplié par les signaux d'oscillateur en quadrature, les sorties de multiplicateur passant à travers des filtres passe-bas dont les sorties actionnent les thermocouples.

Six amplificateurs opérationnels sont utilisés pour assurer un degré élevé de précision. Les méthodes numériques sont employées pour fournir les signaux d'oscillation intérieurs pour les deux canaux de quadrature.

On peut également contrôler les signaux à tous les points clefs de l'instrument.

EE 79 763 *pour plus amples renseignements*

ÉLÉMENTS DE CIRCUITS LOGIQUES

Ferranti Ltd, Ferry Road, Edinburgh 5, Scotland

(Illustration à la page 199)

Trois nouvelles séries d'éléments de circuits logiques sont maintenant fournies par la société Ferranti Ltd, étendant ainsi l'emploi de ce type de circuit à un nombre d'utilisateurs plus étendu. Les nouveaux éléments qui sont de forme rectangulaire offrent de plus grandes vitesses de commutation et des frais plus réduits sans sacrifier la qualité ou la sûreté du fonctionnement.

La série 400 constitue une variante de la série cylindrique 300 mais les éléments qui la composent sont de forme rectangulaire. La densité d'entassement est comparable à celle de la série 300 et, dans de nombreux cas, la nouvelle forme permet une plus grande superposition modulaire de circuits imprimés. Les modules comprennent des éléments actifs au silicium dont les vitesses de commutation dépassent 3 MHz et qui fonctionnent dans la gamme de température de —40 à +100° C.

La série 500 se compose de modules rectangulaires à bas prix dont les vitesses de commutation dépassent 5 MHz. Ils comprennent des transistors au silicium et des diodes au germanium à agglomérant d'or d'une fiabilité éprouvée. Leur gamme de température est de —20 à +90° C.

Les modules de la série 700 que l'on voit dans notre gravure sont de conception plus simple afin de fournir une gamme à bon marché d'éléments pour le réalisateur industriel. Les vitesses de commutation dépassent 1 MHz et la gamme de température s'étend de −20 à +90° C.

Un soin tout particulier a été apporté au maintien de la fiabilité à un niveau élevé, des transistors planaires au silicium étant utilisés à cet effet. Tous les éléments sont encapsulés dans de la résine d'époxyde, ils ne pèsent qu'environ 10 g et sont mécaniquement robustes.

Les modules sont présentés sous forme rectangulaire et mesurent 37,6 mm × 11,43 mm × 14 mm. Les sorties de connecteurs sont sous forme de câbles volants et le nombre maximum de sorties est de 14, sur une seule ligne de 2,54 mm d'espacement. Tous les circuits fonctionnement sur rails d'alimentation standard de +12 V et −6 V; la tolérance sur toutes les lignes est nominalement de ±0,5 V. Les niveaux logiques de cette série de modules sont semblables à ceux du reste de la gamme Ferranti, c'est à dire: état 1 = +6 V (suivant la charge) et état 0 = 0 V (suivant la charge) et état 0 = 0 V (approximativement).

La société Ferranti Ltd offre un service d'applications et dans de nombreux cas elle peut fournir des modules montés sur plaquettes de circuit imprimé pour répondre à un certain nombre de fonctions standard telles que les compteurs de décades réversibles et non-réversibles, les enregistreurs de décalage, les machines à additionner, les décodeurs, etc.

EE 79 764 *pour plus amples renseignements*

CHANGEUR AUTOMATIQUE D'ÉCHANTILLONS

Isotope Developments Ltd, Bath Road, Beenham, Reading, Berkshire

(Illustration à la page 199)

Une nouvelle gamme de systèmes changeurs automatiques constitués de corps solides pour usages médicaux et vétérinaires ainsi que pour d'autres applications cliniques et de recherche nécessitant la mesure automatique de l'énergie de radiation a été réalisée par la société Isotope Developments Ltd, qui fait partie du groupe Elliott-Automation. Cette gamme comprend les systèmes de comptage automatiques I.D.L. 'Alphamat,' 'Betamat' et 'Lobetamat' pour la mesure des radiations alpha, béta et béta à faible énergie respectivement.

Le nouveau système changeur automatique constitué de corps solides 'Lobetamat', qui est déjà utilisé en médecine, injecte automatiquement jusqu'à 50 échantillons séparés consécutivement à un assemblage à double tube Geiger-Muller protégé au plomb. Cet assemblage se compose d'un tube à fenêtre à extrémité mince entouré d'un compteur de protection qui fournit, en liaison avec un circuit anti-coïncidence, une protection supplémentaire afin d'éliminer les effets de radiations de fond et cosmiques. Un élément de lecture et à programme fonctionne conjointement avec l'échelle et un imprimeur numérique peut fournir un enregistrement continu donnant le numéro d'échantillon du malade, le comptage et le temps.

Le système 'Lobetamat' peut être pré-programmé de trois manières, c'est à dire qu'il peut s'éteindre lorsque tous les 50 échantillons ont été comptés, de réempiler les échantillons dans leur ordre original avant de s'arrêter et de réempiler et compter à nouveau, répétant ce cycle automatiquement aussi souvent qu'il est nécessaire.

EE 79 765 *pour plus amples renseignements*

SOURCE D'ÉVAPORATION

Thin Film Components Ltd, 26 Mead Lane, Chertsey, Surrey

Cet appareil est une source d'évaporation d'une grande souplesse d'emploi pouvant être montée à l'intérieur de n'importe quel système à vide poussé pour le dépôt de matériaux métalliques et non-métalliques tels que les films minces. Il peut être utilisé pour la production de microcircuits de circuits intégrés ainsi que pour les recherches générales et la mise au point de composants microminiatures.

Le système fait usage d'électrons émis par un filament chaud pour bombarder le matériau évaporant placé dans un creuset refroidi à l'eau et placé au-dessous du filament.

La tête évaporante se compose de 4 creusets montés sur un bloc de cuivre refroidi à l'eau qui est fixé à son tour à un scellement au vide rotatif permettant les dépôts consécutifs de 4 matériaux différents sans contamination. On peut également effectuer le chauffage à résistance classique de l'évaporant en fixant des barres de montage.

La puissance nécessaire à la mise en oeuvre de la source provient d'une console mesurant 53,34 cm de large x 55,88 cm de hauteur x 40,64 cm de profondeur, des inter-connexions appropriées étant également fournies. L'appareil fournit une alimentation entièrement variable (faible tension) pour le filament émetteur ainsi qu'une alimentation variable de 0 à 10 kV c.c., permettant de contrôler l'entrée de chaleur à la source. La puissance maxima est de 1,5 kVA.

L'appareil est capable, en refroidissant à l'eau le conteneur d'évaporant, de produire des films minces ultra-purs avec un minimum d'effort et une fiabilité maxima. Ces avantages, joints à l'extrême souplesse d'emploi de l'appareil en font l'outil idéal pour la recherche et la production.

EE 79 766 *pour plus amples renseignements*

INSTRUMENT DE MESURE

Brookdeal Electronics Ltd, Myron Place, Lewisham, London, S.E.13

(Illustration à la page 200)

Il s'agit ici d'un élément compact et autonome pouvant être employé comme indicateur dans les systèmes équilibrés ayant une impédance de source de 2 kΩ, tel que le détecteur sensible aux phases Brookdeal, type PD629. La précision de base de l'instrument est déterminée par un mouvement à cadre mobile de qualité supérieure ayant une largeur d'échelle de 6,5 cm. L'appareil comprend un filtre qui élimine tout effet de flottement à la fréquence d'échantillonnage. Ce dispositif ainsi que l'échelle de grand format assurent des indications d'une fiabilité et d'une répétabilité élevées.

On a particulièrement veillé au détail dans la construction de cet appareil. Ainsi l'alimentation est stabilisée aux diodes Zéner afin d'éliminer la dérive de zéro; des courants d'appoint supérieurs à la déviation sur la totalité de l'echelle peuvent être obtenus sur toutes les gammes sans perte de précision. Une sortie à faible impédance est prévue pour entraîner un enregistreur à plume. Un commutateur spécial peut inverser tant les indications de l'instrument de mesure que celles de l'enregistreur à plume sans toutefois modifier le calage du zéro ou le détalonnage. L'instrument offre un choix de trois constantes de temps et comprend un atténuateur à plots. Il est logé dans un "demi-coffret" standard Brookdeal mais il peut être ajouté au réseau de déphasage Brookdeal dans un bâti de 47,5 cm.

EE 79 767 *pour plus amples renseignements*

ALIMENTATION DE TÉLÉ-IMPRIMEUR

Industrial Instruments Ltd, Stanley Road, Bromley, Kent

(Illustration à la page 200)

Le Transipack 309/B est un bloc transistorisé d'alimentation stabilisée conçu pour l'emploi avec les télé-imprimeurs. Il fournit une tension de +80 et −80 V à une charge maxima de 1 A.

Le Transipack 309/B remplit une fonction analogue à celle du redresseur No. 66C et il peut être monté sur bâti de type postal de 47,5 cm. Il est particulièrement indiqué pour les applications où les considérations d'espace et de poids sont de prime importance car il ne mesure que 8,75 cm de hauteur sur 20 cm de profondeur et ne pèse qu'un peu plus de 6 kg.

La disjonction à maxima est assurée par un coupe-circuit à action rapide réenclenchable comprenant une bobine de fonctionnement à chaque ligne de sortie qui se déclenche à 3 A; la protection électronique est assurée contre les court-circuits de faible durée.

EE 79 768 *pour plus amples renseignements*

SYSTÈME D'INSTRUMENTS DE MESURE

Dymar Electronics Ltd, Rembrandt House, Whippendell Road, Watford, Hertfordshire

(Illustration à la page 200)

La société Dymar Electronics Ltd. a créé une gamme entièrement nouvelle d'instruments de mesure à fiches universels et de laboratoire.

La base de ce système est constituée par l'enregistreur type 70 qui peut recevoir n'importe quel instrument de mesure à fiches.

L'enregistreur, livrable en deux modèles, soit la version pour montage sur banc d'essai et celle pour bâti de 47,5 cm, comporte un mouvement de mesure de précision de 12,5 cm et des possibilitées d'alimentation étendues.

Il est muni d'un miroir antiparallaxe et il est compensé contre les variations de température.

