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Commentary

THE Council of Engineering Institutions, formerly known as the Engineering Institutions Joint Council, was last month granted a Royal Charter and has been honoured by His Royal Highness the Duke of Edinburgh who has agreed to become its Founder President for a period of five years.

The Council was set up in 1962 as a federation of thirteen chartered Engineering Institutions with a corporate membership of some 130 000.

The object for which the council was originally set up is to promote and co-ordinate in the public interest, the development of the science, art and practice of engineering, in short, the general improvement of British engineering in all its aspects, and with this in view the council proposes to:—

(1) Establish, uphold and advance the standards of qualification, competence and conduct of the professional engineer.
(2) Advance the aims and objectives of its members, so far as they relate to the science, art or practice of engineering.
(3) Establish relations with Government, national and international bodies and with the public; and to co-operate with other bodies at all levels of technical and professional competence whose objects and purposes may be related to those of the Council.
(4) Encourage co-operation with Universities and other educational institutions.
(5) Encourage co-operation between its members on matters relating to the science, art and practice of engineering.

For professional engineers who were already corporate members of one (or more) of the constituent Institutions on 3 August this year, the granting of the Royal Charter provides the new title of 'Chartered Engineer' which may be abbreviated to 'C.Eng'.

Since the Council was set up considerable progress has been made towards unified standards of assessing professional competence through the evolution of common examinations but it is to the future that the council is looking with its proposals for the up-and-coming engineer who wishes to become 'Chartered'.

To start with he must have acquired corporate membership of one of the Constituent Institutions and then he will have to pass the council's examination which will be established at the level of a British University degree or hold a qualification accepted by the Council as giving exemption from that examination. It is the intention of the Constituent Institutions to adopt the council examination as setting the minimum academic standard for entry to corporate membership against which degrees, diplomas and other awards will be judged for exemption.

It is not proposed to change the present system whereby the thirteen Institutions conduct their own methods of establishing professional competence before admission to corporate membership and there is every intention of maintaining the many links between the Institutions and their counterparts in the Commonwealth and other countries as these are regarded as the main and most efficient means of dealing with the specific disciplines.

The council hopes to set up a two part examination, Part I of which will, it is intended, to be common entranceship to the thirteen Institutions although it is difficult at this stage to reconcile the varying requirements of the individual Institutions each with its own specialized branch of science and technology.

As far back as 1953 certain related chartered engineering Institutions joined together to establish a joint common Part I examinations and later, an Engineering Institution Examinations Standing Committee was formed to carry out this duty which has since been taken over by the Engineering Institutions Joint Council—now the Council of Engineering Institutions—who over the past two years have been studying reports from various sub-committees.

At the moment the proposed syllabus of the common Part I Examination comprises papers in the following subjects: Mathematics, Properties of Materials, Principles of Electrical Engineering, Applied Mechanics, Fluid Mechanics, Thermodynamics; which may well suit the candidate entering what one may call the broad old established mechanical, civil or electrical engineering fields, but is outside the scope of the aspiring electronic engineer.

It is not surprising therefore to learn that one of the constituent members—the Institution of Electronic and Radio Engineers has raised this point and would have preferred an examination, equivalent to the level of the first year of a degree course, in which a candidate would be allowed the choice of a group of subjects appropriate to his particular field of engineering, leaving four compulsory papers of broader basic coverage with the suggested titles of Mathematics, Engineering Science, Principles of Mechanical Engineering and Principles of Electrical Engineering.

The Institution of Electronic and Radio Engineers feels that while a common entrance examination may broaden a candidate's approach it precludes the degree of specialization which characterizes present day engineering studies, particularly in universities where newly-established departments in electronics are a post-war feature, and to quote from a recent statement the Institution "feels that while it must continue to give the fullest support to the principle of a common examination it can no longer agree that this demands a Part I examination with no optional papers."

Details of the common entrance examination are to be issued later this year by the Council of Engineering Institutions when no doubt this particular problem will have been solved.

A Tunnel Diode Quantizer

By R. V. L'Archeveque*, B.Sc., Ph.D., and B. McA. Sayers†, M.Sc., Ph.D., D.I.C.

A fast-stepping tunnel diode staircase generator operating in decades is described. A stepping rate of the order of 10^7 steps per second allows the use of the unit for quantization, analogue to digital conversion and other operations with fair precision and potentially high speed.

(Voir page 634 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 641)

RECENT work in the authors' laboratory has required detailed study of fast-stepping tunnel diode chains mainly for the generation of rapidly-adjustable and precise voltage reference levels. One particular use for this operation is in a quantizer recently developed for analogue to digital conversion and other functions. In the essence of the scheme interchange between speed and resolution is readily achieved; for the type of physiological signals with which the laboratory is concerned high resolution is not required, but economy, circuit simplicity and potentially fast sampling are important.

Quantization is effected by comparing the input signal with a reference level potential; when the difference between input and reference exceeds the selected quantum magnitude in either direction, a level detector generates a pulse of polarity corresponding to the sign of the instantaneous difference voltage. This pulse marks the instant at which the input signal departs from the reference level by a single quantum interval, and causes the reference level voltage to step up or down by one quantum interval. Thus a single two-level detector, in conjunction with a stepping reference, can be used to define an input range of quantum intervals equal in number to the number of available reference steps.

The stepping reference has been realized very easily by means of tunnel diode chains, and it has been found that precision and long-term stability appropriate to the requirements can be met with a system of this type. In the prototype instrument to be described here, four chains have been employed, of ten diodes each, to define 100 levels on each side of zero level.

The ten tunnel diodes making up each decade are sequentially switched by a succession of level-detector output pulses from a low voltage to a high voltage stable state and the total chain voltage rises by ten successive steps; the chain is re-set after the tenth diode is switched. In a manner of a decade counter the pulse initiating the resetting action effectively serves as trigger pulse to a second decade of tunnel diodes; the voltage steps due to the first decade chain are arranged to be weighted to a tenth the contribution of steps due to the second decade chain. Thus by summing the suitably weighted output of the two decade chains a series of 100 voltage levels may be set up, spaced proportionally to the difference between low and high voltage stable states of each tunnel diode.

While in the prototype instrument four decades have been employed, two for each polarity, measurements indicate that a third decade on each signal polarity could be incorporated, thus accommodating 2 000 levels. Rapid adjustment of the reference level may be effected and an input signal followed with relatively high speed. Simplifications of the system allow its use also as a sampling analogue to digital convertor.

With the present system the input signal is represented by the constantly updated state of the reference level which is formed from the algebraic sum of instantaneous voltages from the positive and negative decade pairs. An algebraic count of the level detector output pulse train allows digital indication of the instantaneous input signal magnitude.

This technique depends upon fast reliable switching of tunnel diode chains in single steps, and on the precision and stability with which the total chain voltage defines a set of reference levels. Since the switching of tunnel diodes in chains is accomplished here on principles somewhat different from some previous methods, it is convenient initially to discuss the switching action in qualitative terms.

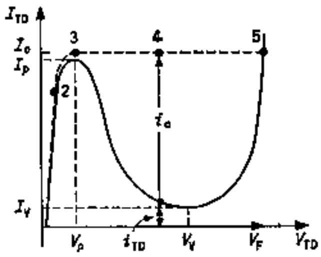


Fig. 1 (left). Simplified operating point trajectory superimposed on a tunnel diode characteristic for switching of a tunnel diode by current drive increasing from 0 to I_o

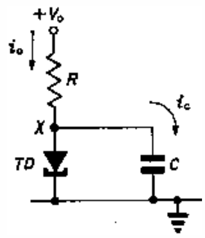


Fig. 2 (right). Basic switching circuit, the capacitor mainly representing junction capacitance

Single Tunnel Diode Switching

Consider first the single tunnel diode characteristic, Fig. 1, and the circuit of Fig. 2. The explanation of switching action to be given will mainly involve the initial static conditions and the influence of signal-induced variation from the static operating point. Hence the variational or dynamic resistance of the tunnel diode under various conditions may be expected to prove a valuable guide to circuit behaviour. The essential ideas are as follows. First, the tunnel diode variational resistance is infinite at the point of peak current, and thus any immediate subsequent change of circuit current (by increase of i_o or by injection of current otherwise into node X) must then flow in parallel pathways only. In the circuit of Fig. 2, the diverted current will flow into the shunt capacitance (essentially the junction capacitance of the diode; it is assumed for simplicity of discussion, except where stated, that this capacitance is linear and the small diode inductance is neglected). This effective diversion of current initiates a corresponding change of capacitor voltage. Second, if the diversion of current into parallel pathways

* *Now at Atomic Energy of Canada Ltd.*
† *Imperial College of Science and Technology.*

causes a change in voltage which further diverts current from the diode, a regenerative switching action will result. Similar arguments suitably rephrased apply to deviations about the point of minimum diode current, the valley point.

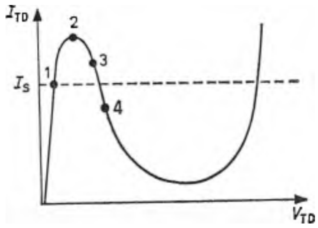


Fig. 3. Tunnel diode characteristic with standing current, I_S, and quiescent operating point 1

In practice a current drive situation could be realized with a constant current source, magnitude I_s, with an additional current pulse (diode-coupled, say) to displace the diode operating point above the peak current. Note that there is a minimum pulse width required for full forward switching. If the input current pulse takes the diode from static point 1 (Fig. 3) past the peak current, point 2, regenerative capacitor charging and switching will be initiated. But, if at point 3, say, with the instantaneous diode current still greater than the original standing value I_s, the input triggering pulse was removed, the diode could revert to point 1. If the input pulse duration exceeds the interval until the diode current falls below its standing value, say to point 4, full forward switching will be successfully initiated.

In a voltage drive situation the supply resistance R is small. To set up a circuit with two stable states, say, a choice of series resistance as for the circuit of Fig. 2 must be such that intersection of the load line (No. 1, say, in Fig. 4(a)) with the diode characteristic defines two points, each corresponding to a region of positive variational resistance. If the load resistance is chosen less than the smallest magnitude of negative variational resistance (No. 2 in Fig. 4(a)), the circuit will always be stable; this can be seen by considering the equivalent circuit in Fig. 4(b). If $R < |R_{DN}|$ where R_{DN} is the minimum negative variational resistance exhibited by the diode, the parallel combination of R and R_{DN} will present an invariably positive variational resistance to the current source $i_o = V_o/R$. Assume, however, that a suitably large value of R is chosen and V_o be allowed to increase slowly from zero.

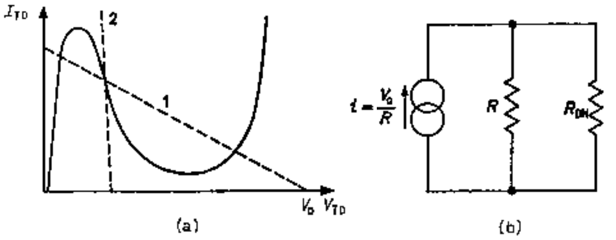


Fig. 4. (a) Load lines for a voltage drive situation: line 1 for bistable operation, line 2 for theoretically full stability
(b) D.C. equivalent circuit for a tunnel diode with negative resistance R_{DN} and load resistance R (line 2)

The load line is displaced to the right as V_o increases and its intercept with the diode characteristic, defining diode current, ascends towards the peak current point. The relationship between circuit conditions and the load line-diode characteristic diagram of Fig. 5 is important. Since V_o is given and the voltage drop across R is fixed by Ohm's law and described by the load line, the diode voltage must lie on the load line; the total circuit current is also at all times defined by the load line, but the instantaneous division of total current between diode and shunt capacitor is determined by the diode characteristic. For a voltage drive V_o, rising relatively slowly just beyond the level required for switching, the successive load lines and operating points are shown in Fig. 5. In practice, Fig. 6(a) and 6(b), a standing current I_s is arranged to flow through the diode so that a small external drive only is required for triggering. The input pulse is then connected to the diode through the coupling resistance and a small capacitance. (One way of preventing the trailing edge of the trigger pulse from resetting the diode is to use a coupling diode.) The magnitude of driving pulse required depends upon the coupling circuit resistance, $R = R_A$ or R_B in Fig. 6(b).

If the diode negative resistance is $-R_D$ and $R_A < |-R_D|$ the pulse height required to shift the operating point from

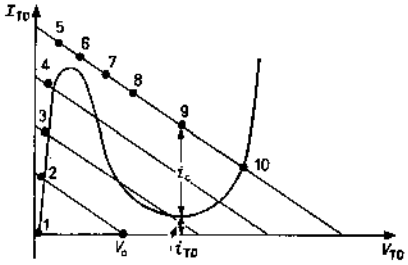


Fig. 5. Successive load lines for increase of voltage drive V_o, series resistance R as in Fig. 2 being small; the simplified switching trajectory is shown by points 1 to 10
Note current division at intermediate point 9

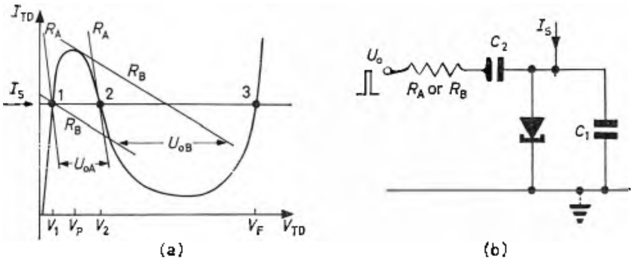


Fig. 6. (a) load line construction for two low resistance drive conditions showing minimum voltage drive U_{oA} or U_{oB} required to initiate full forward switching. *I_S is the standing current*
(b) The relevant circuit specifying the drive coupling arrangement

1 beyond 2 is U_{oA}; if then the input pulse is removed, switching will occur as previously explained. Hence $U_{oA} = V_2 - V_1$. If the coupling resistance $R = R_B$ say, and $R_B > |-R_D|$ the input pulse voltage must exceed $U_{oB} = V_p - V_1 + R_B(I_p - I_s)$ to initiate switching. In each case the second stable point is at 3.

The Sequential Switching of Tunnel Diodes in Series

Sequential switching of two or more (up to certainly more than ten) diodes by drive pulses can be utilized for a number of purposes, including scaling and the generation of a multi-level reference voltage. The main interest here is in the latter usage.

The salient requirement for this purpose is to ensure that only one diode switches at each input pulse. This may be arranged by various methods of varying complexity but a new scheme will be described which is potentially very satisfactory. The chain of tunnel diodes is as usual selected to have different peak currents and a relatively low impedance drive is used. An input pulse causes the diode of lowest peak current (still unswitched) to switch first; the consequent rapid change of voltage across the chain can itself be used to reduce total current during the switch-

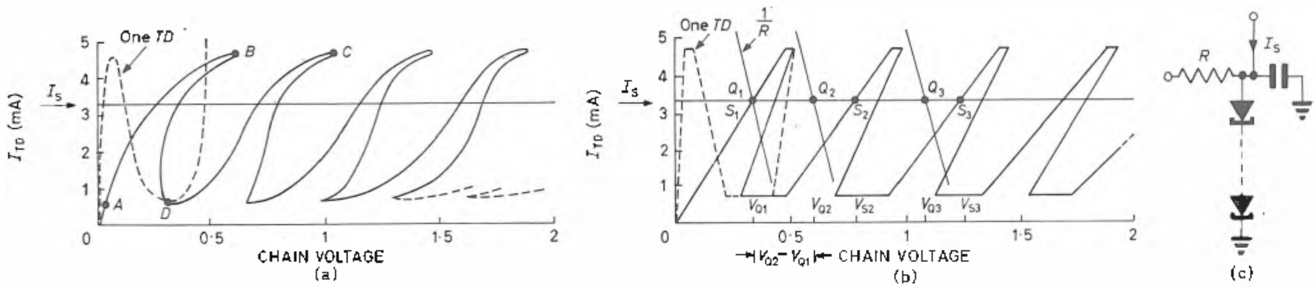


Fig. 7. (a) The complete static characteristic for a tunnel diode chain as in Fig. 7(c), one diode being assumed to switch at a time

(b) Linear approximation assuming that only one diode at a time approaches the peak-current plateau prior to switching S_1, S_2, S_3 *represent operating points at fixed standing current for sequential switching*

(c) The diode chain arrangement I_S *is obtained from a constant current source*

ing action, thus moving each other diode away from its peak point during this critical phase. This action occurs automatically within the chain without the use of auxiliary circuits. In order to ensure that the switching of the single diode (switching time $<$ 10nsec) causes this decrease in total diode chain current, it is essential to limit the rate of rise of the drive pulse to below that of the diode switching action and to use a low impedance drive. The action described does not necessarily prevent more than one diode from switching; however, a much larger pulse would be required to switch two diodes with a single pulse. In practice, the separation required between values of peak current for any two diodes in the chain can be made quite small and an entirely reasonable tolerance on driving pulse amplitude can be permitted. Adding shunt capacitance across a tunnel diode acts in effect to increase the peak current for triggering purposes.

Provided that only one diode is thus triggered at a time, driving pulse amplitude tolerance can be calculated from the static characteristics of a chain. The characteristic is shown in Fig. 7(a) for one situation and for another in linearized form in Fig. 7(b).

The first of these static characteristics for an N-diode chain could be hypothetically tracked in the form shown using a current source provided that while all diodes tend to approach their peak points together, one—having a slightly lower peak current than all others—switches alone to its high voltage state. The trajectory of the operating point moves through A to B, switches to C and if current is reduced, moves to D.

If the slope of the low-voltage branch of each diode corresponds approximately to a resistance R_{D1}, V_p is the peak voltage, V_F the forward-peak voltage and V_v the valley voltage as previously defined, the approximate voltages are as follows: at A, NI_VR_{D1}; at B, NV_P; at C, $(N-1)V_p + V_F$; and at D, $(N-1)\ I_VR_{D1} + V_v$. Similar arguments apply to the remainder of the hypothetical trajectory. It will be noted that as more diodes are switched and more peaks have been traversed in the characteristic the slope of the approach to successive peaks alters correspondingly to the increased variational resistance of those diodes in the high voltage state.

A slightly more realistic assumption is that account be taken of the varying peak characteristics. Using the linear approximation for a single diode, this is represented by the hypothesized characteristic as shown in Fig. 7(b), with only one diode at each peak reaching the approximately infinite resistance state, each other diode remaining at a current slightly below its corresponding peak point. This linearized static characteristic has been found to give easily calculated results well approximating all observations; its valid application requires that the appropriate conditions previously mentioned are met, i.e., with respect to selection of peak currents (in these circuits $\Delta I_p \geqslant 40\mu$A) and to rate of rise of each driving pulse.

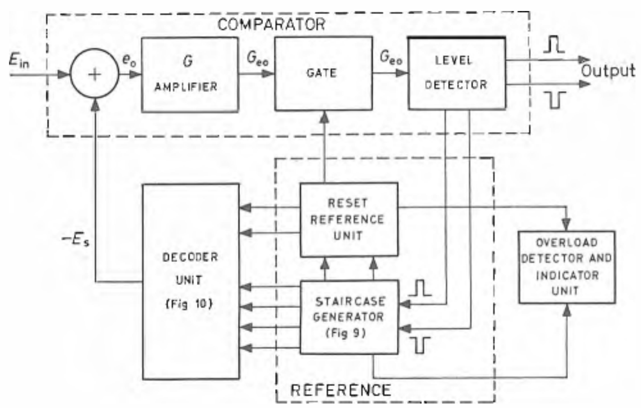
It is convenient to supply a standing current I_s to the chain. The driving pulse is to be coupled to the chain from a low-impedance source, and its rate of rise is to be limited. The circuit is shown in Fig. 7(c) and R is assumed to represent the total source resistance and to correspond to a voltage drive. The situation is specified by the load line on the first segment of Fig. 7(b), intersecting at point Q_1. A pulse of sufficiently small rate of rise and magnitude $(V_{Q2} - V_{Q1})$ moves the load line to the position intersecting Q_2; when the pulse terminates, the chain operating point settles at point S_2. If, however, starting from S_1, the chain were driven by a single pulse of magnitude $(V_{Q3} - V_{Q1})$, two diodes would switch. It will be seen that a stylized characteristic as in Fig. 7(b) allows calculation of a suitable drive pulse amplitude and tolerance, providing R is known.

Resetting the chain can be effected by connecting a transistor (normally cut-off) across the chain and driving it into saturation, thus returning all diodes to their low voltage state. Connecting the base of the resetting transistor across the bottom diode of the chain is usually sufficient to ensure that this is the last to switch. However, commencement of collector current in the resetting transistor initiates diversion of the chain current, tending to inhibit full switching of the transistor; in fact, resetting may not occur. In this event a series (delaying) inductance may be used between the bottom diode and transistor (thus reducing maximum switching rate) or a monostable circuit interposed in the same place.

The Practical Quantizer

Circuits of the type described above, and others, have been used in the prototype quantizer now to be discussed (Fig. 8). The input signal is compared with a reference level, Fig. 9, the difference being applied to

Fig. 8. Arrangement of quantizer: *The input signal* E_{in} *is compared with a reference potential* E_s *which is stepped up or down to follow* E_{in}



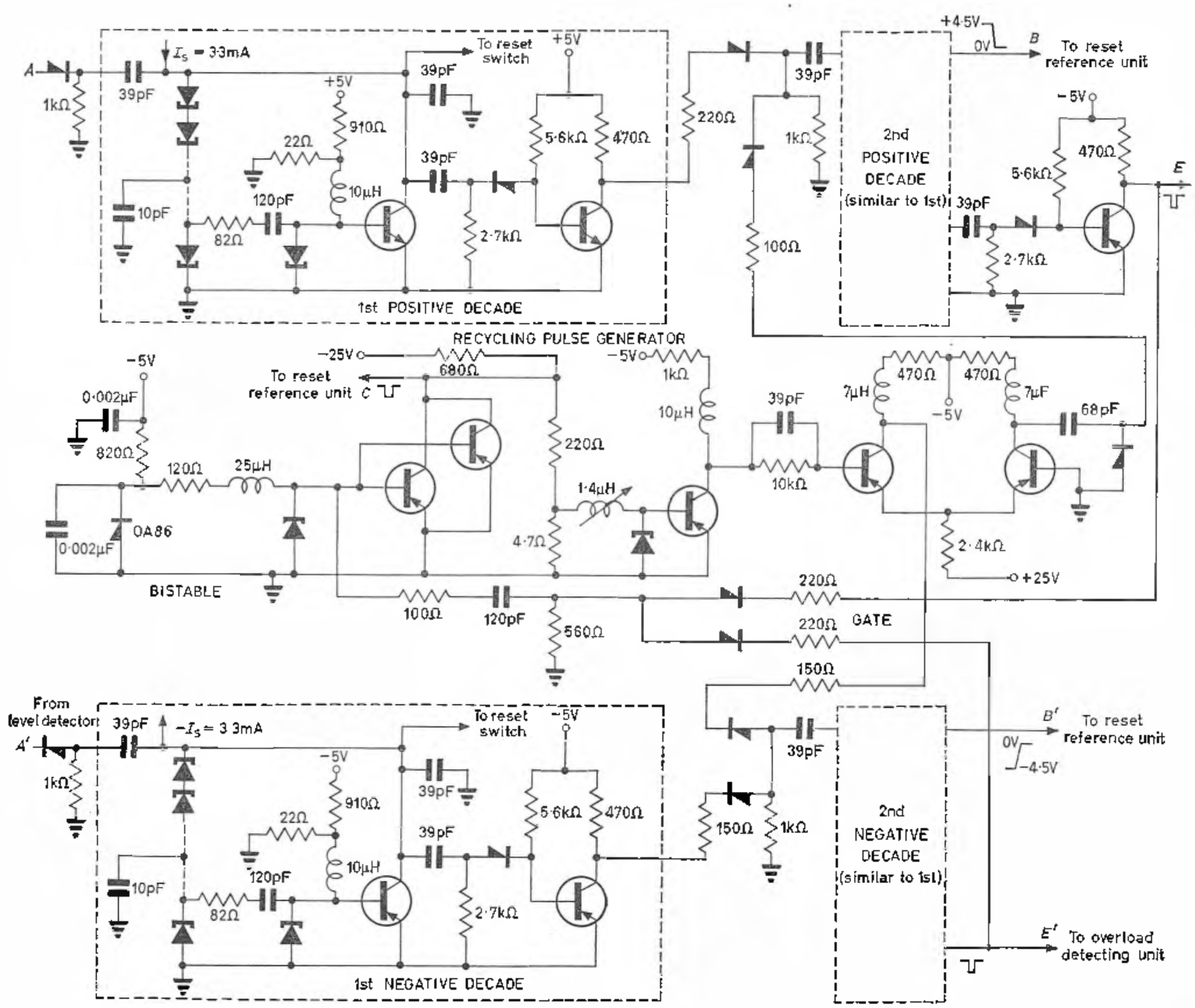


Fig. 9. The staircase generator unit defining ± 100 levels *The four staircase voltages are weighted and summed to generate the reference potential. All tunnel diodes are 1N2941, all other diodes OA47, all transistors, pnp ASZ21 and npn 2N797*

an amplifier supplying two complementary and symmetrical level-detectors (positive and negative) set to produce a continuous train of output pulses (at 1·25Mc/s in the present circuit) if the difference exceeds a quantum interval. Each detector (using a new tunnel diode transistor hybrid circuit, to be described elsewhere) produces pulses of appropriate polarity, either positive or negative, positive pulses being applied to a positive staircase generator, negative pulses being applied to a negative staircase generator. These cause the staircase voltage to step up positively or negatively, thus altering the total reference level and reducing the input to the level detector below threshold before a second pulse is produced at the level detector output (i.e. within 0·8μsec at present).

The two (positive and negative) two-stage decade staircase waveforms are summed algebraically by a decoding resistive network to produce the net sum defining the total reference level potential, Fig. 10. The scheme of handling positive and negative levels independently is very convenient, but it produces problems related to resetting. An example follows. Depending upon its prior history, an input signal which has reached a level of +60 units at the instant of interest, could feasibly be presented by any possible combination of positive and negative staircase levels summing algebraically to +60 units, viz., +60 and −0; +80 and −20; +100 and −40, etc. Overloading of either positive- or negative-polarity staircases can and does occur frequently during processing of a signal, at levels in this case of +100 and −40 units. The positive staircase must be reset and a net +60 units retained; the following method has been used.

The first overloading pulse causes the positive staircase to reset to zero, and initiates a 10Mc/s pulse generator which simultaneously supplies pulses of required polarity to positive and negative staircases, stepping up both until the latter overloads and resets, thus switching off the 10Mc/s pulse generator. The positive staircase then contains +60 units while the negative is at zero. Signal processing can then continue. In fact, 10Mc/s pulses have been applied only to the most significant (second) decades of each polarity. At worst, then, nine pulses will suffice to effect resetting and 1μsec for this cycling operation will be required.

A difficulty of this scheme is illustrated in this example, in that when the positive staircase resets, the total reference level instantaneously applied to the difference amplifier changes greatly and the amplifier becomes grossly out of balance. There is a risk of amplifier overload with consequent blocking, and of triggering the level detector due to this occurrence. The latter problem is resolved by using a gate on the level-detector, operated during re-cycling. The former needs careful amplifier design to allow smooth overloading and fast recovery. It is also convenient to switch into circuit a positive or negative fixed-level voltage equivalent to 100 quantum levels, to replace the resetting staircase level temporarily until cycling is complete. This is accomplished by the reset reference unit, brought into action only during the cycling process.

Some details of the complete staircase generator are shown in Fig. 9. Positive pulses from the level detector are applied to the first positive decade at A. The last pulse of each 10 from the level detector triggers the monostable multivibrator, saturating the reset transistor across the chain; this initiates reset of the chain voltages and transmits a further pulse through a common emitter stage to the second positive decade. A similar sequence applies to negative pulses at A' from the level detector.

The cycling action is as follows. When either of the two higher order chains produces an overload output pulse (at E or E'), an OR-gate passes pulses to a tunnel diode bistable, the switching of which cuts off collector current in two parallel transistors, normally diverting supply current from a tunnel diode relaxation oscillator. The diode oscillator then produces 10Mc/s pulses which are applied to a long-tailed pair; the positive and negative output pulses are taken respectively to positive and negative higher order decades, thus stepping both together until either overloads, resets and thereby passes a terminating pulse to the bistable controlling the pulse generator.

Staircase Generator Precision

In an incremental system of this type the stepping reference is the only unit required to maintain accuracy over a large dynamic range; in fact, this unit determines almost entirely the accuracy of the convertor. Errors introduced by the comparator amplifier and by the level detector unit represent normally a small fraction of one quantization interval (the dynamic range of these units covering only one such quantum interval).

Experience with tunnel diode chains indicates that the necessary stable precision and linearity can readily be provided. Using 1N2941 diodes, it has been found that some 95 per cent of tunnel diodes in chains generate voltage steps certainly within 6 per cent of the average. Discarding a few diodes with extreme step values improves this figure appreciably and simple selection of diodes in sets of 10 to make up chains with their own average step values produces step deviations of better than 1 per cent. Selection of diodes to produce steps better than 2 per cent is adequate since trimming of any particular diode step with a shunt resistance can reduce the step value by up to 2 per cent. Thus an accuracy to ± 1 per cent between steps (i.e. of typically about 470mV with the 1N2941) can easily be obtained for up to 10 diodes in a chain biased with a constant current. This figure can certainly be improved by adequate trimming to better than $\pm 0{\cdot}1$ per cent of each step; in this event, however, some thermal stabilization of the tunnel diodes is required.

The chain voltage linearity is determined by the tolerance on individual steps and the number of times the step tolerance can be repeated with the same polarity in successive steps. For the prototype unit, defining 100 levels on either side of zero with two decade chains twice, it was assumed that the non-linearity of chain voltage could be as high as three times step tolerance. It is clear that a reference unit defining ± 1000 levels in three decades can be built. However, if a decade system is not required, similar results can be obtained with four pairs of chains using fewer diodes.

The prototype unit steps at a rate of 1·25Mc/s even though the staircase units can step at above 10Mc/s (as used during re-cycling for example), the present limitation, imposed to accommodate the relatively long recovery time of the comparator amplifier after overload, is of little consequence since a comparator circuit using very high gain and feedback over many stages (Fig. 10) is not essential and an amplifier with much faster recovery can be employed.

The sensitivity of the present unit with a quantization error of $\pm 0{\cdot}5$ per cent corresponds to a minimum quantization interval of 4mV; the lower limit envisaged is

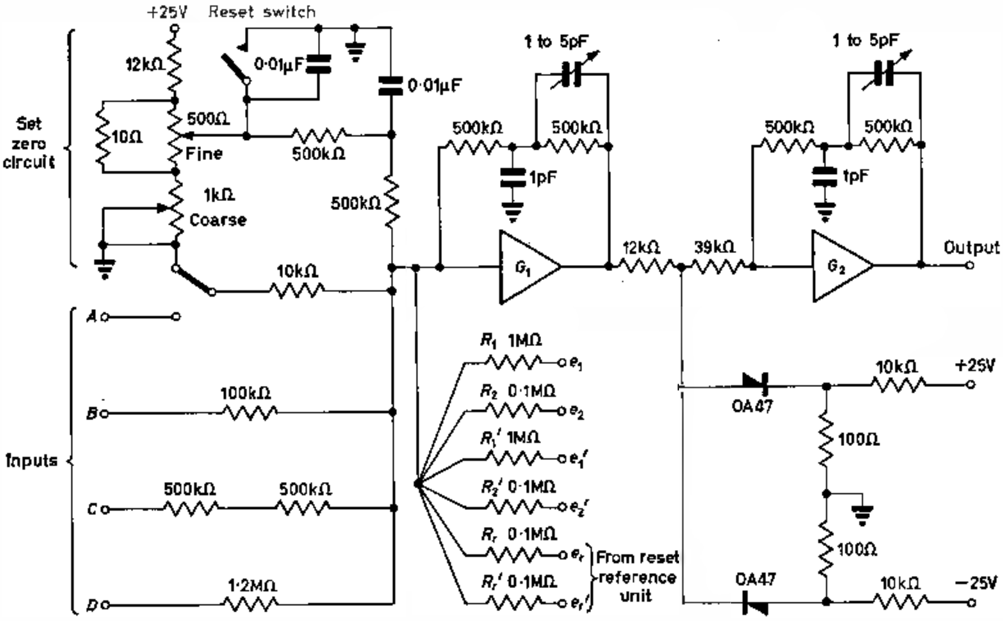


Fig. 10. Summing amplifier unit. *Fixed resistors are metal oxide. Resistors R_1 R_2 R_1' R_2' R_r R_r' can be adjusted effectively to equalize differing average step voltages from the four staircase chains*

below 1mV. The full scale range of 200 quantization intervals can be increased from $\pm 0{\cdot}4$V conveniently up to, say, ± 100V. With a minimum stepping interval of $0{\cdot}8\mu$sec the unit will follow a full-scale signal altering at a maximum rate of 100 steps in $100 \times 0{\cdot}8\mu\text{sec} = 80\mu\text{sec}$; this corresponds to a full scale sinusoid at about 2kc/s and maximum slope of input signal on the most sensitive range of about 5×10^3V/sec. It will be noted that one important feature of this system is the capability for trading resolution with speed; if the dynamic range of the signal is restricted, higher speed is immediately available.

Applications of the Quantizer and its Basic Circuits

A unit of this type can be used as a continuously-following analogue to digital convertor (a.d.c.), with the addition of a simple forward-backward counter. In use as a sampling a.d.c., the scheme becomes simpler since no re-cycling is required and the encoding counter need only count forward. Each of these systems depends upon the fast stepping reference generator; the comparator and display systems can clearly be realized in various ways. The availability of this stepping reference provides some extremely interesting possibilities. Since tunnel diodes in a chain can be made to switch in turn according to peak current, it is feasible to arrange that the sequence of switching takes place according to physical position in the chain. By circumspect choice of monitoring networks, it is therefore possible to obtain a parallel-coded output derived from the chain itself without detracting from the precision of the system.

However, with the aid of a forward-backward counter the system can be used as the basis of an a.d.c., a digital integrator or perhaps a non-linear functional transformer, and an analogue multiplier (for example, by using the coded output representing one signal to control the gain of a virtual-earth scaling amplifier by gating an associated resistive network). It will also be noticed that a peak of the input signal is indicated at the quantizer output by a reversal of output pulse polarity. Hence the quantizer can be used also as the basis of a direct-coupled kick-sorter or a level-preserving pulse stretcher. An application particularly using tunnel diode chains is in use in the Engineering in Medicine Laboratory for measurement of signal statistics—especially the simple and joint probability distributions of intervals between spikes occuring in neural electro-physiological signals. Here the first spike opens a gate applying standard pulses to the staircase generator; a second spike initiates a reading of the staircase level (several decades may be summed into a virtual earth) as an output pulse of height proportional to interval duration, prior to resetting the staircase chains to zero. This interval-to-height conversion circuit is very simple and a dynamic range of 100:1 can conveniently be used; extension of the scheme for conditional probability measurements is quite simple.

Using the quantizer in its present form, as a sampling a.d.c., the prototype unit effects 6 000 conversions per second with a resolution of $\pm 0{\cdot}5$ per cent. Returning all chains to zero after each conversion, thus obviating the need for cycling, allows up to 12 000 conversions per second. With a similar scheme, still using a decade system stepping up at the currently acceptable rate of 10Mc/s, and with a fast recovery amplifier, a sampling rate up to 100 000 conversions per second would be available. However, if a system of successive approximations is used, a maximum conversion time of 2μsec for $\pm 0{\cdot}5$ per cent quantization error can be expected. A scheme of this kind has been devised using the same 10Mc/s stepping reference and level detectors but with much simpler ancillary circuits; development is proceeding.

Acknowledgment

This article contains some material taken from a University of London Ph.D. thesis by R. V. L'Archeveque.

Electronic Speed Indicating Equipment for Railways

For the past thirty years, the method of indicating the speed of a locomotive has been by means of an axle-driven generator and a voltage sensitive instrument calibrated in miles per hour, located in the driver's cab. Mileage was recorded by a mechanical mileage counter mounted on another axle. This arrangement has given reliable service over very large distances but its chief disadvantage is poor low-speed indication and pointer oscillation.

Speedometer equipment is now required to control automatic field weakening of the main traction motor, and in the near future to provide completely automatic low speed control of locomotives used with liner trains and 'merry-go-round' trains at power stations.

These requirements have led to the development by Associated Electrical Industries Ltd of the type S electronic tachometer equipment, which is made up of a transmitter, control box and indicating instruments.

The transmitter is mounted on the axle and is fitted in place of the normal axle box cover. It has no bearings of its own and incorporates an electromagnetic probe in which pulses are generated by the rotation of a toothed wheel attached to the locomotive axle.

In the control box are plug-in electronic units utilizing static transistorized circuits throughout. The box is energized from the 110V d.c. supply available on the locomotive, but has its own power unit which provides adequately smooth supply for the electronic components, at the necessary voltages.

The speed indicating and mileage recording instruments are mounted in the driver's cab.

The complete system is sensitive to the frequency of the pulses generated.

Pulses from the transmitter are fed into the control box and after amplification and change of wave shape, these are utilized to carry out the required functions.

The indicating instrument is a milliammeter calibrated in miles per hour. The pulses reaching the instrument are integrated by its moving-coil movement, and the figure indicated will equal the mean current. Since the pulses are all of the same magnitude, the mean current is directly proportional to the frequency of the pulses and a linear scale is obtained on the instrument.

A calibrated shunt across the instrument compensates for wheel wear. Adjustment is provided for wheel diameters between 36in and 48in. When two instruments are required they are connected in series, thus avoiding any problems of unequal current sharing.

To record mileage a capacitor discharge circuit is used to energize an electromagnetic counter.

Speed sensitive relays are used for automatic field weakening; premature operation of the relays due to wheelspin is prevented by built-in delay circuits.

The new equipment gives clearer low speed indication than any previous form of speed indicating device. Accurate readings down to 0·2mile/h are obtained because of the complete absence of pointer flutter.

It is of interest that a similar type of equipment to the one described above has been designed for recording the number of hours an engine has run. In this equipment any form of engine-initiated signal can be used to actuate a small control box which in turn operates an electromagnetic counter.

The speedometer and field diversion system are at present being installed on some type 2 diesel-electric locomotives now being built for British Railways. Type 1 locomotives now on order for B.R. will be fitted with this speedometer system.

A Square Law Current Control System

By C. I. Sach*

This article describes a unit which controls a current I_2 in a load so that it varies according to the current I_1 flowing in an identical load in such a way that the sum of the squares of the two currents remains substantially constant. When these loads are windings on a common core the net ampere turns are controlled while maintaining the dissipation constant.

(Voir page 634 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 641)

THE requirement for a current control unit with constant dissipation arose during a project to obtain well controlled small movements of cutting tools by making use of the magnetostrictive property of nickel alloys. This

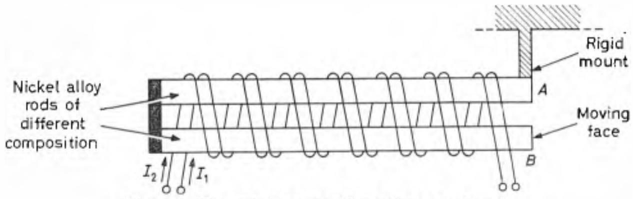


Fig. 1. The winding configuration employed

work was initiated by S. K. Dean and is being reported elsewhere[1]. The differential magnetostrictive movement in two rods is used, but as the resultant movement is quite small (0·0003in max.), it is important that thermal expansion effects be kept to a minimum. This cannot be achieved using a single coil around the compound rod since the dissipation, and hence temperature, will change with the magnetizing current supplied to the coil. The configuration in Fig. 1 was therefore used.

In this system two identical coils of 4·5Ω resistance are wound over the rods. The current I_1 in the first coil is manually set between -2 and $+2$A, while the current I_2 in the second coil is constrained by a control unit to be such that $I_1^2+I_2^2 = 8$. For this, I_2 must be variable between $+2$ when $I_1 = \pm 2$A and 2·82 when $I_1 = 0$. With one direction of current flow of I_1 the magnetizing force produced adds to that produced by I_2. The maximum resultant magnetomotive force is $4n$ ampere turns where n is the number of turns in each coil. When I_1 is flowing in the other direction its effect subtracts from that of I_2 and in the limit produces a resultant of zero when I_1 and I_2 are each 2A.

System Description

An arrangement for the current controller is shown in Fig. 2. Here E represent a variable d.c. supply which produces the primary current I_1. The value of this current is monitored by inserting a small stable resistance R_s in series with the coil, thus producing a voltage E_1 equal to R_sI_1. A similar system is used with the second coil which

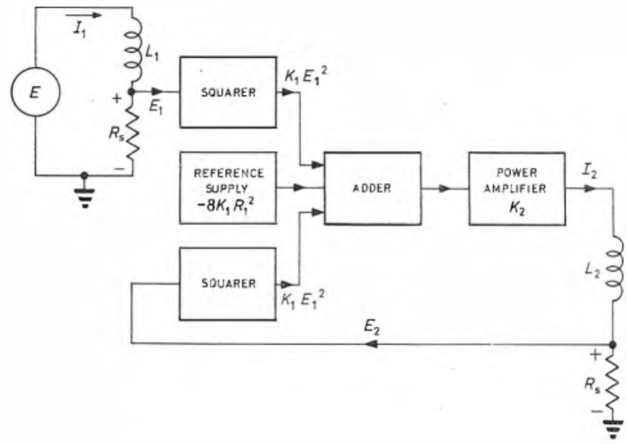
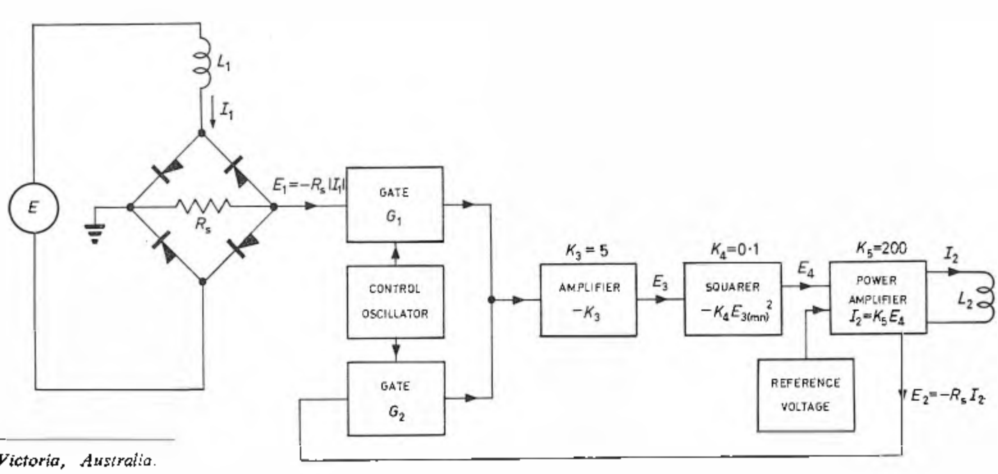


Fig. 2. Arrangement of current controller

derives its current I_2 from a power amplifier, producing a potential E_2 equal to R_sI_2. The two voltages E_1 and E_2 are then fed to separate squaring circuits which give outputs of $K_1R_s^2I_1^2$ and $K_1R_s^2I_2^2$ respectively. These voltages are fed to a three input summing unit together with a fixed reference potential of $-8K_1R_s^2$. The output of the adder then drives a high gain power amplifier which provides the secondary current I_2.