Les instruments à fiches suivants, appartenant à la gamme initiale, sont maintenant offerts à la vente: le millivoltmètre à large bande, type 701, le microvoltmètre c.c., type 721, le kilovoltmètre c.c., type 722, le générateur de signaux a.f., type 741, et le décibelmètre, type 761. D'autres appareils sont actuellement en cours de mise au point et seront livrés prochainement. Le millivoltmètre à large bande type 701 a été conçu pour mesurer les tensions alternatives dans une gamme étendue d'amplitudes et de fréquences. Il se distingue par une impédance d'entrée de 10 MHz sur toutes les gammes, des sensibilités sur la totalité de l'échelle de 1 mV à 300 V et une gamme de fréquence s'étendant jusqu'à 5 MHz. Il emploie un amplificateur à quatre étages et un circuit redresseur stabilisé au moyen d'environ 50 dB de réaction totale assurant ainsi une linéarité excellente et une précision parfaite en dépit du vieillissement des lampes. Il comprend des bornes de sortie permettant de l'employer comme pré-amplificateur sensible pour commander des oscilloscopes ou des amplificateurs de puissance.

Le microvoltmètre à tension continue, type 721, a été prévu pour la mesure de faibles tensions continues avec charge négligeable sur le circuit soumis à l'essai. Il offre un choix de deux bornes d'entrée donnant soit une sensibilité totale de 100 μV pour une impédance d'entrée de 1 MΩ ou une sensibilité totale de 10 mV pour une impédance d'entrée de 100 MΩ. Un chopper synchrone à faible bruit est utilisé et empêche pratiquement toute dérive. Un détecteur sensible aux phases fournit une indication de polarité automatique de la tension mesurée. La tension d'étalonnage, provenant d'une diode Zéner, est d'une précision de ±1 %.

Le kilovoltmètre à tension continue, type 722, mesure avec précision des tensions s'élevant jusqu'à 30 kV avec une charge négligeable. En raison de la résistance d'entrée très élevée, c'est à dire 3000 mΩ, la consommation de courant sur la tension soumise à l'essai est d'un maximum de 10 μA à 30 kV et proportionnellement inférieure sur d'autres gammes.

L'emploi de hautes tensions stabilisées et d'alimentations de chauffage dans l'amplificateur à courant continu prévient toute dérive. Un potentiel de référence de 100 V ±1 % est fourni par un tube au néon pour l'étalonnage.

Le générateur de signaux universel type 741 a une couverture de fréquence de 30 Hz à 300 kHz et un niveau de sortie contrôlé avec précision. Il emploie un oscillateur à pont de Wien avec stabilisation d'amplitude automatique au moyen d'un thermistor. Le niveau de sortie maximum est de 10 V et il peut être attenué de 80 dB par plots de 10 dV à l'aide d'un atténuateur de précision de 600 Ω. Une commande d'amplitude précise permet l'interpolation du calage entre plots de manière à ce que la tension de sortie puisse être réglée avec précision au niveau voulu sans qu'il y ait lieu d'employer des instruments de contrôle supplémentaires.

Le décibelmètre type 761 permet une évaluation rapide du comportement du bruit de réception dans la bande de 1 à 220 MHz. Il est muni d'un générateur de bruit utilisant une diode saturée et un détecteur de bruit qui comprend un atténuateur précis de 3 dB, éliminant ainsi toute possibilité d'erreur due à la non-linéarité. Il existe deux versions de cet instrument dont l'impédance de sortie est de 50 ou 75 Ω.

EE 79 769 *pour plus amples renseignements*

AMPLIFICATEUR D'IMPULSIONS

Research Electronics Ltd, Bradford Road, Cleckheaton, Yorkshire

(Illustration à la page 201)

La société Research Electronics Ltd, vient d'annoncer la mise au point d'un nouvel amplificateur d'impulsions couplé au courant alternatif, à savoir l'amplificateur d'impulsions linéaires à large bande et de haute précision, modèle 9040 A.

Cet instrument comporte tous les perfectionnements, particulièrement en ce qui concerne la stabilité, pour pouvoir répondre aux conditions les plus rigoureuses de la spectroscopie gamma et des études de hauteur d'impulsions. Il a été prévu pour être utilisé en liaison avec l'analyseur de hauteur d'impulsion à une voie de grande précision, Research Electronics modèle 9050, mais il peut être employé également avec n'importe quel autre analyseur de hauteur d'impulsions de qualité supérieure.

Divers amplificateurs de tête permettent d'utiliser l'instrument avec les chambres d'ionisation rapide, les compteurs à scintillations et les compteurs proportionnels, y compris les compteurs au trifluorure de boron pour le comptage des neutrons. Une caractéristique intéressante de l'appareil est constituée par l'instrument de mesure qui donne une indication de la hauteur d'impulsion instantanée maxima fournie à la sortie. Cette possibilité simplifie considérablement le processus de réglage et donne une indication visuelle de la surcharge.

EE 79 770 *pour plus amples renseignements*

RELAIS DE PUISSANCE

Londex Ltd, 207 Averley Street, London, S.E.20

(Illustration à la page 201)

Une gamme entièrement nouvelle de relais à tension continue ou à tension alternative ayant une durée de vie mécanique d'environ 20 millions d'opérations a été réalisée par la société Londex Ltd, appartenant au groupe Elliott-Automation. La durée moyenne des contacts d'argent utilisés dans ces relais est d'au moins 5 millions d'opérations sur charges non-inductives à des puissances nominales de 6 A 240 V/c.a. ou 12 Vc.c.

Ces nouveaux relais, Londex type TOP, sont remarquablement robustes, sûrs et d'un fonctionnement silencieux. Il existe des modèles à permutation bipolaire et tripolaire sous forme enfermée ou non-montée. Les relais enfermés sont munis de housses Makrolon et ils sont fixés à des bases à fiches. Ces dernières sont octales, du type standard, dans le cas des relais bipolaires et à 11 broches standard pour les types tripolaires. Un schéma des connexions est imprimé sur le haut du couvercle qui porte également des cosses permettant de loger des pinces à ressort. Les contacts à relais non-montés comprennent de cosses à souder fixées dans un panneau à contact.

Les valeurs maxima de contact non-inductif sont les suivantes: courant alternatif de 6 A-440 V, 1500 W; courant continu de 6 A-250 V, 75 W jusqu'à 50 V et 20 W à 100 V ou davantage. Les bobines qui actionnent le relais dans la gamme de +6 % à −15 % de la tension nominale consomment 4 VA de courant continu, 7,5 VA d'entrée à 50 Hz c.a. ou 2 à 3 W sur tension continue.

Les temps de fonctionnement et de déclenchement sont de 3 à 10 msec sur tension alternative et de 15 à 30 msec sur tension continue.

EE 79 771 *pour plus amples renseignements*

TRANSDUCTEURS DE PRESSION

Westland Aircraft Ltd, Saunders-Roe Division, Osborne Works, East Cowes, Isle of Wight

(Illustration à la page 201)

La Division Saunders-Roe de la société Westland Aircraft Ltd a réalisé trois nouveaux transducteurs de pression pour gaz ou liquides, comprenant tous des extensomètres à feuilles agglomérées dans une configuration de pont complète.

Ces nouveaux transducteurs se distinguent par leur faible dimension, leur changement volumétrique réduit en fonction de la pression, leur fréquence naturelle très élevée et leur faculté de résister à des valeurs d'accélération élevées sans répercussion aucune sur la sortie électrique.

Trois types couvrent les gamme de 0 à 30lb/pouce carré jusqu'à 0 à 40 000 lb/pouce carré, l'un des modèles pouvant être utilisé comme transducteur différentiel dans les gammes de 0 à 50 jusqu'à 0 à 1500 lb/pouce carré.

La tension appliquée maxima est de 10 V dans un pont de 220 Ω et une sensibilité supérieure à 2 mV/V peut être obtenue sur la plupart des gammes. D'autres valeurs de résistance peuvent être fournies sur demande.

EE 79 772 *pour plus amples renseignements*

CONTRÔLEUR ET INDICATEUR DE TEMPÉRATURE

Fords (Finsbury) Ltd, Chantry Avenue, Kempston, Bedford

(Illustration à la page 201)

Un nouveau contrôleur de température, muni d'un indicateur de température visuelle, a été crée par la division d'électronique de Fords (Finsbury) Ltd.

Une caractéristique spéciale de cet instrument, vendu sous le nom de Dioford, est sa sonde semiconductrice à distance. Cette dernière ne mesure que 1,9 cm de longueur × 0,47 cm de diamètre et peut être utilisée à des distances allant jusqu'à 22,86 m de l'instrument. La télécommande s'effectue par bobinage non compensé sans perte d'efficacité. Lorsque cet instrument est muni de son indicateur, il peut effectuer le contrôle serré et l'indication de températures s'élevant jusqu'à 150° C. Sans l'indicateur, il peut contrôler des températures jusqu'à 260° C. Lorsqu'il est fourni sans l'indicateur, ce dernier étant indépendant du circuit de contrôle, il convient particulièrement à l'emploi dans les conditions comportant des vibrations considérables. Un relais à fiches renfermé dans l'instrument permet la commutation de puissance allant jusqu'à 1,25 kW sans l'emploi d'autres contacteurs.

L'instrument est de construction compacte et robuste, ne mesurant que 11,43 cm de hauteur × 22,86 cm de longueur × 15,24 cm de profondeur. Il existe des modèles pour montage sur banc d'essai ou sur panneau. L'appareil est entièrement muni de semiconducteurs et les composants sont d'une valeur nominale suffisante pour donner un ordre de fiabilité élevée.

La société Fords (Finsbury) Ltd fait partie du Tube Investments Group.

EE 79 773 *pour plus amples renseignements*

CONDENSATEUR AU MICA À DÉCADES

H. W. Sullivan Ltd, Murray Road, Orpington, Kent

(Illustration à la page 201)

Il s'agit ici d'un condensateur au mica à trois décades entièrement blindé avec un condensateur à air supplémentaire à variation continue. Il est nécessaire d'ajouter un quatrième cadran variable afin d'obvier à l'inconvénient et au manque de précision qui pourrait être provoqué par l'addition d'un condensateur variable indépendant à un instrument à trois cadrans.

Le coefficient de température est de l'ordre de ±20 parties dans $16^6/°$ C et des tensions de régime de pointe allant jusqu'à 500 V peuvent être utilisées.

Le condensateur est réglé pour pouvoir fournir une indication entièrement directe sur toute la gamme avec une précision de 0,1 %. En outre, il a été rigoureusement étudié pour avoir une faible inductance et résistance en série, de sorte que son comportement aux fréquences supérieures est excellent tant en ce qui concerne la capacité effective que le facteur de puissance. Les limites supérieures de fréquence sans perte de précision sont comme suit:

C (μF)	f (kHz)
0,001	150
0,01	50
0,1	20
0,1	7

Jusqu'à ces limites de fréquence, le facteur de puissance n'augmente pas au-dessus de l'ordre de 0,0005. Si des défauts de précision de l'ordre de 1 % sont admissibles, les limites de fréquence supérieures peuvent être portées de 20 kHz à 500 kHz selon la valeur de capacitance.

EE 79 774 *pour plus amples renseignements*

Résumés des Principaux Articles

Le LOPAD: Un système de traitement linéaire logarithmique des données analogiques par T. K. Cowell et M. Gordon

Résumé de l'article aux pages 146 à 151

Les auteurs décrivent un système de calcul pouvant effectuer une gamme étendue d'opérations algébriques par rapport direct à des variables représentées par des signaux électriques.

Le principe de ce système est de traduire les signaux sous forme logarithmique de manière à ce que des multiplications ou des divisions puissent être effectuées par simple addition ou soustraction.

La conversion logarithmique est réalisée par l'emploi de la caractéristique directe d'une diode à semiconducteur du type commercial. Le mode de calcul est, d'autre part, indépendant de la température ambiante, malgré l'effet marqué de cette dernière sur les caractéristiques de la diode.

Fonctionnant directement sur signaux de transducteur, ce système permet de gagner un temps considérable dans le traitement des résultats expérimentaux et il trouve actuellement un grand nombre d'applications dans le domaine biomédical.

Un stimulateur piloté par quartz par J. A. Reynolds

Résumé de l'article aux pages 152 à 155

On a réalisé un stimulateur à double impulsion dont on peut varier la synchronisation entre les impulsions de sortie par incréments de 0,1 millisecondes dans une gamme totale d'une seconde. La précision de la synchronisation est assurée par l'emploi d'un oscillateur à cristal ainsi que d'un certain nombre de Dékatrons vendus dans le commerce.

Le transistor à semiconducteur à oxyde métallique par R. A. Hilbourne et J. F. Miles

Résumé de l'article aux pages 156 à 160

Le transistor à semiconducteur à oxyde métallique est un nouveau dispositif semiconducteur fonctionnant suivant le principe de l'effet de champ mais présentant une résistance d'entrée thermiquement stabilisée beaucoup plus élevée. Ces dispositifs à caractéristique de stabilité ont main-

tenant été suffisamment mis au point pour pouvoir être livrés à la vente et cet article constitute une introduction descriptive.

La construction du transistor à semiconducteur à oxyde métallique, ainsi que la théorie de son fonctionnement et ses caractéristiques sont analysées en détail. Plusieurs applications des dispositifs tant dans les circuits linéaires que comme commutateurs sont indiquées.

Compteurs et enregistreurs de décalage à commutateurs statiques par J. F. Young

Résumé de l'article aux pages 161 à 163

L'un des avantages des compteurs à commutateurs universels statiques à diodes-transistors à haute tension est qu'ils peuvent actionner directement des indicateurs numériques à cathode froide. Un code de comptage décimal-binaire approprié permet l'emploi du transfert de dizaines à couplage direct sous forme de réaction tout en gardant une redondance suffisante pour permettre l'économie et un certain degré d'égalisation de la charge dans la matrice à diodes utilisée pour le décodage. Un enregistreur de décalage utilisant les mêmes commutateurs est également décrit.

L'étude de dispositifs de couplage équilibreurs entre deux lignes à haute fréquence de 50/600Ω pour hautes puissances
par J. Wood

Résumé de l'article aux pages 164 à 168

Cet article traite de l'étude de transformateurs à haute puissance à large bande de haute fréquence et souligne en particulier l'emploi de bobinage équilibré/déséquilibré pour adapter des lignes coaxiales à faible impédance à un systéme d'alimentation équilibré ou à des antennes à haute impédance.

L'emploi de ferrite en nickel zinc contribue de manière décisive à l'étude du transformateur et à son comportement qui se caractérise par une très grande efficacité (plus de 97%) et un taux d'ondes stationnaires très réduit.

Les conditions d'étude sont examinées en détail et des propositions sont présentées pour une étude mi-théorique mi-empirique.

Des résultats expérimentaux se rapportant à des transformateurs typiques sont indiqués avec des tracés de taux d'onde stationnaire sur coordonnées cartésiennes.

Dispositifs constitués de corps solides pour redresseurs E.H.T. et diodes d'efficacité pour télérécepteurs par F. D. Bate

Résumé de l'article aux pages 169 à 173

Cet article traite des problèmes de l'emploi des dispositifs constitués de corps solides pour redresseurs E.H.T. et des diodes d'efficacité dans les récepteurs de télévision. Il étudie l'effet de la charge emmagasinée et de la résistance directe des dispositifs constitués de corps solides sur la régulation des diodes E.H.T. ainsi que de l'échauffement résultant du flux de courant de capacité lorsque de longues chaines de diodes sont utilisées.

Un système multiplex de démonstration à division de temps basé sur la technique de la modulation d'amplitude des impulsions
par R. Barrett

Résumé de l'article aux pages 174 à 177

Il s'agit ici d'un système de laboratoire de conception simple réalisé pour montrer aux étudiants les principes de base des techniques de division de temps multiplex, ainsi que le partage dans le temps d'un moyen de communication moyen entre plusieurs canaux d'information séparés.

Un chercheur d'extrémité simple et automatique par P. Zorkoczy

Résumé de l'article aux pages 178 à 179

L'auteur décrit une méthode pour obtenir une valeur maxima de la sortie d'un système. Des données sont ainsi extraites de la différence entre la sortie $y(t)$ et le retard différé $y(t-\tau)$. Ces renseignements sont employés dans une combinaison d'éléments de commutation et analogiques simples pour contrôler les valeurs de l'entrée de manière à atteindre la sortie maxima.

Le multiplicateur d'électrons de canal par J. Adams et B. W. Manley

Résumé de l'article aux pages 180 à 181

Le multiplicateur d'électrons de canal se compose d'une dynode résistive continue formée sur la surface intérieure d'un tube dont le rapport longueur: diamètre est d'environ 50. Lorsqu'un potentiel d'environ 2,5kV est appliqué entre les extrémités du tube, un électron pénétrant par l'extrémité à faible potentiel déclenche une cascade d'électrons secondaires à partir du mur. Le nombre d'électrons émergeant de l'extrémité à potentiel élevé du tube peut dépasser 10^5.

Ces dispositifs peuvent être petits et robustes et sont potentiellement utiles dans les applications de l'espace.

Divers montages de canaux ont été étudiés aux laboratoires de recherches Mullard et il a été constaté qu'en donnant au multiplicateur de canaux une forme en spirale on évite ainsi complètement les effets de pression.

Contrôle de fréquence de circuits multivibrateurs astables à transistors par signaux de tension continue variable
par S. L. Hurst et C. L. Seal

Résumé de l'article aux pages 182 à 185

La fréquence des circuits multivibrateurs astables à transistors accouplés au collecteur du type courant est normalement indépendante de la tension de façon très marquée. Grâce à certaines modifications légères du circuit on a pu remédier à ce défaut et la fréquence d'oscillation qui en résulte est maintenant presque linéaire par rapport à la tension. Des utilisations possibles pour ces circuits modifiés sont suggérées dans cet article.

Relais modulateurs à transistors à interrupteurs simples ou en paires accordées par A. Lydén

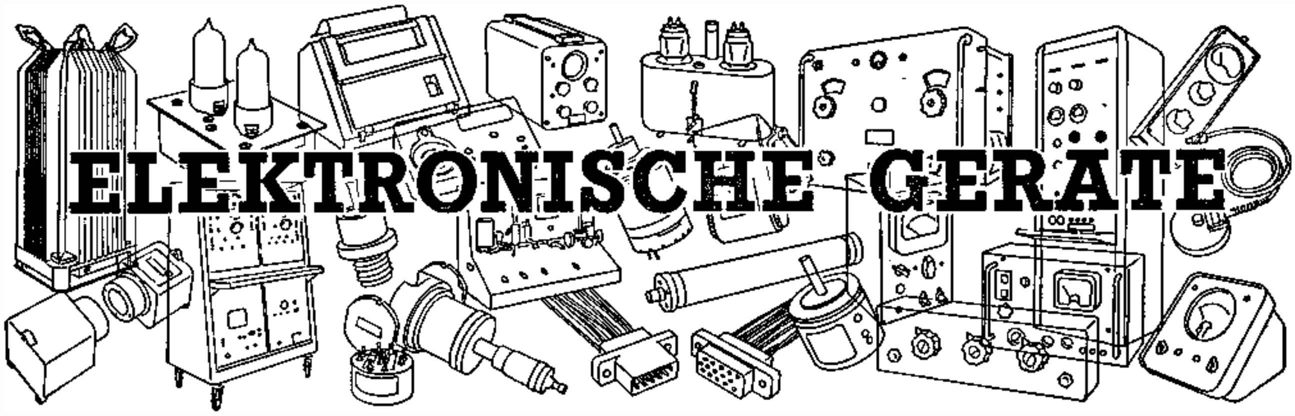
Résumé de l'article aux pages 186 à 189

L'auteur examine la théorie avancée qui utilise les rapports modifiés de Ebers et de Moll pour le relais modulateur à transistors. Il donne des équations tant pour l'état stable que pour le comportement transitoire. Les interrupteurs en paires accordées ont une tension décalée et des fuites plus réduites ainsi qu'une moindre dérive que le transistor simple. Le montage à deux transistors en parallèle a la plus faible dérive inhérente.

L'étalonnage dynamique du gain d'un intégrateur électronique par K. Taylor

Résumé de l'article à la page 190

Cet article décrit une méthode précise pour mesurer et étalonner le gain d'un intégrateur électronique. Il fournit en outre une estimation expérimentale du degré de précision.



ELEKTRONISCHE GERÄTE

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 196 bis 201

Rohrtrimmer

Vertrieb: Standard Telephones & Cables Ltd, Capacitor Division, Brixham Road, Paignton, Devonshire

(*Abbildung Seite 196*)

Die STC-Capacitor Division vertreibt nunmehr in Grossbritannien die preisgünstigen Stangard - Glasrohr - Kolbentrimmer. Diese Trimmer sind für Geräte wie beispielsweise Oszillografen, Messbrücken, Messender, Frequenznormale, Eichgeräte, Spektrumanalysatoren, FM-Tuner und viele andere Verwendungszwecke geeignet, für die bisher nur Keramik - Scheibentrimmer, Keramik-Rohrtrimmer und Lufttrimmer infrage kamen.

Stangard-Trimmer werden von der J.F.D. Electronics Corporation in New York hergestellt. Sie sind robust gebaut und haben eine Vinyleinkapselung, die das Glasdielektrikum gegen Stösse schützt. Niedrige Erstellungskosten werden durch Festelektrodenbänder und automatisierte Fertigungsstrecken erzielt.

Diese Trimmer kommen in zwei Ausführungen: eine für Bestückung gedruckter Schaltungen und die andere für Montage auf Frontplatten oder Chassis. Die Betriebsspannung ist 1 kV—, und die vier Pikofarad-Regelbereiche sind: 0,5 ... 3,0; 0,8 ... 5,0; 1,0 ... 8,0 und 1,0 ... 12,0.