Fig. 3. Arrangement of practical system



* *Defence Standards Laboratories, Victoria, Australia.*

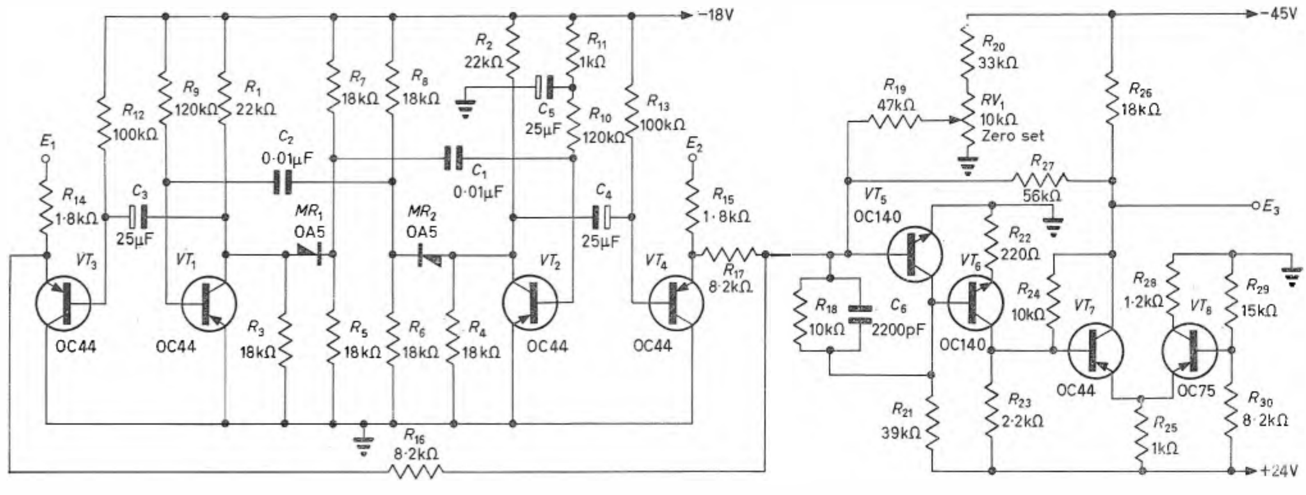


Fig. 4. Gates and ×5 amplifier

If the amplifier gain is high its input will be constrained by the feedback loop to be very close to zero.

This means that:

$$K_1R_s^2I_1^2 + K_1R_s^2I_2^2 - 8K_1R_s^2 = 0 \quad \ldots\ldots\ldots \quad (1)$$

This reduces to the required system control equation:

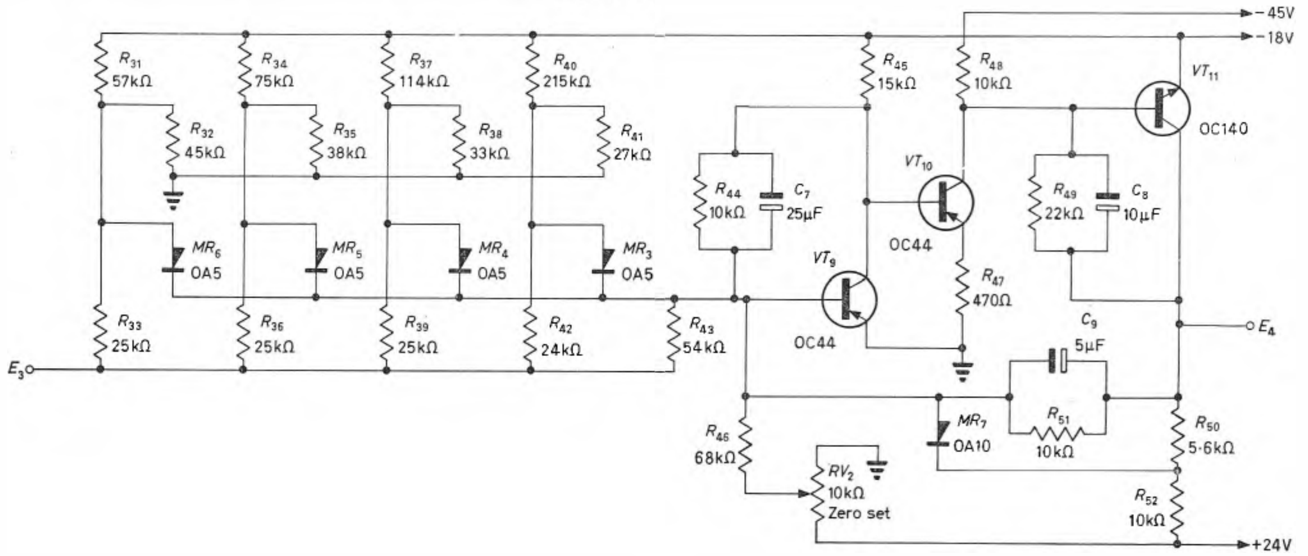
$$I_1^2 + I_2^2 = 8 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

Certain modifications of this basic system are necessary for a practical system. Since the squarer uses a diode shaping network situated at the amplifier input it is desirable that the voltages E_1 and E_2 should be in the range 0 to 10V so that the non-ideal forward characteristics of the diodes are not greatly significant in determining the shape of the curve. The large values of input voltages require R_s values of the order of 5Ω and these waste a large amount of power, e.g. 20W. Thus it is desirable to have smaller series resistances (0·5Ω is used) with some amplification prior to the squaring circuits. Thus the diagram of Fig. 2 should now include two extra gain stabilized amplifiers.

To overcome the problems of having two identical channels, each containing an amplifier and a squarer, a sharing system is used which allows each voltage, E_1 and E_2, to be fed alternately to the same amplifier and squarer. The schematic of this new system is shown in Fig. 3. Here a 500c/s symmetrical multivibrator operates two gates G_1 and G_2 alternately. The input to gate G_1 is taken from a series resistance $R_s = 0{\cdot}5\Omega$ which is arranged in a bridge circuit so that it always gives output equal to $-R_s|I_1|$. It is not necessary to observe the sign of I_1, since it is squared subsequently so the sign is lost. Also this method means that the squaring circuit is required to handle signals of one polarity only and this simplifies its design. The input to gate G_2 is not obtained from R_s in series with the second coil L_2 but is taken from a point inside the amplifier which produces a voltage closely equal to $-R_sI_2$. For the currents required, E_1 varies from 0 to −1 while E_2 varies from −1·4 to −1V. The amplifier following the gates has a voltage gain of $-K_3$ ($K_3 = 5$) so that its output is a square wave E_3 whose two levels are 0 to +5V when gate G_1 is open and +5 to +7·0V when G_2 is open. The squarer consists of a diode shaping circuit and a gain stabilized amplifier which filters out the 500c/s ripple and produces a smooth output signal of:

$$E_4 = -K_4\langle E_3^2\rangle \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

Fig. 5. Squaring unit



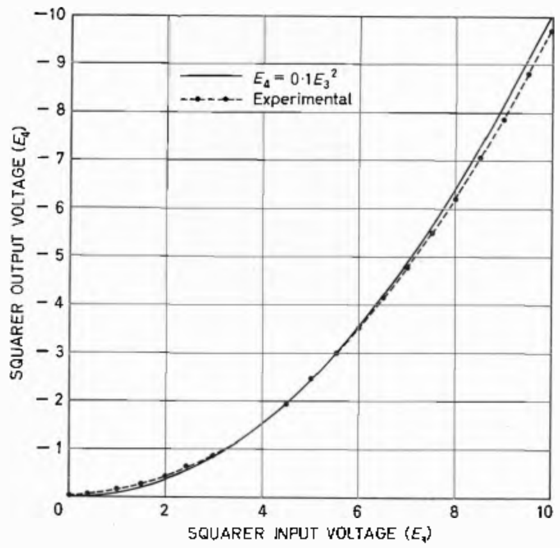


Fig. 6. Squarer characteristic

where $\langle E_3^2 \rangle$ is the mean value of the E_3 waveform after it has been squared. K_4 in this unit is 0·1. Substituting block diagram symbols, it may be shown that:

$$E_4 = \frac{-K_4 R_3^2 K_3^2}{2} (I_1^2 + I_2^2) \quad \ldots\ldots\ldots\ (4)$$

This reduces, on substituting figures, to:

$$E_4 = -0{\cdot}312\,(I_1^2 + I_2^2) \quad \ldots\ldots\ldots\ldots\ (5)$$

This signal is fed to the input of a power amplifier where it is added to an effective reference potential of +2·5V. The amplifier has a gain of K_5 = 200A/V and its output drives the second coil L_2. Due to this high gain and the negative feedback loop, the resultant of the two inputs must be closely equal to zero.

used in a multivibrator circuit which uses diodes MR_1 and MR_2 to improve the negative going edge of the waveforms occurring at the collectors. The base supply resistance of each transistor is 120kΩ, but it was found necessary to decouple that associated with VT_2 by means of R_{11} and C_5. Without this, the circuit had a tendency to go into a stable state with both transistors saturated and therefore not having sufficient loop gain to start oscillation. The decoupling circuit ensures that VT_2 does not immediately saturate when the supplies are turned on. The outputs of the multivibrator are used to switch transistors VT_3 and VT_4 which act as gates in the following manner. When VT_3 is saturated and VT_4 is off, a current flows from the virtual earth point at the base of VT_5 through R_{15} and R_{17} to the voltage point E_3. This current is amplified by transistors VT_5 to VT_7 which forms an amplifier with an accurate and stable transfer impedance of 5V/mA. Due to VT_3 being saturated it effectively clamps the input to the amplifier via R_{16} to 0 volts independent of the actual signal E_1. On the alternate half cycle, E_1 produces an input current while E_2 is isolated. The amplifier forward transfer impedance is closely equal to 450V/mA as defined by the three stages whose separate gains are controlled by local feedback. This figure for forward gain in conjunction with the feedback resistor R_{27} of 56kΩ gives a gain of 5V/mA. Any trimming of gain can be carried out by adjusting R_{27}. The output is adjusted by RV_1 to be 0 when the input is zero and the output is capable of swinging between 0 and 10V while supplying a current of up to 5mA. A single lag element C_6 is used to prevent instability in the amplifier.

The circuit diagram of the squaring unit is shown in Fig. 5. As explained before, this has a transfer function $E_4 = -0{\cdot}1\,\langle E_3^2 \rangle$ and can handle input signals in the range 0 to +10V. Straight line approximations are used for the quadratic function and these are obtained with

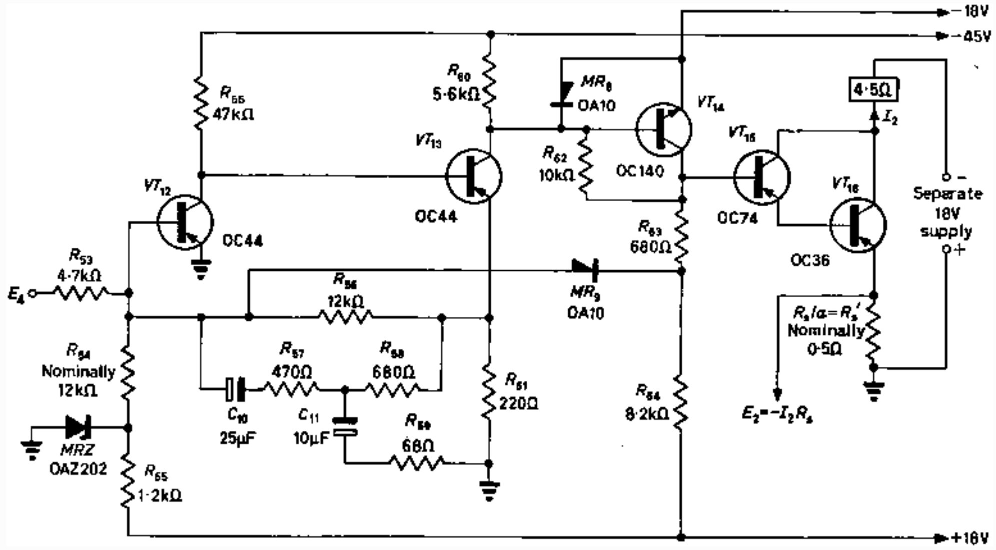


Fig. 7. Power amplifier

Thus $2{\cdot}5 - 0{\cdot}312\ (I_1^2+I_2^2) = 0$, which reduces to the desired system equation.

Description of Units

The circuit diagram of the gating unit and its following amplifier is given in Fig. 4. Transistors VT_1 and VT_2 are

a biased diode network at the input of an amplifier which produces a virtual earth termination for the shaping network. For inputs in the range 0 to +2V, the amplifier input current is obtained solely through R_{43}. At +2V, diode MR_3 becomes conducting and forms an additional current source for the amplifier. The other diodes, MR_4 to MR_6 conduct

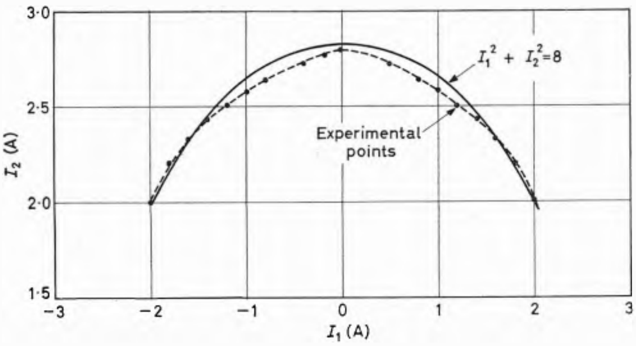


Fig. 8. Plot of I_2 as a function of I_1

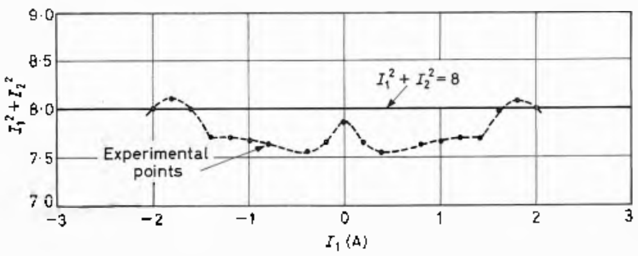


Fig. 9. Plot of $I_1^2 + I_2^2$ as a function of I_1

at successive 2V increments of the input, thus producing the desired approximation. The amplifier, as for the previous one, uses both local and overall feedback to achieve a stable gain figure of 10V/mA. The shunt feedback stages are used to introduce two poles in the transfer function while the zero introduced by C_9 in parallel with the feedback resistor R_{51} enables a frequency response having a 40dB/decade roll off at 5c/s to be achieved. This is adequate for removing most of the 500c/s components from the output of E_4. The zero control RV_3 is used to set the output to 0 volts when the input is zero while diode MR_7 is used to eliminate a spurious d.c. bias condition where VT_9 is saturated.

The characteristic of the squarer is shown in Fig. 6 where it shows a good approximation to the desired one. The scale on the E_3 axis of the curve can be altered by adjusting the −18V supply line since this is the one which sets the bias potentials for the diodes.

The power amplifier and reference potential are shown in Fig. 7. Transistors VT_{12} and VT_{13} form a current amplifier of gain 50 which is defined by the ratio R_{56}/R_{61}. The components in parallel with R_{56} are used for shaping the frequency response and will be dealt with in the section on stability. The negative going signal E_4 is added to the positive reference potential produced by the Zener diode MRZ_1 using the adding resistors R_{53} and R_{54} connected to the virtual earth input of the current amplifier pair. The value of R_{54} is adjusted to set the quiescent conditions so that when E_4 equals −2·5V there is a small input current sufficient to produce an output current I_2 of about 2·5A. Following the current amplifier is a shunt feedback stage VT_{14} with a gain of 10V/mA. This drives the compound series feedback stage VT_{15} and VT_{16}. The output from the collector of VT_{16} drives the second load. The current through the load is measured by the resistance R_s in the emitter circuit. This is an approximation since the load current I_2 is only α times that through the resistance, where α is the grounded base current gain of VT_{16}. Thus the voltage E_2 produced is really equal to $-I_2 R_s/\alpha$ introducing an error of perhaps 3 per cent. In practice the value of the emitter resistance is reduced to a value R_s' such that R_s'/α equals R_s to overcome this error.

Diode MR_9 is used to limit the negative excursion at the base of VT_{15} to −2·5V. This limits I_2 to a maximum value of about 3A which prevents VT_{15} and VT_{16} from saturating. If this does happen, VT_{14} still tries to supply the much larger base current required and can destroy itself. Diode MR_8 is employed to eliminate ambiguity in setting the output current. Without it, it is possible to obtain the correct value of output current under a condition where VT_{14} is cut off with its base considerably more negative than its emitter. However, the inclusion of the diode prevents this. The overall gain of the power amplifier is 200A/V which means a variation of E_4 of only 5mV is needed to swing I_2 over its full range.

Stability

Owing to the use of gates in the system, oscillations can only build up for those periods when G_2 is operating. Considering this case, the open loop frequency response must be shaped to prevent oscillation. The d.c. loop gain is variable due to the presence of the squarer. This results in an incremental loop gain of 10^4 when $E_3 = 10$V, dropping to 200 when E_3 is less than 2V. The filter in the squarer amplifier introduces a 40dB/decade cut at 5c/s but this on its own is not enough to prevent oscillation. Hence an additional network is used in the power amplifier. Initially this operates as a lag-lead network by virtue of C_{10} shunting R_{56}, thus giving a 20dB/decade cut off at 0·5c/s down to 5c/s where it levels off again due to R_{57} and R_{58} in series with C_{10}. The effect of this is to drop the loop gain by 20dB before the filter cut off becomes important and this is sufficient to render the loop stable. However, there was still some ringing present so a lead-lag network was introduced to increase the phase margin near the point where the loop gain was about one. The lead response is introduced by capacitance C_{11} in conjunction with R_{57} and R_{58} which reduces the amount of feedback and hence increases the forward gain of the amplifier. The presence of R_{59} in series with C_{11} terminates the lead characteristic after a decade. With this arrangement the step response of the system has an overshoot of 15 per cent and reaches 90 per cent of its final value within 30msec.

Performance

The plot of I_2 versus I_1 is shown in Fig. 8 where reasonable agreement with the theoretical curve is obtained. The variation of $I_1^2 + I_2^2$ as a function of I_1, is given in Fig. 9 where it can be seen that the agreement is good. The maximum excursion from the design value of 8 is 5 per cent.

Conclusions

A system has been described which is capable of holding the sum of the squares of two currents constant within 5 per cent while the effective current available for energizing coils to magnetize magnetostrictive rods varies from 0 to 4A.

Acknowledgment

This article is published by permission of the Chief Scientist, Australian Defence Scientific Service, Department of Supply, Melbourne, Victoria, Australia.

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Design Data for Voltage Dependent Resistor Function Generators

By G. Brown*, A.M.I.E.E., and R. A. B. Bond†, M.Sc.

An operational amplifier circuit, comprising a padded voltage dependent resistor input network and a linear resistor feedback, is analysed as a power law function generator. Tabular data is given with which optimum padding resistor values can rapidly be computed for power law exponents in the range 1·5 to 4. The optimizing criterion is minimum square error. The table also gives the resulting function generator errors. Error and temperature compensation is briefly discussed.

(Voir page 634 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 641)

A SUITABLE voltage dependent resistor (also known as a silicon carbide varistor) incorporated in the input or feedback circuit of an operational amplifier results in a cheap and reliable non-linear function generator. This technique has been quite widely used in analogue computation and has also found application in servo systems and electronic circuit design[1,2,3,4]. The technique is particularly valuable in analogue computer problems where large numbers of function generators are required, e.g. fluid flow network investigations.

Since manufacturing tolerances on individual voltage dependent resistors (v.d.r.) are comparatively wide, the design of circuits using these elements requires some experimental procedure[1,2]. In analogue computation, it is most convenient to use a special design programme on the computer itself, but this can occupy much valuable computer time, particularly if large numbers of v.d.r. circuits are to be designed. Also, the method can leave doubts as to whether the optimum accuracy has been achieved. Thus, it is desirable to establish a rapid numerical design method and to analyse, quantitatively, the accuracy of this technique of function generation. In this article, results are given of a digital computer based investigation into a simple, but versatile, form of v.d.r. function generator. This investigation leads to the required rapid design method together with data giving the magnitude of errors in the function generators.

It is to be noted that the non-linearities involved in the design of v.d.r. function generating circuits preclude any form of standard mathematical analysis and hence recourse to graphical techniques[4] and for this article digital computation.

Voltage Dependent Resistors

The characteristics of voltage dependent resistors have been described elsewhere[1,2] and also in manufacturers' literature, for example Ref. 5. From this it is seen that v.d.r.'s are:

(a) Virtually insensitive to polarity.

(b) Stable with respect to current loading provided they are operated well within their rated dissipation.

(c) Described approximately by the relationship

$$E = KI^n \quad \ldots\ldots\ldots\ldots\ldots\ldots (1)$$

where E is the potential driving current I through the v.d.r. K and n are parameters characterizing each resistor and are approximately constant for a wide range of E (Brown and Walker[2]).

(d) Temperature sensitive in parameter K.

(e) Not significantly temperature sensitive in parameter n.

The available ranges of parameters K and n given by one manufacturer[5] are:

$$14 \leqslant K \leqslant 1\,100$$
$$0{\cdot}14 \leqslant n \leqslant 0{\cdot}3$$

Admittedly not all v.d.r.'s grouped in these ranges are suitable for function generators, but these values were used as approximate limiting values in this investigation.

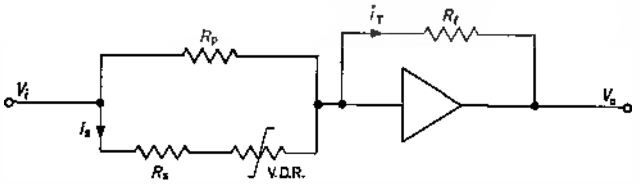


Fig. 1. Arrangement of v.d.r. function generator

V.D.R. Function Generation

The circuit considered here is shown in Fig. 1.

The investigation was restricted to power law functions of the form:

$$V_o = -(\text{sgn } V_i)\,\frac{|\,V_i^r\,|}{100^{r-1}}\,;\; r > 1 \quad \ldots\ldots\ldots (2)$$

where (sgn V_i) means 'sign of V_i' and V_i and V_o are input and output voltages respectively. The working voltage range was taken to be ± 100V. Such power law functions have wide physical interest and in function generator form can be used to construct analogues of other functions which can be expressed as power series.

In Fig. 1 the input and feedback networks can be interchanged resulting in a function generator for which

$$V_o = -(\text{sgn } V_i)\,\frac{|\,V_i^{1/r}\,|}{100^{(1/r)-1}} \quad \ldots\ldots\ldots\ldots (3)$$

In Fig. 1, R_p, R_s and R_f may be varied and the design problem is to determine values which give the closest approximation to the desired function.

Determination of R_s, R_p and R_f

In the following analysis it is assumed that the relationship given in equation (1) is precise. In fact, at low values of current through the v.d.r. both K and n increase[2]. However, when this occurs in the circuit of Fig. 1 (at low V_i) the current through R_p will be much greater than i_s and therefore the assumption of constant K and n values is considered acceptable. Experimental work subsequently confirmed this.

The behaviour of the circuit of Fig. 1 is conveniently

* *African Explosives & Chemical Industries Ltd., Johannesburg.*
† *Department of Mathematics, University of Natal, Durban.*

described by the equations:

$$V_i = QI_s + I_s^n \quad \ldots\ldots\ldots\ldots\ldots \ (4)$$

and e = error in $V_o = V_o - (V_i^r/100^{r-1})$

$$= (P_2/P_1)V_i + (I_s/P_1) - (V_i^r/100^{r-1}) \quad \ldots\ldots \ (5)$$

where

$$P_1 = C/R_f,\ P_2 = C/R_p \text{ and } Q = R_s/C \quad \ldots\ldots \ (6)$$

in which $C = K^{1/n}$ and $I_s/C = i_s$, the actual current through the v.d.r.

In minimizing e, the three parameters P_1, P_2 and Q may be varied, i.e. there are three degrees of freedom. To simplify the analysis an extra constraint was imposed, namely:

$$e = 0 \text{ at } V_i = V_o = \pm 100\text{V} \quad \ldots\ldots\ldots \ (7)$$

This constraint does not degrade the attainable accuracy excessively. Without this constraint it was found that resistor settings for optimum accuracy became rather critical. Therefore from this point of view the constraint is practical. Also an error free point at $V_i = \pm 100$V is a useful feature for checking the function generator.

At this juncture it is of value to consider a graphical representation of equations (4) and (5) to illustrate the form of e and the effect of altering P_1, P_2 and Q. Firstly, it should be remembered that in the input network is required, *ideally*, to have a characteristic such that:

$$-V_o = R_f\, i_T' = V_i^r/100^{r-1}$$

The ideal input network characteristic is therefore:

$$i_T' = (V_i^r/100^{r-1}) \,.\, 1/R_f$$

or

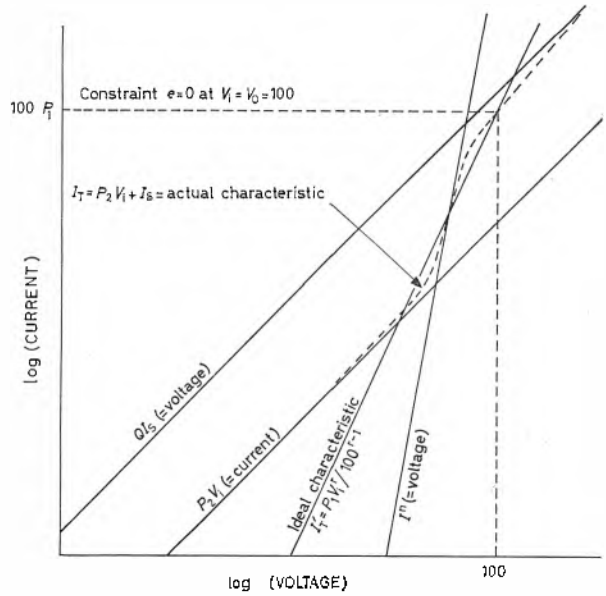
$$I_T' = P_1 V_i^r/100^{r-1} \quad \ldots\ldots\ldots\ldots\ldots \ (8)$$

where $I_T'/C = i_T'$ = ideal current through R_f.

Fig. 2 gives the appropriate graphs which are conveniently plotted on logarithmic co-ordinates. This graph is not drawn to scale, so as to exaggerate the effects described below. In Fig. 2, the *actual* total current (I_T) through the input network is formed by addition of the voltages QI_s and I_s^n at constant current followed by the addition of this characteristic to P_2V_i at constant voltage. The resultant curve is then $P_2V_i + I_s = I_T$. Multiplication by $1/P_1$ then sets V_o at the desired level.

From Fig. 2 the error in V_o is seen to be $(I_T - I_T')/P_1$.

Fig. 2. Graphical representation of v.d.r. function generator behaviour



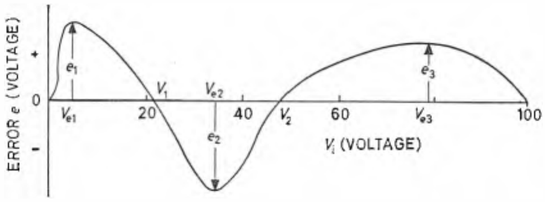


Fig. 3. Typical error curve

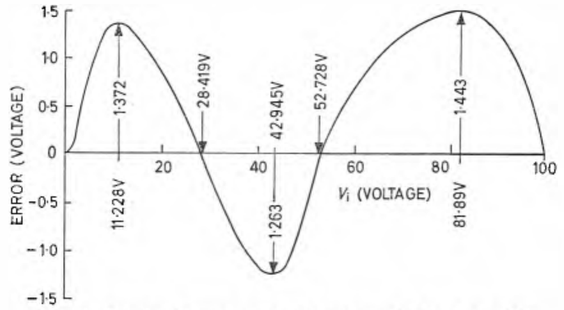


Fig. 4. Sketch of error curve of 'squarer' when $n = 0{\cdot}21$

This curve, over the input voltage range 0 to 100V, is plotted in Fig. 3 where e_1, e_2 and e_3 are local maxima in the magnitude of e. The behaviour of these maxima as a result of varying P_2 or Q can be deduced from Fig. 2. An increase in the value of Q, together with adjustment of P_1 to maintain the constraint (7), results in an increase in the magnitude of e_1 and e_3 and a decrease in magnitude of e_2. An increase in P_2 has the same effect.

In Fig. 2, the 'sandwich' effect of I_T between QI_s and P_2V_i is noteworthy. This shows that at low input voltages (and currents) the network characteristic is assymptotic to R_p and at high input voltages is assymptotic to R_s and infers the importance of the selection of these padding resistors.

If resistor values are chosen so that $e_1 = e_2 = e_3$, any deviation from these values must increase the maximum error of the function generator. For any set of parameters K, r and n it seems most unlikely that there will be more than one such optimum choice of resistor values. This was confirmed by an error 'contour plotting' investigation on a digital computer.

In view of the constraint (7), any two resistor values can be considered as being independently variable. In order to exploit the fact that the error e can be considered as a function of two independent variables, for any given V_i, parameters V_1 and V_2 were introduced as follows:

(a) Suppose $Q = 0$ and $P_2 = 0$, then choose P_1 to make $e = 0$ at $V_i = V_2$.

Thus

$$P_1 = 100^{r-1} \,.\, V_2^{(1/n)-r} \quad \ldots\ldots\ldots\ldots \ (9)$$

(b) Retaining this value of P_1 and with $Q = 0$ choose P_2 to make $e = 0$ at $V_i = V_1$.

Thus

$$P_2 = P_1 \,.\, (V_1/100^{r-1}) - V_1^{(1/n)-1} \quad \ldots\ldots\ldots \ (10)$$

(c) Now choose Q to make $e = 0$ at $V_i = 100$V.

$$Q = 1/(P_1 - P_2) - [100\,(P_1 - P_2)]^{n-1} \quad \ldots\ldots \ (11)$$

(obtained by substituting the value of I_s when $V_i = 100$V from equation (5) into equation (4)).

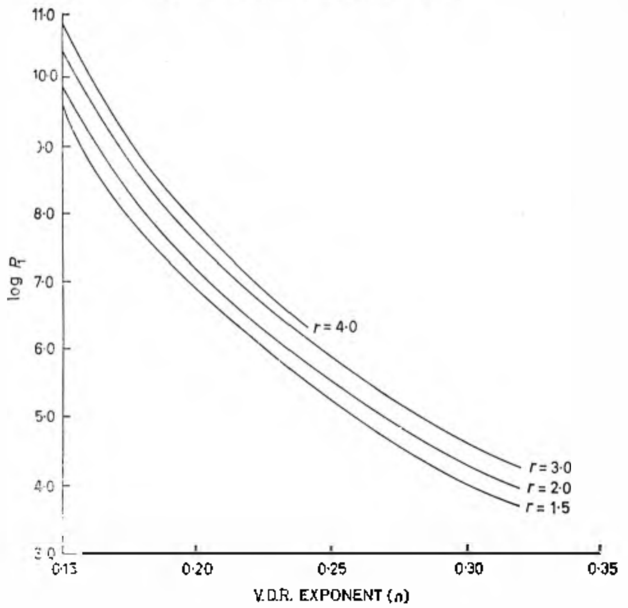
The error e is now a function of V_i, V_1 and V_2. In practice it was found that e did take zero values in the vicinity of V_1 and V_2.

The digital computer (IBM 1620) was programmed so that, with input data r and n and arbitrarily chosen starting values of V_1 and V_2, the calculation proceeded from equation (9) through equations (10) and (11) to (4). At equation (4) a sub-routine calculated 25 values of I_s to give 25 equi-spaced values of V_i (i.e. 4V intervals in range 0 to 100V) and thence 25 values of e were computed from equation (5). Initially it was hoped to apply the error criterion $e_1 = e_2 = e_3$ but it was found that overall computation times were excessively long and thus a least square error criterion was finally adopted. Therefore from equation (5) the computation proceeded to calculate:

$$\sum_{i=1}^{i=25} e_i^2$$

At this point a standard 'valley descending' sub-routine was used in which the above calculation was repeated for three sets of V_1, V_2 values centred on the arbitrarily chosen values of V_1 and V_2. From the resulting sum square error values the line of steepest descent was determined and the programme then incremented V_1, V_2 in this direction and recommenced the calculations. This process continued until such time that the optimum V_1, V_2 had been located to within 0·1V in the V_1, V_2 plane. The computer finally printed out values for r, n, P_1, P_2, Q, r.m.s. error, e_1, e_2, e_3, and the location of the peak errors with respect to V_i.

Fig. 5. Graph of n against $\log P_1$



Results

Table 1 gives the results of the calculations described above for the power law exponents 1·5, 2, 3 and 4. The application of this table to v.d.r. function generator design is clearly a simple matter. For example if a particular v.d.r. characteristic has been experimentally measured to give:

$$E = 290\, I^{0 \cdot 27}$$

and it is required to construct a function generator:

$$V_o = -(\text{sgn}\, V_i) \,.\, V_i^2/100$$

(Note that the relatively wide tolerances on v.d.r. manu-

TABLE 1

Design Data for V.D.R. Function Generators

NOTE: *In this table numerical values for $P_1 P_2$ and Q are written in the form aEb. This is to be interpreted as $a \times 10^b$. e.g. 0·3210E07 = 0·3210 × 10⁷*

3/2 POWER LAW

n	P_1	P_2	Q	e_{rms}	e_1	e_2	e_3	V_{e1}	V_1	V_{e2}	V_2	V_{e3}
0·15	·4563E10	·2231E10	·2112E—09	0·932	+1·729	−1·049	+1·138	10·113	26·401	37·313	47·423	77·867
0·18	·5222E08	·2315E08	·1702E—07	0·849	+1·293	−1·195	+1·069	7·944	21·866	35·486	45·338	79·132
0·21	·3210E07	·1423E07	·2570E—06	0·707	+1·300	− ·975	+ ·778	8·007	27·532	37·361	48·767	80·582
0·24	·3019E06	·1242E06	·2535E—05	0·643	+1·045	− ·962	+ ·707	7·329	22·850	35·593	47·860	80·427
0·27	·4901E05	·1872E05	·1445E—04	0·595	+ ·846	− ·943	+ ·617	5·361	22·218	35·272	47·272	81·283
0·30	·1073E05	·3717E04	·6162E—04	0·531	+ ·642	− ·873	+ ·643	4·913	19·629	32·727	45·001	79·271
0·32	·4623E04	·1496E04	·1364E—03	0·504	+ ·534	− ·828	+ ·644	4·387	18·061	31·597	43·647	79·133

SQUARE LAW

n	P_1	P_2	Q	e_{rms}	e_1	e_2	e_3	V_{e1}	V_1	V_{e2}	V_2	V_{e3}
0·15	·8750E10	·2194E10	·6219E—10	1·403	+1·574	−2·217	+1·994	11·532	26·368	42·229	50·334	81·520
0·18	·1227E09	·2996E08	·4070E—08	1·162	+1·500	−1·656	+1·716	12·026	27·233	41·944	51·582	81·987
0·21	·5703E07	·1326E07	·7938E—07	0·988	+1·372	−1·263	+1·443	11·228	28·419	42·945	52·728	81·891
0·24	·5772E06	·1261E06	·6935E—06	0·796	+1·236	−1·081	+1·063	10·784	30·972	43·995	54·452	83·110
0·27	·8830E05	·1662E05	·4053E—05	0·680	+ ·942	− ·996	+ ·920	9·990	30·401	42·492	53·599	83·125
0·30	·1954E05	·3086E04	·1624E—04	0·594	+ ·689	− ·896	+ ·818	8·191	26·630	41·424	52·269	83·809
0·32	·8551E04	·1208E04	·3356E—04	0·526	+ ·572	− ·841	+ ·681	7·764	26·690	40·470	52·165	84·173

CUBE LAW

n	P_1	P_2	Q	e_{rms}	e_1	e_2	e_3	V_{e1}	V_1	V_{e2}	V_2	V_{e3}
0·15	·3647E11	·4228E10	·7689E—11	1·452	+1·531	−2·022	+2·527	19·515	37·035	52·782	61·605	86·671
0·18	·3754E09	·3511E08	·6269E—09	1·101	+1·125	−1·712	+1·955	17·069	34·948	52·683	61·667	86·755
0·21	·1560E08	·1436E07	·1163E—07	0·802	+1·141	−1·080	+1·187	18·921	40·133	55·483	64·918	88·416
0·24	·1288E07	·8948E05	·1093E—06	0·569	+ ·798	— ·681	+ ·965	17·758	42·921	54·394	64·380	87·586
0·27	·1899E06	·9040E04	·4916E—06	0·358	+ ·507	− ·539	+ ·487	16·333	42·353	55·234	65·634	88·590
0·30	·4003E05	·9344E03	·1281E—05	0·189	+ ·220	− ·280	+ ·283	14·249	40·468	53·441	64·185	88·458
0·32	·1684E05	·1572E03	·1210E—05	0·072	+ ·083	− ·121	+ ·094	14·216	40·372	54·364	64·713	89·626

QUARTIC LAW

n	P_1	P_2	Q	e_{rms}	e_1	e_2	e_3	V_{e1}	V_1	V_{e2}	V_2	V_{e3}
0·15	·8902E11	·6724E10	·1633E—11	1·105	+1·549	−1·432	+1·651	26·402	49·166	63·576	71·789	89·686
0·18	·7624E09	·3818E08	·1365E—09	0·680	+ ·939	− ·927	+1·191	24·318	47·440	62·001	71·263	89·321
0·21	·2611E08	·7854E06	·2230E—08	0·379	+ ·534	− ·502	+ ·563	24·355	48·677	63·047	72·375	90·566
0·24	·2042E07	·1536E05	·7188E—08	0·087	+ ·124	− ·116	+ ·133	23·324	50·281	63·018	72·501	91·178

facture require K and n to be experimentally measured before this design method can be applied. This measurement is a simple task.)

Entering Table 1, 'square law' section, with $n = 0{\cdot}21$ the following values are obtained:

$$P_1 = 0{\cdot}5703\times10^7,\ P_2 = 0{\cdot}1326\times10^7,\ Q = 0{\cdot}7938\times10^{-7}$$

now $C = K^{1/n} = 290^{1/0\cdot21} = 5{\cdot}32\times10^{11}$.

Therefore

$$R_f = C/P_1 = 93{\cdot}3\text{k}\Omega$$

$$R_p = C/P_2 = 400{\cdot}0\text{k}\Omega$$

and

$$R_s = CQ = 42{\cdot}2\text{k}\Omega$$

In practice it is sufficient to set these resistance values with a tolerance of ±1 per cent.

From Table 1 the error data can be used to sketch the error curve of this function generator. This is shown in Fig. 4, in which the maximum error is seen to be $+1{\cdot}443$V at $V_i = 81{\cdot}89$V. As previously mentioned, the location of zero errors at V_1 and V_2 is approximate.

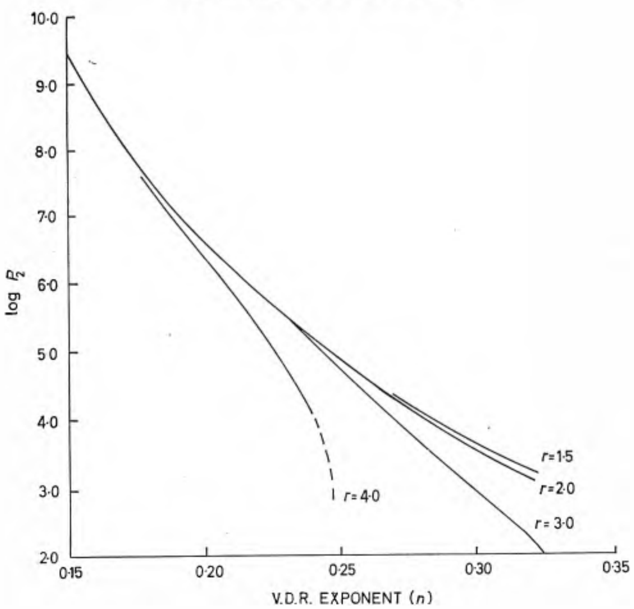
It is possible (and necessary) to interpolate between r and n values in Table 1. For P_1, P_2 and Q this is best achieved graphically. Figs. 5, 6 and 7 show graphs of the common logarithms of P_1, P_2 and Q against n. These graphs, if plotted on a scale of about 2in/decade are sufficiently accurate for interpolation purposes. Note that P_2 and Q tend asymptotically to zero as $1/n \to r$ corresponding to $R_p \to \infty$ and $R_s \to 0$. That is, at $1/n = r$ no padding resistors are required.

Selection of a V.D.R.

The form of a particular error curve was shown in Fig. 4. For all cases considered here the error curve is similar, i.e. two maxima and one minimum. It is, however, more instructive to examine a graph of r.m.s. error against r with n a parameter. This is shown in Fig. 8 where it is apparent that there is a value of r, for each n, where the r.m.s. error is a maximum.

In qualitative terms, at values of $r \to 1$ the function generator input circuit is nearly linear in characteristic and the contribution required from the v.d.r. is small and hence small errors. At values of $r \to 1/n$ the v.d.r. requires

Fig. 6. Graph of n against log P_2



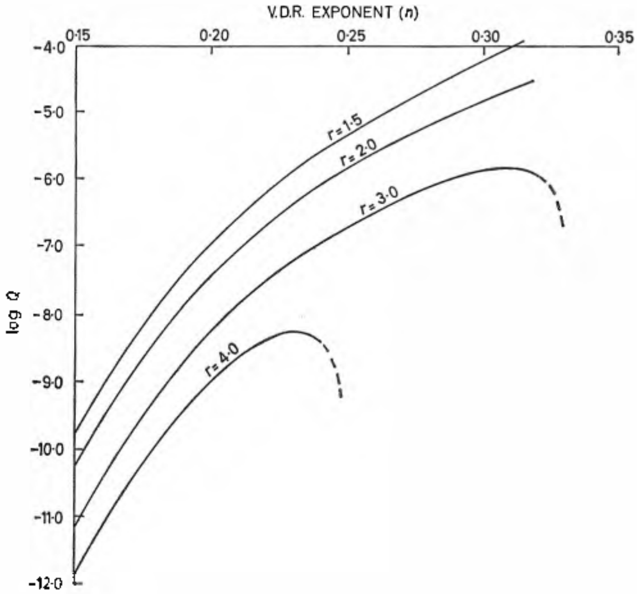


Fig. 7. Graph of n against log Q

little padding from R_p and R_s and thus errors are again small. Between these extremes is a zone, where both v.d.r. and padding resistors equally contribute to forming the function and, since their individual characteristics are far removed from that of the desired function, the errors are at a maximum.