Elektrische Eigenschaften sind u.a.: Q besser als 1000 bei 1 MHz, Isolationswiderstand 10^6 MΩ bei 500 V— und ein Dielektrikum, das über 2 kV widersteht. Ausserdem haben Stangard-Trimmer ein reibungsloses Einstelldrehmoment und mehrgängige Einstellung für Feinabstimmung. Diese Eigenschaften werden über den Temperaturbereich —55° C ... +125° C voll aufrechterhalten.

EE 79 751 *für weitere Einzelheiten*

Hochleistungs-Wanderfeldröhren

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

(*Abbildung Seite 196*)

Die von Mullard vorgestellte neue Wanderfeldröhre LB4-20 wird als das fortschrittlichste Modell für den Mikrowellen-Fernmeldedienst bezeichnet. Sie gibt eine Sättigungsleistung von 20 W ab, d.h. 100 Prozent mehr als ältere, speziell für den Nachrichtenverkehr entwickelte Typen, und hat ausserdem einen niedrigen AM-PM-Konversionsfaktor von nur 1,5°... 2° pro dB.

40 dB Verstärkung und ein Frequenzbereich von 3,7 ... 4,2 GHz gestatten entweder grössere Bandspreizung für mehr Kanäle oder eine längere Übertragungsstrecke ohne Verzerrungszunahme.

Die Röhre ist in die Halterung P4L5 eingebaut, aber nicht geschlossen konstruiert, d.h. eine Ersatzröhre kann in eine vorhandene Halterung eingebaut werden, wobei die letztere lediglich auszurichten ist. Zur Erleichterung der Ausrichtung sind die Fokussierelemente vorn auf der Halterung angeordnet. Die Röhre wird durch einen Spezialsockel, der sperrige Kabelbäume überflüssig macht, zwangsläufig positioniert.

Ergänzend zur LB4-20 wurde die Wanderfeldröhre LB6-20 entwickelt, die auch eine Sättigungsleistung von 20 W abgibt, aber mit 37 dB Verstärkung im Frequenzbereich 5,9 ... 6,5 GHz arbeitet. Die LB6-20 hat trotz dieser hohen Ausgangsleistung einen AM-PM-Konversionsfaktor von 1,5°... 2° pro dB, ihr Rauschpegel ist also äusserst niedrig.

Für die LB4-20 sowohl wie die LB6-20 ist bei horizontalem Einbau Eigenkühlung ausreichend, bei vertikalem Einbau muss sie jedoch durch einen Konvektionskanal oder langsame Luftdurchströmung unterstützt werden. Für Sonderanwendung ist eine Ausführung mit Ableitkühlung lieferbar.

EE 79 752 *für weitere Einzelheiten*

Wirbelstromtester

C.N.S. Instruments Ltd, 61 Holmes Road, London, N.W.5

(*Abbildung Seite 196*)

Das Modell 600 ist für die folgenden Aufgabenbereiche bestimmt: Kontrolle von Schraubenlöchern auf Ermüdungs-, Spannungs- oder Korrosionsrisse; Vergleichsmessung von Wandstärken, z.B. an Tragflügelprofilen; Sortieren gemischter Werkstoffe, vorausgesetzt sie haben unterschiedlichen spezifischen Widerstand. Das volltransistorisierte Prüfgerät hat folgende technische Daten: einen umschaltbaren Frequenzbereich von 4 kHz, 10 kHz und 25 kHz (für gewisse Verwendungszwecke bis zu 100 kHz); drei durch Schalter wählbare Betriebsarten, und zwar Amplitudendemodulation, Phasenaufnahme und Phasenunterdrückung, wodurch unerwünschte Variable vom Personal vernachlässigt werden können; eine 152 mm lange Messskala zur Erleichterung des Ablesens; ein breites Sortiment von Messköpfen, einschliesslich solcher für Flächen, Durchführungsspulen und Nachweis von Rissen in Schraubenlöchern.

Sonderzweckmessköpfe lassen sich einzelnen Anforderungen entsprechend herstellen, und Druckkarten werden auf Lager gehalten, um gegebenenfalls den Service zu erleichtern.

EE 79 753 *für weitere Einzelheiten*

Dehnungsmessbrücke

The M.E.L. Equipment Co. Ltd, 207 Kings Cross Road, London, W.C.1

(*Abbildung Seite 197*)

Eine neue tragbare Dehnungsmessbrücke Philips PR9205 ist nunmehr von M.E.L. Equipment Co Ltd zu beziehen. Das nur gerade über 4,5 kg wiegende Gerät ist robust konstruiert und für Aussendienst in den verschiedensten Industriezweigen geeignet. Es kann mit Dehnungsmessstreifen zwischen 120 Ω und 2 kΩ Widerstand in Halb- oder Vollbrückenschaltung eingesetzt werden.

Bei Einrichtung des Gerätes für einen bestimmten Verwendungszweck lässt es sich unabhängig vom Messbereich nullen. Mit eingerichtetem Gerät ist der Nullabgleichbereich ±10 000 Mikrodehnung und der Messabgleichbereich ±8250 Mikrodehnung, was einen geeichten

Gesamtbereich von ±18 250 Mikrodehnung ergibt.

Nullabgleich wird auf einem Drehspulinstrument angezeigt und durch einen Wendelpotentiometerregler bewirkt, dessen 3600°-Skala in Mikrodehnung geeicht ist. Kurzzeitmessungen können daher schnell und genau vorgenommen werden, während die Genauigkeit von Langzeitmessungen durch hohe Konstanz und genaue Wiedereinstellbarkeit des Nullpunktes gewährleistet ist. Schwankungen der Umgebungstemperatur haben wenig Einfluss auf das Gerät, dessen Drift zwischen −10° und +40° C unter 1 Mikrodehnung/° C liegt.

Das Gerät ist volltransistorisiert und wird aus aufladbaren internen Batterien gespeist. Es kann aber ebenso gut mit fünf gewöhnlichen Batteriezellen betrieben werden, wobei Spannungsänderungen zwischen 5,5 und 7,5 V die Arbeitsweise nicht beeinflussen.

EE 79 754 *für weitere Einzelheiten*

Mehrspuriger Aufnahmekopf

Gresham Lion Electronics Ltd, Lion Works, Hanworth Trading Estate, Feltham, Middlesex

(*Abbildung Seite 197*)

Gresham Lion Electronics Ltd hat einen 33-Spuren-Aufnahme-Magnetkopf für 1″-Magnetband entwickelt. Dieser neue Kopf gibt die höchste in Grossbritannien kommerziell greifbare Spurendichte und ist vor allem für Digitalanwendungsmöglichkeiten gedacht, bei denen es auf hohe Spurendichte ankommt.

Die 33 Spuren sind in einem kompakten Kopf untergebracht, der 38,1 mm hoch, 31,8 mm tief und 15,2 mm breit ist. Die Spurenteilung ist 0,76 mm und die Spurbreite 0,25 mm.

Der Kopf hat eine Spaltbreite von 6,4 μm, und die Wicklungen für jede Spur haben vier Anschlüsse, so dass die beiden symmetrischen Wicklungen entweder einzeln, in Serie oder parallel geschaltet werden können. Er gibt bei 300 Bits pro Zoll und 38 cm/s eine Spannung von 1,5 mV_{ss} ab.

Für Rechteckwellensignale von 10 kHz oder darunter und 38 cm/s Bandgeschwindigkeit ist die Übersprechdämpfung 20 dB. (Das Übersprechen wird durch gleichzeitiges Aufzeichnen gleichphasiger Signale auf zwei alternativen Spuren und Messen des in den zwischenliegenden Spuren induzierten Signals bei Wiedergabe bestimmt.)

EE 79 755 *für weitere Einzelheiten*

Zeitgeber-Zähler

Elesta Electronics Ltd, 34 Cambridge Park, London, E.11

(*Abbildung Seite 197*)

Dieses tragbare transistorisierte Gerät mit Neonlichtanzeige kann mit einer Zählrate von bis zu 3 kHz bis zu 9999 zählen. Als Zeitgeber eingesetzt und mit dem internen Oszillator entweder auf 1000 oder 100 Hz geschaltet, hat es Höchstzeitbereiche von 9,999 s bzw. 99,99 s. Wahlweise steht auch eine Frequenz von 10 Hz zur Verfügung; andere Frequenzen können gegen Sonderauftrag geliefert werden. Das Eingangssignal soll Rechteckwellenform und mindestens 7 V_{ss} Amplitude haben. Auf Wunsch kann ein interner Schmitt-Trigger eingebaut werden, der die Empfindlichkeit auf 3 V_{ss} steigert und zuverlässigen Betrieb mit allen Eingangswellenformen sicherstellt.

EE 79 756 *für weitere Einzelheiten*

Spannungsteiler

Cambridge Instrument Co. Ltd, 13 Grosvenor Place, London, S.W.1

(*Abbildung Seite 197*)

Dieser Kelvin-Varley-Spannungsteiler wurde entwickelt, um bei Aufrechterhaltung eines konstanten Eingangswiderstandes zwischen Null und x1 Vervielfachungsverhältnisse in Schritten von ×0.0001 zu ermöglichen. Er ist in drei Ausführungen für Eingangswiderstände von 1 kΩ, 10 kΩ und 100 kΩ und Höchstbelastbarkeit mit 75, 250 und 750 V lieferbar.

Der Teiler hat vier Verhältnisskalen mit Verhältnissen von ×0.1, ×0.01, ×0.001 und ×0.0001, von denen die drei höheren je neun Stellungen haben, die unterste noch eine zehnte, die das Höchstverhältnis auf ×1 bringt. Die Schalter haben Schichtbürsten und versilberte Kontakte; ihre Kontaktwiderstandsschwankungen sind nicht grösser als ±0.00002 Ω/Dekade.

Alle Widerstände bestehen aus sorgfältig gealtertem Manganindraht und—mit Ausnahme derjenigen für die unterste Skala—Ayrton-Perry-Wicklungen auf Glimmerkarten in spannungsfreien Halterungen. Sie werden auf ±0,05 Prozent ihres Nennwertes abgeglichen, und Verhältnisfehler sind innerhalb ±0,1 Prozent der Skaleneinstellung.

EE 79 757 *für weitere Einzelheiten*

Modulare Gestelleinheiten

Vero Electronics Ltd, South Mill Road, Regents Park, Southampton

(*Abbildung Seite 197*)

Mit Einführung einer 5¼″ (133 mm) Modul-Gestelleinheit kann Vero Electronics Ltd nunmehr das volle Sortiment der gängigen Grössen in 5¼″, 7″ und 8¾″ Höhe und 1″, 2″, 4″ und 8″ Modul-Frontplattenbreite liefern. Die Gesamtbreite des Einschubes ist 19″ bei 11″ (280 mm) und 14″ (356 mm) Tiefe. Nach diesem System kann man in wenigen Minuten und lediglich mit einem Schraubenzieher eine der über 400 möglichen Kombinationen in starrer und dauerhafter Struktur aus Standardelementen erstellen.

EE 79 758 *für weitere Einzelheiten*

Integrierter Funktionsverstärker

SGS-Fairchild Ltd, 23 Stonefield Way, Ruislip, Middlesex

(*Abbildung Seite 197*)

Aus ihrer Serienfertigung kann SGS-Fairchild Ltd nunmehr den μA702, einen kompletten, in Planar-Epitaxial-Technik aus einem einzigen Siliziumplättchen hergestellten Funktionsverstärker liefern.

Der μA702 ist ein Funktionsverstärker mit hoher Verstärkung für Einsatz mit externen Gegenkopplungselementen, die die Betriebseigenschaften bestimmen. Kennzeichnende Eigenschaften sind niedrige Gegengleichspannung und Drift, grosse Bandbreite, niedrige Leistungsaufnahme, grosser Ausgangshub und Betrieb innerhalb eines breiten Speisespannungsbereiches.

Dieser Verstärker kann—je nach Anschlussweise der Gegenkopplung—als Gleichspannungsverstärker in verschiedenen Verknüpfungen benutzt werden. Unter den Anwendungsmöglichkeiten ist die Verstärkung der Ausgangsspannung von Wandlern wie Ohmsche Brücken, Halleffekt-Elemente, Fotodioden usw. zu nennen. Mit geeigneten Gegenkopplungselementen kann der μA702 die in Analogrechnern erforderlichen Integrations-, Differenzierungs-, Summier- und Subtraktions-Funktionen ausführen. Ausserdem kann man den Verstärker mit reaktiven Gegenkopplungselementen für genaues Frequenzformen der Verstärkungskurve und mit einem überbrückten T-Filter als Bandpassverstärker einsetzen. In den Gegenkopplungsweg können Dioden eingeschaltet werden, um Detektoren mit niedrigem Schwellenwert oder Instrumentgleichrichter zu bilden.

Der μA702 hat einen Spitzenausgangsstrom von 10 mA, eine Höchstverlustleistung von 250 mW, eine typische Leerlaufverstärkung von 63 dB und eine typische Wärmedrift von 5 μV/° C. Der Verstärker ist sowohl in militärischer Ausführung μA702 (−55° C…+125° C) wie in professioneller Form μA702C (−25° C…+100° C) lieferbar.

Abgebildet ist eine Fotomikrografie des Funktionsverstärkers μA702, die seine Geometrie zeigt.

EE 79 759 *für weitere Einzelheiten*

Thyristor

British Brown-Boveri Ltd, Glen House, Stag Place, London, S.W.1

(*Abbildung Seite 198*)

Die British Brown-Boveri Ltd hat vor kurzem einen neuen Thyristor in ihr Fertigungsprogramm aufgenommen. Dieser mit CS.30 bezeichnete Typ hat einen mittleren Durchlassstrom von 70 A bei einer Bolzentemperatur von 80° C und 180° Leitungswinkel. Er ist für Nennspannungen zwischen 100 V und 600 V in 100-V-Abständen lieferbar, und der Mindestdurchbruch in Durchlassrichtung wird nicht unter Spannungen auftreten, die 40 Prozent über diesen Nennwerten liegen. Der Einperioden-(20 ms)-Spitzenstossstrom des CS.30 ist 1300 A.

EE 79 760 *für weitere Einzelheiten*

Tonfrequenzwobbler

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(*Abbildung Seite 198*)

Dawe Instruments Ltd hat einen neuen

Wobbler vorgestellt, der den Tonfrequenzbereich von 20 Hz...20 kHz mit nur einer logarithmischen Skala überstreicht. Das mit Typ 443B bezeichnete Gerät gibt über den gesamten Frequenzbereich—und selbst bei Anschluss an nichtlineare oder frequenzabhängige Netzwerke—eine konstante Ausgangsspannung ab. Die Frequenzskala kann voll durchgedreht und die Frequenzsteuerung zur automatischen Aufzeichnung des Frequenzganges mit dem Dawe-Pegelschnellschreiber 1406 gekoppelt werden. Das Wobbeln wird entweder durch Betätigung eines Frontplattenschalters oder durch Fernsteuerung zwecks Gleichlauf mit dem Schreiber 1406 eingeleitet. Ein einstellbarer Frequenzmarker ist vorhanden.

Der Oszillator arbeitet nach dem Überlagerungsprinzip, und die Ausgänge von zwei HF-Oszillatoren (von denen einer mit Festfrequenz, der andere mit über einen kleinen Bereich veränderlicher Frequenz arbeitet) werden in einen Mischer gespeist. Die Differenzfrequenz im Tonfrequenzbereich wird gefiltert und verstärkt; eine übertragergekoppelte Endstufe gibt Ausgangsimpedanzen von 6, 60, 600 oder 6000 Ω. Der Ausgang ist von 12,5 V bis zu 125 μV in 10-dB-Stufen regelbar; die Ausgangsspannung an den Lastklemmen oder zum Eingang des Abschwächers ist kontinuierlich regelbar und wird auf einem Frontplatteninstrument angezeigt.

Die Frequenzänderung erfolgt mittels eines speziell entwickelten Kondensators mit echt logarithmischer Skala innerhalb ±1 % ±1 Hz. Für manuelle Frequenzeinstellung gibt es Grob- und Feinantrieb auf der Frontplatte oder extern vom Pegelschnellschreiber. Die Frequenz kann bei jeder Einstellung intern oder extern mit bis zu 200 Hz moduliert werden.

Eine Schaltung für Spannungspressung regelt den Ausgang des Oszillators auf konstanten Ausgangsstrom oder Ausgangsspannung. Der Klirrfaktor liegt von 200...2000 Hz unter 0,1 Prozent und steigt zwischen 20 Hz und 20 kHz auf nicht mehr als 1 Prozent bei einer Abschwächer - Ausgangsspannung von 10 V. Das für Gestelleinbau oder in Tischausführung lieferbare Gerät hat einen Verbrauch von 70 W am 50- oder 60-Hz-Netz und wiegt einschliesslich Gehäuse etwa 25 kg.

EE 79 761 *für weitere Einzelheiten*

Impuls-Zeitgeber

Telefon Fabrik Automatic A/S, 7 Amaliegade, Kopenhagen, K, Dänemark

(*Abbildung Seite 198*)

Der Impuls-Zeitgeber EN5-01 ist in erster Linie als Präzisionsprüfgerät für Bauelemente und Ausrüstungen des automatischen Fernsprechdienstes gedacht, hat aber auch überall dort Anwendungsmöglichkeiten, wo genaue und konstante Impulse mit bekanntem Tastverhältnis erforderlich sind.

Der Generator hat zwei "statische Kontakte" (Festkörper), die Betrieb mit Arbeits- und Ruhekontakten gestatten. Ausserdem werden ein Zeitbasissignal sowie Markerimpulse für die Vorder- und Hinterflanke der Kontaktfunktion abgegeben.

Obwohl der Generator hauptsächlich als genauer Zeitgeber entwickelt wurde, kann er gleichzeitig als genauer Amplitudengeber und Stromversorgung verwendet werden. Durch Verbinden der statischen Kontakte mit einer externen Stromversorgung, z.B. dem TFA-Typ EC3-01, kann ein Ausgangssignal mit einer Amplitude erreicht werden, das genau dem der Stromversorgung entnommenen Amplitude entspricht.

Die eindeutigen Kenndaten des Öffnens und Schliessens der statischen Kontakte gestatten sehr genaue Kontrolle elektrischer Aufladungen in kleinen Inkrementen.

Ein quarzgesteuerter Oszillator liefert Impulse mit 10 μs Periode. Im nachgeschalteten Frequenzteiler oder Zeitbasisvervielfacher können Perioden von 10 μs, 100 μs oder 1 ms gewählt werden.

Dem Zeitbasisvervielfacher ist ein Impulswähler nachgeschaltet, der für jede gewünschte Anzahl von Impulsen (N) zwischen 1 und 99 einstellbar ist.

Sowie der Wähler N Impulse aufgenommen hat, wird er auf Null zurückgestellt und gibt einen Impuls an einen 100-Stufen-Ringzähler ab.

Der Ringzähler gibt für jede Gruppe von 100 aufgenommenen Impulsen einen Ausgangsimpuls ab, der mit 0-Impuls bezeichnet wird und dessen Periode der gewünschten Folgezeit des Endsignals entspricht. Der Ringzähler gibt auch einen Extraimpuls ab, der p Stufen später als der 0-Impuls auftritt. Dieser p-Impuls kann zwischen 1 und 99 Prozent der Periode und in Stufen von 1 Prozent verzögert werden.

Zwei statische Kontakte werden durch die zwei Ausgangsimpulse des Ringzählers betätigt. Der O-Impuls öffnet einen der Kontakte (den Ruhekontakt) und schliesst den anderen (den Arbeitskontakt). Der p-Impuls stellt die Kontakte auf ihren normalen Zustand zurück.

Aus dem oben erwähnten ist zu ersehen, dass die Verzögerung des p-Impulses direkt das Tastverhältnis als Prozentsatz der Periode bestimmt.

Zwischen Ringzähler und statischen Kontakten gehen die O-Impulse durch eine Torschaltung, die durch einen Dreiweg-Hebelschalter auf der Frontplatte gesteuert wird. In der oberen arretierten Stellung des Schalters werden alle O-Impulse durchgelassen, und die Kontakte arbeiten kontinuierlich. In der unteren, nicht arretierten Stellung wird nur eine auf der Frontplatte vorgewählte Anzahl von O-Impulsen durchgelassen, und die Kontakte werden so oft wie vorgewählt betätigt. In der mittleren oder Bereitschaftsstellung ist das Tor gesperrt, keine O-Impulse werden zu den Kontakten durchgelassen, die deshalb in ihrem normalen Zustand bleiben. In der Bereitschaftsstellung des Schalters kann die Torschaltung jedoch durch eine externe negative Spannung angesteuert werden, wodurch sich ein Impulssignalgemisch erzeugen lässt, das für einen Augenblick ein mehrstelliges Wählersignal eines Fernsprechteilnehmers nachbildet.

Ein Spezialgerät zum Programmieren eines mehrstelligen Wählersignals wird in Kürze lieferbar werden.