Fig. 8 is, therefore, valuable as an aid to selection of v.d.r.'s for function generation. In general terms v.d.r.'s whose exponents are greater than about 0·21 are obviously desirable in this application.

The selection of the v.d.r. parameter C (or K) can be made from Table 1. In a given analogue computer there are limiting values on input and feedback resistors, the upper limit being possibly 10MΩ and the lower limit determined by the maximum load current which can be drawn from an operational amplifier. These limits are particularly important when values are being obtained for R_f and R_p. With n initially selected and P_1 found, the selection of C can be stated as:

$$R_{f(\max)} \,.\, P_1 \geqslant C \geqslant R_{f(\min)} \,.\, P_1\ (= (100/i_{\max}) \,.\, P_1)$$

and similarly for P_2. In practice the lower limit condition, $C \geqslant R_{f(\min)} \,.\, P_1$, is the one most likely to be violated.

Error Compensation

As a result of the analysis of the v.d.r. function generator circuit the user knows, *a priori*, the nature and magnitude of the errors. Therefore a compensating circuit can be included to reduce these errors in cases where high precision is required. One method of doing this is to construct, with additional v.d.r. function generators, a polynomial approximation to the error curve and subtract this from the principal function generator output.

It was stated earlier that the v.d.r. parameter K was temperature dependent, i.e.

$$E = K(1+\gamma T)I^n$$

where T is temperature

and γ is the temperature coefficient.

It is stated[5] that γ lies between $-0{\cdot}6$ and $-0{\cdot}9$ per cent/°C at constant voltage and $-0{\cdot}14$ to $-0{\cdot}20$ per cent/°C at constant current. For the v.d.r. padding circuit of Fig. 1 the constant current value of γ is the more applicable. Table 1 describes the optimum settings of padding resistors

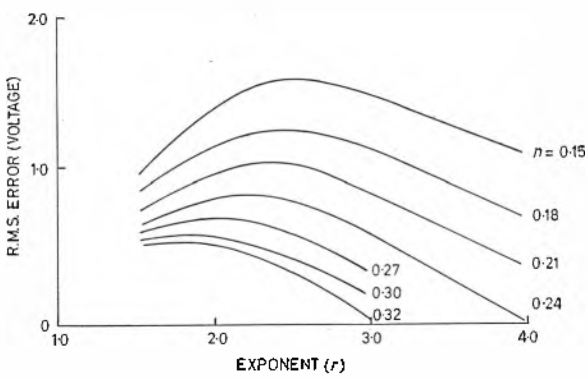


Fig. 8. Variation of r.m.s. error against r

in terms of ratios C/R_f, C/R_p and R_s/C. To maintain accuracy these ratios must remain constant with temperature. This can be achieved with the aid of negative temperature coefficient resistors, e.g.

$$C/R_f = \text{constant} = \frac{K^{1/n}(1+\gamma T)^{1/n}}{R_f(1+\delta T)}$$

$$\simeq \frac{K^{1/n}}{R_f}\left[\frac{1+(1/n)\gamma T}{1+\delta T}\right]$$

Therefore, to maintain a constant ratio C/R_f:

$$(1/n)\gamma = \delta$$

where δ is the required temperature coefficient of R_f. The other ratios are treated in a similar manner.

Temperature compensation of this v.d.r. function generator is only necessary in applications where the range of ambient temperature is large. For normal laboratory environments it is unnecessary since the error introduced is small compared to e_1, e_2 and e_3.

The padding resistor circuit used in this function generator is of the simplest form. More complex padding circuits are possible and with careful choice can lead to an error curve having more than one and a half cycles of error as shown in Fig. 3. In this case, the magnitude of errror would be reduced. Optimization would then take place, not in a space defined by the two variables V_1 and V_2, but in a multi-dimensional space say $V_1 V_2 \ldots . V_n$. This leads to a considerable complication in calculating the optimum condition but is well within the capabilities of modern digital computers.

Conclusions

A method is given whereby a v.d.r. function generator circuit operating in the range ± 100V can be rapidly designed for minimum error. To use this method, the v.d.r. parameters K and n must be known and should be experimentally measured. The overall result is a cheap, simple function generator of modest, but useful, accuracy which has application to analogue computer and control systems.

Acknowledgments

The authors are indebted to the Directors of African Explosives and Chemical Industries Ltd, Johannesburg, for permitting the Company's digital computer to be used to solve this problem. They also gratefully acknowledge the assistance given by Mr. G. J. H. Evans, A.E. & C.I. Ltd, who gave much of his spare time to programme this problem for the computer.

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A New Flight Data Recorder

A new flight data recorder is now in production by Plessey-UK Ltd. Known as the PU710 it is, basically an instrument for flight incident data recording, capable of assisting in tracing the causes and circumstances of equipment faults and human errors arising during flight. In this primary role it meets the Ministry of Aviation's mandatory requirement for providing a permanent record, which will survive shock, fire, and chemical attack, of the aircraft's height, heading, airspeed, pitch attitude and acceleration.

For the maximum usefulness to airline operators, however, such a recorder should have sufficient capacity and expansion potential for the inclusion of additional channels to investigate the possibilities of performance and maintenance recording, and for the provision of parallel recording facilities for quick access play-back purposes. It should also be capable of expansion to provide full maintenance recording, should this prove economically desirable.

The PV710 Series is, therefore, a range of equipments designed around a simple 24-channel digital recording system, which itself forms part of a broader concept offering considerable additional capacity and providing for both wire and tape recording media. Expansion to 70 channels, at a sampling rate appropriate to the investigation and monitoring of aircraft systems, is provided for in the original equipment.

A higher standard of accuracy than that demanded for a simple accident recorder was considered desirable for this equipment for two reasons:

(1) Standards of resolution accuracy, particularly of transducers, are constantly rising and equipment designed merely to meet a minimum specification would quickly become obsolete.

(2) Maintenance investigations require considerably higher accuracy if they are to be of value.

From the start therefore, the Plessey-UK engineers who developed the equipment set themselves a performance target of one-third of a degree of synchro shaft angle, or one part in 1 000 of the full rotation. In practice, the PV710 employs a 10-bit digital word format which gives a resolution of 1 part in 1 024, and this same order of accuracy of 0·1 per cent is maintained right from transducer input to the ultimate ground processor results.

Furthermore, the comprehensive and continuous self-checking system employed will not permit a calibration error or fault in the recording system to continue undetected.

The basic airborne installation consists of (1) the Recorder Unit, (2) a central Processing Unit which accepts and processes the various signals fed to it from transducers mounted at various points in the aircraft, and (3) the pilot's Control Unit.

The Control Unit is mounted in the pilot's console. Besides giving continuous indication of satisfactory system operation it provides simple test facilities for aircrew or ground crew use. Operation of the entire equipment is fully automatic, with the sole exception of setting and inserting a flight identification number. Data recording is started automatically as soon as the air speed reaches 50 knots and comprehensive performance self-monitoring normally obviates the need for aircrew attention at any time during a flight.

The Recorder assembly is manufactured by S. Davall and Sons and comprises two units: the Drive Unit, which incorporates a rigid chassis for airframe attachment, and the Cassette, which latches into the Drive Unit and is firmly secured by steel straps. The actual changeover of cassettes, every 200 hours, is accomplished very easily in a few seconds.

A 'doll's eye' indicator and test switch, duplicating those in the Control Unit, are included in the Recorder Unit so that maintenance crews can replace cassettes and make functional checks on the equipment without communication with the flight deck.

The Processing Unit, which includes the air speed and altitude transducers, is built into a ¾ ATR unit for mounting in the radio rack. The electronic circuits are in plug-in modules and are separately replaceable without recalibration or other adjustment.

A 12V d.c. to 230V a.c. Silicon Controlled Rectifier Invertor

By D. I. Ross* and B. E. G. Goodger*

A 12V d.c. to 230V a.c. silicon controlled rectifier (s.c.r.) invertor is described. The frequency of the invertor is controlled by a 1kc/s fork oscillator to produce a stable 50c/s output. The circuit works efficiently under loads varying from zero to 70W. A safety circuit is incorporated to protect the s.c.r.'s from accidental overload.

(Voir page 634 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 641)

A NUMBER of d.c. to a.c. invertors using silicon controlled rectifiers and operating at frequencies in the vicinity of 400c/s have been described in the literature[1-4]. However at frequencies around 50c/s difficulties in commutation are experienced and the efficiency of the invertor drops. This article describes an s.c.r. invertor which will operate very efficiently at 50c/s under loads varying from zero to its full operating load of 70W.

The S.C.R. Invertor

The basic invertor circuit[1-3] is shown in Fig. 1. Operation is as follows. When SCR_1 is switched on current flows through half the primary of the output transformer

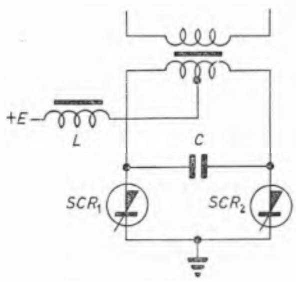

Fig. 1. Basic s.c.r. invertor circuit

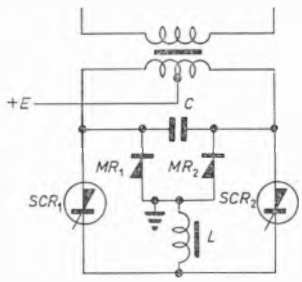

Fig. 2. Improved s.c.r. invertor circuit

charging the capacitor C to $2E$ by auto-transformer action. If now SCR_2 is switched on C will discharge through the two s.c.r.'s producing a reverse voltage across SCR_1 thus cutting it off. C then charges in the opposite direction ready for the next switching cycle. The inductance L in the positive supply line limits the capacitor charging current during the switching period. The time-constant of L and C should be an appreciable fraction of one cycle of the operating frequency. This circuit is very sensitive to the load it is supplying, the required commutation capacitance varying considerably with the load. An improved circuit is shown in Fig. 2. The feedback diodes MR_1, MR_2 compensate for inductive loads and reduce the peak voltage across the s.c.r.'s. The inductance L is in the cathode circuit of the s.c.r. and forms an oscillatory circuit with C thus producing the required reverse voltage across an s.c.r. during the switching period. The value of L is appreciably less in this case since the resonant period of L and C is of the order of the switch-off time of the s.c.r. This circuit is not so sensitive to varying load conditions as the first and appears to work satisfactorily above about 100c/s.

However at lower frequencies the circuit becomes critical, design of the output transformer particularly being important in this region. The reason for this is that during the interval between switching periods (an interval of practically half a cycle) the commutating capacitor C must hold sufficient of its charge to produce the revers voltage required to cut the conducting s.c.r. off when th other is switched on by the next trigger pulse. At fre quencies of 50c/s this means that either C must be larg or the transformer primary impedance must be kept hig so that no appreciable charge is lost through the trans

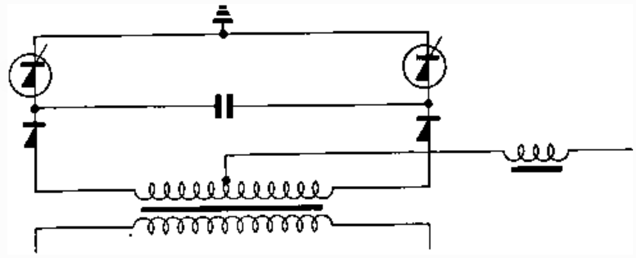

Fig. 3. Low frequency s.c.r. invertor circuit as given by Ward

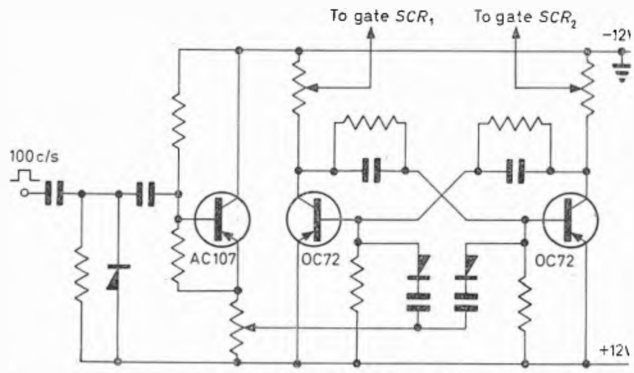

Fig. 4. S.C.R. trigger circuit

former primary. In converting 12V to 230V the trans former primary impedance is necessarily limited so tha C must be large to obtain reliable commutation. In th prototype built a capacitance of around 80μF was requirec the actual value being very dependent on the load th invertor was driving. With such capacitances the chargin time becomes long, and the increased losses reduce th efficiency considerably. This problem could be overcom if after having been charged to its full voltage $2E$, th commutating capacitor could be isolated from the rest o the circuit until the next switching pulse fires the secon s.c.r. Ward[5] has done this by inserting series diode between each side of the transformer primary and th commutating capacitor as shown in Fig. 3. In this circui the only discharge path for the capacitor is through th s.c.r.'s so no appreciable loss can occur between switchin periods. By incorporating series diodes in the circuit, leavin in the shunt diodes of Fig. 2 to limit the peak voltag across the s.c.r.'s, the required commutating capacitanc was reduced to 7μF and the invertor functioned satis factorily under all load conditions from zero to full loa whether resistive or inductive. In the case of inductiv

* *Victoria University of Wellington, New Zealand.*

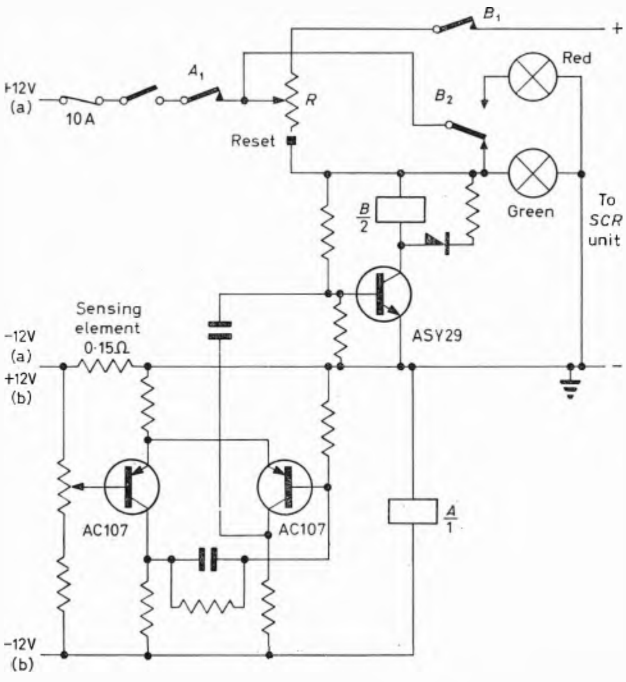

Fig. 5. S.C.R. safety circuit

loads some compensating capacitance was required across the load to reduce harmonic distortion of the output waveform.

S.C.R. Trigger Circuit

Transistorized firing circuits for s.c.r.'s have been published in the literature[6]. In this particular application a standard 50c/s a.c. supply was required to run seismic recorder motors. A stable 50c/s trigger pulse was therefore required to drive the s.c.r. supply.

A 1kc/s fork oscillator was used as the signal source since this was available and had the required stability. The 1kc/s signal was passed through a pulse shaper and decade counter to obtain 100c/s pulses. Further decade counters supplied 1sec, 1min, and 10min pulses for calibration of other equipment. Philips circuit blocks were used for the pulse shaper and decade units. The 100c/s pulses from the first decade were then fed into the trigger circuit shown in Fig. 4. This consists of an emitter-follower, to avoid loading of the decade counter, followed by a flip-flop unit which divides by two and supplies the required 50c/s gate pulses for the s.c.r.'s. The s.c.r. gates are direct coupled to the 250Ω potentiometers in the collector circuits of the OC72's. The use of potentiometers allows individual adjustment of the voltages required by each s.c.r. gate.

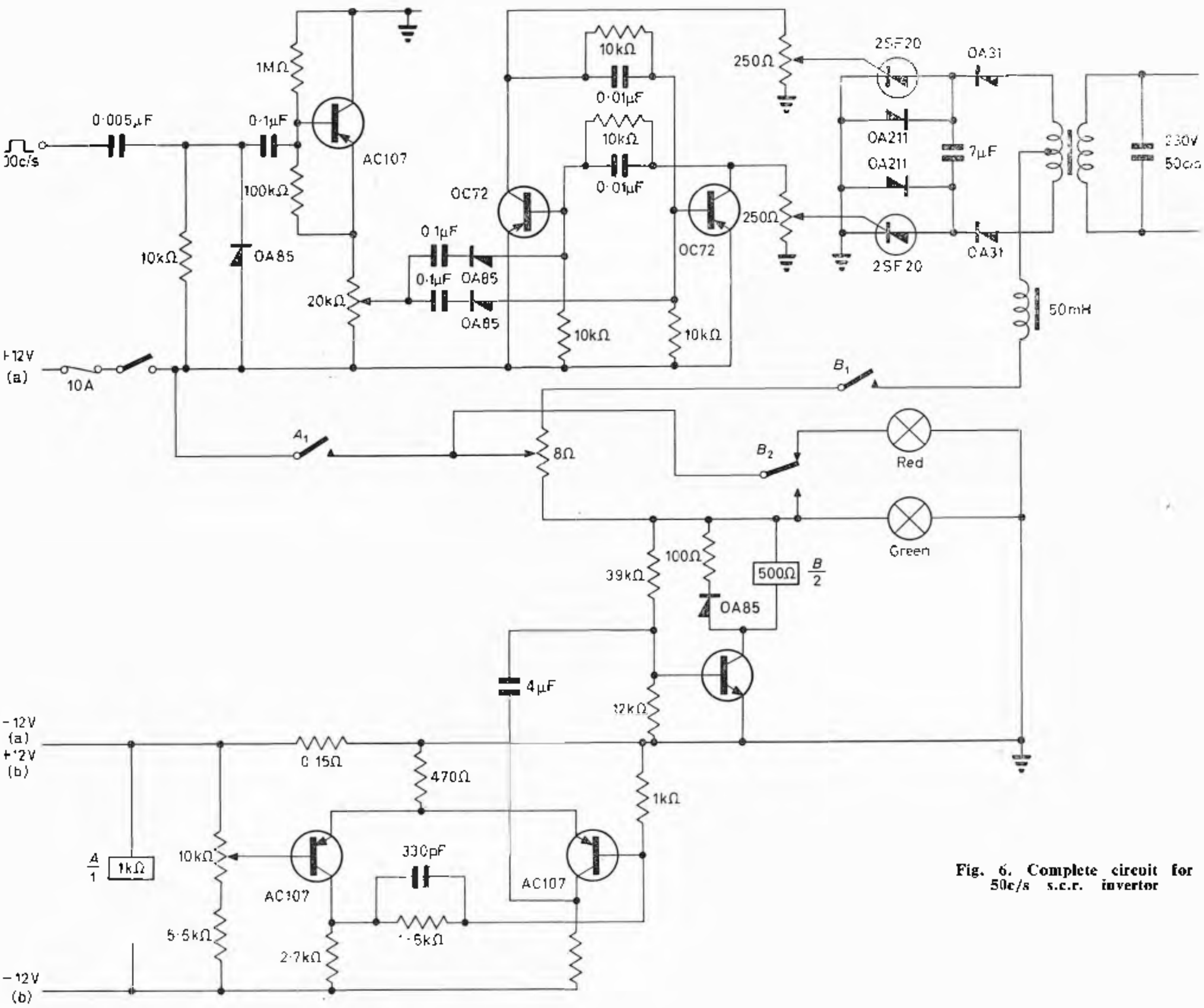

Fig. 6. Complete circuit for 50c/s s.c.r. invertor

S.C.R. Safety Circuit

The current through the s.c.r.'s will rise rapidly if triggering or commutation fail for any reason. Some form of safety circuit is therefore necessary to safeguard the s.c.r.'s if this were to happen. The circuit used in this particular case is shown in Fig. 5 and consists of a Schmidt trigger feeding a power transistor and relay. The Schmidt trigger is tied across a sensing element in the negative line to the invertor. If the current through this element rises above a specified level the Schmidt trigger switches, cutting off the power transistor and hence opening the relay *B* and breaking the power to the s.c.r.'s. The use of a relay instead of a transistor in the power line to the invertor eliminates the possibility of damage to the safety circuit by transients while still being sufficiently fast to safeguard the s.c.r.'s. Relay *A* cuts the power to the invertor should the power to the Schmidt trigger be cut off for any reason. Resistance *R* was used to adjust the supply current to the invertor and was arranged to act as a manual reset should the safety circuit cut the supply voltage. In this way the invertor could not be started until *R* was returned to its maximum position. The lights indicate the state of the relay *B*.

Conclusions

A reliable 12V d.c. to 230V a.c. invertor working at 50c/s has been constructed which functions satisfactorily under a wide variety of load conditions, the maximum output power with the s.c.r.'s used being 70W. The use of series diodes between the primary of the output transformer and the commutating capacitor allows a considerable reduction in the required commutating capacitor and makes the design of the transformer much less critical.

The complete circuit of the invertor is shown in Fig. 6. As already mentioned the invertor was constructed to run the synchronous motors of two seismic recorders. The circuit was sufficiently insensitive to load conditions that switching either or both the motors had no effect on the operation of the invertor. With the values shown the safety circuit was designed to open the s.c.r. supply line when the current rose above 7A. The efficiency of the invertor was very high, the only appreciable losses occurring in the choke, output transformer and safety circuit sensing element.

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A Four-Output Transistorized Amplifier for Communal Television Distribution

The amplifier shown in Fig. 1 is designed for the distribution of a television video signal of bandwidth up to 12Mc/s and of amplitude 1V peak-to-peak of either polarity. The circuit uses six type BCY42 transistors introduced recently by Standard Telephones & Cables Ltd.

Four outputs are provided, each suitable for feeding into a 75Ω coaxial line. In these conditions, the amplifier has an overall gain of unity. There is almost perfect isolation between the outputs, since the voltage gain is defined by overall feedback and the output impedance of the amplifier is very low. A 75Ω resistor provides matching for each output.

PERFORMANCE

Maximum operating temperature without heat sinks for the transistors: +50°C.

Sinusoidal Signals

Power gain is unity (this can be varied slightly by adjusting the feedback resistors R_F) from the 75Ω input line to each 75Ω output line.

Maximum variation in output level if three out of the four outputs are removed:

0·2dB over a bandwidth of 12Mc/s.

Bandwidth under all load conditions:

±0·2db from below 10c/s to 12Mc/s.

±1db maintained up to 20Mc/s.

Low frequency input time-constant:

0·5sec.

Delay time:

4·6nsec approx., all outputs connected

3·4nsec approx., one output connected

High frequency phase response under all load conditions:

<10° phase change up to 20Mc/s

Total r.m.s. distortion (input level 1 volt peak-to-peak, all load conditions):

<1% below 5Mc/s

<4% below 10Mc/s

<8% below 12Mc/s

Pulse Signals

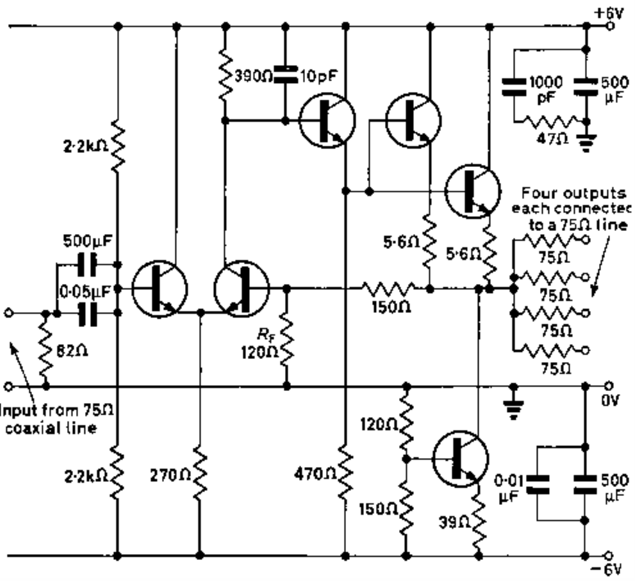
Maximum overshoot is less than 5% for an input pulse having a 20nsec rise time.

Rise time: 10nsec approx. one output connected
12nsec approx. four outputs connected

L.F. Response: 10% drop on output pulse for an input pulse of 50msec duration.

The high frequency performance of this amplifier depends on two conditions: The supply lines must be of low inductance. Also, the electrolytic decoupling capacitors must be shunted by ceramic capacitors with leads of low inductance so that their resonant frequency is above the highest frequency to be used.

Fig. 1. The television distribution amplifier



The Design of Minimal NOR/NAND Logical Circuits

By D. Zissos* and G. W. Copperwhite*

A systematic method is described of designing minimal NOR/NAND *logical circuits, when the complements of the input signals are not available.*
A minimization technique for NOR/NAND *circuits is introduced and a method of classifying logical circuits is mentioned.*

(Voir page 635 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 642)

THE design technique described here always results in the minimal circuit realization(s) of a given switching function. Minimal circuit realizations are at present essential for economy and reliability, although with the advent of microminiaturization this may not be so in the future. A circuit is minimal if it uses the minimum number of modules and if the number of connexions is minimum, provided the first condition is satisfied.

Throughout the article the NOR module is considered as producing the AND function of the complements of the input signals and the NAND module their OR function. This is because it is easier to think in terms of AND and OR functions rather than in terms of NOR and NAND functions.

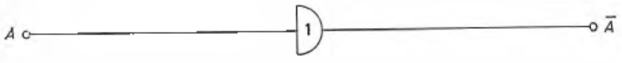


Fig. 1. A NOR module used as an invertor

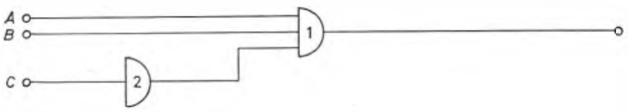


Fig. 2. Circuit realizing the AND function $\overline{A}\,\overline{B}\,\overline{C}$

The design technique consists of first designing non-minimal circuits and then minimizing them using the simplification technique explained in Appendix 1.

The circuit configurations shown are those of NOR modules and negative logic (negative level of signals corresponds to the logical '1' and the positive level of signals to the logical '0') which are identical to the circuit configurations of NAND modules and positive logic. In appendix 2 a method for realizing NAND circuits and negative logic is suggested.

In Appendix 3 a method of classifying logical circuits is explained.

Realization of the NOR Function

A NOR module with a single input generates the complement of the applied signal, as shown in Fig. 1.

Realization of the AND Function

To realize the AND function, f, of given signals the comcomplement of each signal is fed into a NOR module. The output of this module is the required AND function, f.

Example To realize the function $\overline{A}\,\overline{B}\,\overline{C}$, one feeds A, B and $\overline{C}$ into module 1 of Fig. 2.

* *City of Liverpool College of Technology.*

Fig. 3 shows the general NOR module arrangement for generating the AND function of three signals A, B and C.

Realization of the OR Function

The OR function, f, of specified signals is realized in two steps. In the first step the complement of the function

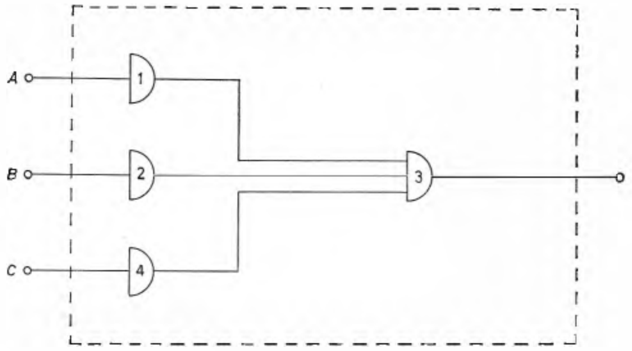


Fig. 3. NOR module arrangement equivalent to a three-input AND gate

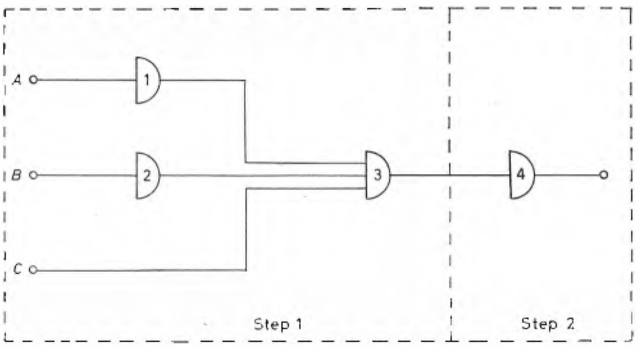


Fig. 4. Circuit realizing the OR function $\overline{A}+\overline{B}+C$

to be realized, $\overline{f}$ (AND form), is generated as explained above. In the second step f is derived from $\overline{f}$ by feeding it in to a NOR module.

Example Fig. 4 shows the realization of the OR function $\overline{A}+\overline{B}+C$.

The general circuit arrangement for generating the OR function of three signals A, B and C is shown in Fig. 5.

Realization of the OR-Function-with-a-Common-Factor

The OR function is first realized as explained above. The complement of the common factor is next fed into the final module, the output of which is the required signal.

Example The circuit realization of the function $A(B+C)$ is shown in Fig. 6.

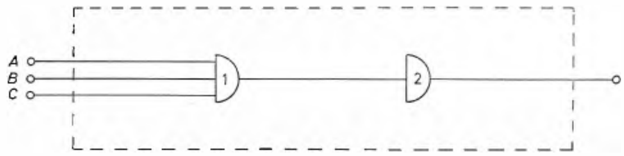


Fig. 5. NOR module arrangement equivalent to a three-input OR gate

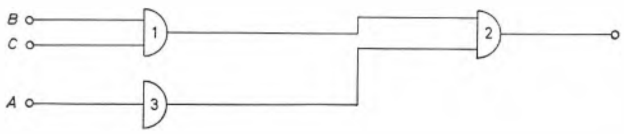


Fig. 6. Circuit realization of the function A(B + C)

Realization of AND/OR and OR/AND Functions

The realization of AND/OR and OR/AND functions is accomplished systematically in three steps.

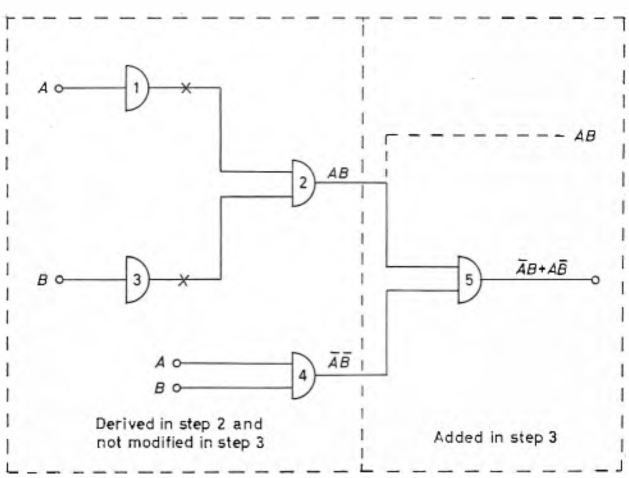
Step 1 In this step the minimal AND/OR expression of the complement of the function to be realized is derived either by applying De Morgan's theorem to it or from its excitation table directly, see Appendix 4. If there are common factors, the expression is rewritten with the common factors, taken outside the brackets. All forms of the minimal expression thus derived are noted.

Step 2 In the second step all the signals specified by the AND and the OR-with-a-common-factor terms derived in Step 1 are generated separately, using the methods already explained.

Step 3 In the third and final step the circuits derived in Step 2 are examined for redundant modules using the minimization technique explained in Appendix 1. The design is completed by feeding the output signals of the minimal circuit to the final module, the output of which is then the required function. The non-minimal circuits are ignored.

The design technique will now be illustrated with six selected examples. In examples 1, 2, 3 and 4 because the functions are relatively simple, their complements are derived using De Morgan's theorem. The functions to be realized in examples 5 and 6 contain more than two product terms, therefore their complements are derived by considering the '0' entries in their respective excitation tables. See Appendix 4.

Fig. 7. Minimal circuit realizing function $\overline{A}$ B + A $\overline{B}$



EXAMPLE 1

In this example the Exclusive OR function (either but not both) of two variables A and B, viz. $\overline{A}\,B + A\,\overline{B}$ will be realized.

Step 1 $f = \overline{A}\,B + A\,\overline{B}$

$$\overline{f} = (A + \overline{B})(\overline{A} + B) = A\,\overline{A} + A\,B + \overline{A}\,\overline{B} + B\,\overline{B}$$

$$A\,B + \overline{A}\,\overline{B} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

Step 2 Each of the two 'product terms' in equation (1) is separately generated, as shown in Fig. 7.

Step 3 Modules 1 and 3 have each a signal input. Therefore the circuit is examined with a view to eliminating them. A signal $\overline{A}\,B$ would make module 1 redundant, but such a signal is not available. Similarly a signal $A\,\overline{B}$ would make module 3 redundant but this also is not available. Therefore Fig. 7 is the minimal circuit realization of the function.

Note If a second output is taken from module 2, then the circuit becomes a half-adder, a circuit commonly used in digital computers.

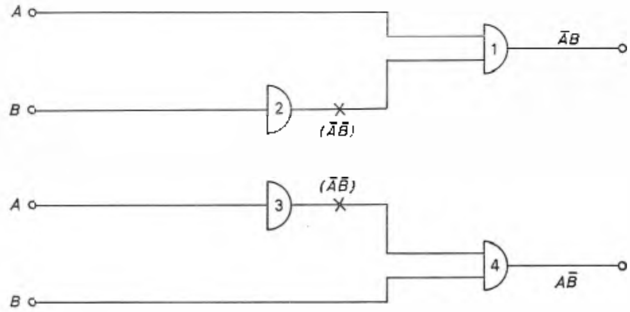


Fig. 8. Non-minimal circuit generating the signal $\overline{A}$ B and A $\overline{B}$

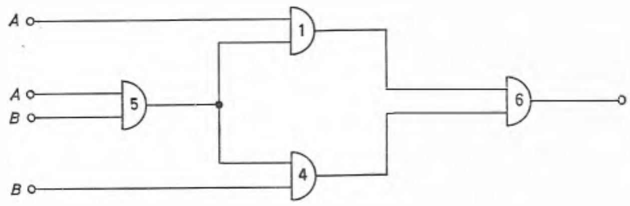


Fig. 9. Minimal circuit realization of the function A B + $\overline{A}\,\overline{B}$

EXAMPLE 2

In the second example the complement of the Exclusive OR function (both or none) of two variables A and B, viz. $A\,B + \overline{A}\,\overline{B}$ will be realized.

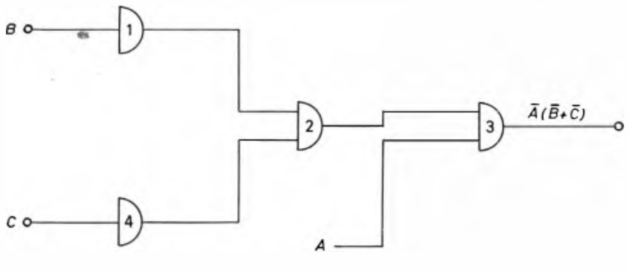
Step 1 $f = A\,B + \overline{A}\,\overline{B}$

By De Morgan's theorem

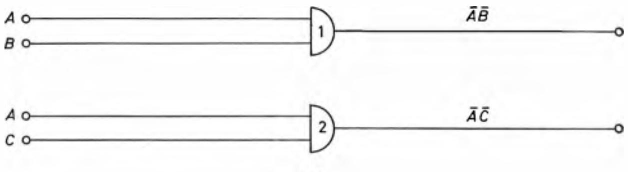
$$\overline{f} = \overline{A}\,B + A\,\overline{B} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

Step 2 The two signals specified by 'the product terms' $A\,\overline{B}$ and $\overline{A}\,B$ are separately generated by the non-minimal circuit of Fig. 8.

Step 3 Modules 2 and 3 are both used as invertors, therefore the possibility of eliminating them is now examined. The alternative signals that may be used instead of $\overline{A}$ and $\overline{B}$ are shown in brackets in Fig. 8. They are both the same, viz. $\overline{A}\,\overline{B}$. Although this signal is not being generated in the circuit of Fig. 8, it can be produced by the addition of an additional module, module 5. The addition of module 5 makes, therefore, module 2 and 3 redundant and the design is finalized by the introduction of the final module 6. (Fig. 9).



(a)



(b)

Fig. 10 (a). Circuit corresponding to equation (3)
(b). Circuit corresponding to equation (4)

EXAMPLE 3

In this example the AND/OR function $A + B\,C$ will be realized.

Step 1 $f = A + B\,C$

By De Morgan's theorem

$\overline{f} = \overline{A}\,(\overline{B} + \overline{C})$ .... (3)

$= \overline{A}\,\overline{B} + \overline{A}\,\overline{C}$

Step 2 The 'OR function-with-a-common-factor' is realized in Fig. 10(a) and the two product terms $\overline{A}\,\overline{B}$ and $\overline{A}\,\overline{C}$ in Fig. 10(b).

Step 3 The circuit of Fig. 10(a) cannot be minimized, since the two signals that would eliminate modules 1 and 4, viz. $\overline{B}\,C$ and $B\,\overline{C}$ respectively are not readily available. The circuit of Fig. 10(b) is already in its minimal form. The second circuit is, therefore, chosen to generate the signals to feed the final module, as shown in Fig. 11.

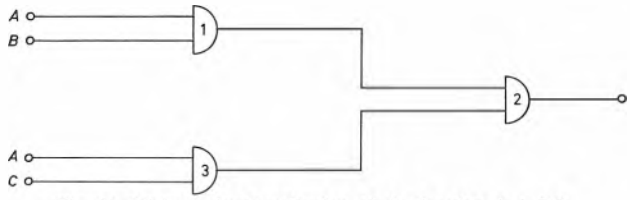
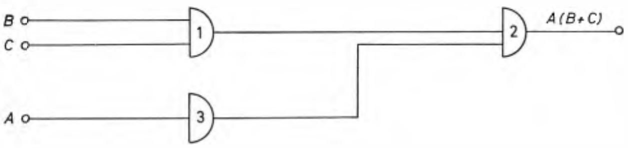
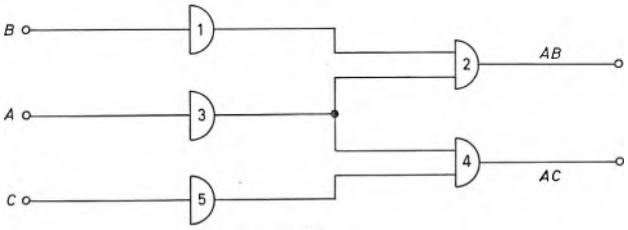


Fig. 11. Minimal circuit realization of the function A + B C

Fig. 12 (a). Circuit corresponding to equation (5)
(b). Circuit corresponding to equation (6)



(a)



(b)

TABLE 1

Excitation table of the function $A\,\overline{C} + B\,\overline{C} + A\,B$

C \ AB	00	01	11	10
0	0	1	1	1
1	0	0	1	0

EXAMPLE 4

In the fourth example the function $\overline{A} + \overline{B}\,\overline{C}$ is chosen for circuit realization.

Step 1 $f = \overline{A} + \overline{B}\,\overline{C}$

By De Morgan's theorem

$\overline{f} = A\,(B + C)$ .... (5)

$= A\;B + A\,C$ .... (6)

Step 2 The OR-function-with-a-common-factor $A(B+C)$ is realized by Fig. 12(a). Fig. 12(b) generates separately the two product terms $A\,B$ and $A\,C$.

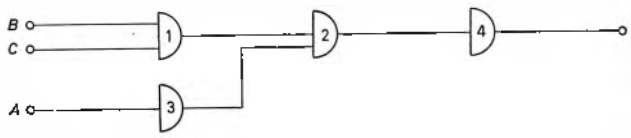
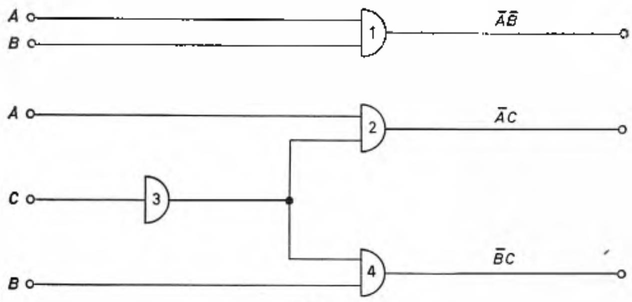
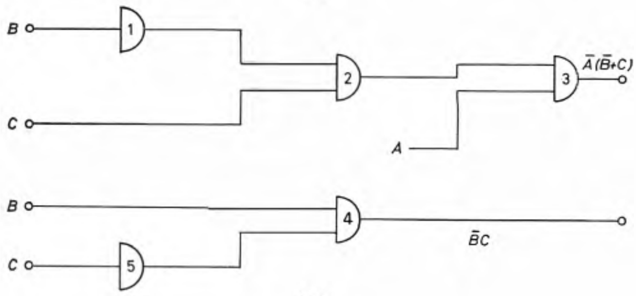


Fig. 13. Minimal circuit realisation of the function $\overline{A} + \overline{B}\,\overline{C}$

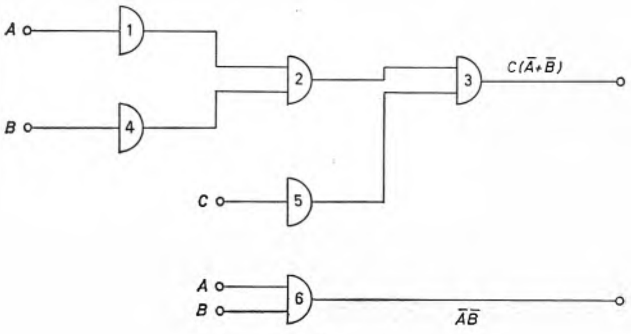
Fig. 14 (a). Circuit corresponding to equation (7)
(b). Circuit corresponding to equation (8)
(c). Circuit corresponding to equation (9)



(a)



(b)



(c)

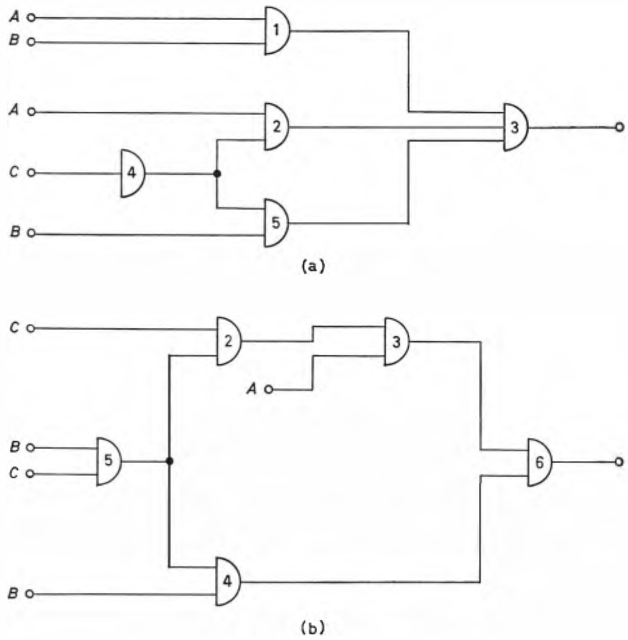


Fig. 15. Minimal circuit realizations of the switching function $A\overline{C} + B\overline{C} + AB$

Step 3 None of the circuits can be minimized, therefore the circuit corresponding to equation (1) is used to feed the final module, as shown in Fig. 13.