EE 79 762 *für weitere Einzelheiten*

Spektrumanalysator

Fenlow Electronics Ltd, Springfield Lane, Weybridge, Surrey

(*Abbildung Seite 199*)

Der Fenlow-Spektrumanalysator SA3 ist ein neues Gerät, das die spektrale Dichte von Wellenformen im Frequenzbereich 1,5 Hz...5 kHz bestimmt. Festfilterbreiten von 0,3, 1,5, 37,5 und 187 Hz sind lieferbar. Die von den Filtern durchgelassene Rauschleistung wird mit Hilfe von Thermoelementen gemessen, deren Ausgang auf einem Messgerät angezeigt wird und zum Treiben eines automatischen Kurvenschreibers zur Verfügung steht.

Der Analysator arbeitet nach dem Überlagerungsprinzip: das aufgenommene Signal wird mit den Oszillatorsignalen in Quadratur multipliziert, und der durch ein Tiefpassfilter gespeiste Ausgang der Multiplizierschaltung treibt die Thermoelemente.

Die benutzten sechs Funktionsverstärker ergeben grosse Genauigkeit; Digitalmethoden werden für die Erzeugung der internen Oszillatorsignale für die beiden Quadraturkanäle angewendet.

Das Gerät hat Einrichtungen zur Überwachung der Signale an allen Schlüsselpunkten.

EE 79 763 *für weitere Einzelheiten*

Logische Schaltelemente

Ferranti Ltd, Ferry Road, Edinburgh 5, Schottland

(*Abbildung Seite 199*)

Das Ferranti-Programm für logische Schaltelemente wurde durch drei neue Serien erweitert, die einen breiteren Benutzerkreis ansprechen sollen. Die neuen Elemente sind rechteckig ausgeführt, haben hohe Schaltgeschwindigkeiten und sind billiger, ohne dabei Qualität oder Zuverlässigkeit einzubüssen.

Serie 400 ist eine Alternative zur zylindrischen Serie 300 in rechteckiger Form, die mit der Serie 300 vergleichbare Packdichte gibt. In vielen Fällen gestattet die neue Form bessere modulare Schichtung gedruckter Schaltungen. Die Moduln enthalten aktive Siliziumelemente, deren Schaltgeschwindigkeiten über 3 MHz liegen und die über einen Temperaturbereich von −40° C... +100° C arbeiten.

Serie 500 besteht aus einer Reihe preiswerter rechteckiger Bausteine mit

Schaltgeschwindigkeiten von über 5 MHz. Sie sind mit Silizium-Transistoren und Germanium-Golddrahtdioden bewährter Zuverlässigkeit bestückt und für einen Temperaturbereich von $-20°$ C... $+90°$ C ausgelegt.

Die abgebildeten Bausteine der Serie 700 sind in vereinfachter Ausführung gebaut, um dem industriellen Konstrukteur ein preiswertes Baumuster zu bieten. Schaltgeschwindigkeiten liegen über 1MHz, der Temperaturbereich zwischen $-20°$ C und $+90°$ C.

Besondere Aufmerksamkeit wurde der Aufrechterhaltung der hohen Zuverlässigkeit gewidmet, und es werden durchweg Silizium-Planartransistoren verwendet. Alle Bausteine sind vollständig in Epoxydharz gekapselt und bei geringem Gewicht (<10 g) mechanisch robust.

Die Abmessungen der rechteckig ausgeführten Bausteine sind 37,6 × 11,4 × 14 mm. Die Anschlüsse für Steckverbindungen sind als freie Zuleitungen ausgeführt, und die Höchstanzahl der Ausgänge ist 14 in einer Linie mit 2,54 mm Abstand. Alle Schaltungen arbeiten mit Standard-Speisespannungen von $+12$ V und -6 V mit einer Toleranz von $\pm 0{,}5$ V. Die logischen Pegel dieser Bausteinserien sind dieselben wie für die anderen des Ferranti-Programmes, d.h. Zustand 1 = +6 V (lastabhängig) und Zustand 0 = 0 V (ungefähr).

Ferranti hat einen technischen Beratungsdienst, der in vielen Fällen auf Druckschaltungskarten montierte Bausteine zur Ausführung einer Anzahl von Standardfunktionen liefern kann, wie z.B. umkehrbare und nicht umkehrbare Dekadenzähler, Schieberegister, Addierer, Decoder usw.

EE 79 764 *für weitere Einzelheiten*

Automatischer Probenwechsler

Isotope Developments Ltd, Bath Road, Beenham, Reading, Berkshire

(Abbildung Seite 199)

Ein neues Programm automatischer Festkörper-Wechslersysteme für medizinische, tierärztliche und andere klinische und Forschungszwecke, die automatisches Messen von Strahlungsenergie erfordern, wurde von Isotope Developments Ltd, einem Mitglied der Elliott-Automation-Gruppe, entwickelt. Es umfasst auch die automatischen I.D.L.-Zählsysteme "Alphamat", "Betamat" und "Lobetamat" für das Messen von Alpha-, Beta- bzw. weichen Betastrahlungen.

Das neue automatische Festkörper-Wechslersystem "Lobetamat", das bereits in der Medizin Verwendung findet, schiebt bis zu 50 getrennte Proben nacheinander zu einer mit Blei abgeschirmten GM-Doppelröhrengruppe. Diese besteht aus einem Rohr mit dünnem Endfenster, das von einem "Schutz"-Zähler umgeben ist, der zusammen mit einer Antikoinzidenzschaltung zusätzliche Abschirmung gegen die Einwirkung von Hintergrund- und Höhenstrahlungen gibt. In Verbindung mit dem Untersetzer wird ein Programmier- und Anzeigegerät betrieben, und ein Digitaldruckwerk erstellt ein kontinuierliches Protokoll mit der Probennummer des Patienten, dem Zählresultat und der Zeit.

Das "Lobetamat"-System kann für drei Arbeitsweisen vorprogrammiert werden: Abschalten nach Beendigung der Zählung aller 50 Proben; Wiederaufstapeln der Proben in der ursprünglichen Reihenfolge vor Abschalten; Wiederaufstapeln und Wiederzählen sowie automatisches Wiederholen dieses Zyklus so oft wie gewünscht.

EE 79 765 *für weitere Einzelheiten*

Verdampfungsquelle

Thin Film Components Ltd, 26 Mead Lane, Chertsey, Surrey

Die Ausrüstung ist eine vielseitige Verdampfungsquelle für Einbau in jedes Hochvakuumsystem für Niederschlag metallischer und nichtmetallischer Stoffe als dünne Filme. Sie eignet sich besonders für Einsatz in der Herstellung von Mikroschaltungen und integrierten Schaltungen, sowie in der allgemeinen Forschung und Entwicklung von Mikrominiatur-Bauelementen.

In diesem System bombardieren die von einem heissen Heizfaden emittierten Elektronen das aufzudampfende Material, das sich unterhalb des Heizfadens in einem wassergekühlten Tiegel befindet.

Der Aufdampfkopf besteht aus vier Tiegeln auf einem wassergekühlten Kupferblock, der seinerseits mit einer drehbaren Vakuumdichtung so zusammengebaut ist, dass vier verschiedene Stoffe nacheinander ohne Verunreinigung aufgedampft werden können. Durch Anbau von hochbelastbaren Stromschienen lässt sich das aufzudampfende Material mittels herkömmlicher Widerstandsheizung verdampfen.

Die für Betrieb der Quelle erforderliche Energie wird über geeignete Verbindungskabel einem ca. 53 cm breiten, 56 cm hohen und 40 cm tiefen Schrank entnommen. Zur Steuerung der Wärmezufuhr zur Quelle steht eine vollregelbare Niederspannungsversorgung für den Emitterheizfaden sowie eine von 0...10 kV— regelbare Speisespannung zur Verfügung. Die Höchstleistung ist 1,5 kVA.

Durch Wasserkühlung des Trägers des aufzudampfenden Materials ist die Ausrüstung in der Lage, mit geringstem Aufwand und höchster Zuverlässigkeit ultrareine Dünnfilme zu erzeugen, was in Verbindung mit ihrer aussergewöhnlichen Vielseitigkeit die Ausrüstung zu einem sehr wertvollen Werkzeug für Forschung ebenso wie Fertigung macht.

EE 79 766 *für weitere Einzelheiten*

Messgerät

Brookdeal Electronics Ltd, Myron Place, Lewisham, London, S.E.13

(Abbildung Seite 200)

Es handelt sich hier um ein kompaktes, in sich geschlossenes Gerät, das als Anzeiger in ausgeglichenen Systemen mit 2 kΩ Quellenimpedanz Verwendung findet, z.B. mit dem phasenempfindlichen Brookdeal-Gleichrichter PD629. Die Grundgenauigkeit des Gerätes wird durch das hochwertige Drehspulmesswerk mit 66 mm Skalenlänge bestimmt. Ein eingebautes Filter beseitigt Schwankungen bei der Schaltfrequenz, was in Verbindung mit der langen Skala zuverlässige Ablesungen und Wiederholbarkeit gewährleistet.

Allen Einzelheiten wurde in diesem Gerät besondere Aufmerksamkeit gewidmet: die Kompensationsspannung ist zenerstabilisiert, um Nulldrift zu beseitigen, und grössere Kompensationsströme als der Skalenendwert lassen sich in allen Bereichen ohne Minderung der Genauigkeit erreichen. Zum Treiben von Schreibern sind niederohmige Ausgänge vorhanden. Ein Schalter für Richtungsumkehr kehrt sowohl Messgerät- wie auch Schreiberanzeige um, wodurch aber weder Kompensation, noch Nullung geändert werden. Das Gerät bietet die Wahl von drei Zeitkonstanten und hat einen Stufenabschwächer. Es ist in einem Standard - Brookdeal - "Halbgehäuse" untergebracht und kann mit einem Brookdeal-Phasenschieber in ein volles Gehäuse für 19"-Gestelleinbau eingebaut werden.

EE 79 767 *für weitere Einzelheiten*

Fernschreiber-Stromversorgung

Industrial Instruments Ltd, Stanley Road, Bromley, Kent

(Abbildung Seite 200)

Transipack 309/B ist eine transistorisierte, besonders für Fernschreiber entwickelte Konstant-Stromversorgung, die bei Höchstbelastung mit 1 A +80 V und −80 V abgibt.

Das Modell führt dieselben Funktionen wie das Gleichrichtergerät No 66C aus und kann in ein 19"-Postgestell eingebaut werden. Es ist für Verwendungszwecke geeignet, in denen Volumen und Gewicht ausschlaggebend sind, und ist bei 6,35 kg Gewicht nur 89 mm hoch und 203 mm tief.

Ein rückstellbarer Schnellabschalter mit einer Erregerspule in jeder Speiseleitung gibt Überlastungsschutz, der bei 3 A ausgelöscht wird; gegen kurzzeitige Kurzschlüsse schützt eine elektronische Schaltung.

EE 79 768 *für weitere Einzelheiten*

Messgerätesystem

Dymar Electronics Ltd, Rembrandt House, Whippendell Road, Watford, Hertfordshire

(Abbildung Seite 200)

Dymar Electronics Ltd hat ein völlig neues Programm von Labor- und Mehrzweckmessgeräten in Einschubtechnik herausgebracht.

Die Grundlage des Systems bildet das Messinstrument Typ 70, das jedes Einschubmessgerät des Programmes aufnimmt.

Das Messinstrument ist in Ausführun

gen für Tischgebrauch oder für Einbau in 19"-Gestelle lieferbar und enthält ein 127-mm-Präzisionsmesswerk sowie umfangreiche Stromversorgungen.

Das Messinstrument ist mit einem Ablesespiegel ausgerüstet und für Temperaturschwankungen kompensiert.

Aus dem geplanten Programm wurden folgende Einschubmessgeräte bekanntgegeben: Breitbandmillivoltmeter Typ 701, Gleichstrom-Mikrovoltmesser Typ 721, Gleichstrom-Kilovoltmesser Typ 722, NF-Messender Typ 741 und Rauschfaktormesser Typ 761. Weitere Einschübe sind in der Entwicklung und werden in Kürze produktionsreif.

Das Breitbandmillivoltmeter ist für das Messen von Wechselspannungen über einen breiten Bereich von Amplituden und Frequenzen ausgelegt. Es hat für alle Bereiche einen Eingangswiderstand von 10 MΩ, Vollausschlagempfindlichkeiten von 1 mV bis zu 300 V und einen über 5 MHz hinausgehenden Frequenzbereich. Die vier Verstärkerstufen und die Gleichrichterschaltung werden durch etwa 50 dB Gesamtgegenkopplung stabilisiert, was ausgezeichnete Linearität und volle Genauigkeit mit Röhrenalterung gewährleistet. Vorhandene Ausgangsklemmen gestatten Einsatz des Gerätes als empfindlicher Vorverstärker zum Treiben von Oszillografen und Leistungsverstärkern.

Der Gleichstrom-Mikrovoltmesser 721 ist zum Messen kleiner Gleichspannungen ohne nennenswerte Belastung der zu prüfenden Schaltung bestimmt. Zwei Eingangsklemmen erlauben wahlweise einen Skalenendwert von 100 μV bei 1 MΩ Eingangsimpedanz oder 10 mV Skalenendwert bei 100 MΩ Eingangsimpedanz. Ein rauscharmer Zerhacker im Eingang sorgt für praktisch driftfreies Arbeiten und ein phasenempfindlicher Demodulator für automatische Polaritätsanzeige der gemessenen Spannung. Die einer Zener-Diode entnommene Eichspannung hat eine Unsicherheit von ± 1 Prozent.

Der Gleichstrom-Kilovoltmesser 722 misst Spannungen bis zu 30 kV genau und ohne nennenswerte Belastung. Bei dem sehr hohen Eingangswiderstand von 3 000 MHz ist die Stromentnahme bei einer zu messenden Spannung von 30 kV nur 10 μA und entsprechend niedriger für andere Bereiche. Driftfreiheit wird durch stabilisierte Anoden und Heizspannungen im Gleichstromverstärker gewährleistet. Für Eichzwecke gibt eine Glimmröhre eine Bezugsspannung von 100 V ± 1 %.

Der Mehrzweck-Messender 741 gibt erweiterte Frequenzüberstreichung von 30 Hz ... 300 kHz und genau kontrollierten Ausgangspegel. Der Wien-Brückenoszillator hat einen Thermistor für automatische Amplitudenkonstanthaltung. Der Höchstausgangspegel ist 10 V und kann mittels eines 600-Ω-Präzisionsabschwächers in 10-dB-Stufen um 80 dB abgeschwächt werden. Eine Amplitudenfeinregelung gestattet Interpolation der Einstellung zwischen Stufen, so dass man die Ausgangsspannung ohne zusätzliche Kontrollinstrumente genau auf den gewünschten Pegel einstellen kann.

Der Rauschfaktormesser 761 erlaubt schnelle Bestimmung der Rauschkenndaten eines Empfängers im Band 1 ... 220 MHz. Der Rauschgenerator arbeitet mit einer gesättigten Diode und einem Rauschdetektor mit genauem 3-dB-Abschwächer, wodurch alle möglichen auf Nichtlinearität zurückzuführenden Fehler behoben werden. Der Einschub ist in zwei Ausführungen mit 50 Ω oder 75 Ω Ausgangsimpedanz lieferbar.

EE 79 769 *für weitere Einzelheiten*

Impulsverstärker

Research Electronics Ltd. Bradford Road, Cleckheaton, Yorkshire

(*Abbildung Seite 201*)

Ein neuer hochwertiger, wechselstromgekoppelter Impulsverstärker Modell 9040A wird von Research Electronics Ltd als linearer Breitband-Impulsverstärker hoher Genauigkeit vorgestellt.

In das Gerät wurden alle erdenklichen Verfeinerungen eingebaut, speziell in bezug auf die schwersten Anforderungen von Gammaspektroskopie und Impulshöhenuntersuchungen an die Stabilität. Es wurde für Einsatz in Verbindung mit dem hochgenauen Einkanal-Impulshöhenanalysator 9050 der Research Electronics entwickelt, kann jedoch auch zusammen mit jedem anderen hochwertigen Impulshöhenanalysator verwendet werden.

Verschiedene Vorverstärker erlauben Einsatz des Verstärkers mit Schnellionisationskammern, Szintillationszählern und Proportionalzählern einschliesslich Borfluoridzählern für das Zählen von Neutronen. Eine interessante Einrichtung ist das eingebaute Messgerät, das die grösste momentane Impulshöhe anzeigt, die im Ausgang ankommt. Das vereinfacht das Rüsten erheblich und ermöglicht visuelle Überlastungsanzeige.

EE 79 770 *für weitere Einzelheiten*

Hochleistungsrelais

Londex Ltd. 207 Averley Street, London, S.E.20

(*Abbildung Seite 201*)

Londex Ltd, ein Mitglied der Elliott-Automation-Gruppe, hat Relais für Gleich- oder Wechselstrom in völlig neuer Bauart und mit einer mechanischen Mindestlebensdauer von 20 Millionen Betätigungen herausgebracht. Die durchschnittliche Lebensdauer der in diesem Relais benutzten Silberkontakte ist bei induktionsfreier Last und typischen Nennwerten von 6 A, 240 V∼ oder 12 V– mindestens 5 Millionen Betätigungen.

Diese als Londex-Typ TOP angebotenen neuen Relais sind ungewöhnlich robust, zuverlässig und lautlos in Betrieb. Modelle können mit zwei- oder dreipoligem Umschalter in gekapselter oder Einbauausführung geliefert werden. Gekapselte Relais haben Staubhauben aus Makrolon und sind mit Einsteck-sockeln ausgerüstet, und zwar Standard-oktalsockeln für zweipolige und elfpoligen Standardsockeln für dreipolige Typen. Ein Anschlussdiagramm ist auf die Haubenoberseite gedruckt, auf der auch die Nasen für die Halteklammer angeordnet sind. Kontakte der Einbaurelais sind an Lötfahnen in einer Kontaktplatte angeschlossen.

Maximale ohmsche Kontaktbelastungen sind für Wechselstrom 6 A, 440 V, 1 500 W, für Gleichstrom 6 A, 250 V und 75 W bis zu 50 V, 20 W bei 100 V und darüber. Spulen, die das Relais bei +6 Prozent bis −15 Prozent der Nennspannung erregen, nehmen kontinuierlich 4 VA∼, bei Einschalten 7,5 VA∼ 50 Hz oder 2 ... 3 W– auf. Ansprech- und Abfallzeiten sind 3 ... 10 ms mit Wechselstrom und 15...30 ms mit Gleichstrom.

EE 79 771 *für weitere Einzelheiten*

Druckgeber

Westland Aircraft Ltd. Saunders-Roe Division, Osborne Works, East Cowes, Isle of Wight

Abbildung Seite 201)

Die Saunders-Roe-Abteilung der Westland Aircraft Ltd hat drei neue Druckgeber für Flüssigkeiten und Gase vorgestellt, die alle mit aufgekitteten Dehnungsmessstreifen in vollständiger Brückenschaltung arbeiten.

Diese neuen Messwertwandler haben kleine Abmessungen, geringe Raumänderung mit Druck, sehr hohe Eigenfrequenz und die Fähigkeit, hohe Beschleunigungen ohne Einwirkung auf den elektrischen Ausgangswert auszuhalten.

Die drei Typen überstreichen die Bereiche 0 ... 2 kg/cm² bis zu 0 ... 2 812 kg/cm², und ein Modell kann als Differenzdruckgeber in den Bereichen 0 ... 3,5 kg/cm² bis zu 0 ... 105 kg/cm² dienen.

Die höchste an eine 220-Ω-Brücke angelegte Spannung ist 10 V, und für die meisten Bereiche ist eine Empfindlichkeit von über 2 mV/V erzielbar. Andere Widerstandswerte sind auf Wunsch lieferbar.

EE 79 772 *für weitere Einzelheiten*

Temperaturwächter und -anzeiger

Fords (Finsbury) Ltd. Chantry Avenue, Kempston, Bedford

(*Abbildung Seite 201*)

Vom Bereich Elektronik der Fords (Finsbury) Ltd wurde die Fertigung eines neuen, mit Sichtanzeige ausgerüsteten Temperaturwächters aufgenommen.

Ein wesentliches Konstruktionsmerkmal des unter dem Namen "Dioford" vertriebenen Gerätes ist der Halbleiter-Fernmesskopf, der bei einem Durchmesser von 4,76 mm nur 19 mm lang ist. Er kann mit nichtkompensierten Kabeln ohne Einbusse von Leistungsfähigkeit bis zu rund 23 m vom Instrument entfernt eingesetzt werden. Das Gerät mit dem Anzeiger ist für genaue Regelung und Anzeige von Temperaturen bis zu 150° C geeignet, regelt ohne Anzeiger jedoch Temperaturen bis zu

260° C. Wenn es ohne den von der Regelschaltung unabhängigen Anzeiger geliefert wird, kann man es mit besonders gutem Erfolg unter Bedingungen einsetzen, in denen es starken Schwingungen unterliegt. Ein Einsteckrelais im Gerät ist für Schalten von Leistungen bis zu 1¼ kW ohne andere Kontaktgeber bemessen.