EXAMPLE 5

The function $f = A\,\overline{C} + B\,\overline{C} + A\,B$ will next be realized.

Step 1 From the function's excitation table (Table 1):

$$\overline{f} = \overline{A}\,\overline{B} + \overline{A}\,C + \overline{B}\,C \quad \ldots\ldots\ldots\ldots \quad (7)$$

$$= \overline{A}\,(\overline{B} + C) + \overline{B}\,C \quad \ldots\ldots\ldots\ldots \quad (8)$$

$$= C\,(\overline{A} + \overline{B}) + \overline{A}\,\overline{B} \quad \ldots\ldots\ldots\ldots \quad (9)$$

Step 2 The circuits generating the product terms specified by equations (7), (8) and (9) are shown in Figs. 14(a), 14(b) and 14(c).

Step 3 The circuit of Fig. 14(a) cannot be minimized. Modules 1 and 5 in Fig. 14(b) can be eliminated if a module generating the signal $\overline{B}\,\overline{C}$ is introduced in its final form. Fig. 14(c) cannot be minimized.

This means that the circuits of Figs. 14(a) and 14(b) when connected to their final modules will produce two circuits with the same number of modules (5) and the same number of connexions (10), as shown in Fig. 15.

EXAMPLE 6

A more difficult circuit realization will now be attempted, viz. the realization of the function

$$f = \overline{A}\,B\,\overline{C} + \overline{A}\,\overline{B}\,C + A\,B\,C + A\,\overline{B}\,\overline{C}.$$

Step 1 All the different algebraic forms that the complement of the function, *f*, takes are derived directly from the excitation table of the function (Table 2) by considering the '0' entries as in example 5.

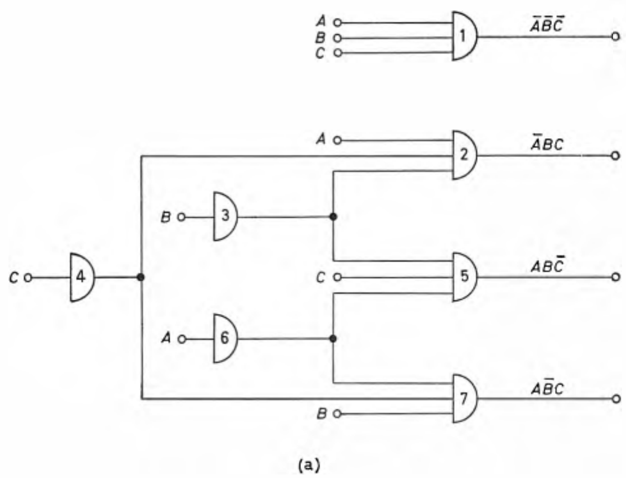
$$\overline{f} = \overline{A}\,\overline{B}\,\overline{C} + \overline{A}\,B\,C + A\,B\,\overline{C} + A\,\overline{B}\,C \quad \ldots\ldots \quad (10)$$

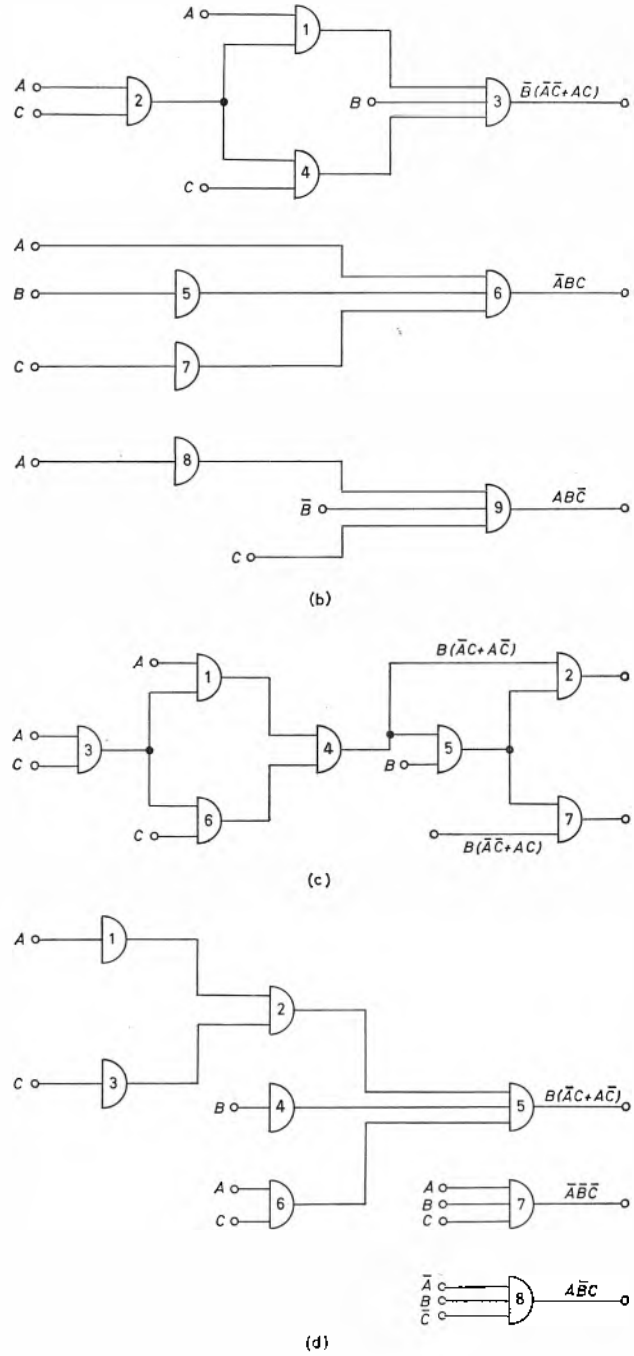
$$= \overline{B}\,(\overline{A}\,\overline{C} + A\,C) + \overline{A}\,B\,C + A\,B\,\overline{C} \quad \ldots\ldots \quad (11)$$

$$= \overline{B}\,(\overline{A}\,\overline{C} + A\,C) + B\,(\overline{A}\,C + A\,\overline{C}) \quad \ldots\ldots \quad (12)$$

$$= \overline{A}\,\overline{B}\,\overline{C} + A\,\overline{B}\,C + B\,(\overline{A}\,C + A\,\overline{C}) \quad \ldots\ldots \quad (13)$$

Fig. 16 (a). Circuit corresponding to equation (10)
(b). Circuit corresponding to equation (11)
(c). Circuit corresponding to equation (12)
(d). Circuit corresponding to equation (13)





Step 2 The circuits corresponding to equations (10), (11), (12) and (13) are shown in Figs. 16(a), 16(b), 16(c) and 16(d) respectively.

Step 3 Fig. 16(a) is first examined for redundant modules. The six signals, any of which can be used to feed module 2 instead of $\bar{B}$, are: $\overline{A}\,\overline{\overline{C}}$, $B\,\overline{C}$, $(A+\overline{B})\,\overline{C}$, $(\overline{A}+\overline{B})\,\overline{C}$, $(\overline{A}+B)\,\overline{C}$ and $\overline{A}\,B\,\overline{C}$. None of these signals is available in the circuit, therefore module 4 cannot be eliminated. Because of the symmetry of the circuit, modules 3 and 5 cannot be eliminated either, i.e. Fig. 16(a) is in its minimal form.

Fig. 16(b) is next examined for redundant modules.

Module 7 is used to generate the signal $\overline{C}$. But a signal containing $\overline{C}$ as a factor viz. $\overline{A}\,\overline{C}$, is already present in the circuit (output of module 2). The possibility of using this signal in lieu of $\overline{C}$ is now considered as a first attempt at eliminating module 7. On examination, this is found to be possible, since $\overline{A}\,\overline{C}$ is one of the six alternative signals that can replace module 7 in the circuit.

The same signal, $\overline{A}\,\overline{C}$, can also be used to eliminate module 8.

TABLE 2

Excitation table of the function $A\,\overline{B}\,\overline{C} + \overline{A}\,B\,\overline{C} + \overline{A}\,\overline{B}\,C + A\,B\,C$

C \ AB	00	01	11	10
0	0	1	0	1
1	1	0	1	0

The designer's attention is next directed to the possibility of eliminating module 5. The output of module 3 contains a term with $\overline{B}$ as a factor, but examination of the circuit shows that this signal cannot be used to eliminate module 5. The outputs of modules 1 and 4 can be modified to include $\overline{B}$ as a factor, since both signals are fed into module 3, into which B is also fed. Therefore by feeding B into modules 1 and 4, their outputs are modified to $\overline{A}\,\overline{B}\,\overline{C}$ and $A\,\overline{B}\,\overline{C}$ respectively. But $\overline{A}\,\overline{B}\,C$ can be fed into module 6 instead of $\overline{B}$ and $A\,\overline{B}\,C$ into module 9 instead of $\overline{B}$. The final circuit, therefore, based on the minimization of Fig. 16(b) will contain 7 modules, and 20 connexions.

Fig. 16(c) was realized by considering equation (12) as representing the exclusive OR function of the signals B and $\overline{A}\,C + A\,C$. This circuit is not subject to minimization, producing a final circuit that uses 8 modules and 16 connexions.

An examination of Fig. 16(d) reveals that it cannot be minimized. Therefore a final circuit based on this circuit will contain 9 modules and 19 connexions.

Therefore, the minimal circuit realizing the functions $A\,\overline{B}\,\overline{C} + \overline{A}\,B\,\overline{C} + \overline{A}\,\overline{B}\,C + A\,B\,C$ is that based on equation(11), the final form of which is shown in Fig. 17.

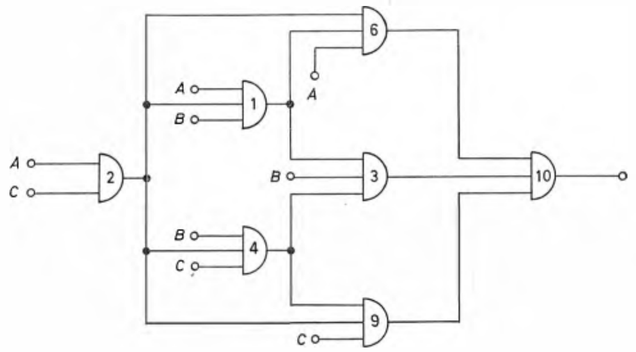


Fig. 17. Minimal circuit realization of the function $A\,\overline{B}\,\overline{C} + \overline{A}\,B\,\overline{C} + \overline{A}\,\overline{B}\,C + A\,B\,C$

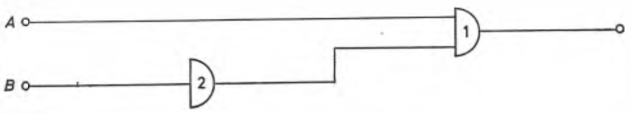


Fig. 18. Circuit realization of the function $\overline{A}\,B$

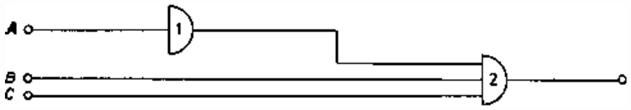


Fig. 19. A NOR circuit configuration realizing the function $A\,\overline{B}\,C$

Analysis of the Examples

The examples were selected to illustrate various points of the design procedure.

Example 1 shows that the circuits derived in step 2 may be in their minimal form.

Design example 2 has been included because it is one of the simplest circuits that illustrates the application of the minimization technique.

The third circuit shows that realizing the AND-OR form of $\overline{f}$ resulted in the minimal circuit, whereas in the fourth circuit taking common factors produced the minimal circuit design.

The fifth example illustrates that a function may have more than one minimal circuit realization.

The last circuit indicates that in certain cases a given signal may be modified to generate the signal needed to eliminate a module.

Appendix

(1) A Minimization Technique

The design technique as explained in the introduction consists of first designing non-minimal circuits and then minimizing them by eliminating redundant modules.

Module 2 in Fig. 18 would be redundant if the signal $A\,\overline{B}$ were readily available. This is easily shown to be true, since in this case the output is

$$(\overline{A}\,\overline{B})\,\overline{A} = (A+B)\,\overline{A} = \overline{A}\,B.$$

In the case of Fig. 19, module 1 would be redundant, if any of the following signals were available:

$\overline{A}\,\overline{B}$ or $A\,\overline{C}$ or $\overline{A}\,(\overline{B}+C)$ or $\overline{A}\,(B+\overline{C})$ or $\overline{A}\,(\overline{B}+\overline{C})$ or $\overline{A}\,\overline{B}\,\overline{C}$.

(2) NAND/NAND Circuit Realizations and Negative Logic

To realize a given switching function using NAND modules and negative logic, interchange the 'pluses' and 'dots' in the algebraic expression. The NOR/NOR circuit

TABLE 3

Excitation table of the function $\overline{A}\,\overline{B}\,\overline{C} + \overline{A}\,B\,C + A\,B\,\overline{C}$

C \ AB	00	01	11	10
0	1	0	1	0
1	0	1	0	0

configuration corresponding to the derived function will realize the initial function if the NOR modules are replaced by NAND modules. A useful point to keep in mind when so transforming algebraic functions is to include all product terms in brackets before the transformation. The brackets are not affected by the transformation.

Example Assume the function $A\,B + A\,C$ is to be realized using NAND/NAND circuit configuration and negative logic.

$$f = A\,B + A\,C = (A\,B) + (A\,C)$$
$$\overline{f} = (A + B)(A + C) = A + B\,C$$

Fig. 11 shows the required circuit configuration of NAND modules that realizes the function $f = A\,B + A\,C$.

(3) Classification of Logic Circuits by Index Numbers[1]

An index number may easily be derived from the minimal AND/OR expression of a given switching function by allocation +1, +2, +4, +8, ... to $A, B, C, D, \ldots$ and −1, −2, −4, −8, ... to $\overline{A}, \overline{B}, \overline{C}, \overline{D}, \ldots$ respectively. The numbers corresponding to a product term are next added together algebraically. The numbers corresponding to the product terms are next re-arranged in ascending order starting from left, the numbers being separated from each other by strokes.

Example

$$f = \overline{A} + A\,\overline{B}\,\overline{C}$$
$$= \overline{A} + \overline{B}\,\overline{C}, \text{ after simplification}$$
$$\overline{A} = -1,\ \overline{B}\,\overline{C} = -2\ -4 = -6$$
$$\therefore \text{ index number } -6/-1.$$

In some instances two different switching functions may have the same index number. For example both the functions $\overline{A}\,\overline{C} + B\,\overline{C}$ and $\overline{B} + \overline{A}\,\overline{C}$ have the same index number −5/−2. In such a case a suffix is introduced at the end of the index

$$\overline{A}\,\overline{C} + B\,\overline{C} \ \ldots\ldots\ -5/-2/a$$
$$\text{and } \overline{B} + \overline{A}\,\overline{C} \ \ldots\ldots\ -5/-2/b$$

(4) Derivation of the Complement of Switching Functions

The complement of a given switching function may be derived either by applying De Morgan's theorem or by considering the '0' entries in the excitation table of the function.

De Morgan's theorem is usually preferred if the function is simple, but otherwise the excitation table is to be preferred.

De Morgan's theorem states that to find the complement of a given function replace every letter in the original expression by its complement and interchange all the 'dots' and 'pluses'. When applying De Morgan's theorem it is advisable to simplify first the original expression and include all the product terms in brackets. The brackets are not affected by the complementing procedure.

Example

$$f = A\,.\,B + \overline{A}\,.\,C + B\,.\,C$$
$$= A\,.\,B + \overline{A}\,.\,C$$
$$= (A\,.\,B) + (\overline{A}\,.\,C)$$
$$\overline{f} = (\overline{A} + \overline{B})\,.\,(A + \overline{C})$$
$$= A\,.\,\overline{A} + A\,.\,\overline{B} + \overline{A}\,\overline{C} + \overline{B}\,\overline{C}$$
$$= A\,.\,\overline{B} + \overline{A}\,.\,\overline{C}$$

Conclusion

A method of designing minimal NOR/NAND logical circuits has been explained. The method consists of partly manipulating expressions and partly working with the actual circuit itself. It is hoped at this stage that, as engineers gain more insight into the logical design/problem of such circuits, a method limited to the manipulation of the switching functions only will evolve itself.

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A Gunn Effect Epitaxial Device for Microwave Applications

Standard Telecommunication Laboratories Ltd has developed an experimental microwave oscillator with a volume of only 0·1in³ that produces several milliwatts of c.w. power at 1Gc/s. It is the first Gunn effect device to use epitaxial construction and shows promise for use as an amplifier as well as an oscillator.

Two years ago an English scientist J. B. Gunn observed that the application of a steady electric field above a certain threshold level to low resistivity gallium arsenide caused the charge carriers to break up into domains moving along the potential gradient at the carrier drift velocity. The transit time in short enough samples is such that the resulting currents are at microwave frequencies, but to date the development for c.w. operation in the gigacycle per second region has been restricted by the problem of removal of heat from the active area.

The epitaxial construction used in the STL device enables the frequency and the threshold power of the device to be determined independently of the resistivity of the active region or the mechanical and thermal properties. It uses a substrate of semi-insulating gallium arsenide about 100 microns thick on which is grown a 15 micron thick layer of gallium arsenide whose properties are optimum for the active region of the device. The effective cross-sectional area for the current path is determined by removing part of the layer to form a narrow track 100 microns wide. The anode and cathode are formed by converting two parts of the track to n+ regions, leaving a gap of the original n type material between them, the length of which determines the self-oscillating frequency. The semi-insulating substrate is bonded to a heat sink and electrical connexions to the anode and cathode are made by micro-welding tapes to metal contacts on the top surface of the n+ regions.

The shape of the substrate determines the mechanical and thermal properties of the device. The electrical properties, determined by the dimensions of the epitaxial n type track, can be varied independently of the shape of the substrate. The configuration is suitable for mass production using precision masking techniques similar to those used for the manufacture of semiconductor integrated circuits.

Samples of these epitaxial devices have been operated at frequencies above 1Gc/s with c.w. output powers of a few milliwatts. Much high powers are possible under pulse conditions and the c.w. output power should improve with further development.

The STL team has also shown that the device can be operated in a triggered mode. By applying a biasing field of a value just below the threshold value for self oscillation, a single domain can be launched by means of a relatively small pulse that takes the field just over the threshold value. An output pulse of much larger amplitude can result and a technique of amplification at microwave frequencies using this phenomenon is under active investigation at the laboratories. A frequency modulated signal can be amplified provided that the maximum frequency does not exceed the self-oscillating frequency.

Attention is also being given to the idea of three terminal devices in which domain formation can be influenced by the potential of an isolated gate electrode similar to that in a metal oxide semiconductor transistor, with the cathode and anode in the position of source and drain respectively.

A D.C. Differential Amplifier Using Field Effect Transistors

By J. C. S. Richards*, Ph.D., A.M.I.E.E.

Field effect transistors biased to give low temperature drift form the input stage of a stable differential amplifier having a bandwidth of 100kc/s, a rejection ratio of more than 5 000, and a differential gain stabilized at 100. The zero drift is less than 100μV/°C, and the input current less than 10^{-9}A. The usefulness of the amplifier is limited by the low frequency noise generated in the transistors (about 50μV r.m.s.).

(Voir page 635 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 642)

IT has been shown that field effect transistors (f.e.t.) can be used to make single ended d.c. amplifiers having high input impedance and low voltage drifts (<50μV/°C say). For many applications differential amplifiers are necessary, and the present article discusses how these may be designed using f.e.t. in the input stage.

The two main problems in such amplifiers are to keep the drifts small (<100μV/°C, say) and to make the rejection ratio sufficiently high (>5000, say). A single ended output is usually necessary, and in that case a third difficulty arises in applying negative feedback to stabilize the gain.

Temperature Drifts in F.E.T.

The temperature drift[1,2] in a f.e.t. operated in the pinch-off region may be made positive or negative by adjusting the gate bias V_g to be greater or less than the value V_{gT} at which temperature drifts are negligible. Suppose the drain current I_d has the value I_{dT} and the mutual conductance g has the value g_T when the gate bias is V_{gT}; then I_{dT} is given by $g_T V_T$, where V_T is a parameter of the f.e.t. which usually lies between 0·3 and 0·8V. If the gate bias V_g differs by ΔV_g from V_{gT}, then the equivalent input drift is about $1{\cdot}2\Delta V_g/V_T$ mV/°C; if I_d differs by a small amount ΔI_d from I_{dt}, then the drift is about $1{\cdot}2\Delta I_d/I_{dT}$ mV/°C.

The basic form of amplifier to be discussed is shown in Fig. 1. By suitable choice of R_{sA}, R_{sB}, R_{ss} and V_{ss}, each f.e.t. can be biased so as to have zero temperature drift. If this is done, the temperature drift problem is in principle solved, and the problem of high rejection ratio can be tackled independently.

In differential amplifiers using bipolar transistors (or valves) the usual way of avoiding excessive drifts is to match the two halves of the differential pair. This can be done with f.e.t.[3], but the manufacturers' tolerances on f.e.t. characteristics are at present so large that it is difficult and expensive to choose suitable pairs. The present technique offers some minor advantages: small differences of temperature between the two f.e.t. are not important, and small temperature drifts elsewhere in the amplifier can be compensated by changing the current in one of the input f.e.t.

Rejection Ratio of Simple Circuit

For the purposes of calculation it is assumed that a f.e.t. can be adequately represented by a mutual conductance g, and that the drain slope resistance ρ is effectively infinite. This is nearly true for ZFT12, though not for all f.e.t. It is also assumed that the succeeding stages of the amplifier have a large rejection ratio and a low in-phase gain, and that the relative values of R_{dA} and R_{dB} do not greatly affect the output, so that Z_A is nearly equal to Z_B.

If R_{ss} and V_{ss} in Fig. 1 are made very large, so that they act as a current source, the overall rejection ratio

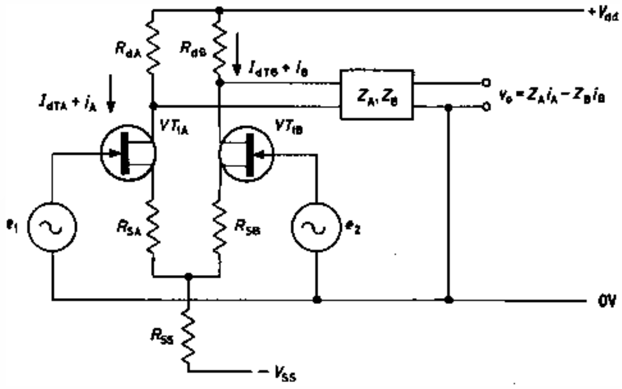


Fig. 1. Basic differential amplifier

would be infinite if the above assumptions were exact. Experimentally a rejection ratio of over 300 is obtained even with badly matched components. This is about a factor of 10 too small, and some means of balancing the circuit is required.

A More Elaborate Circuit

The circuit shown in Fig. 2 can be adjusted to have a very large rejection ratio, and the gain is established by feedback. This circuit can be developed from that of Fig. 1 by first adding R_1, R_2, R_3 and R_4, with one end of R_4 taken to earth instead of to the output of the amplifier. R_1 and R_2 are approximately equal, and the rejection ratio is adjusted by altering the relative values of R_3 and R_4.

Although R_3 and R_4 are much larger than R_1 or R_2, they may be small enough to make the in-phase gain appreciable and to make the rejection ratio too dependent on the mutual conductances of the f.e.t. To avoid this, an auxiliary amplifier feeds into the junction of R_1 and R_2 a current proportional to any change $(i_A + i_B)$ in the total drain current of the two f.e.t. The negative feedback loop thus formed tends to hold the total f.e.t. current constant and has the same effect on the in-phase gain as increasing R_3 and R_4 by a factor $(A_1 + 1)$, where $A_1(i_A + i_B)$ is the output of the amplifier. This technique has been fully discussed in connexion with differential amplifiers employing thermionic valves[4] and bipolar transistors[5].

If one end of R_4 is now taken to the output of the main amplifier instead of to earth, overall feedback can be applied without affecting the rejection ratio A_r.

The exact expressions for A_r and the differential gain

* *The University of Aberdeen.*

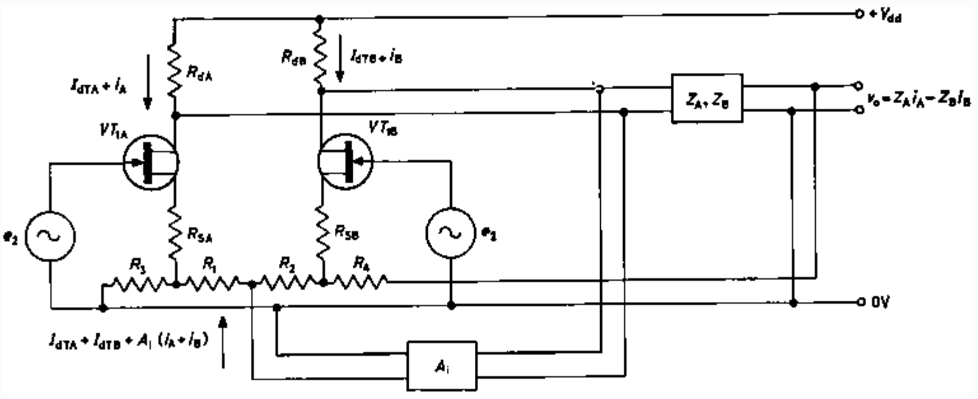


Fig. 2. Differential amplifier with adjustable rejection ratio and stabilized gain

A_d are rather lengthy, and the less important terms are simplified by assuming that $R_3 \gg R_1$, $R_4 \gg R_2$, $g_1 R_3 \gg 1$, and $g_2 R_4 \gg 1$. It is convenient to write r_A for $(1/g_A + R_{5A})$, r_B for $(1/g_B + R_{5B})$, $2R_1'$ for $(R_1 + R_2)$, $2/R_3'$ for $(1/R_3 + 1/R_4)$, $2Z_A'$ for $(Z_A + Z_B)$ and $2r_A'$ for $(r_A + r_B)$.

The rejection ratio A_r is then given by

$$1/A_r = (R_1'/R_4)\left((R_4/R_3) - (Z_B/Z_A)\right) + \frac{A_i R_1}{(1+A_i)R_3'}\left((Z_B/Z_A) - (R_2/R_1)\right) + \frac{r_A}{(1+A_i)R_3'}\left((r_B/r_A) - (Z_B/Z_A)\right)$$

The differential gain A_d is given by

$$-(1/A_d) = \frac{A_i Z_A' R_2 + Z_B R_1' + \frac{1}{2}(Z_B r_A - Z_A r_B)}{Z_A(A_i + 1)R_4} + \frac{R_1' + r_A'}{Z_A'}$$

The condition for a large overall feedback factor is $Z_A \gg R_4$, and in this case the second term in the expression for A_d becomes negligible. If $A_i \gg 1$, so that the total current in the first stage is sensibly constant, and $Z_B \simeq Z_A$, then A_d is R_4/R_2. The differential gain depends only slightly on R_3, so that if R_4 is first adjusted to give the required gain, R_3 may then be adjusted to give a good rejection ratio without appreciably altering the gain.

Dependence of A_d on A_i

If the current feedback loop associated with the first stage is omitted so that $A_i = 0$, then A_d becomes approximately $2R_4/(R_1 + R_2 + r_A - r_B)$ instead of R_4/R_2. This result, which is perhaps a little surprising, can be understood from the small signal equivalent circuits given in Fig. 3. In these it is assumed that $R_3 \gg r_A$, $R_4 \gg r_B$.

In Fig. 3(a) A_i is zero. If $Z_B(\simeq Z_A)$ is effectively infinite, but v_o is finite, then i_A must be equal to i_B. It follows that

$$v_o = -2R_4 i_A = 2(e_2 - e_1)R_4/(R_1 + R_2 + r_A - r_B)$$

In Fig. 3(b) if A_i is assumed infinite, then $(i_A + i_B)$ must be zero. Since infinite Z_A constrains $(i_A - i_B)$ to be zero, both i_A and i_B must be zero. Hence $v' = e_1$, and $v'' = e_2$, and it immediately follows that

$$v_o/R_4 = (v'' - v')/R_2 = (e_2 - e_1)/R_2$$

as required.

Practical Circuit

A complete circuit embodying the above principles is given in Fig. 4. To hold the total current in VT_{1A} and VT_{1B} constant, their mean drain potential, which is essentially the potential of the emitters of VT_{2A} and VT_{2B}, is fed via R_7 to the base of VT_{5B}. The collector current of VT_{5B} is almost exactly equal to the total source current of VT_{1A} and VT_{1B}. This negative feedback loop has a current gain A_i of about $\frac{1}{2}\beta_5 R_d/R_5$ where R_d is the average value of R_{dA} and R_{dB}, and β_5 is the average current gain of VT_{5A} and VT_{5B}.

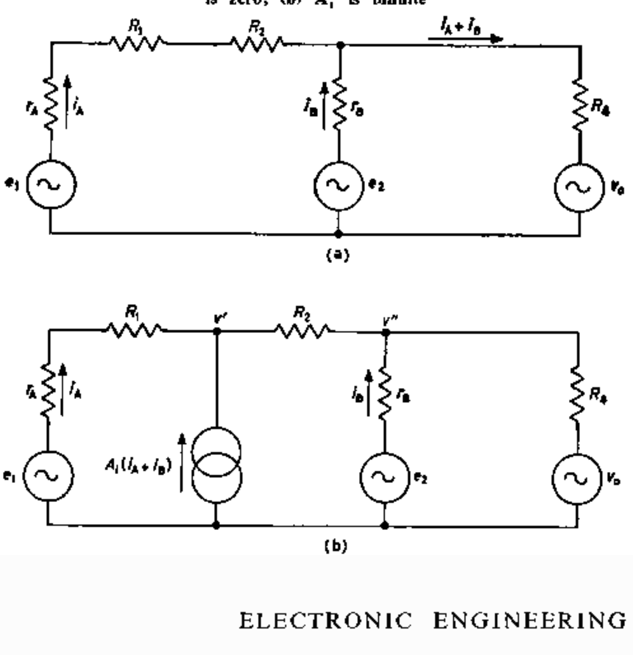
Strictly speaking this feedback loop tends to hold constant the potential between the VT_1 drains and the negative power supply line. Since the VT_1 current is determined by the potential between the VT_1 drains and the positive power supply line, the value of VT_1 current is very sensitive to power supply variations unless these are balanced out in some way. The Zener diode MR_1, resistors R_5 and R_6, and the transistor VT_{5A} effectively do this. The diode MR_2 compensates for temperature variations in MR_1, and MR_3 compensates for changes in the base-emitter potential of VT_2. With this system power supplies stable to 0·1 per cent are adequate.

Drifts

If their currents are set correctly, drifts associated with temperature changes in VT_{1A} and VT_{1B} should be small ($<30\mu V/°C$). Any subsequent change ΔI_{5B} in the collector current of VT_{5B} is unlikely to change the working point of VT_{1A} and VT_{1B} sufficiently to make them temperature sensitive, but directly produces an equivalent input voltage change of about $\Delta I_{5B}(r_B - r_A)$. To keep such drifts sufficiently small, VT_{1A} and VT_{1B} must be selected so that $(r_B - r_A)$ is less than 1kΩ, say. This matching condition is easily met. Out of a sample of eight ZFT12, five gave values of r which differed by less than 1kΩ; another two could be satisfactorily paired. Only one transistor could not be matched even in this small sample.

Gate leakage currents at 20°C are of the order of 10^{-9}A which is comparable with the grid current of many thermionic valves. With a 200kΩ source impedance this

Fig. 3. Equivalent small signal circuits for estimating gain when (a) A_i is zero, (b) A_i is infinite



implies a voltage drift of less than $20\mu V/°C$ at 20°C even if the two gate currents are dissimilar.

Drift in the second stage transistors VT_{2A} and VT_{2B} is not likely to have serious effects, provided these are selected to have current gains of at least 80, and provided $(R_{dA} - R_{dB})$ is not too large ($<5k\Omega$, say). This latter condition is usually satisfied if the f.e.t. have been selected to have nearly equal values of r.

The large number of factors involved makes it difficult to assess the total temperature drift. Experimentally it is found possible to achieve drifts of $<100\mu V/°C$ fairly readily. If need be, the bias on one of the f.e.t. can be adjusted to offset drifts elsewhere in the amplifier.

Performance

The differential gain, which is over 3 000 without feedback, is set at 100 by means of R_{11}, and an output of over $\pm 5V$ to $2k\Omega$ is available. The 3dB bandwidth is about 100kc/s. With the input short-circuited, the noise referred to the input is about $50\mu V$ r.m.s. (or about $300\mu V$ peak-to-peak); this implies an equivalent noise resistance of about $2M\Omega$. About three-quarters of the noise consists of frequencies below 1kc/s, so that reducing gain at high frequency is not very helpful in reducing noise.

The rejection ratio can easily be made greater than 5 000 at low frequencies by adjusting R_8, and if C_1 is correctly set it does not fall below 1 000 at 100kc/s. In-phase signals of up to 2V peak-to-peak can be handled; appreciably larger in-phase inputs give an output which is largely second harmonic.

Overall sensitivity can conveniently be varied by about 20 to 1 by varying the sensitivity of the oscilloscope or recorder. To allow the amplifier to handle a wide range of both in-phase and differential signals, the very simple decade attenuator shown in Fig. 5 can be used. It reduces the overall bandwidth to about 50kc/s, and a rejection ratio of over 3 000 can be obtained at frequencies up to 10kc/s by adjustment of the trimmers. At frequencies greater than 30kc/s the input capacitance of the amplifier is appreciably frequency dependent, and the rejection ratio is unlikely to be more than 1 000 at 30kc/s.

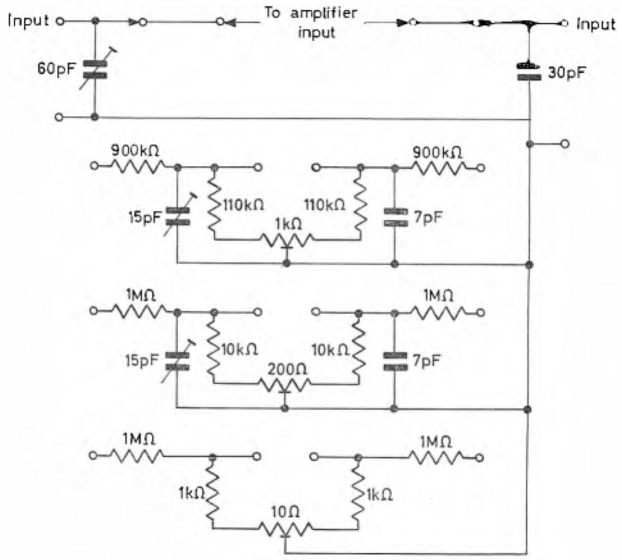


Fig. 5. Simple decade attenuator

Because of the relatively large value (about 30pF) of the ZFT12 gate to source capacitance, the high frequency performance of the amplifier is adversely affected unless the signal source impedance is less than $100k\Omega$ if it is purely resistive, or has a capacitive component of more than 50pF. The cables connecting the source to the amplifier will normally have a capacitance of more than 50pF —provision for balancing cable capacitance is made at the attenuator.

Fig. 4. Complete circuit of differential amplifier

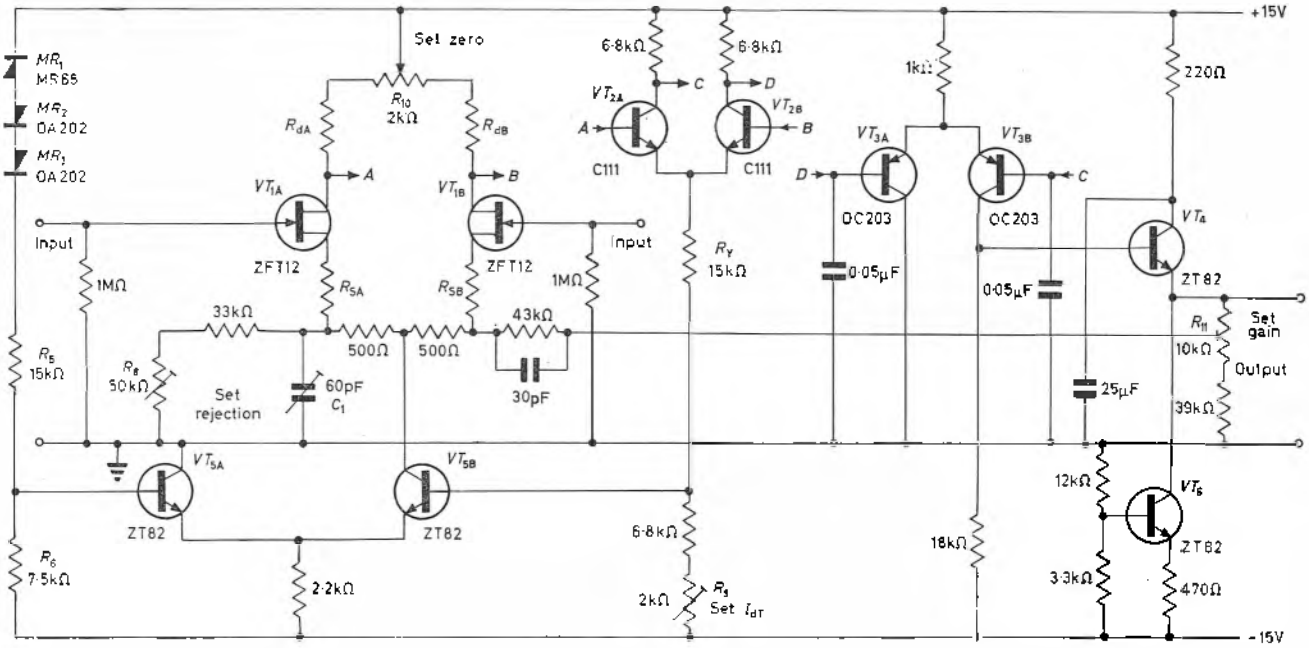
Unmarked components are given by

$$(R_{sA} + 500)\, I_{dTA} - (R_{sB} + 500)\, I_{dTB} = (V_{gTA} - V_{gTB})$$

Either R_{sA} or R_{sB} can be made zero

$$(R_{dA} + 1\,000)\, I_{dTA} = (R_{dB} + 1\,000) \quad I_{dTB} = 7{\cdot}5V$$

R_{10} is a 10 turn Helipot



Conclusion

A differential amplifier using f.e.t. in the input stage can be made to drift less than a valve amplifier or a bipolar transistor amplifier of comparable input impedance. Good rejection ratio and stable differential gain are readily obtained. Preliminary testing of f.e.t. to find the values of current and bias which give zero temperature drift is necessary, but very close matching of characteristics is not essential.

Compared with an amplifier using suitably chosen valves, the amplifier described has the disadvantages of a high noise level at low frequencies, and a lower input impedance; for these reasons it is unsuitable for refined biophysical measurements. It is to be hoped that these defects will be much reduced as f.e.t. are further developed. Meanwhile for many purposes they are small disadvantages to set against compactness, short warm-up time, freedom from microphony and the ability, if need be, to run from primary batteries.

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A Delay Device and A.C. Logic Switch in the Resistor-Capacitor-Coupled-Transistor-Logic

By G. Philokyprou*, Ph.D.

The resistor-capacitor-coupled-transistor logic system of circuit logic introduced by P. L. Cloot[1,2] possesses many attractive features for low speed logic. The basic circuit of this system, when connected in the a.c. configuration (input through capacitor only) behaves as a delay device. At the same time by making conventions for the 0 and 1 states, the same circuit can be used as an a.c. logic switch. In the following the delay (monostable) character of the basic circuit is explained and the associated delay time is calculated, while the switching properties are exploited for the construction of AND *and* OR *gates. The delay and a.c. switching properties explained make this system of circuit logic much more powerful and flexible.*

(Voir page 635 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 642)

THE basic circuit is shown in Fig. 1(a). Input to this circuit can be effected in two different ways. The first corresponds to the case when the input is through C and R_B connected in parallel (Fig. 1(b)), and the second when the input is through C with R_B returned to the negative supply or earth (Fig. 1(c)). In the first case, in which the coupling is d.c., the basic circuit will be called 'd.c. basic circuit', while in the second, in which the coupling is a.c., the basic circuit will be called 'a.c. basic circuit.'

Basic circuits of the above configurations can be connected together to form a computer system. For this reason the behaviour of the basic circuit of each configuration must be examined in conjunction with the basic circuit to which its input is connected. The one will be called the 'driving', and the other the 'driven' circuit. P. L. Coot[1,2] has only examined the case when both the driving and driven circuits are in the d.c. configuration wherefrom he derived his first design rule. Of the remaining combinations of driving and driven circuits the ones that present practical interest are examined below.

Driving Circuit: *D.C. Basic Circuit.*

Driven Circuit: *A.C. Basic Circuit with R_B Returned to the Negative Supply.*

In this case with suitable values of R_A, R_B, and R_C the driven circuit is always in the 'on' state. The driven circuit will be affected only when the driving circuit produces a positive voltage step at its output. In this case, if the value of capacitor C is suitably chosen, the driven circuit will be cut off for a time t_d depending on the transistor parameters and component values. The driven circuit, therefore, behaves as a monostable device which changes its state only when positive voltage steps are applied to its input.

* *Nuclear Research Centre "Democritus", Athens, Greece.*

Suitable values for C can more easily be found by using the charge control parameters[3,4] of the transistor. The condition which must be satisfied in order to cut off the driven transistor is:

$$CV \geqslant Q_{OFF} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

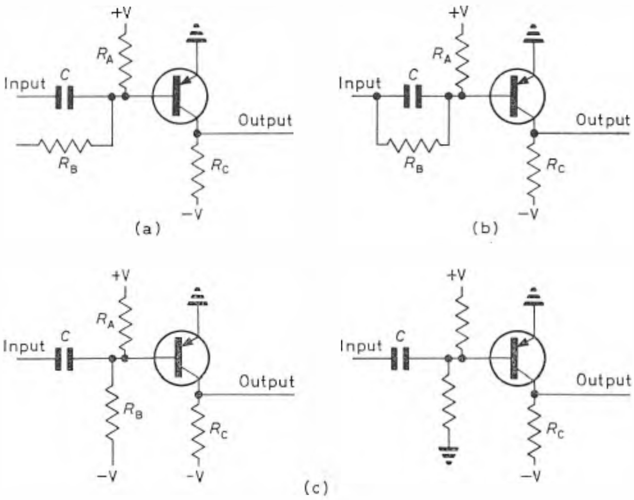


Fig. 1. The basic circuit configurations
(a) The basic circuit
(b) The d.c. basic circuit
(c) The a.c. basic circuit

where Q_{OFF} is the charge which must be removed from the base of a saturated transistor in order to cut it off, and V the positive voltage step applied to C.

Calculation of the Delay Time

Once condition (1) is satisfied the delay time t_d of the circuit can be calculated. Relation (1) may be written in he form:

$$CV = Q_{off} + CB_B \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

vhere V_B is the voltage to which the base of the transistor vill be raised by the positive step; V_B will decay exponen-ially with time as the equivalent circuit of Fig. 2 shows

$$\left(\text{where: } R = \frac{R_A R_B}{R_A + R_B} \text{ and } V' = V\frac{R_A - R_B}{R_A + R_B}\right.$$

\s long as $V_B \geqslant V_{BE(sat)}$ the transistor will be in its off tate. When V_B drops below $V_{BE(sat)}$, base current will tart flowing through R (Fig. 2) and when a charge[3,4] Q_{on} iccumulates in the base region, the transistor will revert o its on state. Fig. 3 shows the voltage, V_c, at the collector; $_{d1}$ is the time required for V_B to reduce to $V_{BE(sat)}$ ($\simeq 0$), vhile t_{d2} is the time required for Q_{on} to accumulate in the

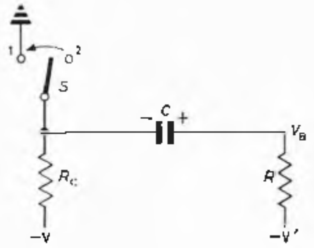


Fig. 2. Equivalent circuit for a positive voltage step applied to an a.c. basic circuit

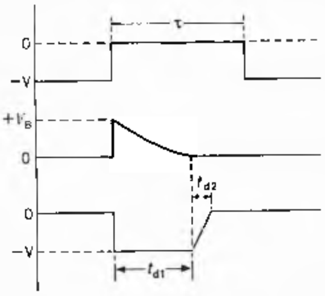


Fig. 3. Waveforms for an a.c. basic circuit with $t_{d1} < \tau$

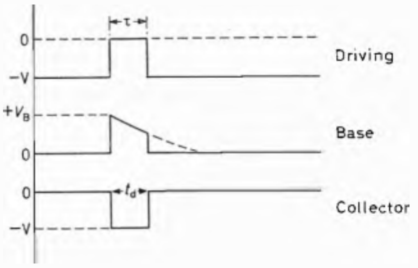


Fig. 4. Waveforms for an a.c. basic circuit with $t_{d1} < \tau$

ɔase region. From the equivalent circuit of Fig. 2, t_{d1} and $_{d2}$ are found as:

$$_{d1} = 2{\cdot}3\frac{R_A R_B}{R_A + R_B}C\log\left[1 + \frac{R_A + R_B}{R_A - R_B}(1 - (Q_{off}/CV))\right] \quad \ldots\ldots\ldots\ldots \quad (3)$$

ınd:

$$t_{d2} = \frac{R_A R_B}{R_A - R_B}[(Q_v/V) + (\tau_c/R_c)] \quad \ldots\ldots\ldots \quad (4)$$

Relations (3) and (4) are valid only when $t_{d1} < \tau$, where · is the time during which the driving transistor is in the ɔn condition (Fig. 3). If t_{d1}, as calculated from relation (3), s $t_{d1} \geqslant \tau$ (Fig. 4), V_B will become zero in time τ; relation 3) therefore is not valid. t_{d2} will also be smaller since the ıegative step at the input of the basic circuit will con-ribute to the action of R in accumulating Q_{on} in the base ·egion, and thus (Fig. 4):

$$t_d \simeq \tau \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (5)$$

Maximum Number n' of Driven Basic Circuits

The condition for maximum n' may be found by con-sidering a negative voltage step applied at the input of :he driven basic circuits. The equivalent circuit for this :ase is that of Fig. 5(a) with the switch S moving from ɔosition 1 to 2. Potential V_c will fall exponentially with time. The value $V_c\tau'$ which V_c will reach until the moment when the switch S will be closed is:

$$V_c\tau' = -V[1 - \exp(-(\tau'/n'CR_c))] = -kV \quad \ldots \quad (6)$$
$$0 \leqslant k \leqslant 1$$

where τ' is the duration of the negative part of the driving waveform. The positive step at the input of the n' driven circuits, therefore, will be of amplitude kV, and if this value satisfies relation (1) the n' basic circuits will be cut off: otherwise the driving action will fail. Relations (1) and (6) give:

$$n' \leqslant \frac{\tau'}{2{\cdot}3\,CR_c\log\dfrac{1}{1 - (Q_{off}/CV)}} \quad \ldots\ldots\ldots\ldots \quad (7)$$

From this it is seen that n' is an increasing function of C and therefore one should arrive at the conclusion that a large value of C is required. A large value of C, however, will have as a result that t_{d1} of relation (3) will be large whenever τ' is large or n' small (or when both these happen); this means that in a computer system the delay time of the delay device will not be uniform. For this reason, when choosing the switching transistor and the component values for the basic circuit, it is more reason-able to have the difference ($CV - Q_{off}$), i.e. the delay time, fixed first and then proceed investigating the conditions under which the value of n' will be maximum without affecting the delay time. It is advisable to accept $k_{min} = 0{\cdot}9$ and solve equation (6) for n':

$$n' \leqslant \tau'/2{\cdot}3\,CR_c \quad \ldots\ldots\ldots\ldots\ldots \quad (8)$$

in which case the delay time will not be affected by more than 10 per cent when n' is maximum.

From relation (8) is is seen that n' may be improved by having small values of C, which in turn means that Q_{off} must be small. Q_{off} now depends on the values of R_A, R_B, R_C and the transistor charge control parameters[3,4] β_o, Q_v, τ_s and τ_{co} and, since R_A, R_B, R_C and β_o will be fixed by taking into account other factors[1,2] except the ones in dis-cussion here, it is clear that n' may be improved by choos-ing transistors having small values of Q_v, τ_s and τ_{co}.

Relation (8) is a new design rule for resistor-coupled transistor logic.

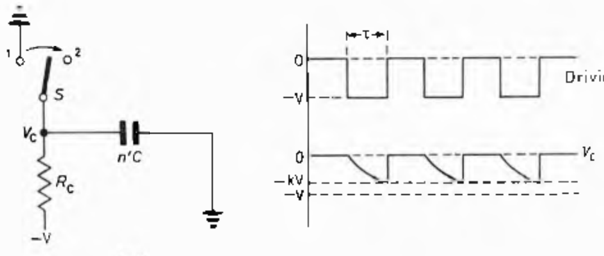
Both Driving and Driven Circuits A.C. with R_B returned to the Negative Supply

This case becomes identical with that previously examined when the duration of the pulse at the input of the driven circuit is the one given by relation (3). The con-dition that this driving can be effected may be found by combining relations (1), (3) and (5). This is given by:

$$n^* \leqslant -\frac{R_A R_B\log\left[1 + \dfrac{R_A + R_B}{R_A - R_B}(1 - (Q_{off}/CV))\right]}{R_o(R_A + R_B)\log(1 - (Q_{off}/CV))} \quad \ldots\ldots\ldots\ldots \quad (9)$$

where n^* is the number of driven circuits. If $n^* < 1$ the

Fig. 5. Equivalent circuit for a negative voltage step applied to n' a.c. basic circuits and the integration effect at their input



(a) (b)

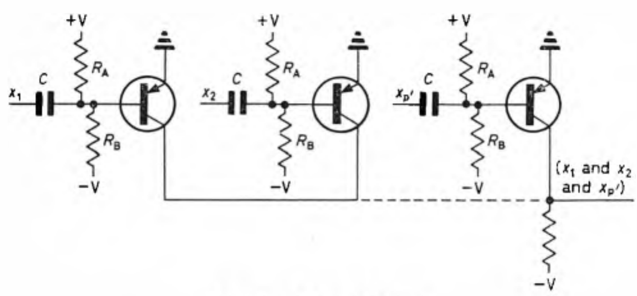


Fig. 6. A.C. AND gate

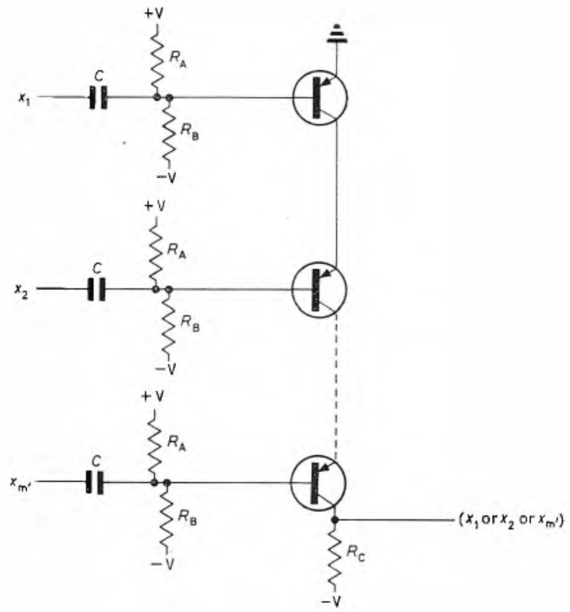


Fig. 7. A.C. OR gate

driving will not be possible which means that a.c. basic circuits cannot be cascaded. If on the other hand $n^* > 1$ advantage can be taken of the driving circuit under discussion to introduce in a system a delay longer than t_d of equation (3). Delay times are added in this way which means that delay times equal to $n_1 t_d$ can be introduced, where n_1 is the number of cascaded basic circuits.

When n^* is near its maximum value as obtained from relation (9) the delay time of the driven circuits will be less than the one obtained from relation (3) owing to the integration effect already discussed. In order to avoid this, condition (9) must be imposed on n^* in which case:

$$n^* \leqslant t_d/2{\cdot}3\, CR_c \quad \ldots\ldots\ldots\ldots \quad (10)$$

Switching Properties of the A.C. Basic Circuit

So far only the delay properties of the a.c. basic circuit have been examined. In order to use this circuit as a logic switch its 0 and 1 states must be confined; these are:

(a) For the input: 1 is represented by a positive voltage step.
0 is represented by the absence of a positive voltage step.

(b) For the output: 1 is represented by the presence of a pulse.
0 is represented by the absence of a pulse.

These conventions, though they appear complicated, are compatible between themselves as well as with the conventions made for the d.c. basic circuit. In this way not only is it possible to construct gates with a.c. basic circuits (a.c. gates) but it is possible to construct mixed d.c.-a.c. gates which in many cases prove to be simple and most efficient.

A.C. AND Gates

Fig. 6 shows an a.c. AND gate having p' input terminals All p' transistors are conducting and the only case under which a pulse will appear at the output is when positive voltage steps occur at the input of all p' basic circuits. The duration of the gate output pulse is equal to the shortest delay time associated with each individual a.c. basic circuit forming the gate.

A.C. OR Gates

Fig. 7 shows an a.c. OR gate having m' input terminals. All m' transistors are conducting unless a positive voltage step is applied to the input of one or more of them. The duration of the output pulse is equal to the longest delay time associated with each individual circuit forming the gate and receiving a 1 at its input.

Maximum Number of Input Terminals to A.C. AND and A.C. OR Gates

Comparing the a.c. gates already described with the d.c. gates described by Cloot[1,2] it is seen that the a.c. OR gate corresponds to the d.c. AND gate, while the a.c. AND corresponds to the d.c. OR gate. The formulae given by Cloot[1,2] are therefore valid for the a.c. gates with the above correspondence.

Some Experimental Figures

Table 1 contains details and results from two basic circuits. The GET871 transistor was used in the one circuit and the AC151 in the other. With the first basic circuit a digital correlator was constructed while with the second a desk function calculator is under construction. The values given for t_{d1}, t_{d2}, n'_{max} and n^*_{max} are typical values obtained from over 500 basic circuits in each case. These values correspond to typical component and transistor parameter values substituted in the proposed formulae.

TABLE 1

TRANSISTOR	R_A (kΩ)	R_B (kΩ)	R_C (kΩ)	C (pF)	t_{d1} (μsec)	t_{d2} (μsec)	n'max $\tau > 8$ (sec)	n'max $\tau > 3$ (sec)	n^*max
GET871	100	18	1·8	200	2·5	0·5	10	4	4
AC151	100	22	1·8	470	8	3	4	—	4

Acknowledgments

The author wishes to thank Mr. P. L. Taylor M.A., A.M.I.E.E., for his valuable guidance during the first stage of this work; Mr. C. Lascaris, Head of the Electronics Division of the "Democritus" N.R.C., for providing the opportunity to continue the work, and Mr. D. Doukas for his many suggestions apart from the burden of experimental work which he has carried out.

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Distortion in Feedback Oscillators

By P. F. Ridler*, B.E., A.M.I.E.E.

A theory predicting the distortion in feedback oscillators is developed, relating the distortion in the output waveform to the distortion in the amplifier section of the oscillator. A practical circuit having waveform distortion of less than 0·01 per cent is given supporting the theory.

(Voir page 635 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 642)

THE usual analysis of a Wien bridge oscillator considers the (almost) balanced output of a Wien bridge fed into a dual input amplifier, the output of this amplifier being fed back as the input of the bridge (Fig. 1). One of the arms of the bridge is usually a temperature sensitive resistor for stabilization purposes. While this analysis is perfectly valid, it is felt that it does not show up, as clearly as might be desirable, the relationship between the oscillator waveform distortion and the distortion in the amplifier section alone, and that generally the present analysis is clearer and simpler.

If an amplifier (Fig. 2), which has a gain $A(\omega)$ at any

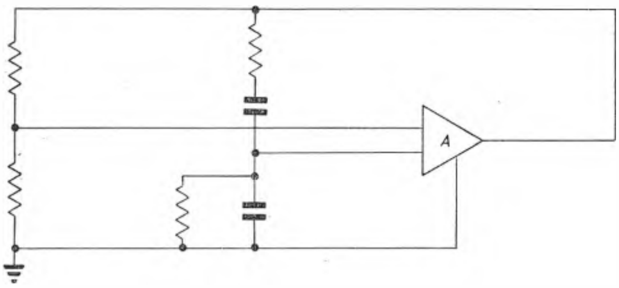

Fig. 1. Wein bridge oscillator drawn with amplifier having dual input

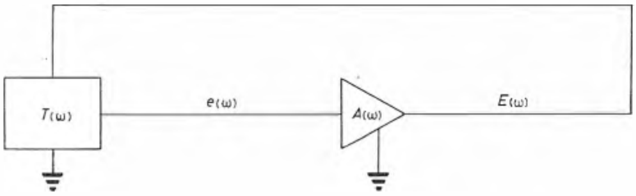

Fig. 2. Amplifier/oscillator with three terminal feedback network

frequency ω, receives an input voltage $e(\omega)$ (which may be a complex wave containing several frequencies in harmonic relationship) then the output voltage $E(\omega)$ will be:

$$E(\omega) = A(\omega)\,e(\omega) + \sum_{2}^{\infty} D(\omega_n) \quad \ldots\ldots\ldots \quad (1)$$

where the term $\sum_{2}^{\infty} D(\omega_n)$ represents those distortion product terms which are generated in the amplifier. No account is taken of intermodulation terms as any amplifier used for this purpose would be sufficiently linear that intermodulation terms would be of considerably lower magnitude than harmonic distortion terms.

If this output voltage is now fed back to the input of the amplifier through a network having a ratio of output to input voltage $T(\omega)$ then the input voltage to the amplifier will be:

$$e(\omega) = T(\omega)\,E(\omega) \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

* *Salisbury Polytechnic, Rhodesia.*

Substituting equation (2) into equation (1):

$$E(\omega) = T(\omega)\,A(\omega)\,E(\omega) + \sum_{2}^{\infty} D(\omega_n) \quad \ldots\ldots \quad (3)$$

but one may write for this output voltage:

$$E(\omega) = E(\omega_1) + \sum_{2}^{\infty} E(\omega_n) \quad \ldots\ldots\ldots\ldots \quad (4)$$

where $E(\omega_1)$ is the output at fundamental frequency and the term $\sum_{2}^{\infty} E(\omega_n)$ represents the harmonic (distortion) content of the output waveform.

$$\therefore E(\omega_1) + \sum_{2}^{\infty} E(\omega_n) = T(\omega)\,A(\omega)\,E(\omega) + \sum_{2}^{\infty} D(\omega_n) \quad \ldots\ldots \quad (5)$$

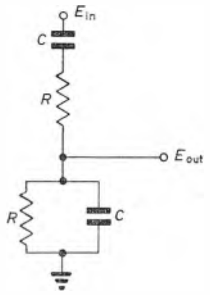

Fig. 3. ' Wein network ' or reactive side of Wein bridge

$$= T(\omega_1)\,A(\omega_1)\,E(\omega_1) + \sum_{2}^{\infty} T(\omega_n)\,A(\omega_n)\,E(\omega_n) + \sum_{2}^{\infty} D(\omega_n) \quad \ldots\ldots\ldots\ldots\ldots \quad (5a)$$

where the fundamental terms have been separated from the harmonic terms.

Now for stable oscillation at frequency ω_1:

$$T(\omega_1)\,A(\omega_1) = 1 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (6)$$

$$\therefore E(\omega_n) = T(\omega_n)\,A(\omega_n)\,E(\omega_n) + D(\omega_n) \quad \ldots\ldots\ldots\ldots \quad (7)$$

yielding:

$$E(\omega_n) = \frac{D(\omega_n}{1 - T(\omega_n)\,A(\omega_n)} \quad \ldots\ldots\ldots\ldots \quad (8)$$

This last equation shows that there is a direct relationship between the distortion of the amplifier section of the oscillator and the distortion of the oscillator output waveform, the relationship containing the transfer function of the frequency determining network. If the distortion of the amplifier can be measured and the transfer function of the feedback network is known, then the distortion of the oscillator waveform can be predicted.

If the feedback network is that shown in Fig. 3, called the 'Wien network' for convenience, then its transfer function will be (see Appendix):

$$T(\omega) = \frac{1}{3 + j(\omega/\omega_1 - \omega_1/\omega)} \quad \ldots\ldots\ldots \quad (9)$$

and for the n^{th} harmonic of ω_1:

$$T(\omega_n) = \frac{1}{3 + j(n - (1/n))} \quad \ldots\ldots\ldots \quad (10)$$

In the mid-frequency range of a variable frequency oscillator, the gain of the amplifier will be the same for the fundamental and for the lower order harmonics, and in the case of the Wien oscillator this gain will be 3 times for stable oscillation. This gives a ratio of oscillator distortion to amplifier distortion:

$$E(\omega_n)/D(\omega_n) = \frac{1}{1 - 3T(\omega_n)}$$

$$= \frac{1}{1 - \dfrac{3}{3 + j(n - (1/n))}} \quad \ldots\ldots \quad (11)$$

and:

$$\frac{|E(\omega_n)|}{|D(\omega_n)|} = \sqrt{\left[1 + \frac{9}{(n + (1/n))^2}\right]} \quad \ldots\ldots \quad (12)$$

Table 1 gives the numerical ratio for the first few harmonics.

TABLE 1

n	2	3	4	5
$\dfrac{\lvert E(\omega_n)\rvert}{\lvert D(\omega_n)\rvert}$	2·25	1·50	1·28	1·18

A transistor oscillator was designed to verify this analysis and the circuit diagram is given in Fig. 4. The output stage VT_3 was designed to draw from the supply minimum direct current consistent with delivering 2·2V r.m.s. to the feedback network, amplitude stabilizing thermistor, and a 600Ω external load. This voltage was arbitrarily chosen to suit a monitored attenuator which was available. The common collector configuration is used for low distortion and a transistor VT_4 is used as a high incremental impedance emitter load to minimize the required supply voltage.

The driver stage is designed in similar fashion, while the first stage is run at a current which will ensure low noise operation. Negative feedback through the thermistor from output to first stage emitter keeps the gain at that necessary to maintain stable oscillation, while also serving to raise the input impedance so that the frequency determining network is not unduly loaded.

The distortion of this oscillator over the range 200c/s to 10kc/s is less than 0·01 per cent as measured on a Hewlett-Packard 302A wave analyser, but unfortunately equipment was not readily available to cover the rest of the range of 1·5c/s to 150kc/s.

Quite obviously, lower distortion figures than these could have been obtained by a different choice of feedback network, but the over-riding considerations were ease of construction and economy of components, the final article being destined to be built for use in a teaching laboratory.

Measurements to verify the ratios given by equation (12) have been made for the 2nd and 3rd harmonics of 1 600c/s, and are in good agreement with the theory. Low distortion filters were not available to reduce source oscillator distortion to below the level of that in the amplifier, so that parallel-T resistance capacitance networks had to be used for this purpose, as well as to reduce the fundamental component at the wave analyser input which otherwise would overload its input amplifier. Nevertheless, agreement was within the limits of experimental error (4 per cent) and it is felt that the analysis is correct, only time and teaching load precluding further verification.

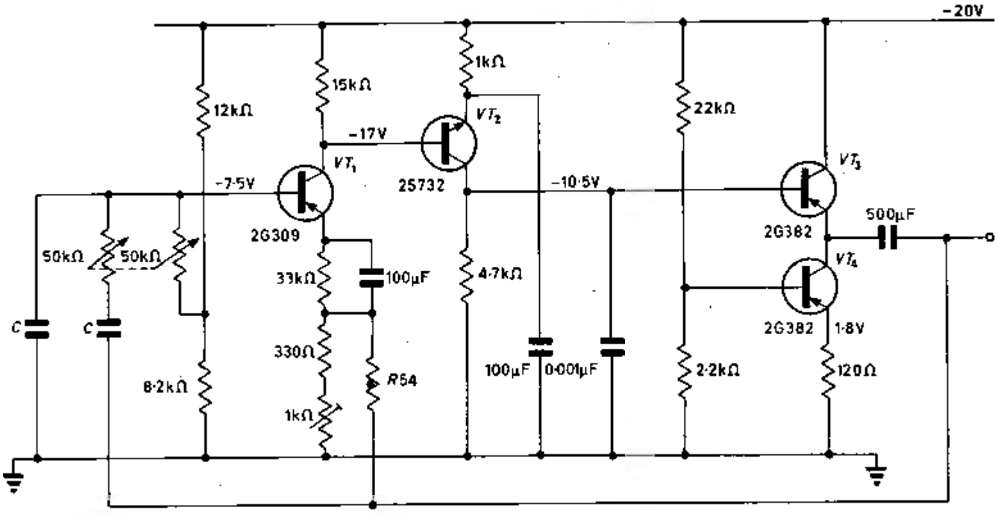

Fig. 4. Arrangement of low distortion (0·01 per cent) variable frequency audio oscillator

Acknowledgments

The author wishes to acknowledge the assistance of Messrs. N. A. Lee and B. D. Redmile who carried out the distortion measurements.

Appendix

Transfer Function of the 'Wien Bridge' (Fig. 3)

$$E_{out}/E_{in} = \frac{R\,(1/j\omega C)}{R + (1/j\omega C)} \div \left[\frac{R\,.\,(1/j\omega C)}{R + (1/j\omega C)} + R + (1/j\omega C)\right]$$

$$= \frac{R}{1 + j\omega T} \div \left[\frac{R}{1 + j\omega T} + \frac{1 + j\omega T}{j\omega C}\right]$$

where $T \equiv RC$

$$= \frac{j\omega T}{j\omega T + (1 + j\omega T)^2}$$

Let $\omega_1 T = 1$, then:

$$E_{out}/E_{in} = \frac{j\,(\omega/\omega_1)}{j\,(\omega/\omega_1) + (1 + j\,(\omega/\omega_1))^2}$$

$$= \frac{1}{3 + j\,(\omega/\omega_1 - \omega_1/\omega)}$$

Silicon Field-Effect Transistors

By R. Cullis*, B.Sc.

The proliferation of types of field-effect transistors (f.e.t.) now available may have caused some confusion in the minds of would-be users. In this article the fabrication of commonly encountered variants of silicon f.e.t. is described, and the properties associated with the different structures outlined in the hope that some of the obscurities of the new component may be resolved. Finally an indication is given of the future developments which are likely to occur in this new branch of transistor engineering.

(Voir page 635 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 642)

THERE has been a recent revival of interest in the field-effect transistor although it appears improbable that a large commercial demand for discrete devices will arise until 1966-7. The f.e.t.'s properties of high input impedance, low noise and low power consumption make it an attractive component for incorporation in solid state

Fig. 1. Double-diffusion f.e.t.
(a) Initial n-type channel diffusion
(b) Localized p-type gate diffusion

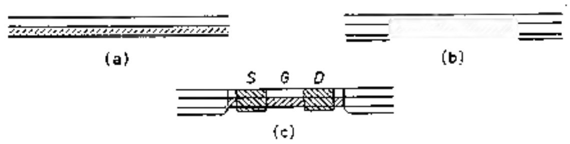

Fig. 2. Double-epitaxial f.e.t.
(a) Growth of n- and p- epitaxial layers
(b) P-type isolation diffusion
(c) N-type source and drain contact diffusion

circuits however, and so its future seems assured.

Although the field-effect or unipolar transistor was proposed by Shockley in 1952, it was not until the introduction of planar and epitaxial technologies that the production of such transistors with characteristics which were compatible with those of other semiconductor devices became practicable. Thus to see how the properties of the f.e.t. arise one must consider production methods. F.E.T.'s can be made by a number of combinations of diffusion, oxidation and epitaxial deposition processes, and it is the purpose of this article to review the different characteristics resulting from each approach.

Production Techniques

Figs. 1 to 4 show in cross-section the stages in production of various silicon unipolar transistors. Fig. 5 shows the schematic layout of a discrete device. In this, the large areas provide lands for applying contacts while the width of the narrow gate strip determines the channel length. For clarity the n-channel device (a term applied to f.e.t.'s) will be described in all cases, although the concepts apply equally to the analogue of the pnp transistor, the p-channel device.

F.E.T.'s Produced by Diffusion

The double-diffused planar f.e.t. (Fig. 1) was the first of these new types to be developed. In fabrication, an n-type diffusion is made in a p-type substrate. A p-type diffusion is then made within the n-type diffusion, forming a gate electrode and defining the channel. The substrate acts as

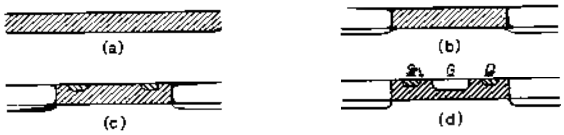

Fig. 3. Diffused-epitaxial f.e.t.
(a) Growth of n-type epitaxial layer
(b) P-type isolation diffusion
(c) N+ source and drain diffusions
(d) P-type localized gate diffusion

Fig. 4. Metal-oxide-silicon f.e.t.
(a) N+ source and drain diffusions
(b) Growth of thin oxide layer and evaporation of gate contact

a second gate electrode. An inverse bias on the two pn junctions sweeps the depletion layers associated with them across the channel, causing it to be "pinched-off."

By the nature of diffusion processes, the distribution of the dopant impurities must be asymmetric with respect to the channel. This means that the two gate electrodes will have different mutual conductance-voltage characteristics. The geometry of the electrodes required by this process also means that the magnitudes of the two gate-channel capacitances will be in a ratio of approximately 3 : 1. Despite the similarity of processing of the double-diffused f.e.t. and the double-diffused bipolar transistor, production yields of the former are much lower, since with this device the second diffusion must be stopped at a much more critical stage. This is reflected in high unit costs.

Improved Device Processing Control through Epitaxy

The development of epitaxy has provided a means of growing layers of either conductivity type on a substrate crystal while maintaining tight limits on resistivity and thickness. This has paved the way for the double epitaxial field-effect transistor (Fig. 2), a device which may be produced with much better controlled final characteristics. A thin n-type resistivity layer and a low resistivity p-type layer are grown successively on a low resistivity p-type substrate. The device area is then defined by either a p-type isolation diffusion or a mesa etching process. Finally, the

* *Standard Telephones & Cables Ltd.*

source and drain electrodes are applied by means of non-critical n-type diffusions, whose sole purpose is to make contact to the n-type channel layer and to raise the surface concentration sufficiently to make the contacts ohmic. An advantage of the double-epitaxial f.e.t. structure is that the resistances which appear in the equivalent circuit in series with the source and drain electrodes are reduced to a minimum. Since both gate electrodes can be of the same resistivity, and the impurity profile across the channel thickness is symmetric, both gate electrodes have similar mutual conductance –voltage characteristics and contribute equally to absolute value of the mutual conductance. Oscillograms of the output characteristics of a double-epitaxial f.e.t. are displayed in Fig. 3. These show the level of pinch-off voltage, mutual conductance and breakdown voltage typically achieved using this technique.

An F.E.T. Suitable for Solid State Circuits

An approach which combines features of both of the above devices is the diffused-epitaxial f.e.t. (Fig. 4). An n-type channel layer is grown on a p-type substrate and the active area is defined by a p-type isolation diffusion. In this device, too, the source and drain contacts are made by non-critical n-type diffusions, but the gate region is formed by a localized p-type diffusion.

Since diffusion processes are less controllable than epitaxial growth processes, it is slightly more difficult to make the diffused-epitaxial than the double-epitaxial f.e.t. However, the advantage is gained that the processes are compatible with those at present used in making solid-state circuits.

Manufacture of F.E.T.'s Controlled by Surface Phenomena

A second type of field effect transistor which may conveniently be incorporated in solid-state circuits is the

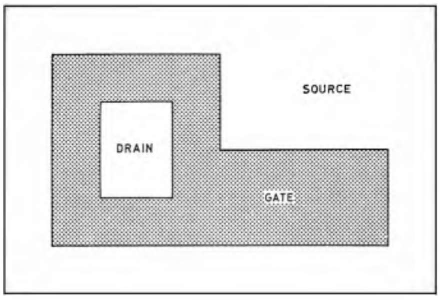

Fig. 5. Layout of typical f.e.t.

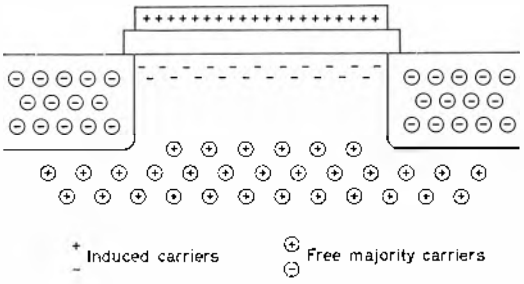

Fig. 6. Principle of operation of m.o.s.t.

metal-oxide-silicon, insulated-gate transistor, called the m.o.s.t. (Fig. 5). In contrast to the diffused and epitaxial f.e.t.'s which depend on bulk properties of the semiconductor, this device is controlled by surface effects. It has, too, a very simple construction. Source and drain are formed by heavily-doped diffusions in a lightly-doped crystal. The gate electrode is an evaporated metal contact

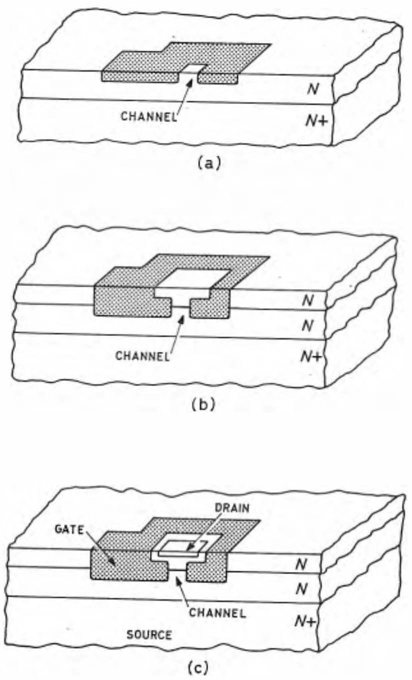

Fig. 7. High frequency f.e.t.

(a) A p-type diffusion is made in a high resistivity epitaxial layer on a low resistivity substrate. A small aperture is left for the channel

(b) A further high resistivity layer is grown and a second p-type diffusion made to contact the sub-epitaxial diffusion

(c) An n+ diffusion is made to provide and ohmic drain contact

bridging the region between source and drain and spaced from the silicon by a thin oxide layer. The principle of its operation may be most readily understood by reference to Fig. 6. A positive voltage is applied to the drain electrode, inversely biasing the junction associated with it. In the absence of any potential on the gate, the only current which will flow is the junction leakage current. However, if a positive voltage is applied to the gate, free negative charges will be induced on the surface of the crystal, and these will act as a channel carrying current from the source to the drain. Increasing the potential applied to the gate has the effect of increasing the free surface charge, and hence the source-drain current. This is referred to as operation in the enhancement mode and is another difference from the bulk f.e.t.'s which all operate in the so-called depletion mode.

A second type of m.o.s.t. also exists. By suitable control of the oxidation conditions a thin n-type layer can be formed on the surface of lightly-doped p-type silicon. This will act as a permanent channel between the source and drain. When a positive voltage is applied to the gate the channel conductance is increased, but the application of a negative voltage will induce a free positive surface charge, thus reducing the channel conductance. The permanent channel m.o.s.t., therefore, operates in both enhancement and depletion modes.

Special precautions have to be taken during the oxide formation of metal-oxide-silicon f.e.t.'s because among other things the solubilities of substances used as dopants are not the same in silicon and silicon dioxide. For example, boron tends to segregate in the oxide, depleting the doping level at the surface of a strongly p-type crystal and causing the formation of an inversion layer on high resistivity material. On the other hand, phosphorus is rejected by the oxide and so a permanent accumulation layer is formed on the surface of n-type material. This has the effect of reducing the mutual conductance of p-channel devices.

Differences in Electrical Characteristics Resulting from the Different Structures

Having outlined the structures of various field-effect transistors, it is appropriate to compare their advantages and drawbacks. There are few fundamental differences between the bulk junction devices, although it may be remarked that double-diffused f.e.t.'s have in general a low gate-channel breakdown voltage comparable with the emitter-base breakdown voltage of a double-diffused bipolar transistor, and the epitaxial-diffused f.e.t. has a slightly higher resistance in series with source and drain electrodes than the other two. However, most of the junction area is well below the surface and, since the unipolar transistor is a majority carrier device and not greatly affected by changes in minority carrier lifetime, the long term stability of bulk f.e.t.'s is excellent.

With the m.o.s.t., on the other hand, the picture is somewhat different. At the surface of a crystal the termination of the lattice structure gives rise to a change in the internal electric field. This change may be caused by preferential adsorption of ions of one sign, alignment of ions or molecules having a dipole moment or simply free bonds in the lattice. These free bonds and adsorbed ions and dipoles give rise to localized energy levels within the band structure of the semiconductor, which are referred to as surface states. Charge carriers can be transferred between the crystal and these surface states. The time taken for this process may vary from a few microseconds, if, for example, the surface state is caused by a free bond (a 'fast' state), to several hours if the state arises from an ion adsorbed in the oxide film (a 'slow' state).

As the charge in the surface states reduces the mutual conductance (g_m) of the m.o.s.t., it is clear that the g_m will vary with time under the influence of an applied potential. The surface states are affected by the previous history of the crystal and their effect is more pronounced when the crystal is lightly doped. Depletion-mode insulated-gate f.e.t.'s are made in nearly intrinsic material and so more instability is encountered with this type. Although the deleterious effects of slow states will be reduced as control of processing improves it appears unlikely that the presence of fast states, which arise from the crystalline nature of the semiconductor material, can be eliminated.

The dielectric strength of fused silica is 570/micron. As the oxide thickness commonly employed in insulated-gate f.e.t.'s is about $0{\cdot}1\mu$, precautions must be taken, in particular with mains operated equipment, to prevent voltage spikes greater than about 50V from appearing at the gate electrode, since these are liable to cause catastrophic failure of the device. With enhancement-mode devices input spikes are also to be avoided since they may cause the power dissipation to exceed the safe limit.

Insulated-gate devices have an input impedance which is typically $10^{13}\Omega$ in parallel with a capacitance of about 1pF. The d.c. input resistance of the bulk devices is of the order $10^{9}\Omega$ in parallel with a voltage-dependent capacitance of about 10pF.

Shot noise in the channel of the bulk f.e.t.'s is low and at a gate current of about 1nA partition noise is typically 0·5dB at 1kc/s. However, charge exchange between surface and bulk states is an inherently noisy process, so insulated-gate f.e.t.'s tend to have a poorer noise performance.

The mobility of charge carriers at the surface may be as little as a tenth of the value attained well within the crystal. Although this would appear to favour junction f.e.t.'s, in practice the frequency response of both types of device is limited by interelectrode capacitances.

By making use of techniques which have been developed for solid-state circuits, it is possible to produce a junction f.e.t. giving an order of magnitude improvement over existing devices. The method of fabrication is shown in Fig. 7. A high resistivity n-type epitaxial layer is grown on a low resistivity n-type substrate. A p-type diffusion is then made in this layer, leaving a small aperture which will ultimately form the channel. A second high resistivity layer is then grown and a further p-type diffusion made to contact the sub-epitaxial diffusion. Finally an n-type diffusion is made to provide an ohmic drain contact.

In this manner it is possible to make a device having an input capacitance of less than 0·09pF and a mutual conductance of 1·5mA/V, which would give a cut-off frequency of around 2·5Gc/s.

Conclusion

The trend is towards making devices from polycrystalline semiconductors. It has already been demonstrated that it is possible to make good diodes in a region of a silicon crystal enclosing several grain boundaries, and it appears probable that a field-effect transistor could be made more easily in this material than a bipolar device.

A Television Camera for Educational Use

Fifteen major educational establishments in the United Kingdom have ordered television systems based on a new series of fully transistorized closed-circuit television cameras which was introduced by The Marconi Co. Ltd at the Educational Foundation for Visual Aids Exhibition in February of this year.

The new series has been designed with the particular needs of educationists in mind. Two versions are available. The first, type V322 A, is a basic camera unit, while the second, type V322 B, has an integral high brightness viewfinder, and provides facilities comparable with a high quality broadcast equipment.

Both versions are extremely simple to operate, and the complex of controls associated with earlier television systems has been virtually eliminated, leaving only the 'on-off' switch, a focus control and iris adjustment. No additional control unit is necessary for normal operation. Once the camera has been switched on; the lens iris adjusted for satisfactory contrast and the focus control set to give the sharpest picture, the camera is completely automatic. The automatic circuits will maintain a completely stable picture and will handle changes in light level of as much as 2000:1 without attention from an operator.

A very high focusing flux is used for the vidicon tube to provide a significant increase in the resolution limit of the tube. The camera has been designed to accept the new 'separate mesh' vidicon tubes which provide improved uniformity, and enhance resolution in the corners of the picture.

The camera is fully transistorized throughout and apart from the vidicon tube and its scanning yoke, all the electronic circuits including power supplies are contained on only two printed circuit boards.

These boards are mounted on either side of the camera case, and can be hinged downwards for easy access, or detachment.

The yoke assembly for the standard 1in vidicon tube is moved in the fore-and-aft direction for focusing. This is controlled either by a three spoke capstan at the right hand side of the rear of the camera, or by a lead screw on the camera back plate. A unique method of mounting the vidicon tube ensures that it is mounted rigidly and accurately with respect to the yoke assembly.

The output from the camera may be provided either as a video signal, or in the form of modulated r.f. A choice of three adjustable, pretuned frequencies in the range 50-88Mc/s is provided.

Gain and Cut-off Frequency of a Transistor Amplifier Stag

By S. Turk*

The effect of the internal impedance of the signal source on the gain and on the frequency characteristics of an amplifier stage with transistor is determined. It is shown that the cut-off frequency obtained is always higher than ω_β in common emitter configuration and always lower than ω_α in common base configuration.

(Voir page 635 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 642)

THE conventional treatment of the amplifier stage with a transistor using the h-parameters is very useful in the frequency range where the dependence of the transistor parameters on frequency is not significant. The corresponding expressions for gain are not suitable if the cut-off frequency is determined by the transistor itself and not by the load.

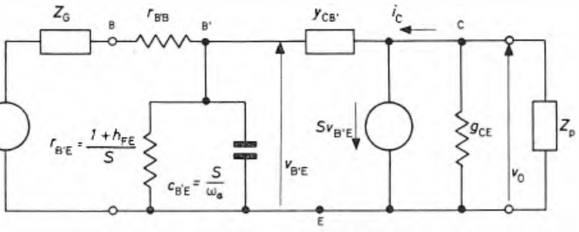

Fig. 1. The π-hybrid equivalent circuit

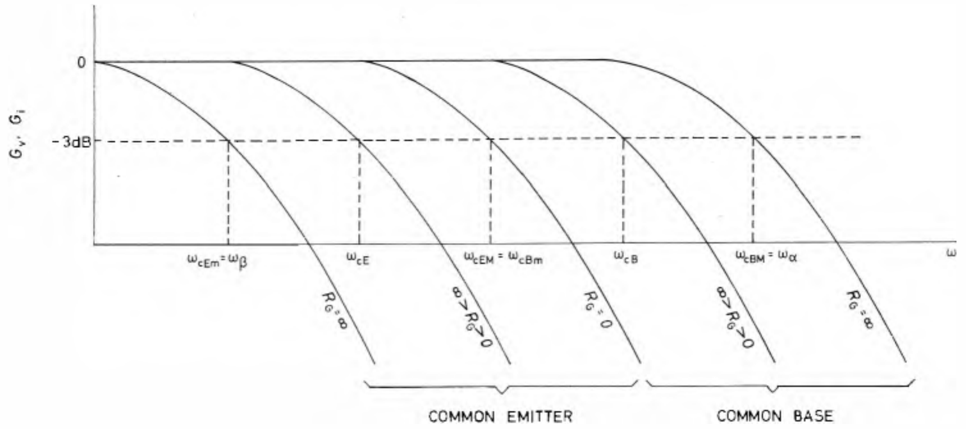

Fig. 2. Dependence of cut-off frequency on source impedance

One of the fundamental reasons why the gain is dependent on the frequency is the fact that the current amplification factor is a function of frequency. This dependence is very often given as

$$h_{fB} = h_{FB}/(1+jf/f_g) \quad \ldots\ldots\ldots\ldots \quad (1)$$

and it is a good approximation in the frequency range below the cut-off frequency f_g. The actual dependence of the current amplification factor on frequency can be described by more elaborate expressions. The actual cut-off frequency is higher and is

$$f_\alpha \simeq 1{\cdot}22\, f_g \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

The phase angle at this frequency is about 57° instead of 45°. The cut-off frequency of the current amplification factor in the common emitter configuration is lower and is

$$f_\beta = f_\alpha/(1 + h_{FE}) \quad \ldots\ldots\ldots\ldots\ldots \quad (3)$$

An amplifier in the common emitter configuration will have, therefore, its 3dB frequency lower than one in th common base configuration. The fact that h_{fB} and h_{fE} have the 3dB point at the frequency f_α, and f_β, respectively does not mean necessarily that the 3dB frequency of th amplifier will be equal to either f_α or f_β.