Die kompakte und robuste Konstruktion des für Tischeinsatz oder Einbau lieferbaren Gerätes gestattet Abmessungen von nur 114 mm Höhe, 229 mm Länge und 152 mm Tiefe. Zur Erzielung hoher Zuverlässigkeit werden durchweg Halbleiter und reichlich bemessene Bauelemente verwendet.

Fords (Finsbury) Ltd gehört zur Tube Investments-Gruppe.

EE 79 773 *für weitere Einzelheiten*

Dekaden-Glimmerkondensator

H. W. Sullivan Ltd, Murray Road, Orpington, Kent

(Abbildung Seite 201)

Es handelt sich um einen vollständig abgeschirmten Dreidekaden-Glimmerkondensator mit einem zusätzlichen Luftdrehkondensator. Eine vierte (kontinuierlich einstellbare) Skala ist notwendig, um die Unbequemlichkeit und Ungenauigkeit zu beseitigen, die die Kombination eines getrennten Drehkondensators und eines Dreiskaleninstrumentes mit sich bringt.

Die Grössenordnung des Temperaturbeiwertes ist $\pm 20 \times 10^{-6}/°C$, Spitzenbetriebsspannungen bis zu 500 V sind zulässig.

Der Kondensator wird über den Gesamtbereich bei direkter Ablesung mit 0,1 Prozent Genauigkeit eingestellt. Da er ausserdem sorgfältig auf niedrigen Serienwiderstand und Induktivität ausgelegt wurde, ist sein Verhalten in bezug auf wirksame Kapazität und Leistungsfaktor bei höheren Frequenzen ausgezeichnet. Die ohne Minderung der Genauigkeit erreichbaren oberen Grenzfrequenzen sind:

C(μF)	f(kHz)
0,001	150
0,01	50
0,1	20
1,0	7

Bis zu diesen Grenzfrequenzen wird der Leistungsfaktor nicht über ca. 0,0005 erhöht. Wenn Ungenauigkeiten von etwa 1 Prozent zulässig sind, können die oberen Grenzfrequenzen je nach Kapazitätswert auf 20 ... 500 kHz erhöht werden.

EE 79 774 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

LOPAD—ein logarithmisches prozessgekoppeltes Verarbeitungssystem für Analogdaten von T. K. Cowell und M. Gordon

Zusammenfassung des Beitrages auf Seite 146-151

Ein beschriebenes Rechnersystem führt an durch elektrische Signale dargestellten Variablen eine grosse Anzahl algebraischer Operationen aus.

Nach dem angewandten Prinzip werden die Signale in logarithmische Form umgewandelt, so dass Multiplikationen und Divisionen als einfache Additionen oder Subtraktionen ausgeführt werden können.

Logarithmische Umwandlung wird durch Ausnutzung der Durchlasskennlinien einer kommerziellen Halbleiterdiode erreicht. Eine entwickelte Technik ergibt eine von der Umgebungstemperatur unabhängige Rechnerarbeitsweise, trotzdem die Temperatur einen ausgeprägten Einfluss auf die Kenngrössen der Diode hat.

Das System arbeitet direkt an Messwertwandlersignalen und kann dadurch viel Zeit bei der Verarbeitung versuchsmässiger Ergebnisse einsparen. Es findet zur Zeit in mehreren Untersuchungen in der Biomedizin Anwendung.

Ein quarzgesteuertes Reizgerät von J. A. Reynolds

Zusammenfassung des Beitrages auf Seite 152-155

In einem entwickelten Doppelimpuls-Reizgerät kann der Abstand zwischen den Ausgangsimpulsen in Stufen von 0,1 Millisekunden über einen Gesamtbereich von 1 Sekunde geregelt werden. Ein Quarzoszillator und eine Anzahl kommerziell greifbarer Dekatron-Einheiten gewährleisten Genauigkeit der Zeitgebung.

Der Metalloxydhalbleiter-Transistor von R. A. Hilbourne und J. F. Miles

Zusammenfassung des Beitrages auf Seite 159-160

Der Metalloxydhalbleiter-Transistor ist ein neues Halbleiterelement, das nach dem Feldeffektprinzip arbeitet, jedoch einen viel höheren, temperaturbeständigen Eingangswiderstand bietet. Die Entwicklung hat jetzt den Punkt erreicht, wo Transistoren mit konstanten Kennlinien kommerziell greifbar sind, und dieser Beitrag ist als Einführung zu diesen Elementen gedacht.

Die Konstruktion der MOS-Transistoren, die Theorie ihrer Arbeitsweise und ihre Kenngrössen werden gegeben. Schliesslich werden verschiedene Anwendungsmöglichkeiten in linearen Schaltungen sowie ein Schalter beschrieben.

Zähler und Schieberegister mit statischen Schaltbausteinen von J. F. Young

Zusammenfassung des Beitrages auf Seite 161-163

Zähler mit Universalschaltbausteinen, die mit Hochvolttransistoren und Dioden bestückt sind, haben den Vorteil, Kaltkathoden-Ziffernanzeiger direkt treiben zu können. Ein geeigneter binärdezimaler ählcode gestattet Verwendung direktgekoppelter Zehnerübertragung als Rückstellung mit ausreichender Redundanz, um in der zum Entschlüsseln benutzten Dioden-Matrix Einsparungen und einen gewissen Belastungsausgleich zu erlauben.

Entwurf von 50/600-Ω-Symmetrieübertragern für hohe Leistungen von J. Wood

Zusammenfassung des Beitrages auf Seite 164-168

Dieser Beitrag befasst sich mit dem Entwurf von Breitband-HF-Übertragern für hohe Leistungen unter besonderer Berücksichtigung der Verwendung unsymmetrischer und symmetrischer Wicklungen für den Sonderzweck der Anpassung niederohmiger Koaxialleitungen an symmetrische Speiseleitungen oder hochohmige Antennen.

Die Verwendung von Nickel-Zink-Ferriten trägt wesentlich zur Konstruktion und Leistung des Übertragers bei, der durch einen hohen Wirkungsgrad (über 97 Prozent) und ein sehr niedriges Stehwellenverhältnis gekennzeichnet ist.

Die Konstruktionserfordernisse werden eingehend besprochen und Vorschläge für eine teils theoretische, teils empirische Behandlung gemacht.

Ergebnisse von Versuchen mit typischen Übertragern werden zusammen mit auf kartesischen Koordinaten aufgetragenen Stehwellenverhältnissen gegeben.

Festkörperelemente für Höchstspannungsgleichrichter und Spardioden in Fernsehempfängern von F. D. Bate

Zusammenfassung des Beitrages auf Seite 169-173

Die mit der Verwendung von Festkörperelementen für Höchstspannungsgleichrichter und Spardioden in Fernsehempfängern zusammenhängenden Probleme werden untersucht. Der Einfluss der gespeicherten Ladung und des Durchlasswiderstandes des Festkörperelementes auf die Regelungen der Höchstspannungsdioden wird besprochen. Die sich bei Verwendung langer Diodenketten vom Kapazitätsstromfluss ergebende Erwärmung wird auch untersucht.

Ein Zeitmultiplex-Vorführungssystem mit Impuls-Amplitudenmodulation von R. Barrett

Zusammenfassung des Beitrages auf Seite 174-178

Der Beitrag beschreibt ein einfaches Labor-System, das gebaut wurde, um Studenten der Fernmeldetechnik die Grundprinzipien der Zeitmultiplextechnik—die zeitanteilige Benutzung eines gemeinsamen Nachrichtenträgers durch mehrere getrennte Informationskanäle—vorzuführen.

Ein einfacher automatischer Extremwertsucher von P. Zorkoczy

Zusammenfassung des Beitrages auf Seite 178-179

Ein Verfahren zum Suchen und Folgen des optimalen Wertes des Ausganges eines Systems wird beschrieben. Die Information erhält man von der Differenz zwischen dem Ausgang $y(t)$ und dem verzögerten Ausgang $y(t-\tau)$; sie wird in eine Kombination einfacher Analog- und Schaltelemente eingegeben, um die Eingangswerte so zu steuern, dass der maximale Ausgang erreicht wird.

Der Kanal-Elektronenvervielfacher von J. Adams und B. W. Manley

Zusammenfassung des Beitrages auf Seite 180-181

Der Kanal-Elektronenvervielfacher besteht aus einer ununterbrochenen ohmschen Dynode auf der Innenoberfläche eines Rohres, dessen Länge zum Durchmesser im Verhältnis von ungefähr 50 steht. Wird an die beiden Rohrenden eine Spannung von etwa 2,5 kV gelegt, so löst ein am Niederspannungsende eintretendes Elektron eine Kaskade von Sekundärelektronen aus der Wand aus. Die Anzahl der sich ergebenden und aus dem Hochspannungsende des Rohres austretenden Elektronen kann 10^8 überschreiten.

Bauelemente dieser Art können klein und robust hergestellt werden und möglicherweise für Weltraumzwecke nützlich sein.

Verschiedene geometrische Kanalausführungen wurden in den Mullard-Forschungslaboratorien untersucht, und es stellte sich heraus, dass Druckeinwirkungen völlig vermieden werden, wenn der Kanalvervielfacher als Spirale ausgeführt wird.

Frequenzsteuerung astabiler Transistor-Multivibratorschaltungen mittels regelbarer Gleichspannungssignale von S. L. Hurst und C. L. Seal

Zusammenfassung des Beitrages auf Seite 182-185

Die Frequenz herkömmlicher kollektorgekoppelter astabiler Multivibratorschaltungen ist üblicherweise fast spannungsunabhängig. Durch geringe Abwandlungen der Schaltung trifft das nicht mehr zu, und die resultierende Schwingungsfrequenz kann jetzt im wesentlichen linear mit der Spannung sein. Anwendungsmöglichkeiten für die modifizierten Schaltungen werden vorgeschlagen.

Zerhacker mit Einzel- und gepaarten Transistoren von A. Lydén

Zusammenfassung des Beitrages auf Seite 186-189

Die verfeinerte Theorie mit modifizierten Ebers- und Moll-Beziehungen für Transistor-Zerhacker wird besprochen. Es werden Gleichungen für den eingeschwungenen Zustand sowohl wie das Einschwingverhalten gegeben. Pärchen-Schalter haben niedrigere Restspannung, kleinere Spitzen und niedrigere Drift als Einzeltransistoren. Die Schaltung mit zwei parallelgeschalteten Transistoren hat ihrer Natur entsprechend die niedrigste Drift.

Dynamische Eichung der Verstärkung eines elektronischen Integrators von K. Taylor

Zusammenfassung des Beitrages auf Seite 190

In diesem Beitrag wird ein genaues Verfahren zum Messen und Eichen der Verstärkung eines elektronischen Integrators beschrieben und eine versuchsmässige Einschätzung der Genauigkeit gegeben.