For the analysis of an amplifier in the region of th cut-off frequency, the π-hybrid equivalent circuit (Fig. 1 is convenient. This equivalent circuit is valid even in th h_{fB} cut-off frequency region. One of the features of th π-hybrid equivalent is the use of the transconductanc $S = dI_E/dV_{B'E} = (q/kT)I_C = 39\, I_C$mA/V. The dependenc of the transconductance on frequency is given by

$$S = S_0(1 - jf/2{\cdot}4\, f_\alpha) \quad \ldots\ldots\ldots\ldots\ldots \quad (4)$$

From equation (4) it is obvious that S can be considere as frequency independent even in the range around f_α.

Assume that the collector impedance and therefor the obtained gain are not too high, so that the Mille effect of the conductanc $y_{CB'}$, is negligible. Further more, in order to avoi dependence of the gain o the load the gain will b expressed as $G_v = i_c/v_i = Sv_{B'E}/v_i$. Including all thes assumptions it can be show that the gain in the commo emitter configuration is

$$G_{vE} = \frac{h_{FE}}{r_{B'E} + r_{B'B} + R_G} \cdot \frac{1}{1 + j\omega/\omega_{cE}} \quad \ldots\ldots\ldots \quad (5)$$

where the cut-off frequenc is given by

$$\omega_{cE} = 1/[r_{B'E} \,\|\, (r_{B'B} + R_G)]\, c_{B'E} \quad \ldots\ldots\ldots\ldots \quad (6)$$

The dependence of the cut-off frequency on the internal impedance of the signal source is readily seen. If R_G is infinite, the cut-off frequency takes its minimum value

$$\omega_{cEm} = 1/r_{B'E}\, c_{B'E} = \omega_\beta \quad \ldots\ldots\ldots\ldots \quad (7)$$

In the case $R_G = 0$ the cut-off frequency is maximum, i.e.

$$\omega_{cEM} = 1/(r_{B'B} \,\|\, r_{B'E})\, c_{B'E} \quad \ldots\ldots\ldots\ldots \quad (8)$$

The frequency ω_{cEM} is thus 10 to 30 times higher than the frequency $\omega_{cEm} = \omega_\beta$.

The gain in the common base configuration can be determined as

$$G_{vB} = \frac{h_{FE}}{r_{B'E}(1 + SR_G) + r_{B'B} + R_G} \cdot \frac{1}{1 + j\omega/\omega_{cB}} \quad \ldots\ldots\ldots\ldots\ldots \quad (9)$$

The cut-off frequency is then

* *University of Zagreb and Institut 'Ruder Boskovic', Zagreb.*

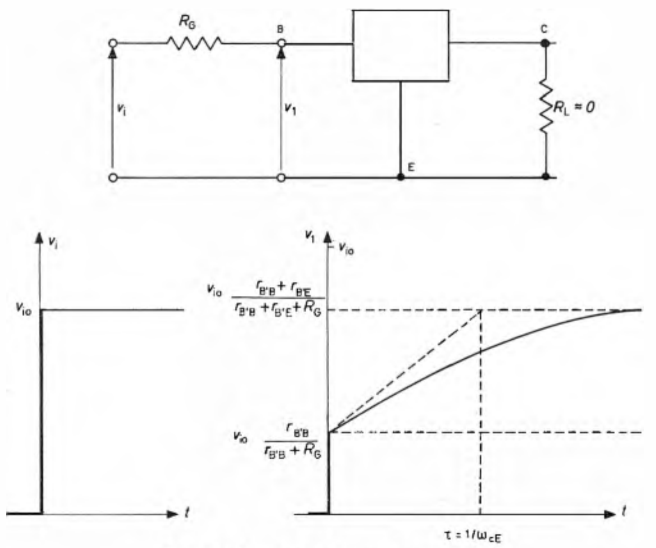

Fig. 3. Basis of oscilloscope method

$$\omega_{cB} = 1/c_{B'E}\,\frac{r_{B'E}\,(R_G + r_{B'B})}{r_{B'E}\,(1 + SR_G) + r_{B'B} + R_G} \quad \ldots\ldots (10)$$

This cut-off frequency is always lower than ω_α. If R_G is infinite, the frequency ω_{cB} obtains its maximum value and is equal to

$$\omega_{cBM} = S/c_{B'E} = \omega_\alpha \quad \ldots\ldots\ldots\ldots (11)$$

On the other hand, if $R_G = 0$ the cut-off frequency has the minimum value $\omega_{cBm} = \omega_{cEM}$. The dependence of the cut-off frequency on the signal source impedance is shown in Fig. 2. Adding the load characteristics to those of the transistor, the overall characteristics can easily be found.

The necessary data of the parameters $r_{B'B}$, $r_{B'E}$ and $c_{B'E}$ can be obtained either by measuring the three cut-off frequencies in the range between ω_β and ω_α or by an oscilloscope method. Such a method is shown in Fig. 3.

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A Stable Frequency D.C. to A.C. Convertor

By G. Smiljanic*

The frequency of a d.c. to a.c. convertor is made independent of supply voltage changes, using Zener diodes to stabilize the voltage amplitude of a transistor base driving transformer, which determines the frequency. The variations of the frequency, obtained with the proposed circuit, are 1 per cent or less for supply voltage changes of 20 per cent.

(Voir page 635 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 642)

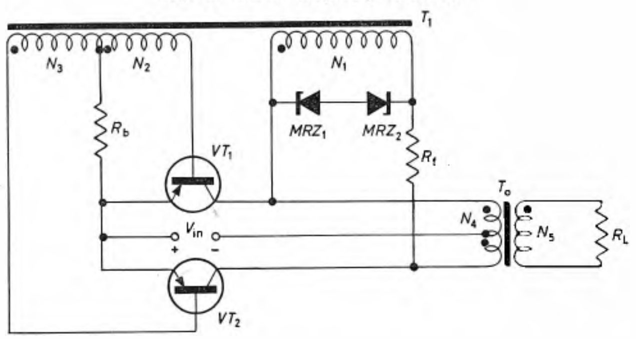
IT is often necessary to obtain an a.c. voltage from a d.c. power source. Transistorized d.c. to a.c. convertors, proposed by Uchrin and Taylor[1] and Royer[2], are very suitable for that purpose, especially if the a.c. voltage is required to deliver power. The frequency of such a d.c. to a.c. convertor is linearly proportional to the d.c. voltage and therefore it varies with the d.c. voltage. However for some purposes it is necessary to have stable frequency. A method for obtaining stable frequency is to use a constant d.c. voltage as the power source. In that case d.c. voltage stabilizers capable of giving the desired power are needed. Another method, suitable for obtaining stable frequency using two small transistor base driving transformers has been proposed[3]. However, a d.c. to a.c. convertor, using only one transistor base driving transformer, as suggested by Bell and Wright[4] and Nowicki[5], may have —in addition to higher efficiency—a relatively stable frequency. For that purpose it is only necessary to modify their scheme in such a way as to stabilize the voltage across the primary winding of the frequency determining transformer. In this article the design of such a convertor is presented and also the experimental results given.

A Convertor Having Stable Frequency

A convertor having stable frequency is shown in Fig. 1. The transformer T_1 is a transistor base driving transformer, which saturates earlier than the output transformer T_o and therefore determines the frequency. It is made with a magnetic core having a rectangular hysteresis loop. The voltage across the primary winding N_1 is stabilized by Zener diodes MRZ_1 and MRZ_2 as shown in Fig. 1. While the transistor VT_1 conducts, the diode MRZ_1 stabilizes the voltage across N_1, MRZ_2 operates as a common forward biased diode, and the magnetic flux in the core T_1 changes in one direction toward saturation. When the core T_1 is saturated, VT_1 cuts off and VT_2 begins to conduct. Now MRZ_2 stabilizes the voltage, and the core T_1 is driven toward saturation in the opposite direction. T_o is a conventional output transformer, capable of transferring the necessary power and giving the required amplitude of the output pulses.

In the case of a particular transformer T_1, having fixed number of turns N_1 and cross-section S, the frequency may change due to a change of the voltage E across the winding N_1, which determines the rate of the transformer flux change, or due to a change in the maximum inductance B_m. For a rectangular hysteresis loop, B_m is more or less constant. The voltage E is stabilized by Zener diodes and the resistor R_f against changes of voltage V_{in}. Moreover the diodes operating as forward biased diodes and Zener diodes may have temperature coefficients of different sign, and thus their total voltage

Fig. 1. Stable frequency convertor



* *Institut 'Ruder Boskovic', Zagreb, Yugoslavia.*

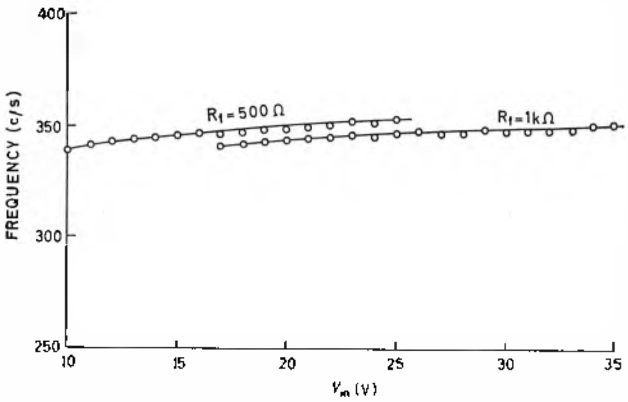


Fig. 2. Frequency as a function of input voltage

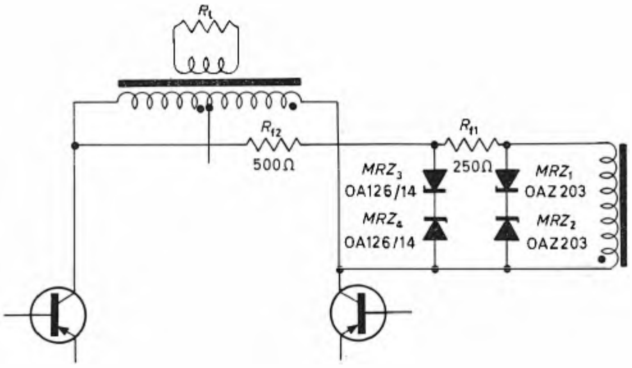


Fig. 3. Use of Zener diode stabilizer

may be temperature compensated. One can also arrange the total temperature coefficient of the diodes to compensate the temperature coefficient of core T_1 and thus obtain a temperature stable frequency.

The frequency change of a convertor, when V_{in} is changed, is shown in Fig. 2. The core T_1 is Permenorm 5000z having a rectangular hysteresis loop. Using a resistor $R_f = 500\Omega$, the frequency changes less than 4 per cent for a change of V_{in} between 10 and 25V. When a resistor $R_f = 1k\Omega$ is used, the change of the frequency is less than 3 per cent for an input voltage change between 17 and 35V. The curves are given to show the performance of the convertor at a large change of the input voltage. In practical cases the voltage V_{in} does not change over such a wide range, and hence the frequency variations are also smaller. If for instance V_{in} changes between 20 and 25V, or 25 and 30V, then with a resistor $R_f = 1k\Omega$ the frequency changes less than 1 per cent.

Using a stabilized voltage $V_{in} = 20V \pm 1$ per cent, the frequency of the convertor, after a warm-up period of approximately half an hour, changes by less than 0·1 per cent for six hours at $R_f = 1k\Omega$ and $R_1 = 2k\Omega$, and less than 0·3 per cent for 24 hours.

A still smaller frequency change than in the above example may be obtained if the voltage E is better stabilized. This can be done using another Zener diode stabilizer in series as shown in Fig. 3. With the stabilizer components given in Fig. 3, and all other elements the same as in the above example, the frequency of the convertor changes less than 0·2 per cent for a change in voltage V_{in} between 25 and 30V.

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Data Transmission for Motorway Control

G.E.C. Electronics' Teledata remote control and indication equipment, used on all emergency telephone systems on British motorways, has now found wider application on the new M4 motorway. Ordered by the Ministry of Transport, the equipment is used in a comprehensive surveillance and control system for the entire length of the elevated section of the new motorway, so that traffic flow can be monitored and controlled from a central point in Hounslow police station.

Probably the most significant new application is its use in conjunction with three remote closed circuit television cameras, mounted on high vantage points overlooking the length of the elevated section. The television pictures are fed back from the cameras to the police station where they are displayed on television monitors. Using Teledata channels the Duty Officer can switch the television cameras on and off, control pan and tilt, operate the zoom lenses and, when necessary, switch on the windscreen wipers on each camera.

Teledata is also used in conjunction with the large direction and diversion signs which straddle the motorway at exit points and in the system of loop detectors embedded in the road surface to transmit information on traffic flow to the central point in the Hounslow police station. This information on traffic flow, in conjunction with the situation displayed on the television monitors, enables the Duty Officer to survey traffic flow along the M4 on a mimic diagram. If for some reason, due to an emergency, a section of the M4 must be closed, the Duty Officer can operate the diversion signs via Teledata circuits.

The emergency telephone system on the M4 also uses Teledata equipment so that the location of any caller can be pinpointed, immediately the telephone hand-set is lifted, on a display panel in the police station. Teledata enables groups of ten telephones (spaced at one mile intervals along the motorway) to share a single cable pair.

Modular remote control equipment, such as is used in this application, is marketed by the Instrumentation & Control Department of G.E.C. (Electronics) Ltd.

A Cryogenic Circulator for Satellite Communications

A new cryogenic stripline circulator has been developed by The Marconi Co. Ltd to operate at temperatures as low as that of liquid helium (−269°C). This type of circulator is an essential component in certain types of parametric amplifier now being developed to improve satellite communications receiving systems.

A number of new types of receiver have been developed to operate at these low temperatures, but problems are encountered when conventional components have to be used. The stripline circulator which has been developed at the Marconi Research Laboratories at Great Baddow, employs a ferrite material which has properties at the temperature of liquid helium, similar to those of a normal ferrite at room temperature.

Special techniques have been employed in the manufacture of this material, which gives excellent results over a wide range of low temperatures, both at liquid helium and liquid nitrogen temperatures (−269°C and −196°C).

The circulator is a four port, stripline device, operating in the frequency range 3600 to 4300Mc/s, normally used for multi-channel telephony and satellite communications. However, the low temperature techniques and the ferrite materials which have been developed could be used to extend the range into other frequency bands if required.

Automatic Solution of Network Frequency Response

By T. Fleetwood*, B.Sc.

A routine method is described for the solution of electrical network response and the use is described of a computer programme which implements the method. Using this type of programme, results for any proposed network configuration can rapidly be obtained.

(Voir page 635 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 642)

A GENERAL method of solving the response of a linear network involves as a first step writing down a set of simultaneous equations that equates to zero the total current leaving each node of the network. The set of equations for a general $n+1$ node network are:

$$\begin{array}{l} \\ \text{Row 0} \\ \text{Row 1} \\ \\ \\ \text{Row } n \end{array} \begin{bmatrix} 0 \\ 0 \\ 0 \\ \vdots \\ 0 \end{bmatrix} = \begin{array}{c} \text{Col. 0 \quad Col. 1 \quad Col. 2} \\ \begin{bmatrix} y_{00} & y_{01} & y_{02} & \cdots & y_{0n} \\ y_{10} & y_{11} & y_{12} & \cdots & y_{1n} \\ y_{20} & y_{21} & y_{22} & \cdots & y_{2n} \\ \vdots & \vdots & \vdots & & \vdots \\ y_{n0} & y_{n1} & y_{n2} & \cdots & y_{nn} \end{bmatrix} \end{array} \times \begin{bmatrix} V_0 \\ V_1 \\ V_2 \\ \vdots \\ V_n \end{bmatrix} \qquad \ldots\ldots\ldots\ldots (1)$$

In order to be able to obtain a solution at least two node voltages must be specified. Usually node 0 will be ground and node 1 the input of the network, thus if node 1 is made 1 volt and node 0 zero volts, solving for the output node voltage will give the response in magnitude and phase.

Each coefficient y_{ij} in the equation is a complex function of frequency and can be represented by a real part a_{ij} and an imaginary part b_{ij}. A node voltage V_i can also be represented as V_{iR} and V_{iI} where R and I denote real and imaginary parts respectively. When these substitutions are made and real and imaginary parts are separated a new relation results:

$$\begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \\ 0 \\ 0 \\ \vdots \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} a_{00} & b_{00} & a_{01} & b_{01} & a_{02} & b_{02} & \cdots & a_{0n} & b_{0n} \\ -b_{00} & a_{00} & -b_{01} & a_{01} & -b_{02} & a_{02} & \cdots & -b_{0n} & a_{0n} \\ a_{10} & b_{10} & a_{11} & b_{11} & a_{12} & b_{12} & \cdots & a_{1n} & b_{1n} \\ -b_{10} & a_{10} & -b_{11} & a_{11} & -b_{12} & a_{12} & \cdots & -b_{1n} & a_{1n} \\ a_{20} & b_{20} & a_{21} & b_{21} & a_{22} & b_{22} & \cdots & a_{2n} & b_{2n} \\ -b_{20} & a_{20} & -b_{21} & a_{31} & -b_{22} & a_{22} & \cdots & -b_{2n} & a_{2n} \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & & \vdots & \vdots \\ a_{n0} & b_{n0} & a_{n1} & b_{n1} & a_{n2} & b_{n2} & \cdots & a_{nn} & b_{nn} \\ -b_{n0} & a_{n0} & -b_{n1} & a_{n1} & -b_{n2} & a_{n2} & \cdots & -b_{nn} & a_{nn} \end{bmatrix} \times \begin{bmatrix} V_{0I} \\ V_{0R} \\ V_{1I} \\ V_{1R} \\ V_{2I} \\ V_{2R} \\ \vdots \\ V_{nI} \\ V_{nR} \end{bmatrix} \qquad \ldots\ldots\ldots\ldots (2)$$

This form contains no complex numbers and thus can be solved using conventional determinant or matrix methods.

The procedure so far described is not practical for hand computation due to the large number of arithmetic steps involved in calculating the complex coefficients and solving the equations. It is, however, well suited to implementation by computer since standard routines are available for solving simultaneous equations. The problem is thus reduced to specifying the structure of the equations and the magnitude of the coefficients. Even this is tedious, however, and it is the purpose of this article to demonstrate how the effort involved can be reduced to writing down a description of the network together with some procedural instructions.

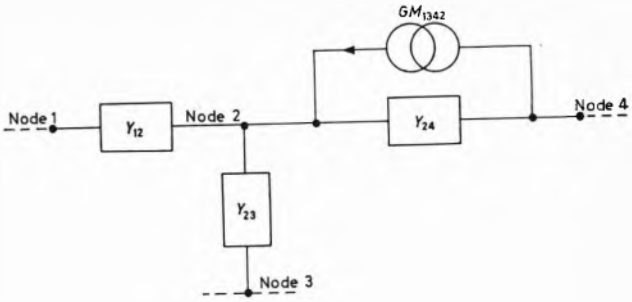


Fig. 1. Part of network showing suffices

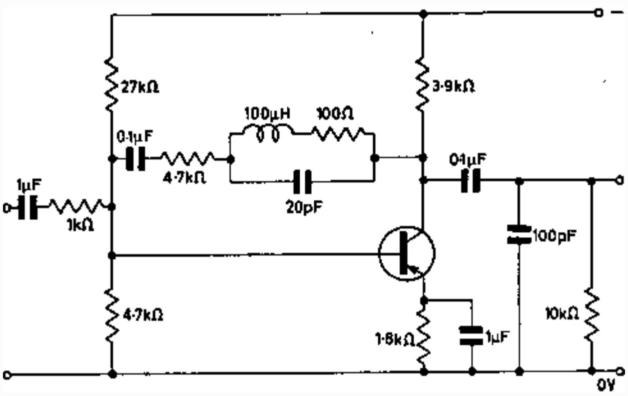


Fig. 2. Amplifier

Determining the Coefficients

Fig. 1 shows part of a network in which the elements are identified by suffices, thus Y_{12} is connected between nodes 1 and 2 and in a similar notation GM_{1342} drives current proportional to V_1 minus V_3 from node 4 to node 2. The admittance element Y_{12} will cause contributions to the current leaving nodes 1 and 2, that is $Y_{12}(V_1 - V_2)$ from node 1 and $Y_{12}(V_2 - V_1)$ from node 2.

* *The General Electric Co. Ltd.*

Rows 1 and 2 of equation (1) define the total currents in these nodes, thus there will be four contributions of Y_{12} to the coefficients in these rows, in columns 1 and 2. In fact for a general admittance Y_{pq} the four additions will be in positions defined by Table 1.

TABLE 1

Row \ Column	p	q
p	$+Y$	$-Y$
q	$-Y$	$+Y$

A current generator GM_{pqrs} will also cause four contributions defined in Table 2.

TABLE 2

Row \ Column	p	q
r	$+GM$	$-GM$
s	$-GM$	$+GM$

Using these rules it is possible to build up a set of equations in a routine manner for any network of admittances and dependant current generators. The size of the skeleton set of equations can be determined from the highest node number involved.

The form given by equation (2) can be derived from the basic set of equations by replacing each complex coefficient $a_{ij} + jb_{ij}$ by the real group $\begin{matrix} a_{ij} & b_{ij} \\ -b_{ij} & a_{ij} \end{matrix}$ in the same relative position. The resultant equations can then be solved for the output node voltage in terms of V_1 and V_0.

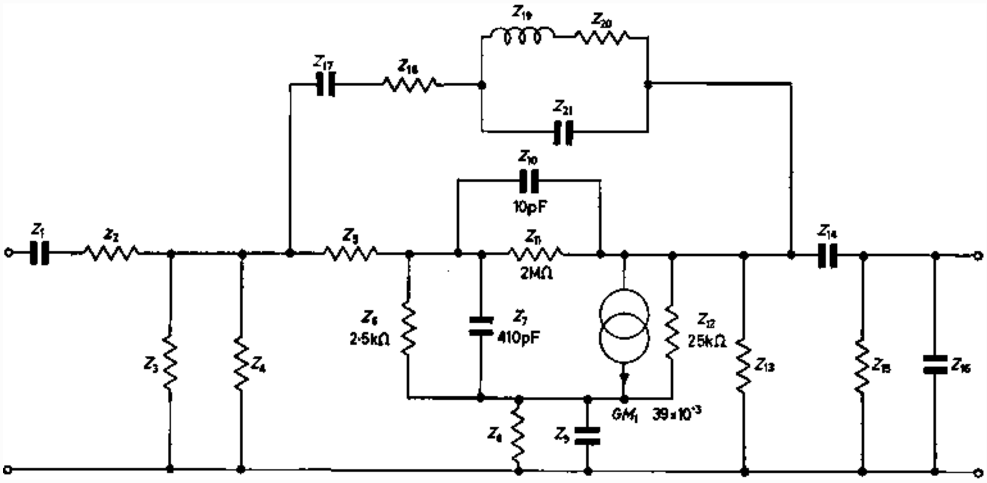
Computer Programme

A programme has been prepared for a Mercury computer implementing the principles described. The programme is not limited to the solution of one network but can be applied to any constructed of admittances and voltage

TABLE 3

```
—55    (Resistors)
2  1,3
3  27,3
4  4·7,3
5  110
6  2·5,3
8  1·8,3
11  2,6
12  25,3
13  3·9,3
15  10,3
18  4·7,3
20  100

−66    (Capacitors)
1  1,−6
7  410,−12
9  1,−6
10  10·5, −12
14  0·1,−6
16  100,−12
17  0·1,−6
21  20,−12

−77    (Inductors)
19  100,−6

−88  (Mutual conductances)
1  39,−3

−11  (Condensations)
1  1  1  2
1  17  17  18
1  19  19  20
2  19  19  21
1  17  17  19
2  3  3  4
2  6  6  7
2  8  8  9
2  10  10  11
2  15  15  16

−22  (Network structure)
1  2  1
2  0  3
2  3  5
2  5  17
3  4  6
3  5  10
4  0  8
4  5  12
5  0  13
5  6  14
6  0  15

—33  (Network structure)
1  3  4  5  4

−44 (Frequencies)
100
250
500
1,3
5,3
1,4
5,4
1,5
8,5
1,6
1·5,6
2,6
2·5,6
3,6
5,6
1,7
—99 (Start computation)
```

Fig. 3. Equivalent circuit of Fig. 2



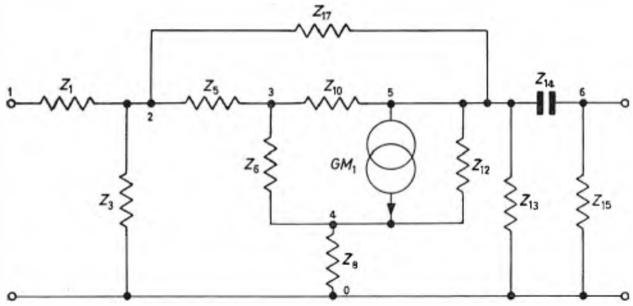


Fig. 4. Re-arrangement of Fig. 3 after condensations

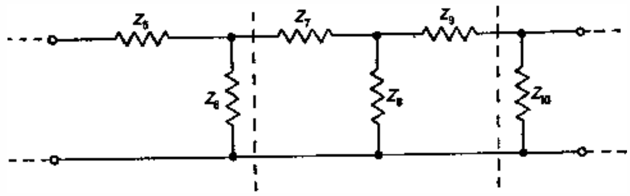


Fig. 5. Part of network to be simplified

dependant generators of up to 20 nodes.

Table 3 shows the form of data which describes a network to the computer. A series of lists are involved, each list being identified by a particular negative code number. The data in fact refers to the amplifier in Fig. 2 and Fig. 3, the latter being the equivalent circuit.

The first four lists specify the values of the components, thus in the first list, 2 1, 3 means the component designated Z_2 is a 1kΩ resistor (1, 3 is short for 1×10^3). In the second list 1 1, -6 means that Z_1 is a capacitor value 1μF.

The fifth list under code -11 details condensations to the basic network. Each entry in the list consists of four numbers. The first indicates the type of condensation and the other three, the components involved. The instruction 1 1 1 2 can be interpreted as: add in series Z_1 to Z_2 and replace the previous value of Z_1 by the resultant value, or in symbolic form. Series addition (1) $Z(1) = Z(1) + Z(2)$. Similarly 2 1 1 2 would mean add in parallel Z_1 to Z_2 and replace the stored value of Z_1 by the resultant.

The condensation facility allows preliminary simplification of the network, thus the number of nodes can be minimized with resultant speeding up of the main calculation.

Fig. 4 shows the form in which the circuit can be redrawn after successive condensations. The correct value of the components of the redrawn network for the frequency concerned will be stored in the computer. At this stage the nodes are numbered and the highest node number is reserved for the output of the network.

The next two lists describe the layout of the network in Fig. 4, thus 1 2 1 signifies that node 1 to node 2 is occupied by Z_1. Under the list heading -33, 1 3 4 5 4 means the mutual conductance GM_1 depends upon $V_3 - V_4$ and passes current from node 5 to node 4.

The remaining list contains the frequencies for which the calculation is to be repeated. No more data is needed and the computer will proceed to calculate the response for each of the frequencies listed.

Table 4 shows the results for the simple network. The data is used with a master programme which is the same for any network being analysed, thus the preparation of the lists as described is the only work involved. More than one network can be processed in a single run on the computer and further lists can be added to the data to assess the effects of systematic changes in structure or component values.

TABLE 4

FREQUENCY (c/s)	GAIN (dB)	PHASE (Degrees)
1·00, 2	−1·98	268·31
2·50, 2	5·10	291·96
5·00, 2	9·06	311·68
1·00, 3	11·28	330·87
5·00, 3	12·37	353·78
1·00, 4	12·41	357·12
5·00, 4	12·42	0·92
1·00, 5	12·43	2·80
8·00, 5	12·86	26·88
1·00, 6	13·05	35·06
1·50, 6	13·27	59·81
2·00, 6	12·35	89·47
2·50, 6	9·77	117·13
3·00, 6	6·02	136·10
5·00, 6	−3·49	146·56
1·00, 7	−14·80	165·03

A star to delta or delta to star transformation facility is available as a condensation instruction. Fig. 5 shows part of a network that can be simplified and an instruction 3 7 8 9 added to the condensation list for the full network will cause the values of Z_7, Z_8 and Z_9 in the computer memory to be replaced by their equivalent delta values. The network can be redrawn (Fig. 6) with one less node than originally. Delta to star transformation can be made in a similar manner by using the code number 4, thus the instruction 4 6 7 8 applied to the network in Fig. 5 would result in the stored values of Z_6, Z_7 and Z_8 being replaced by their star equivalents.

Conclusion

This programme has been used frequently over the past year as a direct aid to circuit development work and the rapid turn round from production of the equivalent circuit to plotting of the results has been found extremely useful.

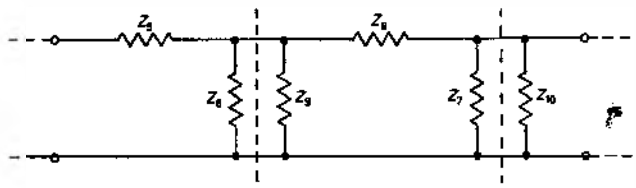


Fig. 6. Re-arrangement of Fig. 5

The design engineer himself prepares the data in the form of a punched tape which is transmitted to a central computer through the G.P.O. Telex system. Results are received through the same system normally the same day and in any case not more than twenty four hours later.

The system which has been described is no more than an introduction to the possibilities of automated analysis. Variants on the basic programme could be evolved to deal with other types of components such as transformers and voltage generators, or to assess the effects of tolerances in a systematic manner. Logically a new computer language might be evolved to deal specifically with network problems including extensions to the solution of transient responses which involve a similar technique.

BOOK REVIEWS

Microwave Circuits

By Jerome L. Altman. 462 pp. Med. 8vo. D. Van Nostrand Co. Inc. 1965. Price £5 16s.

THE subject of microwave circuits is one where there is a considerable dearth of good books, so that a newcomer to the field needs to supplement the standard texts with a large number of original papers. The appearance of a new text is therefore very welcome.

In the preface the author states his object as the presentation of a comprehensive treatment of a limited range of topics, and it is probably fair to say that in some respects he has succeeded. The introductory chapter develops circuit concepts via field theory, and in chapters 2 and 3 the scattering matrix and its applications are discussed at length. Chapter 4 is a treatment of miscellaneous microwave components (broadband transformers, directional couplers, hybrids, phase shifters, and attenuators). Microwave cavities and filters are described in chapters 5 and 6, and chapter 7 deals with periodically-loaded lines. Chapter 8 describes various systems applications of these microwave components, but the material here might well have been used for illustrative purposes in the earlier chapters. The same remark applies to many of the appendices, which are gathered together at the end of the book

The best part of the work is in the treatment of the scattering matrix, where eigenvalue methods are developed logically. However, there is rather too much detail in the applications, e.g. pages 88 to 115 are devoted entirely to the properties of passive T-junctions, and this makes very tedious reading. Might it not have been preferable to allow the reader or student some degree of intellectual exercise by presenting part of this material in the form of problems? The same criticism applies to much of this book, which could be improved by judicious pruning and reorganization. The nomenclature, too, leaves much to be desired. In particular, the use of a script form of type for many symbols is very tiresome, while the employment of numerical superscripts rather than subscripts to denote anything other than powers is inexcusable (this fault occurs throughout chapter 3).

The main omission of this book is the subject of microwave network synthesis on a modern basis using P. I. Richards' theorem. This implies that much of the treatment is outmoded. For example, the multi-section quarter wave transformer is treated by the old approximate method, and there is scarcely any discussion of modern practice in microwave filter design. A minor fault is that the formulae for the Q-factors of waveguide cavities fail to take account of the frequency sensitivity of the obstacle susceptance.

In spite of its limitations the book contains some useful information, particularly on scattering matrices, and should be helpful in the preparation of courses in the subject.

R. LEVY

Dynamic Circuit Theory

By H. K. Messerle. 657 pp. Royal 8vo. Pergamon Press. 1965. Price £5

THE aim of the author of this immense book is to unify the analysis of transducers, electrical machines, transformers, electromechanical systems and servo-mechanisms by using the techniques of linear circuit theory and a common basis of energy storage and conversion. To this end a third of the book is devoted to 'basic electro-dynamic relations', 'transient and linear system analysis' and this is followed by chapters dealing with 'double storage transducers', 'commutator machines', 'two phase machines', 'multiphase systems', 'three phase machines' and 'control systems' (to use the author's chapter headings).

Characteristic equations for each device are derived by precisely similar mathematical processes, thus giving emphasis to common features and avoiding the diversity of approach often found, for example, in books dealing specifically with machines or instruments or transducers.

Matrices are used extensively but their manipulation has been kept at an elementary level covered by a chapter on mathematical techniques. Differential equations are handled primarily with the D-operator, but Laplace transform and root-locus methods are also used.

The reliance on linear mathematical analysis rather than physical argument leads to highly idealized results which might mislead the unwary. (For example, it is stated that all electro-dynamic ammeters have their fixed and moving-coils in series and have square-law scales!) Where problems cannot be treated as linear, the author is forced to give physical explanations which tend to be rather naïve and the treatment descends to an elementary level. The existence of non-linearities is frequently mentioned, but their effects are covered only by tail-end modifications to the analysis (as is the case all too often in many other books treating the generalized machine theory).

A particular feature is the development of fundamental concepts by working from the particular to the general. This may be good educational practise in a lecture course, but here it leads to tiresome repetition and excessive length. The generalized approach to electrical machines is thus obscured by a welter of 'conventional' analysis of transformers, d.c. machines, two-phase and three-phase machines albeit by predominantly mathematical methods.

The book will be of particular interest to those who have learnt their machines and devices by 'conventional' methods and now wish to translate this knowledge into the current mathematical terminology of generalized concepts. In attempting to give both conventional and generalized approaches it may fail to satisfy either camp but it does have value as an integrating text. It could not be recommended for the accuracy or clarity of the physical arguments.

The format is exceptionally clear and diagrams and text are well related. There are numerous worked examples and a set of problems on each chapter. The mathematical treatment is commendably straightforward.

J. W. SKINNER and W. SHEPHERD

Principles of Inverter Circuits

By B. D. Bedford and R. G. Toft. 413 pp. Med. 8vo. J. Wiley & Sons. 1965. Price £4 16s.

ALTHOUGH only two authors are quoted for this book, there are in addition five contributing authors who deal with specialized aspects of the subject. The first chapter is devoted to the theory of transistors and semiconductor devices such as the silicon controlled rectifier. This is a useful introduction to those not familiar with these devices. After a short chapter on invertor principles a long chapter is devoted to phase controlled rectifiers and what are termed line voltage commutated invertors, i.e. invertors operating on to an a.c. system. The work of this chapter applies to any type of controlled rectifier from thyratrons, mercury arc rectifiers to silicon controlled rectifiers and covers much basic standard material. Chapters are then included on parallel capacitor, series capacitor, harmonic and impulse commutated invertors. Voltage and waveform control is then considered.

A long chapter then considers d.c. to d.c. power conversion. One might expect that this would be devoted to invertor-rectifier circuits, but these are dismissed in two pages. The chapter is devoted to most interesting circuits using time ratio control. By this means variable voltage d.c. power supplies are readily produced. Finally a chapter is devoted to modern applications of the silicon controlled rectifier, these varying from large invertors up to nearly 1 000kVA to electric vehicle control, control of motor speed and commutatorless machines.

The book is well written and readily

ınderstood. Mathematical analysis is ;iven of most circuits, but where this s lengthy it is relegated to an appendix. \ number of useful references is given .t the ends of the chapters. The diagrams .re clear and well produced. The book ; to be recommended both to students nterested in the theory of invertors and o engineers seeking up-to-date know-edge on the subject.

G. N. PATCHETT

;eneral Principles of Automation

y J. F. Schuh. 381 pp. Med. 8vo. Phillips 'echnical Library (distributed in U.K. by Cleaver-Hume Press Ltd). 1965. Price 72s.

THE title is quite misleading. Certainly in the 'Detroit' sense, automation ; not the central theme of this book, nd the sub-title, namely, 'what a robot an and cannot do', gives the impression hat the author has solved all the prob-ɜms that lesser mortals have been truggling with for years. And all this ·ithout recourse to even the most lementary non-linear theory! Such aulty titling and sales promotion ıethods are the more to be deplored ince the book is, in its own way, a ound and remarkably readable intro-uction to the subject of *automata*, and as the refreshing European style of uoting original references.

The range of topics included is con-iderable, commencing with a stimulating ccount of the basic concepts of auto-ıata, indicating their apparent corre-pondence with biological structures. :mphasis is placed on the use of ıathematical analysis, and chapter 2 ıtroduces the theory of propositional ɔgic, including aspects of Boolean lgebra. In dealing with codes and ınguage, chapter 3 relies on the pre-eding chapter, especially in dealing ·ith the semantic content of signals. It ; here that Laplace's ghost is intro-luced, which is certainly a haunting liversion. Chapter 4 returns to the real lomain to discuss basic physical com-ɔonents, both singly and in aggregates .s logical networks, memories, signal ransformers, shift registers, control cir-:uits, etc. The signal transfer point of 'iew is taken rather than detailed hard-vare characteristics, although the latter ıre certainly not neglected.

The author gives the impression that ıis first love is for mathematical logic, ɔut in chapter 5 'mechanisms in general' :orms the topic. It is here that linear lifferential equations and Laplace trans-:orm theory are set out. This is followed ɔy an introduction to network theory which, although ably written, omits nany common networks met with in :ontrol system work, notably active ıetworks such as those associated with :he use of operational amplifiers.

Some criticism also applies to the rather superficial section on the prin-:iples of feedback control theory. Surely :his cannot be regarded as a subject so remote from network theory to require ı different jargon in the same book, indeed in the same chapter. In this same hotch-potch chapter the theory of noise and message-separation in digital signal-ling systems is formulated, and it con-cludes with some facts about the structure and programming of digital computers. Three appendices follow, dealing with harmonic functions, stoch-astic processes, and Fourier integrals and series.

In the earlier chapters the author shows a flair for explaining some quite difficult and essential concepts in the understanding of automata. Most of this book is written at a good theoretical level and should interest a wide range of readers, particularly those who prefer the cybernetic approach.

In the original Dutch the book is no doubt unique. The translation into English places it high among surprisingly few other books covering this field in the English language. But please note, the field is *automata* not automation.

K. C. GARNER

Principles of Electron Tubes

By J. W. Gewartowski and H. A. Watson. 655 pp. Med. 8vo. D. Van Nostrand & Co Ltd. 1965. Price £7 4s.

THERE is strong aesthetic appeal and beauty in a clear logical argument. The same sort of satisfaction may be found in such apparently unrelated topics as a well-formulated Euclidean theorem, a well-presented legal discourse, or a well-performed Bach fugue. It is there-fore a joy to find a text-book which, in its presentation, follows carefully considered arguments arranged in an orderly sequence and which, while pre-senting hard scientific facts, does so with such clarity and careful planning.

'Principles of Electron Tubes' is one of the latest additions to the Bell Tele-phone Laboratories Series. Most aspects of electron tubes and their applications are covered. The first three chapters provide an introduction to electron motion, emission, and the focusing and properties of electron beams.

Chapter 4 introduces us to the diode whence we are lead through devices of increasing complexity and higher fre-quencies to magnetrons and carcinotrons.

Chapter 13 is devoted to the important question of noise and the fundamental limitation that this imposes on the per-formance of electron tubes and their associated circuits. Three chapters are then devoted to the conduction of elec-trons through gases, and to cold- and hot-cathode tubes. Finally chapter 17 introduces gas lasers.

Seventeen chapters, sixteen appendices, some 645 pages excluding index cover most aspects of electron tube technology. A good knowledge of mathematics is demanded from the reader, including vectors and advanced calculus. However, by placing emphasis on the basic prin-ciples involved, complicated and rigorous mathematical analyses are avoided when simpler approximate solutions are pos-sible. Numerous problems are also given to enhance the book's usefulness as a text-book.

Engineers engaged on electron tube development should find this book of particular value while those using elec-tron tubes in circuit applications will find the book of interest and help in appreciating the limitations of and some of the problems facing the electron tube designer.

R. A. BONES

Fundamentals of Relay Circuit Design

By A. R. Knoop. 312 pp. Crown 4to. Chapman & Hall Ltd. 1965. Price £6

THE preface to this book epitomizes the philosophical approach of the writer to a work of this nature. Mr. Knoop points out that long practice and experience is essential to the circuit designer.

Tht book commences with fundamen-tals working up in easy steps to complex circuits, making the work an excellent reference book for the student. Unfor-tunately for the British reader, the publication is American and all the symbols, standard voltages and fre-quencies are given as American standards necessitating careful reading and angliciz-ing to his own standards.

The writer has endeavoured to cover too wide a field with one volume, and the chapter dealing with the ancillary circuits for motor starting switchgear is inadequate. This section must also be treated with caution, as in two places the circuits show overload elements in two phases only of the three phase-motor supplies, which is contrary to the standard British and Continental prac-tice and precludes the use of anti-single phasing devices which are becoming a feature of three phase thermal overload motor protection.

The chapter on conveyor control embraces all forms of material hand-ling, and the sequential control tech-niques are fully illustrated.

The section on Railway Signalling gives an interesting history of American railway signalling practice from the early days of the block offices to modern equipment with emphasis on relay inter-locking control.

In reviewing this work it must be acknowledged that it contains a wealth of useful information and there are relatively few books of this type avail-able to the control engineer.

E. M. BUTTERWORTH

Unified Circuit Theory in Electronics and Engineering Analysis

By J. Head and C. Mayo. 174 pp. Demy 8vo. Iliffe Books Ltd. 1965. Price 42s.

The purpose of this book is to show how the output, or response, of a linear system may be formulated, whatever the nature of the input may be. Steady-state conditions present no mathematical difficulties, but decidedly more advanced techniques are required when dealing with the non-steady state.

There are alternative methods of solving problems of this kind, that involving the use of Laplace transforms being currently popular. The authors of the present book prefer, and strongly advocate, the use of operational calculus, which they maintain has the outstanding advantage of directness and simplicity of working that makes it easy for practising engineers and students alike to appreciate the significance of each necessary process.

The early part of the book is concerned with electrical circuits having lumped elements; the latter part is more theoretical, and extends the circuits which are still linear but obtain distributed elements. It also compares operational calculus with symbolic and Laplace-transform calculus more thoroughly. After dealing with two-port networks, the authors show how the handling of four-port networks may be enormously simplified by the use of techniques employing matrices.

High Power Microwave Electronics

By P. Kapitza. 148 pp. Med. 8vo. Pergamon Press. 1964. Price 60s.

This is the first book in a series to be published on high frequency power and its applications, and discusses the foundation of the theory of high power electronic devices of the magnetron type.

Transistors Applied

By H. Kaden. 194 pp. Med. 8vo. Philips Technical Library. 1965. Price 37s. 6d.

A short treatise is given in this book on the physical facts necessary to explain certain specific transistor properties, and presents the principles of the transistor as an amplifying element in combination with the requisite circuit components.

Absolute Stability of Regulator Systems

By M. Aizerman and F. Gantmacher. 172 pp. Med. 8vo. Holden-Day Inc. 1964. Price $8.95

This book gives a historical survey of the problem of absolute stability as applied to automatic control theory and deals also with the new approach developed by the Rumanian scientist V. M. Popov and others.

A Pathfinder

By E. Acheson. 63 pp. Demy 8vo. Acheson Industries (Europe) Ltd. 1965. Price 10s.

This is a reprinted edition of the autobiography written by the author 50 years ago.

Digital Computers in Action

By A. Booth. 146 pp. Crown 8vo. Pergamon Press. 1965. Price 10s.

This small book is intended as an introduction to the digital computer and its programming for those who wish to learn how computers can be of assistance to their particular activities.

Brimar Valve Manual No. 10

416 pp. Demy 8vo. Thorn-AEI Radio Valves & Tubes Ltd. 1965. Price 7s. 6d.

This reference manual contains data on some 630 types of Brimar valves and television tubes. Industrial cathode-ray tubes and switching transistors have been included for the first time.

Mossbauer Effect: Principles and Applications

By G. Wertheim. 116 pp. Demy 8vo. Academic Paperbacks. 1965. Price 63s.

This monograph is intended as a short introduction to the concepts of recoil-free emission and resonant absorption of nuclear gamma rays in solids.

Computer and Information Sciences

By J. T. Tou and R. H. Wilcox. 544 pp. Med. 8vo. Macmillan. 1965. Price £5 10s.

This book contains the papers presented at the first Computer and Information Sciences Symposium (COINS) held at the Northwestern Technological Institute in Evanston, Illinois in June 1963.

Computer Augmentation of Human Reasoning

By M. A. Sass and W. D. Wilkinson. 276 pp. Demy 8vo. Macmillan. 1965. Price 40s.

This book contains the papers presented at the Symposium sponsored by the W.S. Office of Naval Research and The Bunker-Ramo Corporation and held at Washington D.C. in June 1964.

Solar System Radio Astronomy

By J. Aarons. 416 pp. Royal 8vo. Plenum Press. 1965. Price $17.50.

This volume contains the series of lectures presented at the NATO Advanced Study Institute of the National Observatory at Athens in August, 1964 and sponsored by the NATO Scientific Affairs Division.

The Electron in Electronics

By M. G. Scroggie. 276 pp. Demy 8vo. Iliffe Books Ltd. 1965. Price 45s.

The author deals with the new physical concepts of quantum theory, wave mechanics and relativity as applied to electronics. The standard of mathmatics and general physics is set at G.C.E. "A" level and the book is intended to be an introduction and supplement to the more formal and mathematical approach.

Explosion Proof Electrical Equipment

By G. A. Gel'fer, A. V. Ivanov, Ya. G. Medvedev. 334 pp. Crown 4to. Oldbourne Press. 1964. Price £4 10s.

Contains details of explosion proof electrical machinery and equipment manufactured in the U.S.S.R. and is intended for engineers and electricians working in the oil-refining and gas industries.

German-English Electronics Dictionary

By C. J. Hyman. 182 pp. Foolscap 4to Consultants Bureau Enterprises. 1965. Price $14.00.

Contains some 6000 entries in each language and includes specialized terms associated with computers, radar, telephoney and television as well as optics, mathematics and nuclear physics.

Dictionary of Electrical Engineering Vol. 1. German-English-French

By W. Goedecke. 826 pp. Demy 8vo. Isaac Pitman. 1965. Price 63s.

Volume 1 of this series contains some 26000 words and phrases associated with electrical engineering, telecommunications, electronics and related subjects. German is the key language and the English and French equivalents are given. Volume 2 in which French is the key language and Volume 3 for English as the key language are to be published shortly.

Problems in Electronics with Solutions Fourth Editon

By F. A. Benson. 307 pp. Demy 8vo. E. & F. N. Spon Ltd. 1965. Price 45s.

This is the fourth edition of this book originally published in 1958 and which now contains some 350 problems together with answers. The first part of the book sets out the problems and the answers in brief, and the second part the solutions where the steps in the quoted answers are dealt with in more detail.

The Determination of the Time Position of Pulses in the Presence of Noise

By B. N. Mityashev. 190 pp. Demy 8vo. Macdonald & Co. Ltd. 1965. Price 80s.

The author deals with the theoretical and experimental aspects of pulse position registration and describes technical operations when applied to pulse and noise signals in radar and communications.

Signal Flow Analysis

By J. R. Abrahams and G. P. Coverley. 158 pp. Crown 8vo. Pergamon Press. 1965. Price 17s 6d.

The use of signal flow graphs for the analysis of electrical circuits is described in this book which is intended for the undergraduate in the middle of his course or for the practising engineer.

LETTERS TO THE EDITOR

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

A Technique for Probability Density Function Measurement

DEAR SIR,—I would like to confirm that the technique described by the authors of this article (May 1965) is indeed one of great practicability and convenience. A similar method is in use in this Department and has proved to be most successful.

The following additional comments may be of interest to your readers. Firstly, if one of the input channels is omitted the same technique measures the probability *distribution* function of the amplitude of the input waveform. Deviations from a normal or Gaussian distribution can then be assessed by plotting the results on a type of graph paper that is available commercially and known as 'arithmetic probability paper'. A normal distribution would give a straight line on the graph, so deviations from normalcy are easily observed.

My second comment concerns the method of pulse counting. Errors could arise from deviations in the pulse repetition rate and in the period during which output pulses from the gate are counted. These errors can be eliminated by using a commercial counter-timer whose 'clock frequency' is available as a source of input pulses to the gate. In our equipment we set the counter-timer to count for say 100sec. This period is related to the 1Mc/s clock frequency of the instrument, and a sub-harmonic at 100kc/s feeds pulses to the gate in a manner similar to that described in the article. The counter thus reads the required ratio directly, apart from a factor of 10^7.

A final comment concerns the performance of the high gain d.c. amplifiers used as level selectors. These amplifiers are required to have a bandwidth greatly in excess of that of the input signal, since the clipped signal in the output stages has a much wider frequency spectrum than the original. The authors suggest the use of commercial operational amplifiers, but many such amplifiers are unacceptable, both on the score of inadequate bandwidth and because of time-dependent overload characteristics. In view of the possibility of spurious effects in the amplifiers it is advisable to check the performance of the equipment. This can be done by using a periodic test signal of known waveform.

Yours faithfully,

H. SUTCLIFFE,
Royal College of Advanced Technology,
Salford, Lancashire.

The authors reply:

DEAR SIR,—We welcome Mr. Sutcliffe's confirmation of the usefulness of the technique which is described in our article. His other comments are also very much appreciated since they greatly add to the value of our paper. Personal experience also leads us to urge the potential user to heed in particular Mr. Sutcliffe's warning of the need for properly selecting commercially available operational amplifiers.

Yours faithfully,

P. N. NIKIFORUK,
G. SQUIRES,
University of Saskatchewan,
Canada.

A Transistor Tester

DEAR SIR,—With great interest I read Mr. M. M. Winn's article 'A Transistor Tester' in your April issue. Some years ago I was faced with a similar problem: To measure transistor parameters under pulse conditions. For d.c. gain h_{FE} I used a circuit (Fig. 1) which has all the advantages claimed in Winn's article and in addition is much simpler. If $V_{cc} = 0$, i.e. if a and b are short-circuited, then the circuit measures the h_{FE} on the edge of saturation directly when the internal resistance of the meter M_2 is so low, that voltage $V_M \leqslant 50$mV (see Fig. 1) for full scale deflexion.

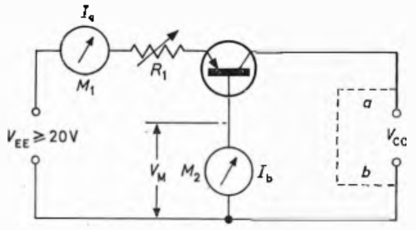


Fig. 1. D.C. gain measuring circuit

The current gain is given by

$$h_{FE} = I_c/I_b = \frac{I_e - I_b}{I_b} \simeq I_e/I_b$$

If the approximate relation $h_{FE} = I_e/I_b$ is used, then the error is $100/h_{FE}$ per cent, i.e. for $h_{FE} \leqslant 30$ it can usually be neglected. If V_{EE} is stabilized and R_1 is formed by a set of calibrated resistors, then the meter M_1 can be omitted and the scale of M_2 directly calibrated in h_{FE}.

Yours faithfully,

O. A. HORNA,
Research Inst. for Math. Machines,
Czechoslovakia.

The author replies:

DEAR SIR,—I was pleased to read about Mr. Horna's circuit. It is a good deal simpler than mine. However, I would like to mention one disadvantage:

The collector to emitter voltage V_{CE} applied to the transistor is equal to $V_{BE} + V_M$. V_M is controllable by altering the resistance R_M of the meter M_2 but depends on h_{FE} of the transistor.

$$V_M \simeq I_E R_M / h_{FE}$$

V_{BE} will vary from transistor to transistor and will rise as I_E is increased. Testing a series of transistors with different V_{BE}'s under the same V_{CE} will take some time as the sum $V_{BE} + V_M$ will have to be manually adjusted either by control of the voltage V_{CC} (see circuit) or by varying V_M. It would be fairly easy to control V_{CE} with a feedback arrangement but the circuit would then be equivalent to mine.

In addition, Mr. Horna's circuit would overload the meter M_2 if a transistor of low β or with an open-circuit collector base junction was connected to the test terminals.

Yours faithfully,

M. M. WINN,
University of Sydney, Australia.

A Fast Relay Sensor for A.C. Power Failures

DEAR SIR,—In the practical use of various a.c. operated electronic equipment, a fast switching operation is often required when unpredicted a.c. power failure occurs. Typical cases of interest are equipment protection and ensuring continuing functioning of the equipment.

A fast response relay, in the 1msec range, requires d.c. for excitation. Such a relay cannot be connected directly to the a.c. line, as it would then be actuated at the line frequency. The use of a full wave rectifier would only double the relay actuating frequency. The addition of a smoothing capacitor would appreciably increase the response time of the circuit.

This note describes a surprisingly simple solution which enables a very fast d.c. relay to be operated directly by the a.c. power line and to function practically instantaneously in case of power line failure. The circuit diagram is shown in Fig. 1. The d.c. power to the fast relay is supplied by a full wave rectifier. The amplitude of the rectified output across the relay is stabilized and limited by a Zener diode. The a.c. amplitude applied to the rectifier is high enough to ensure

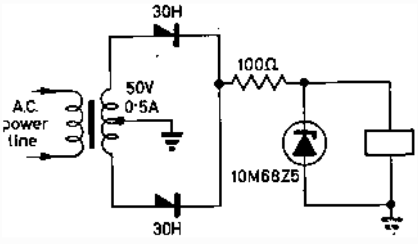
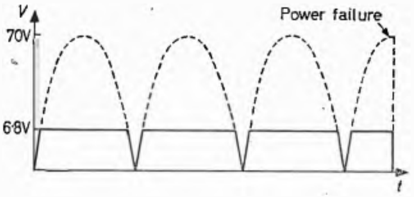


Fig. 1. The fast response relay circuit

Fig. 2. Effect of Zener diode



that the non-conduction period of the Zener will be shorter than the relay response time (Fig. 2). In that case the relay will normally remain energized and function instantaneously in case of an a.c. power failure. The fast response is assured by the absence of a storage capacitor in the circuit. The required windings are a part of the power transformer usually incorporated in a.c. equipment. The proposed circuit is simple, reliable, fail safe and inexpensive.

Yours faithfully,

S. ROZEN,
Atomic Energy Commission,
Israel.

A High Voltage Pulse Generator Using Switched Thyratrons

DEAR SIR,—The above design article by T. Chan and M. W. Gunn in the June issue of ELECTRONIC ENGINEERING revived interest in a previous application[1] of switched thyratrons, which I should like to draw to your attention.

This arrangement which is employed in the Hewlett Packard type 212A pulse generator, comprises a delay-line of value 10µsec which is driven by a thyratron at one end and short-circuited at the other. The pulse duration is adjusted by making a second thyratron conduct during the 10µsec pulse and so short-circuiting it as it passes to the output stage. This and other methods of pulse generation are described in a very useful review[2].

Yours faithfully,

T. J. WYNN,
Derby and District College of Technology.

REFERENCES

1. KAN, G. S. A New 0·07-10 Microsecond General-Purpose Pulse Generator. *Hewlett-Packard J., 1* (Feb. 1950).
2. Pulse and Square-Wave Generators (Review of Modern Instruments). *Electronic and Radio Engr. 36*, 211 (1959).

'D.C.' Amplifiers

DEAR SIR,—I was surprised to read in Mr. Harknett's letter that the Hartley "d.c. amplifier" means "directly coupled amplifier."

After a dozen years in North America, speaking and listening to a strange language, it is difficult to remember accurately the terms I used in England, but I am sure that to me "d.c. amp" has always described the *function* of an amplifier, not any particular design configuration.

I regard a d.c. amplifier as one which translates a d.c. voltage (pardon the term, but it *is* a common one) into a larger d.c. voltage or into a useful d.c. current. The fact that it is normally directly coupled is incidental.

An ad hoc committee of a dozen or so engineers here confirmed that this is the normal Canadian interpretation, and the consensus of opinion was that as long as we go on referring to "d.c. voltage" the term "d.c. amplifier" is acceptable, as it means "an amplifier which responds to a d.c. voltage change at its input."

The fact is that we habitually use the term "d.c." without consciously realizing that "c" stands for current.

While proposing "direct signal," perhaps Mr. Harkness could press for "direct voltage" and "direct power," and thus kill several birds with his stone.

Yours faithfully,

R. C. ELDRIDGE,
Vancouver, B.C.

The correspondent replies:

DEAR SIR,—I confess that I was not aware that the abbreviation d.c. was so generally accepted as an adjective and on reading Mr. Eldridge's letter I sought comfort in British Standard Specifications only to find the term d.c. voltage used therein.

My objection to such usage probably stems from the days when my lecturer would chastize me, it now appears unjustly, for writing d.c. current which he would say meant direct current squared.

It is an unfortunate coincidence that the terms direct coupled and direct current often appear together in the same paragraph together with the abbreviation d.c. and it was to avoid possible confusion that I suggested the term direct signal.

I concede the argument to Mr. Eldridge and agree that it is now accepted to use the abbreviation d.c. as an adjective. To avoid confusion the term direct coupled should always be written in full.

Yours faithfully,

M. R. HARKNETT,
The University, Southampton.

A Transistor Linear Ramp Generator

DEAR SIR,—In the following note a simple transistor ramp generator is described. Waveform duration, rise time and fall time are independently variable and linearity is better than 1 per cent for rise and fall times between 5msec and 5sec. The circuit is shown in Fig. 1.

VT_1, VT_2 form a flip-flop[1], triggered either as here, by push-button, or from any external source providing a 1 or 2V positive going pulse of about 200Ω impedance.

The rectangular, negative going pulse from VT_2 collector, is emitter-followed by VT_3, inverted by VT_4 and emitter-followed by VT_5. The two Zener diodes now d.c. restore the pulse, negatively to the base of VT_6, and positively to the base of VT_7.

In the normal condition, VT_6 is cut off and VT_7 conducting, its emitter being held at zero volts, and its collector circuit being completed through the diode MR_1. In this state, the common collector point will be very near earth potential, and C uncharged.

When the pulse arrives from VT_1, VT_2, transistor VT_7 is cut off, and VT_6 on. VT_6, as a constant current source, will now allow C to charge towards the -9V rail. When VT_1, VT_2 revert to their stable state, VT_6 will be cut off, and VT_7 becomes a constant current discharge path for C.

VT_8, VT_9 are conveniently cascaded emitter-followers, reducing the output impedance to 1kΩ. The 15µF timing capacitor in the flip-flop, gives a pulse, variable between about 0·5sec and 5sec. The values for C_R (coarse rise time control), are:

C_R = 0·5µF Rise-time 5-50msec.
C_R = 5µF „ „ 50-500msec
C_R = 50µF „ „ 500msec-5sec

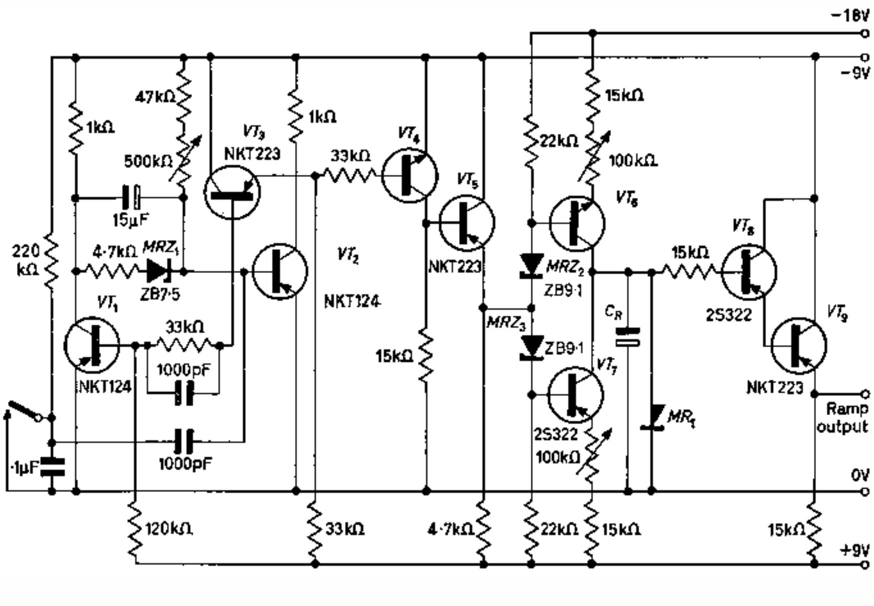
Yours faithfully,

W. J. BANNISTER,
The University Laboratory of Physiology,
Oxford.

REFERENCE

1. BANNISTER, W. J. A Transistor Delay and Pulse Generator. *Electronic Engng. 35*, 746 (1963).

Fig. 1. The linear ramp generator



ELECTRONIC EQUIPMENT

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

(Voir page 629 pour la traduction en français; Deutsche Übersetzung Seite 636)

FREQUENCY STANDARD

The Marconi Co. Ltd, Chelmsford, Essex

(Illustrated below)

An extremely high-stability oscillator, having an accuracy comparable with that of a modern atomic frequency standard, has been developed by The Marconi Co. Ltd. This device, which has a frequency variation of only 30 millionths of a cycle at 100kc/s, is one of the most stable and accurate frequency standards available in production in the world.

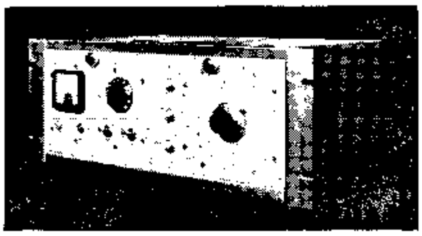
The oscillator, which has a short term frequency stability of better than 3 parts in 10^{10} and an ageing stability of 1 part in 10^9 per month, produces three simultaneous standard frequencies of 100kc/s, 1Mc/s and 2·5Mc/s.

The oscillator uses a highly polished, 2½Mc/s, 5th overtone A-T cut crystal, developed and manufactured by the company's quartz crystal factory at Hackbridge in South London. This polishing process is carried several stages further than in the production of a normal crystal. There are eight perfect interfaces within the crystal, which has a Q value of over 6×10^6. Gold contacts are thermally evaporated on to two of the outside faces of the crystal, to provide the output connexions.

The utmost temperature stability of the crystal and oscillator is ensured by enclosing them in the inner of two ovens, both of which are proportionally controlled and sensitive to temperature changes of as little as one thousandth of a degree. In addition, the crystal is operated at the extremely low power level of 0·5μW.

The crystal outlet of 0·5Mc/s is fed into the frequency division stage where 'staircase divider' circuits are used to obtain the 1Mc/s and 100kc/s outputs with high phase stability. These circuits, together with buffer output stages, use fully-transistorized printed circuit boards.

The complete unit is mounted on a standard 19in panel and contains its own power supplies enabling it to



operate from any normal mains supplies. In case of a mains failure the oscillator will continue to operate from its own built-in emergency batteries with the same frequency stability.

The accuracy and stability of this device makes it ideally suited for navigational guidance systems, computers, satellite aerial guidance, military and defence uses and many other applications in high precision systems.

EE 85 751 *for further details*

DIGITAL INDICATORS

K.G.M. Electronics Ltd, Bardolf Road, Richmond, Surrey

(Illustrated below)

This is a neat slim indicator designed for easy attachment to printed circuit boards, and serves the dual purpose of a handle for board withdrawal together with digital indication.

Known as K.G.M. type M23 indicator, this unit can display up to 10 symbols,

utilizing Inter Service type midget lamps with ratings from 6V to 28V, and incorporates the isolated common terminal feature.

EE 85 752 *for further details*

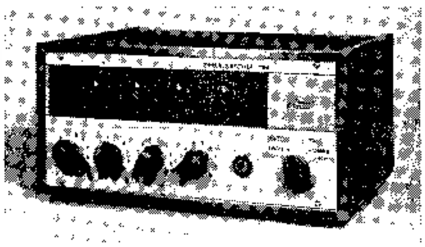
BATCH COUNTER-TIMER

Advance Controls Ltd, Cheltenham, Gloucestershire

(Illustrated above right)

Advance Controls Ltd announces an addition to its range of industrial counters by the introduction of the new TB4 timer-batch counter. This is a versatile instrument, having a maximum counting speed of 5 000 per second and using a combination of transistorized circuits and cold cathode counting tubes to ensure reliability at low cost.

The TB4 has four decades and will batch any quantity up to 9 999 or it can be used for programme control or cycle timing for any period up to 99·99sec in units of 0·01sec. An adjustable pre-batch or pre-time signal provides a warning signal or control function.



A multi-point programme plug is fitted on the rear panel and the input, output and reset arrangements can be varied to suit almost any requirement. Standard or special programme plugs can be supplied, ready for use, or they can be wired by customers from instructions contained in the instrument handbook.

EE 85 753 *for further details*

EDGE CONNECTORS

Plessey-UK Ltd, Wiring & Connectors Division, Ilford, Essex

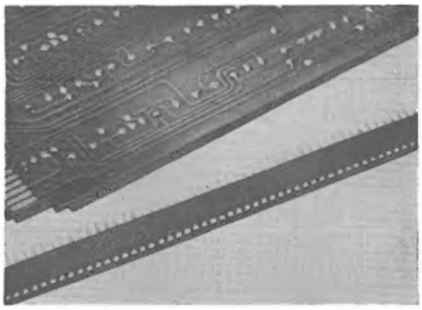
(Illustrated on page 621)

A new range of printed circuit edge connectors manufactured in strips catering for a maximum of 80 contact positions has been introduced by the Wiring and Connectors Division of Plessey-UK Ltd. Designated series 603 the connectors can be cut to the customers' exact requirements, a variety of combinations being available to meet individual printed circuit connector applications.

Series 603 connectors are supplied with fixing dimensions complying with existing standard specifications. Fixing feet are supplied to form closed end receptacles. Alternatively, where open end receptacles are required, mouldings are drilled to accept self-tapping fixing screws.

The new connectors consist of receptacles which mate directly with 1/16in thickness boards. The high performance associated with more expensive connectors is achieved by using simply-formed cantilever contacts and by making optimum use of material and space.

The contacts are specially shaped to provide good mating and contact-force characteristics. This minimizes board wear and contact resistance despite variation in board thickness. Low probe resistance is obtained by the use of brass contacts, which are designed for solder termination. Material finish is optional: hard gold, gold flash on silver or to specific requirements.



The receptacles are moulded in glass-loaded polycarbonate which possesses excellent mechanical and electrical properties and is flame retardant. A polarizing key is supplied to replace any chosen contact.

Rated at 750V a.c. r.m.s. at sea level and 5A continuous current, the connectors are designed for operation between $-40°C$ and $+85°C$, the contact resistance being less than 15MΩ.

EE 85 754 *for further details*

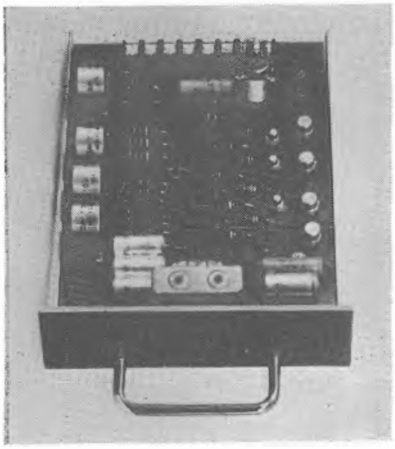
MAINS VOLTAGE MONITOR

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(Illustrated below)

Kent Precision Electronics Ltd has recently developed a new solid state circuit for the monitoring of standard three-phase voltage supplies to provide efficient, accurate and reliable voltage deviation control, protection or warning. The instrument has wide applications for purposes such as stand-by power supplies, voltage regulation and power monitoring for computer installations.

The instrument consists basically of a six-input logic circuit which operates an output trigger to switch a standard electromagnetic relay when any one of the three-phase input voltages are either above or below preset values. The instruments can be set for any nominal mains voltage with predetermined percentage deviations. The output relay contacts will therefore be in the alarm condition when any one of the three-phase voltages deviates from the preset values, to provide absolute under and over voltage protection for any equipment that might be connected to the mains supply being monitored.



The instruments are available in the standard panel-mounting instrument case or, for convenience, in the new KPE modular construction which is shown in the photograph.

EE 85 755 *for further details*

CONDUCTIVITY MEASURING SET

Electronic Switchgear (London) Ltd, 58 Wilbury Way, Hitchin, Hertfordshire

(Illustrated below)

A new portable conductivity measuring set has recently been added to the range of conductivity instruments manufactured by Electronic Switchgear (London) Ltd.

A feature of the type MC3 instrument is a virtually indestructible measuring cell that plugs directly into the side of the instrument case and which requires no maintenance other than occasional cleaning by means of a brush provided. The cell has a tubular, unrestricted bore of 10cm³ capacity, with annular elec-

trodes which permit the conductivities of liquids, pastes and slurries to be measured with equal facility.

The range of measurement is 1 to 100 000 micromhos/cm in three steps, as selected by a switch that is provided with a fourth position which enables the effective operation and accuracy of the instrument to be checked at will.

The transistorized printed circuits are powered from a small, long life, easily replaceable battery that is contained within the die cast case which measures 9½in by 7½in by 3½in.

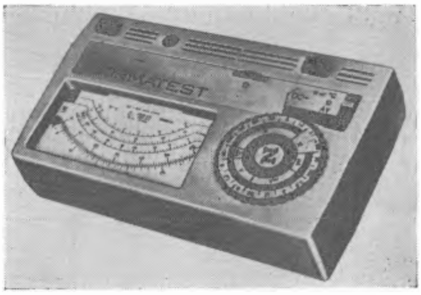
EE 85 756 *for further details*

MULTI-RANGE METER

Distributed by: Croydon Precision Instrument Co., Hampton Road, Croydon, Surrey

(Illustrated above right)

This multi-range meter, manufactured by Norma in Vienna, has forty measuring ranges covering a.c. and d.c. voltage and current, resistance and output. In addition it contains a 12mV d.c. range of 40kΩ/V sensitivity and a temperature scale in degrees Celsius for connexion to iron/constantan thermocouples.



The sensitivity on d.c. is 20kΩ/V and on a.c. 4kΩ/V. D.C. accuracy is ±2·5 per cent and 3·5 per cent on a.c. to 500c/s and ±7·5 per cent at 30kc/s.

The instrument weighs ¾ lb, has a taut suspension movement and is capable of withstanding an overload of 1 000 per cent.

EE 85 757 *for further details*

REGISTRATION DEVICES

Benson-Lehner Ltd, West Quay Road, Southampton

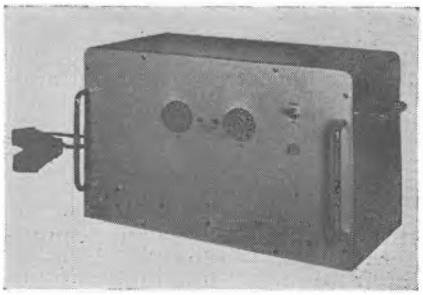
(Illustrated below)

Benson-Lehner Ltd has introduced two new photo-registration devices for use primarily in packaging, wrapping and bagging machines.

Both units are completely transistorized, without relays, contactors or moving parts of any kind, allowing maximum operating speeds to be employed and eliminating much of the down time caused by dirty atmosphere or the need for adjustments.

Type 2C5A immediately responds to any variation in the normal state of registration, either advancing or retarding an auxiliary electric motor or other device to be controlled, depending upon the position of the registration marks.

The other unit is a uni-directional correction system for incorporation in machines where the wrapping material is deliberately overdrawn on each cycle. The correction system performs the task of retarding the feed whenever the registration mark obscures the photo-electric cell.



EE 85 758 *for further details*

TRANSISTORIZED POWER UNIT

Fenlow Electronics Ltd, Springfield Lane, Weybridge, Surrey

(Illustrated on page 622)

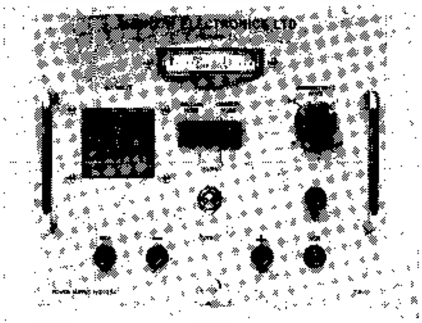
The PU5 power unit is designed for operation from a.c. mains of 200 to 250V or 100 to 150V, 50c/s or 60c/s, the maximum power consumption being 290W. The output is adjustable from

0 to 150V d.c., stabilized at up to 1·5A.

These units can operate in two modes, constant voltage and constant current, and are equipped with 'automatic mode' crossover circuits and visual mode indication and are protected against accidental short-circuiting of the output terminals.

In the voltage mode the output voltage is set in increments of 1V by means of calibrated decade switches and the current drawn by the applied load is indicated on the panel meter. In the current mode the current is set by a panel knob which is marked for quick reference and by means of the current meter. Transition from one mode to the other occurs automatically as the external load is varied through the critical value and two coloured lamps signal the mode change. Provision is made for remote sensing of voltage to avoid the effects of impedance in connecting leads. In addition, voltage may be remotely programmed and units can be connected in series or parallel to augment voltage or current output.

The equipment is supplied as a 'half-rack' unit enabling a pair to be mounted in a standard 19in rack by means of a frame assembly.



EE 85 759 *for further details*

PRINTED CIRCUIT PUSH-BUTTON SWITCHES

Mills & Rockleys (Electronics) Ltd, Swan Lane, Coventry

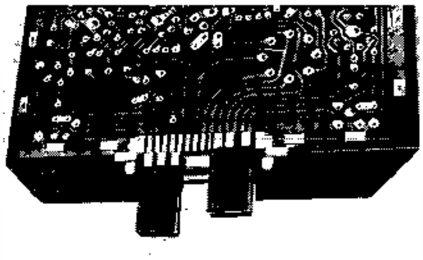
(Illustrated above right)

This switch is designed specifically to be assembled to printed circuit panels in component part form. This obviates the necessity for buying a more complex pre-assembled switch unit which is almost impossible to remove from the printed circuit in cases of failure or other defects common to this type of switch.

The switch relies upon the design of the printed circuit having the appropriate conductor pattern to coincide with the contacts carried by the push-buttons to make or break as required.

The push-buttons, carrying bridging contacts, co-operate with the printed circuit pattern within a slot extending into the push-button which is parallel to the flat surface of the printed circuit panel.

Each push-button is fitted with an internal spring so that on depression of another button any push-button already engaged which is not depressed at the same time returns to its original position.



This is achieved by the design of the shape of the push-button moulding, in conjunction with the retaining bar which is attached to the printed circuit panel.

The assembly of the component parts therefore becomes simply a matter of fitting the push-button mouldings complete with their contacts and springs to the printed circuit panel and being retained by the appropriate length of bar determined by the number of push-buttons required. The design will also permit variable spacing of the buttons if this is preferred on any equipment.

The push-buttons will be standard items, and it will be possible for the user to adopt any number of these since it will only be necessary to select the appropriate length of carrier bar and design the printed circuit to suit. It will therefore be a normal part of Mills & Rockleys' printed circuit production facility to provide the appropriate tooling for the printed circuit, manufacture the circuit panels, and supply or assemble the component parts of the push-buttons to match the panels.

EE 85 760 *for further details*

DIGITAL VOLTMETERS

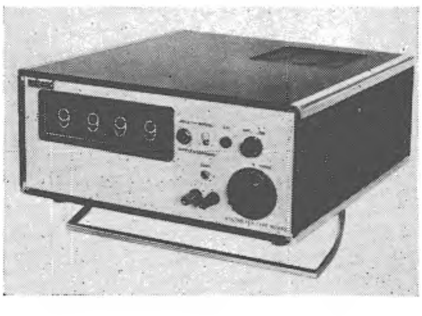
Roband Electronics Ltd, Lowfield Heath Road, Charlwood, Horley, Surrey

(Illustrated below)

Roband Electronics Ltd has introduced two new ranges of compact digital voltmeters. The two basic instruments are the RDV 32, and RDV 42, each of which is available as fully cased, rack mounted or front panel types. Remote read-out units, a serializer and single-range voltmeters for specific duties are also made.

Both instruments are successive approximation types, operating at conversion speeds above 30 per second; they offer decimal or binary coded decimal output signals for control or print-out purposes.

The RDV 42 reads (full scale) from 0·9999 to 999·9V in four ranges with a resolution of 100μV. The RDV 32 reads (full scale) from 1·099 to 999·0V in four ranges with a resolution of 1mV. Both instruments have a minimum accuracy of 0·1 per cent ±1 digit.



Using solid state circuits throughout, these instruments are designed to meet the most arduous industrial conditions while sealed versions are available for use in corrosive atmospheres.

EE 85 761 *for further details*

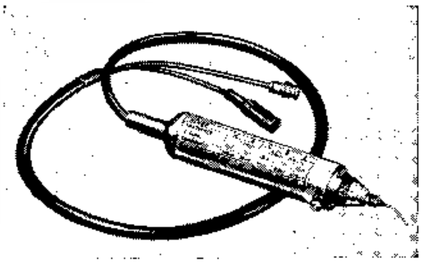
COUNTER PROBE

Racal Instruments Ltd, Crowthorne, Berkshire

(Illustrated below)

Racal Instruments Ltd announce the availability of a new counter accessory—the Active Probe Unit type SA.544. This all-transistorized probe is for use with the Racal direct-reading digital counters types SA.540 and SA.550 for 10Mc/s and 100Mc/s frequency measurement.

Use of the active probe enhances counter performance, typically increasing counter output impedance to 10MΩ in five ranges and sensitivity to 5mV, making possible many measurements over the 100Mc/s frequency range which were hitherto only possible using special external amplifiers and impedance-matching devices. The high inherent



accuracy of the counter is retained for all measurement functions.

The probe, which makes exclusive use of semiconductor devices, is powered directly from the associated counter. It is of rugged construction and designed to withstand the severe environmental operating conditions imposed by military application—N.A.T.O. code reference 6625-99-519-1573 has been allocated.

EE 85 762 *for further details*

SEVEN TRACK RECORDING HEAD

Gresham Lion Electronics Ltd, Lion Works, Hanworth Trading Estate, Feltham, Middlesex

(Illustrated on page 623)

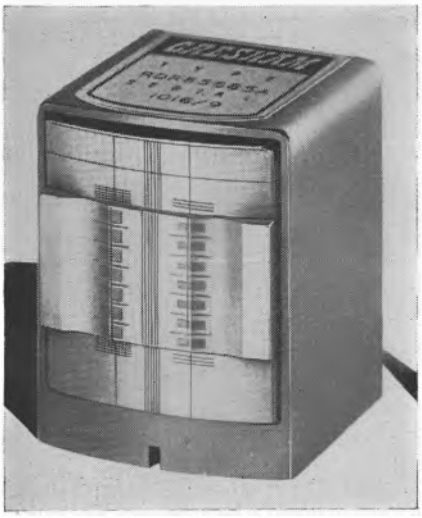
Gresham Lion Electronics Ltd has developed a seven track read after write magnetic recording head. This head is designed for digital applications and gives immediate verification of recorded information.

It is available in two versions; with record elements on the left, for compatibility with IBM equipment, or with record elements on the right. The front face is profiled to minimize long wavelength noise and is cut back beyond the track to prevent grooving by tape edges.

The head measures 1¼in high, 1 1/32in deep and 1in wide; azimuth is held within ±·0002 radians and elevation within ·0005 radians. Gap scatter is contained within a band ·00008in wide and the gap lines are parallel to within ·0002 radians. The read tracks are 0·031in wide at intervals of 0·070in and the write tracks are 0·046in wide at the same interval.

Saturation current on 3M T488 tape at 300 bits to the inch and 99 per cent saturation is 36mA ±15 per cent peak-to-peak. The read output, with a write current of 54mA peak-to-peak, is 44mV peak-to-peak at a tape speed of 120in/sec.

The feed-through when writing on seven tracks simultaneously in phase with a 100kc/s square wave signal having a rise time of 2μsec and a current value



of 60mA peak-to-peak is less than 10mV on to any read track. This figure can be reduced by a ratio of at least 2 : 1 by a screen covering the head face and placed immediately behind the tape.

EE 85 763 *for further details*

HEAT SINKS

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

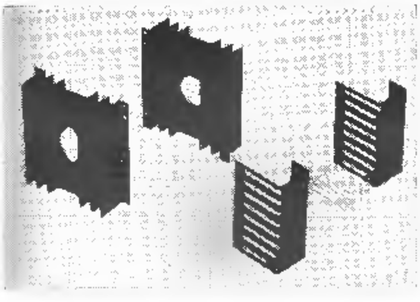
(Illustrated above right)

A.P.T. Electronic Industries Ltd, manufacturers of the Lektrokit rack and chassis systems, has recently added to the Lektrokit range a series of heat sinks. These are available in four models.

Heat sink number 1 (Part No. LK 1111) is designed to mount the internationally standard TO-3 outline transistor, and is shaped to provide a wide range of alternative mounting arrangements, both singularly and in groups. The slotting pattern permits free air circulation throughout the heat sink, thus allowing efficient cooling at any angle. The transistor connexion holes are elongated, enabling two heat sinks to be bolted back-to-back with one transistor, greatly increasing the dissipation rating.

Heat sink number 2 (Part No. LK. 1121) is of identical outline, but with mounting holes for the larger power transistors of the TO-36 outline, also for power Zener diodes and rectifiers.

Heat sinks 1 and 2 are manufactured from 18 gauge high conductivity alumi-



nium, in a folded and slotted form. They are ready pierced for power transistor mounting, and are finished in black to aid heat dissipation. Their thermal resistance in free air is approximately 3·5°C/W.

Heat sinks 3 and 4 (Part Nos. LK.1131/1141) are manufactured from solid aluminium finned extrusions, drilled for transistor mounting and are also finished in black. The thermal resistance of these units is approximately 1·6°C/W, mounted vertically, and 1·75°C/W, mounted horizontally in free air. They are designed for maximum thermal capacity and heat dissipation. Heat sink No. 3 mounts the internationally standard TO-3 outline transistor, and No. 4 covers the larger power transistors of the TO-36 outline, as well as power Zener diodes and rectifiers.

EE 85 764 *for further details*

PRESSURE TRANSDUCERS

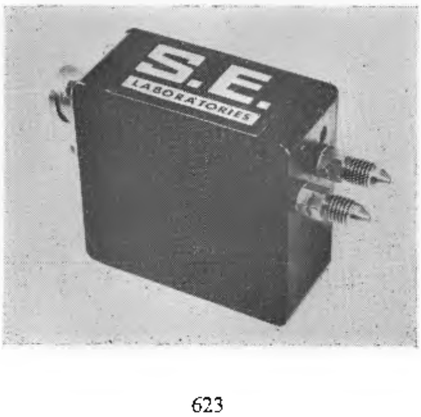
S.E. Laboratories (Engineering) Ltd, North Feltham Trading Estate, Feltham, Middlesex

(Illustrated below)

A range of pressure transducers type S.E. 37 is offered by S.E.L. as self-contained d.c. operated pre-calibrated units. Application is intended primarily for rocket, supersonic and general aircraft systems, and in environments where most other transducers fail due to extreme conditions of vibration.

They have very high accuracy and reliability. Their voltage or current output is for direct connexion to various equipment, e.g. tape recorders, oscillographs, telemetry equipment, high level voltage controlled oscillators, commutators, digitizers or direct reading indicators.

Fully integrated, welded circuits ensure high reliability and continuous operation at temperatures of 150°C with less than 0·015 per cent/°C temperature error, and for three minutes' intermittent opera-



tion, as high as 300°C with a total error of less than 0·35 per cent.

Noise effects due to ground loops and large circulatory currents, e.g. aircraft and ground testing installations, are reduced to a minimum by means of full isolation from d.c. supply lines and output circuits.

The d.c. operated transducer S.E.37 is suitable for differential gauge or absolute pressure applications with corrosive gases or liquids, with pressure ranges between 0 to 2 and 0 to 4 000 lb/in^2. The voltage output of −5V to +5V for a full-scale pressure can be pre-set to customers' own requirements to swing from either zero to maximum or from negative to positive (e.g. 0 to 1 000 lb/in^2, 0 to 5V; or −1 000 to +1 000 lb/in^2, 0 to ±5V). The welded construction and exclusion of all rubber and plastic seals makes this transducer suitable for use with all fluids compatible with stainless steel.

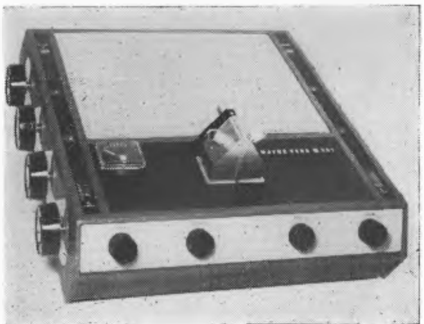
EE 85 765 *for further details*

PRECISION R.F. BRIDGE

The Wayne Kerr Laboratories Ltd, Sycamore Grove, New Malden, Surrey

(Illustrated below)

This new wide-range bridge provides 0·1 per cent accuracy for capacitance and conductance measurements between 100kc/s and 1Mc/s, and it can be operated with reducing accuracy up to 5Mc/s. The B201 is fully portable, including battery (or rectifier), plug-in source and detector units, standards and a null meter. Internal modulation at 1kc/s and a detector outlet jack provide facilities for headphones, and front-panel controls permit the signal level and detector gain to be adjusted. Overall



coverage is 0·0001pF to 0·1μF and 0·001μmho to 1mho (1Ω to 1 000MΩ).

Design of the instrument is based on transformers for accurate voltage and current ratios by means of which the unknown is compared with internal capacitance and conductance standards. Provision of a neutral terminal on the special design of connexion block enables three-terminal measurements to be made. Frequencies of the plug-in source and detector units at present available are 100kc/s and 1Mc/s, but dummy units permit external equipment to be connected for measurements over the range 100kc/s to 5Mc/s.

EE 85 766 *for further details*

MULTI-RANGE METER

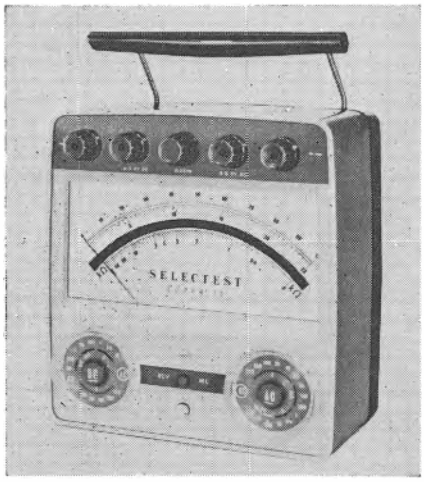
Salford Electrical Instruments Ltd, Peel Works, Silk Street, Salford, 3, Lancashire

(Illustrated below)

The Selectest mark II is a development of the original design with additional refinements and modifications suggested by engineers in the field.

The terminal board now incorporates terminals with sockets to take plug-in leads. The range-change switches have been re-designed to give even greater legibility and simplified range selection. They can be rotated in either direction through 360° and are electrically interlocked.

The automatic overload cut-out and internal components are mounted on a robust printed-circuit board enclosed in a tough melamine case. Two models are available, the 'Selectest Super 50' has a basic movement of 50μA and 32 internal ranges for measuring a.c. and d.c., voltage current and resistance. The 'Select Super K' has a 1mA movement with 39 internal ranges.



EE 85 767 *for further details*

MINIATURE DIGITAL DISPLAY

Counting Instruments Ltd, 5, Elstree Way, Boreham Wood, Hertfordshire

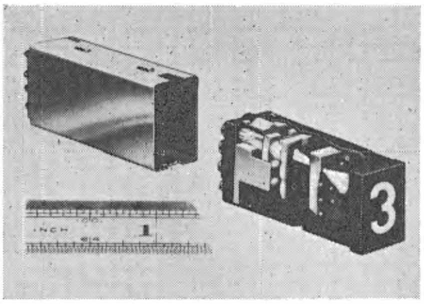
(Illustrated above right)

Counting Instruments Ltd announce a new micro-miniature, rear projection digital display unit measuring ⅞in by ½in wide by 2in which, despite its small size, projects a figure ⅜in high.

Similar in operation to their 3¾in, 1in and ⅝in twelve-lamp displays, this new unit contains an assembly of eleven miniature incandescent lamps at the rear, a series of lenses and a front viewing screen, on to which eleven messages can be projected individually or in combination. As these messages are on film, the unit can display any numbers, letters, words, symbols or colours that can be reproduced photographically.

The display is an insertable unit. This plugs into a housing incorporating the lamp connexions, which can be permanently wired. The insert can be removed from the housing through the front of the panel for easy lamp renewal, or to be replaced by an insert with different messages.

There are four different types of



insert available: sub-panel mounted for use with a continuous screen in multiple assemblies; with individual screen and mounted flush with the panel; or with individual screen and with either full or partial bezel frames. Each type can be grouped to form a multiple assembly.

EE 85 768 *for further details*

STANDARD REFERENCE MAGNETS

Preformations Ltd, Cheney Manor, Swindon Wiltshire

(Illustrated below)

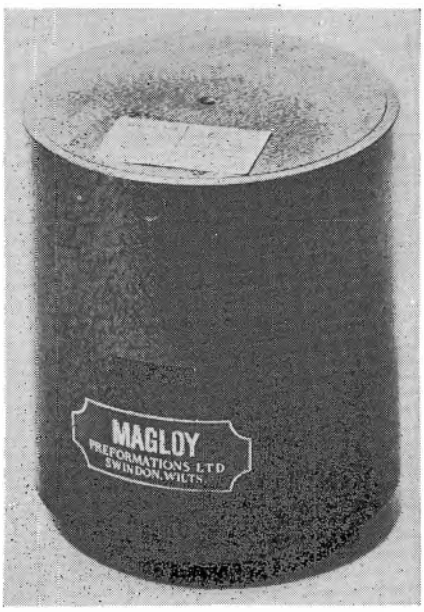
A range of fully screened standard reference magnets is now available from Preformations Ltd, a company in the Plessey Components Group.

Designed for the calibration of search coils, fluxmeters, Hall-effect and similar gaussmeters, these magnets were originally intended for use with a standard size of search coil made by W. G. Pye & Co. Ltd of Cambridge. However, they are equally suitable for the calibration of any search coils or Hall-effect probes with a maximum diameter of 10mm and maximum thickness of 4·75mm.

The reference magnets are fully stabilized, both magnetically and by temperature cycling, during production to guard against irreversible losses resulting from temperature changes and external magnetic fields.

Reference is made to a nuclear magnetic resonance magnetometer for calibration purposes.

Magnets with either transverse or axial



fields are available, with the following flux densities:

Axial—300 and 500 gauss.

Transverse—1 000, 3 000, 5 000, 10 000 and 15 000 gauss.

Accuracy is within ±0·5 per cent of marked flux density at 20°C. Magnetic stability is such that magnets will not be permanently affected by stray fields below 150 oersteds and temperature stability is of the order of 0·2 per cent/°C between −10°C and +90°C. Maximum operating temperature is 100°C.

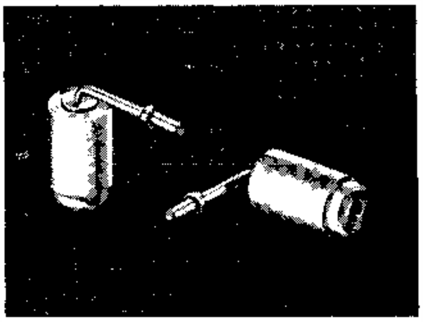
EE 85 769 *for further details*

PRINTED CIRCUIT TEST JACKS

Sealectro Ltd, Walton Road, Farlington, Portsmouth, Hampshire

(Illustrated below)

The SKT-105 PC L1 is a new p.t.f.e. test point jack designed for insertion into printed circuit chassis. The contact end of this new unit is at right-angles to the printed circuit board and is designed to stand higher off the chassis surface than more conventional units. The additional height above chassis permits the interpositioning of test point



jacks such as SKT-103 PC and increasing the number of units that can be mounted to the chassis. The SKT-105 PC L1 features pure p.t.f.e. bushings and the beryllium copper contacts are gold flashed over silver plate.

EE 85 770 *for further details*

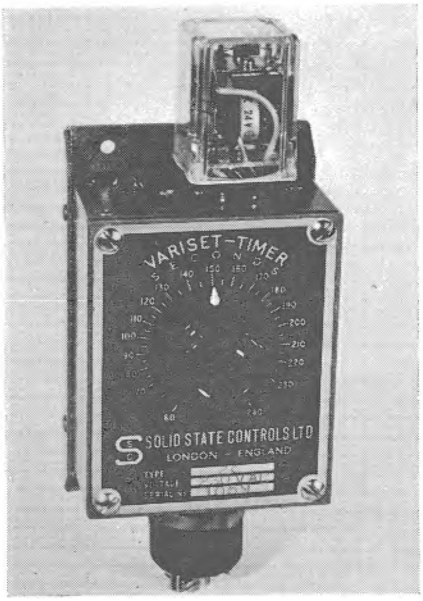
ELECTRONIC TIMERS

Solid State Controls Ltd, 30-40 Dalling Road, London, W.6

(Illustrated on page 625)

The 'Variset' is a completely integral timer with a plug-in relay output capable of switching power loads of 5A as standard. The basic unit is an all-transistor timing network with a life of up to 100 million operations. It is adequately protected against high voltage transients and is not affected by wide variations in voltage and working temperature. The unit is therefore very reliable even in adverse industrial conditions and has a typical repetitive accuracy of 1 per cent. A standard dust-proof plug-in relay is the only wearing part and is readily accessible.

The setting dial is directly calibrated in seconds and specific times may therefore be set with ease. Time delay periods from 0·2sec to 4min are offered in three timing ranges while eight standard operating voltages are available within the range 12 to 110V d.c. and 48 to 415V a.c.



The Variset series is provided with two changeover contacts rated at 5A 250V a.c.; other ratings can also be supplied. Four basic types of operation are provided; delay on energize; delay on de-energize; push-button starting, or re-cycling. Additionally the good recovery characteristic allows instant and accurate re-setting at any time during, or at the end of, the timing period.

EE 85 771 *for further details*

MINIATURE WIREWOUND POTENTIOMETER

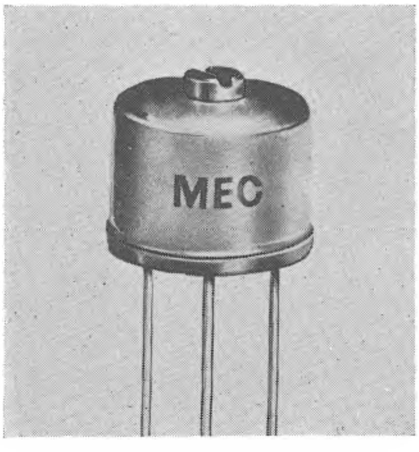
Miniature Electronic Components Ltd, Copse Road, Woking, Surrey

(Illustrated below)

Primarily designed for printed circuit use on a 0·1in module, the T2OP is of TO-5 transistor style and size, the case measuring 0·345in diameter by 0·255in high. It is suitable for aerospace and industrial applications.

It has single-turn vertical adjustment with positive end stops which minimizes setting time; the shaft is self-locking; the metal case and element former provide maximum heat dissipation; low inductance permits use for pulsed or r.f. applications.

Standard Resistance Values: 50, 100, 200, 500Ω; 1, 2, 5, 10kΩ.



Power Rating: ½W at 70°C.
Temperature Range: −55°C to +150°C.

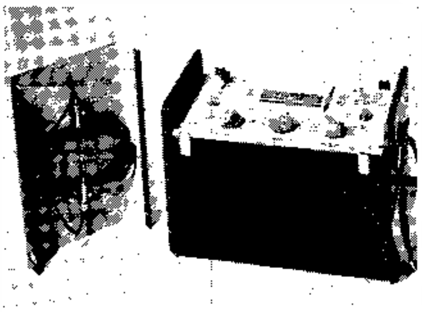
EE 85 772 *for further details*

IMPACT NOISE ANALYSER

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustrated below)

The Dawe type 1412 impact noise analyser is a portable battery-operated instrument for measuring the peak amplitude and analysing the characteristics of impact noise. Used in conjunction with the Dawe range of sound level meters, the type 1412 gives a direct reading of the peak intensity of the noise, and information from which the duration and waveform of a single impact can be estimated. A 'quasi-peak' circuit is used for repetitive sounds, and octave band filters may be connected externally to analyse the frequency composition. In addition, the instrument may be used for vibration work by feeding it from a suitable accelerometer; it can analyse an impulsive waveform from any source capable of supplying 1 to 10V peak across a 50kΩ input impedance within the frequency range 5c/s to 40kc/s.

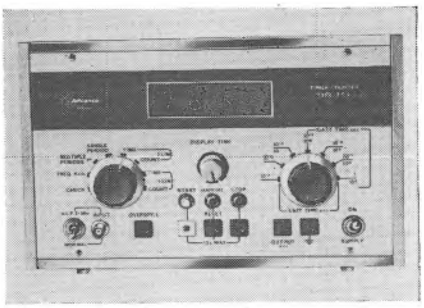


The type 1412 consists of a valve amplifier driving three signal rectifiers and a d.c. valve-voltmeter which can be switched to each of the rectifier circuits in turn. These rectifier circuits provide the peak, quasi-peak and time-average signals from which the shape and magnitude of the waveform can be determined.

The peak circuit has an extremely short rise time; for a rectangular pulse the indication rises to within 1dB of the final value in less than 50μsec. The charge is maintained within 1dB for at least 10sec. The quasi-peak circuit has a fast rise of less than 1msec, and a slow decay time of about 0·6sec. The characteristic is substantially as specified in BS 727:1954 'Characteristics and performance of apparatus for measurement of radio interference'. The time-average circuit gives an indication of the level maintained over a specified period, time-constants between 2 and 200msec being provided by changing a resistor. Under normal conditions the charge is maintained within 1dB for at least one minute.

The instrument is supplied in a portable metal cabinet with detachable cover and neck strap.

EE 85 773 *for further details*



COUNTER-TIMER

Advance Electronics Ltd, Roebuck Road, Hainault, Ilford, Essex

(Illustrated above)

Direct frequency measurement in excess of 5Mc/s is a feature of the specification of the new Advance TC4 timer counter.

Four gate times and seven timing units from 1μsec to 1sec enable accurate measurements to be made of frequency, period and time. The instrument has a four-digit in-line display with a decimal point automatically positioned for frequency measurement. The period facility (up to 10^5 periods) permits accurate measurements of extremely low frequencies.

For time and frequency measurements the basic accuracy of the TC4 is governed by an oven-controlled crystal oscillator set to one part in 10^6 at ambient temperature. Should greater accuracy be required, an external standard producing sinusoidal or pulse signals can also be used.

EE 85 774 *for further details*

MINIATURE WIREWOUND POTENTIOMETER

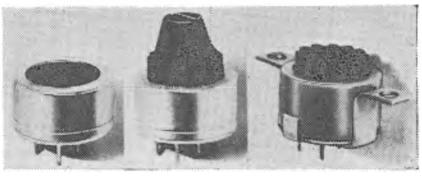
Darstan Ltd, 263 Church Road, Thundersley, Essex

(Illustrated below)

The compact design of this potentiometer makes it suitable for use in printed circuits and in other applications where space is at a premium.

Of wire-wound construction with metal screening case, this component has a slotted cap with positional indication which is easily adjusted by screwdriver. Leads are on standard printed circuit 0·1in module.

The leads provide easy mounting on printed circuit boards—no other form of fixing being necessary. A pre-set knob is available and the potentiometer may be fitted with a front bracket for panel mounting.



Standard resistance values are 5, 10, 20, 50, 100, 200, 300, 500Ω, 1, 2, 5kΩ. Standard resistance tolerance is ±10 per cent and power rating 1W.

EE 85 775 *for further details*

Short News Items

A symposium on electronics, measurement and control in ships and shipbuilding will be held at the University of Strathclyde, Glasgow, on 12 to 15 April, 1966.

The meeting is being organized jointly by the Electronics and Control Section of the Institution of Electrical Engineers Scottish Centre and the Scottish Section of the Institution of Electronic and Radio Engineers.

Contributions within the following categories are invited for inclusion in the programme:

Navigation
Engine control and telemetry
Automatic loading and discharging
Economics, reliability and maintenance
Use of computers and numerical control
Safety aspects
Research and development

Residential accommodation will be available and further information and registration forms can be obtained from the Joint Honorary Secretaries, K. A. Murphy, 50 Holeburn Road, Newlands, Glasgow S.3 and R. D. Pittelo, 35 Crawfurd Road, Burnside, Rutherglen, Glasgow.

The Hungarian Society for Optics, Acoustics and Film Techniques is to hold its Second Conference on Magnetic Recording at Budapest on 11 to 15 October 1966.

Papers are invited for presentation at the Conference dealing with (a) Magnetic Recording on Moving Magnetic Media (b) Static Magnetic Memories and (c) Applications, and should be submitted, together with an abstract of not more than 150 words, not later than 1 December 1965. Official languages of the Conference will be English, Russian, German and Hungarian.

Further details are obtainable on application to the Society at Budapest 196, V. Szabadsag ter 17.

Standard Telephones & Cables Ltd has received an order from the GPO for an extension to the London-West Country microwave communications network.

The network, which is capable of carrying several thousand simultaneous telephone conversations, is at present under construction between London and Bristol and between Bristol and Cardiff. Next year work will begin on the Bristol-Plymouth-Goonhilly leg, of which the new Bristol-Exeter project forms a part. In addition to the circuits already planned between Bristol and Plymouth, two r.f. channels will be added (using the aerials already planned) on the route from Bristol as far as Whitestone, a switching point a few miles outside Exeter.

The equipment to be supplied for the new link will be STC's R16D main-line microwave system in which eight 'narrow-band' radio channels with 40Mc/s spacing can be accommodated on a single Cassegrain aerial, or alternatively sixteen radio channels with 20Mc/s spacing can be accommodated on two aerials using cross polarization.

Each radio channel, operating in the upper 6Gc/s band (6 425 to 7 100Mc/s) carries either a 625 line monochrome or colour television channel, or 960 telephone circuits. The system is therefore capable of expansion to 8 080 telephone circuits using one aerial, or 16 160 circuits using two aerials.

The British Radio Valve Manufacturers' Association (BVA) and the Electronic Valve and Semiconductor Manufacturers' Association (VASCA) announce the following sterling value of their members' sales during the three months ended March 31, 1965:

Valves & Tubes	£11·2M
Semiconductor devices	£ 6·7M
Total:	£17·9M

The total value of the sales for the whole of 1964, published in April last, was £72M.

The Scientific Instrument Manufacturers' Association (SIMA) is holding its first group exhibition in Rumania on 26 October to 2 November this year.

The exhibition will be held in the New Exhibition Centre, Bucharest and 26 companies will be participating. It will be the largest foreign instrument exhibition ever held in Rumania and will feature the latest instrumentation for industry, research and education.

In conjunction with the exhibition there will be a programme of lectures which will be held in the lecture hall at the Academy of Science by kind permission of the President of the Academy.

Ferranti Ltd has entered into an agreement with SAGEM (Societe d'Applications Generales d'Electricite et de Mecanique) of Paris whereby the two firms will pool their knowledge and experience in the design and development of aircraft navigation systems and components.

This co-operation has already borne fruit in the recent award of a contract to Ferranti of the attitude reference system and ground test equipment for the European Satellite Launch Vehicle (Eldo) in which the gyroscopes and accelerometers incorporated in the equipment will be produced by SAGEM. In addition, Sud Aviation (joint constructor of the Concorde with BAC), has chosen to equip the Concorde with an inertial system developed and manufactured by SAGEM and Ferranti.

The International Exhibition of Electronic Components (Salon International des Composants Electroniques) will be held at the Parc des Expositions, Porte de Versailles, Paris on 3 to 8 February, 1966.

Further details are obtainable on application to S.D.S.A., 16 rue de Presles, Paris, 15e.

The International Fair for Electronics, Automation and Instruments at the Forum Exhibition Hall Copenhagen, will now be held from 25 February to 6 March 1966 and not in mid-April as previously arranged.

The Science Research Council has made a grant of £273 250 to the Control Group of the Engineering Department, Cambridge University under Mr. J. F. Coales, for a fundamental investigation into adaptive computer control of complex processes. The grant runs from 1 July, 1965 to 31 December, 1967.

The research programme is aimed at developing general methods for the control of such plants as chemical reactors, oil refineries, steel mills, power stations, paper mills, sugar extraction and food processing plants, using computers of reasonable size. Such control is known, theoretically, to be possible but it requires the determination of an accurate mathematical model and the use of a very large and expensive computer online unless approximate methods of control can be developed.

The plant model-making work cannot be done on an actual plant without interfering with production and safety. Also it would take too long, and too few plants are within reach of sufficiently large computing facilities. Therefore experiments will be carried out at a very much reduced time-scale through simulating the plant as a model on the high speed analogue computer which in turn will be controlled by the on-line digital computer. The development of a mathematical model of the plant would not interfere with the day-to-day running of the industrial plant. Additional instruments will record sufficient data to calculate the accurate mathematical model.

Telefunken has received an order from the Dutch P.T.T. for the first stage of a radio link network in the Netherlands. Equipments of the new radio link system FM 6000-TV/7400 will be supplied which are transistorized throughout with the exception of a klystron in the transmitter stage, and which can handle either 600 telephone channels or one television channel with sound channel.

In addition Telefunken is to install for the Swedish and Danish P.T.T. a radio link across the Oresund between Malmo and Hillerod. In this case equipments of the system FM 1800-TV/6000 will be used which are likewise fitted with transistors throughout with the exception of a travelling wave tube in the transmitter stage. The radio link will be able to carry either 1 800 telephone calls or a television channel with up to four sound channels. Since the path crosses the sea space diversity methods will be used in order to improve the reliability.

The Electronic Engineering Association (E.E.A.) has formed a Divisional Organization with the object of expanding home and overseas markets for British Electronic capital equipment.

Based on the range of potential markets for the equipment for which the E.E.A. accepts responsibility, the Divisions concerned are: Aviation, Space, Maritime, Computers, Industrial Electronics, Radio Communications and Broadcasting.

The British Standards Institution has recently published its recommendations on reverse blocking triode thyristors.

The recommendations cover the essential ratings, characteristics and methods of measurement of thyristors.

The data quoted should make it easier to compare thyristors of the same general class offered by different manufacturers, and, by defining the ratings and characteristics in sufficient detail, enable users to determine the suitability of particular types for circuit design purposes.

Methods of measuring the electrical characteristics and of assessing the working conditions are also given.

Copies of B.S. 3896 *Recommendations on reverse blocking triode thyristors* may be obtained from the BSI Sales Branch, 2 Park Street, London W.1. Price 15/- each (Postage will be charged extra to non-subscribers).

Elliott Traffic Automation has received a contract worth over £60 000 from British Rail for the manufacture and installation of the new electronic system which will display 4-character, alphanumeric train headcodes on cathode-ray tubes inside a signal box. The headcodes, or train descriptions, are transferred automatically from one section of the display to the next as the train moves along the track.

The equipment is to be installed at signal boxes at Eastleigh, Basingstoke and Woking of the Southern Region and will incorporate ferrite core stores and micro-miniature logic elements of the type used in advanced computer systems.

Six smaller signal boxes in the same area are also to be equipped with the new system. The memory systems to be installed in these boxes will consist of bistable electronic elements, mounted on printed circuit boards, as the comparatively small number of displays does not warrant the use of ferrite core stores.

The new Ghana Television Service which has now been brought into operation was designed, built and installed by the Marconi Co. Ltd in collaboration with the Ghana Broadcasting Corporation.

The new service has three television transmitting stations at Accra, Kumasi and Sekondi-Takoradi. These stations cover approximately one quarter of the country including the Atlantic seaboard, the most heavily populated part of Ghana. Each of the three stations is served by a central studio complex in Accra, the capital.

The three television stations use the Marconi 5kW vision transmitter, type B7101. This transmitter will operate on any chosen frequency in Band 1 and has been designed for ease of operation and maintenance. It can be very simply adapted to radiate colour television signals.

The associated 1kW sound transmitters, type B6550, employ frequency modulation and are also designed for ease of operation and maintenance.

Marconi three stack, turnstile aerials mounted on 400ft masts at Accra and Kumasi and a 300ft mast at Sekondi-Takoradi are used. These aerials radiate omni-directionally.

The Digital Systems Department of Ferranti Ltd at Bracknell, Berks, has been advised by the Swiss Government that it is the recommended contractor for the manufacture and supply of data link equipment, spare parts and test equipment, in connexion with the new Air Defence System—Florida—which is to be installed.

Ferranti Ltd commenced the development of digital data link equipment for military purposes in 1954, and has been responsible for some of the major data link projects for the British Government. In addition to this wide range of military projects, a number of civil data links has been developed both in the 1 000 baud range, and the higher speed range of 60 000 bauds. In the high speed range a data link has been installed which connects an adaptive control experiment at the Imperial College of Science and Technology, London, directly to London University's I.C.T. Atlas computer at Gordon Square. The data rate is 60 000 bits/sec checked, with a transmission rate of 500 000 bauds and a line bandwidth of 10kc/s to 3Mc/s. All the civil data links incorporate an error checking system to ensure error free transmission.

The Ministry of Aviation has ordered a further instrument landing system for London Airport from Pye Telecommunications Ltd of Cambridge. When installed, five of Heathrow's six runways will be equipped with i.l.s.

The recent automatic landing by a British European Airways 'Trident' underlines the importance of this international landing aid. Pye Telecommunications, manufactured the i.l.s. equipment on which the first completely automatic landing was made at the Blind-Landing Experimental Unit at Bedford and on which the Trident carried out its test landings at Hatfield. The installation at London will be used for automatic landing of passenger carrying aircraft.

Royston Instruments Limited, a member of the Royston Group of Companies, has received a contract to install Midas data recording systems in the Breguet Atlantic aircraft now being manufactured for NATO.

The Atlantic, a twin-engined, turboprop aircraft for anti-submarine warfare and maritime reconnaissance missions, has been ordered in quantity by the French and German Governments; it is also to be used by Holland and Belgium. It is the seventeenth type of aircraft to be fitted with Midas magnetic tape data recorders which are capable of recording for periods up to sixty hours and up to 1 305 individual channels of information from all parts of an aircraft. They are supplied mainly for maintenance purposes, but are also available as crash recorders.

In addition to its use in RAF Transport Command aircraft, Midas recording equipment is also being supplied to the USAF, Air Canada, and PIA;

Russia and Egypt have also placed orders.

Among the types of aircraft in which Midas is already installed are the BAC One-Eleven, VC-10, Trident, Vanguard, Britannia, Boeing 707, Douglas DC8 and C133, Convair 880, Boeing B52 Bomber, Vulcan, Victor, Avro 748 and Constellation.

The Industrial Control and Electronics Division of the British Electrical and Allied Manufacturers' Association (BEAMA) has recently published a 'Guide to the Specification and Purchase of Electronic Equipment for Industrial Systems.'

The Guide specifies standard types of equipment to meet the requirements for various applications and other facets of operation which need to be considered within the concept of a system.

The Automation Committees of BEAMA, in consultation with representatives from the user industries, will now decide on the supplementary information required for the complete coverage of the particular industry concerned. This supplementary information will be published in subsequent editions.

Copies of the first edition of the Guide are available from the BEAMA Information Division, 8 Leicester Street, Leicester Square, London, W.C.2. (Price 21/-.)

A Conference on Hall Effect Applications will be held at Cambridge, Massachusetts, on 8 to 9 November 1965. Timed so as to follow directly after the large general electronics conference, NEREM, in Boston, the meeting will deal with new applications of the Hall effect, magneto-resistance, and other galvanomagnetic phenomena, and with the problems of materials, theory, device design (including magnetic circuits) which have limited the application of these devices.

Further details are available on application to W. Crawford Dunlap, General Chairman, Conference on Hall Effect Applications, NASA Electronics Research Centre, 575 Technology Square, Cambridge, Massachusetts, USA.

The Solartron Electronic Group Ltd has introduced its first marine radar simulator specifically designed for training applications on board ship. Developed by the Company's Military Systems and Simulation Division, it is a small, fully transistorized equipment suitable for bulkhead mounting by the mercantile marine or world's navies.

It will provide training in anti-collision problems and ship, helicopter or aircraft direction while a ship is on passage or in harbour.

The new simulator can be fitted without the need for modification of the ship's radar equipment, and can be completely isolated from live radar whenever required.

The equipment is capable of feeding one or more of the ship's live radar displays with up to four simulated radar echoes, each of which is fully controllable with respect to course and speed. "Own Ship's" movement can either be that of the real vessel, or that of the simulated Own Ship whose characteristics of manoeuvre can be preset to represent any class of ship. The simulator can, therefore, be used at sea—with the real vessel steady on course or making alterations in course and speed, or in harbour.

A safety device included in the shipborne installation ensures that, when the simulator is working into the live radar displays and an emergency situation arises, the simulator can be instantly isolated from the live radar causing only live radar echoes to be displayed.

The Plessey Electronics Group is now producing a new range of high-grade h.f. communications equipment comprising the PVT 150 Series transmitters and the PVR 800 Series receiving systems.

The equipment is designed for unattending operation in any part of the world by aeronautical, maritime and civil administration authorities and incorporates the very latest system and circuit design techniques.

There are two types in the 1kW PVT 150 transmitter series, the PVT 150A, with a frequency range of 4 to 27Mc/s and the PVT 150B, which covers the range of 1·6 to 27Mc/s and will operate in six different telegraphy and telephony modes.

The British Post Office has placed substantial orders for PVR 800 receivers and both Cable and Wireless and International Aeradio have ordered large numbers of 1kW transmitters for use in their communications networks.

Wayne Kerr Laboratories Ltd has received from the Chinese People's Republic new contracts to the value of more than £40 000 for electronic instruments, consisting mainly of precision a.c. bridges, electrometers and servo performance analysers.

This order follows shortly after the Company's success in the USSR.

For some years Wayne Kerr has maintained exports at a steady 45 per cent of production, of which about half is to the USA.

The Scientific Instrument Manufacturers' Association (SIMA) has recently published a folder entitled 'Careers in the Scientific Instrument Industry.'

Six separate leaflets have been produced covering the educational requirements, the length of apprenticeship and type of training and the prospects for trade apprenticeship, craft apprenticeship, technical apprenticeship for City and Guilds of London Institute courses and National Certificate courses, student apprenticeship and post graduate apprenticeship.

Details of the member companies offering all or any of these apprenticeship training schemes are given in the covering folder.

'Careers in the Scientific Instrument Industry' is being sent by the Association through the Ministry of Labour to Youth Employment Exchanges, appropriate schools, and to technical Colleges and Universities.

Plessey Radar Ltd has received a contract valued at £400 000 from the Bulgarian Government for the supply of three Plessey AR-1 radar systems for installation at Sofia, Varna and Burgas to provide air surveillance for civil aviation over the whole Bulgarian area.

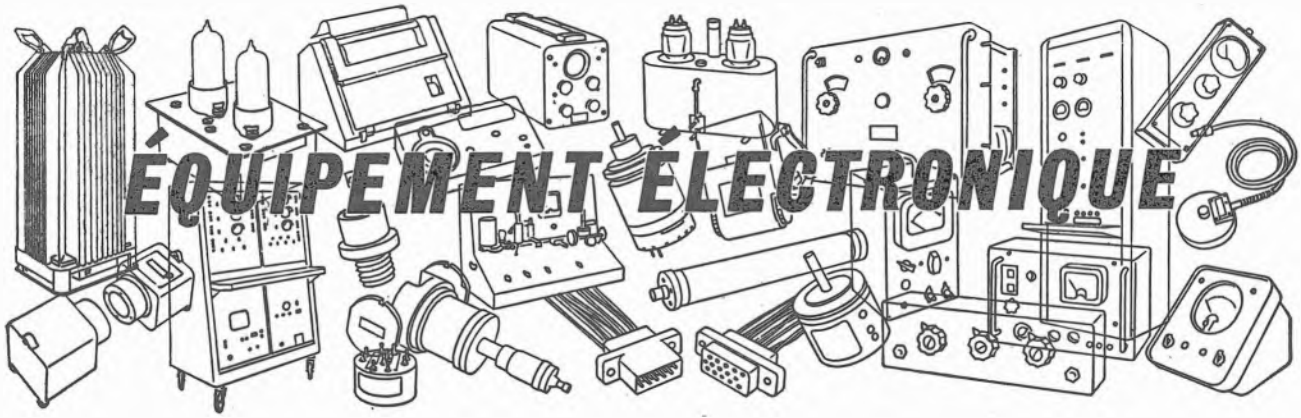
In addition to supplying this equipment and installation services, Plessey Radar will also train a team of 14 Bulgarian engineers at its Isle of Wight Training Centre.

Telcon Metals Ltd of Manor Royal, Crawley, Sussex, a member of the BICC Group of Companies, announce the development of a new magnetic material SATMUMETAL, extending the normal Mumetal magnetization curve up to 15 000 gauss. Losses are approximately one quarter those of silicon iron over the whole operating range of flux density up to 13 000 gauss, and the magnetizing force required over this range is also one quarter that required for silicon iron. Initial and maximum permeabilities are approximately 40 000 and 250 000 respectively, coercive force 0·025 oersteds and remanence 7 000 gauss.

The material which is available in the form of 0·004in toroidal cores fills a gap in the spectrum of magnetic properties of currently available soft magnetic alloys. Applications envisaged for it include cores for numerous types of transformers where losses and magnetizing forces are at a premium, e.g., small distribution transformers, metering and protective devices.

The British Broadcasting Corporation has placed an order with Standard Telephones and Cables Ltd, London for a number of u.h.f. solid-state television translators and 50 watt amplifiers operating in Bands IV and V.

These equipments will be used to improve the reception of BBC-2 programmes in those localities where, due to unfavourable topographical conditions, the received signal strength from the main transmitting stations is too low to provide a satisfactory picture. The equipments, which embody the most modern design techniques, are suitable for either monochrome or colour television. The 50W power output is obtained by the use of a travelling wave tube amplifier driven by the translator.



Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai *Traduction des pages 620 à 625*

ÉTALON DE FRÉQUENCE

The Marconi Co. Ltd, Chelmsford, Essex

(Illustration à la page 620)

Un oscillateur à très grande stabilité, dont la précision est comparable à celle d'un étalon de fréquence atomique moderne, a été mis au point par The Marconi Co. Ltd. Cet appareil, dont la variation de fréquence n'est que de 30 millionièmes par cycle à 100 kHz, est l'un des étalons de fréquence les plus stables et les plus précis produits dans le monde actuellement.

Sa stabilité de fréquence à court-terme est supérieure à 3 parties dans 10^{10} et sa stabilité de vieillissement est supérieure à 1 partie dans 10^9 par mois. Il produit trois fréquences standard simultanées de 100 kHz, 1 MHz et 2,5 MHz.

Il comprend un cristal taillé A-T à 5 ème harmonique supérieur de $2\frac{1}{2}$ MHz, conçu et fabriqué par la fabrique de cristaux de quartz de la société Marconi à Hackbridge, au sud de Londres. Le procédé de polissage perfectionné de ces cristaux est nettement plus avancé que celui employé pour les cristaux ordinaires. A l'intérieur du cristal se trouve huit inter-faces parfaites et son facteur de qualité dépasse 6×10^6. Les contacts en or sont thermiquement évaporés sur les surfaces extérieures du cristal afin d'aussurer les connexions de sortie.

La stabilité maxima aux variations de température est assurée tant pour le cristal que pour l'oscillateur en les enfermant à l'intérieur de deux fours proportionnellement contrôlés et sensibles aux changements de température d'un millième de degré. De plus, le cristal est utilisé au niveau de puissance extrêmement bas de 0,5 μV.

La sortie du cristal de 2,5 MHz est injectée à l'étage de division de fréquence dans lequel des circuits "diviseurs en escalier" sont utilisés pour obtenir des sorties de 1 MHz et de 100 kHz à stabilité de phase élevée. Ces circuits, ainsi que les étages de sortie intermédiaires, comprennent des plaquettes de circuit imprimé entièrement transistorisées.

L'oscillateur complet est monté sur panneau standard de 48,2 cm et il contient ses propres blocs d'alimentation qui lui permettent de fonctionner à partir de n'importe quel courant secteur normal. En cas de panne de secteur, il continue à fonctionner sur ses propres batteries de secours incorporées, avec la même stabilité de fréquence.

La précision et la stabilité de ce dispositif le rendent particulièrement indiqué pour les systèmes de guidage maritime, les calculatrices, le guidage de satellites, les applications militaires et de défense, ainsi que pour de nombreuses autres applications dans les systèmes à haute précision.

EE 85 751 *pour plus amples renseignements*

INDICATEURS NUMÉRIQUES

K.G.M. Electronics Ltd, Bardolf Road, Richmond, Surrey

(Illustration à la page 620)

Il s'agit ici d'un indicateur de format mince et élégant pouvant être aisément fixé à des plaquettes de circuit imprimé. Il sert à la fois de poignée pour le retrait de la plaquette et d'indicateur numérique.

L'indicateur KGM type M23 peut, en effet, afficher jusqu'à 10 symboles, grâce à l'emploi de lampes miniature du type militaire dont la valeur nominale va de 6 V à 28 V. Il comprend, en outre, le dispositif à bornes communes isolées.

EE 85 752 *pour plus amples renseignements*

COMPTEUR-MINUTERIE DE LOTS

Advance Controls Ltd, Cheltenham, Gloucestershire

(Illustration à la page 620)

La société Advance Controls Ltd vient d'ajouter ce nouveau compteur-minuterie TB4 à sa gamme de compteurs industriels. Il s'agit d'un instrument d'une grande souplesse d'emploi dont la vitesse de comptage maxima est de 5000 coups par seconde et qui utilise une combinaison de circuits transistorisés et de tubes de comptage à cathode froide pour garantir la fiabilité à bas prix.

Le TB4 comprend quatre décades et peut compter n'importe quelle quantité jusqu'à 9999 ou être utilisé pour le contrôle de programmes ou encore le minutage de cycles pendant n'importe quelle période jusqu'à 99,99 sec, en unités de 0,01 sec. Un signal réglable de pré-lot et de pré-temps forunit un signal d'avertissement ou de contrôle de fonction.

Une fiche de programme multi-points est fixée sur le panneau arrière et les arrangements d'entrée, de sortie et de réenclenchement peuvent être variés pour répondre à n'importe quel besoin. Des fiches de programme standard ou spéciales peuvent être fournies, prêtes à l'usage, ou elles peuvent être bobinées par le client selon les instructions données dans le manuel de l'appareil.

EE 85 753 *pour plus amples renseignements*

CONNECTEURS DE TRANCHES

Plessey-UK Ltd, Wiring & Connectors Division, Ilford, Essex

(Ilustration à la page 621)

Une nouvelle gamme de connecteurs de tranches de circuits imprimés a été créée par la Wiring and Connectors Division de la société Plessey UK Ltd. Ces connecteurs, fabriqués en bandes pouvant recevoir un maximum de 80 positions de contact, portent la référence de série 603, et ils peuvent être coupés selon la spécification exacte du client, une variété de combinaisons étant prévue pour répondre aux applications individuelles de connexion de circuits imprimés.

Les connecteurs de la série 603 sont fournis en dimensions de fixation conformes aux spécifications normales actuelles. Des pieds de fixation sont fournis pour former des prises de courant à extrémité fermée. En variante, et lorsqu'on exige des prises de courant à extrémité ouverte, on peut obtenir des moulages perforés pour recevoir des vis de fixation à prises automatiques.

Les nouveaux connecteurs se composent de prises de courant qui s'accouplent directement à des plaquettes de 15 mm d'épaisseur. Les qualités de ces connecteurs, qui sont plus chers que les connecteurs ordinaires, sont obtenues en employant des contacts en porte à faux de forme simple et en faisant un usage optimum des matériaux et de l'espace.

Les contacts sont de forme spéciale afin d'assurer un bon accouplement et des caractéristiques de force-contact. On réduit ainsi l'usure de la plaquette et la résistance de contact, malgré les variations d'épaisseur de la plaquette. La résistance à faible pôle s'obtient par l'emploi de contacts en laiton conçus pour les terminaisons de soudure. Le fini du matériau est facultatif: or dur, or brillant sur argent ou fini sur commande spéciale.

Les prises de courant sont moulées en polycarbonate de verre, dont les caractéristiques mécaniques et électriques sont excellentes. C'est également un matériau ayant des propriétés ignifuges. Une clé de polaristaion est fournie pour remplacer n'importe quel contact voulu.

Ces connecteurs, dont la valeur nominale est de 750 V c.a. efficaces au niveau de la mer et de 5 A de courant continu, ont été étudiés pour pouvoir être utilisés entre $-40°C$ et $-85°C$, la résistance de contact étant inférieure à 15 MΩ.

EE 85 754 *pour plus amples renseignements*

CONTRÔLEUR DE TENSION SECTEUR

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(*Illustration à la page 621*)

La société Kent Precision Electronics Ltd a récemment mis au point un circuit constitué de corps solides pour le contrôle d'alimentations normales en en tension triphasée. Ce dispostif assure le contrôle sûr, précis et efficace des déviations de tension, ainsi que la protection ou l'avertissement en cas de pareilles déviations. L'instrument permet de nombreuses applications pour le contrôle des alimentations de secours, la régulation de tension et le contrôle du courant pour calculatrices.

Il se compose essentiellement d'un circuit logique à six entrées qui actionne un déclencher de sortie pour la commutation d'un relais électromagnétique standard, lorsque l'une des tensions d'entrée triphasées est au-dessus ou au-dessous des valeurs préréglées. L'instrument peut être réglé pour n'importe quelle tension nominale de secteur avec des déviations à pourcentage prédéterminé. Les contacts du relais de sortie sont donc en "état d'alarme" lorsque n'importe laquelle des tensions triphasées dévie des valeurs préréglées. Les contacts accomplissent ainsi une protection absolue contre les surtensions ou les sous-tensions pour tout matériel se trouvant relié à la tension de secteur contrôlée.

Le contrôleur peut être fourni dans un coffret standard pour montage sur panneau ou, afin d'en faciliter l'usage, sous forme de construction modulaire nouvelle, ainsi que le montre notre gravure.

EE 85 755 *pour plus amples renseignements*

INSTRUMENT DE MESURE DE LA CONDUCTIVITÉ

Electronic Switchgear (London) Ltd, 58 Wilbury Way, Hitchin, Hertfordshire

(*Illustration à la page 621*)

Un nouvel appareil portatif de mesure de la conductivité vient d'être ajouté à la gamme d'instruments de mesure de conductivité fabriqués par Electronic Switchgear (London) Ltd.

L'instrument de mesure type MC3 comporte une cellule de mesure pratiquement indestructible qu'on introduit directement sur le côté du coffret de l'instrument et qui n'exige aucun autre entretien qu'un nettoyage occasionnel au moyen d'une brosse. La cellule a un alésage tubulaire de 10 cm^3 de capacité, avec des électrodes anulaires qui permettent de mesurer la conductivité de liquides, de pâtes et de boues avec la même facilité.

La gamme de mesure s'étend de 1 à 100 000 micromhos/cm en trois plots pouvant être choisis par un commutateur qui est muni d'une quatrième position permettant de vérifier à volonté le fonctionnement efficace et la précision de l'instrument.

Les circuits imprimés transistorisés sont actionnés par une petite batterie de longue durée, facilement remplaçable et se trouvant à l'intérieur du coffret coulé en coquille et mesurant 24,13 cm × 19,05 cm × 8,89 cm.

EE 85 756 *pour plus amples renseignements*

INSTRUMENT DE MESURE MULTIGAMMES

Distributeurs: Croydon Precision Instrument Co., Hampton Road, Croydon, Surrey

(*Illustration à la page 621*)

Cet instrument de mesure multigammes, fabriqué par la société Norma de Vienne, comprend quarante gammes de mesure couvrant les tensions alternatives et continues, la résistance et la sortie. Il comprend en outre une gamme de tension continue de 12 mV d'une sensibilité de 40 kΩ/V et une échelle de température en degrés centigrades pour thermocouples fer/constantant.

La sensibilité sur tension continue est de 20 kΩ/V et sur tension alternative de 4 kΩ/V. La précision sur courant continu est de ±2,5 % et 3,5 % sur courant alternatif jusqu'à 500 Hz et ±7,5 % à 30 kHz.

L'instrument pèse 400 g environ. Il comprend un mouvement à suspension tendue et il peut résister à une surcharge de 1000 %.

EE 85 757 *pour plus amples renseignements*

DISPOSITIFS ENREGISTREURS

Benson-Lehner Ltd, West Quay Road, Southampton

(*Illustration à la page 621*)

La société Benson Lehner Ltd a introduit deux nouveaux dispositifs photo-enregistreurs, principalement prévus pour les machines d'emballage et de conditionnement.

Les deux éléments sont entièrement transistorisés et ne comprennent ni relais ni contacteurs ni aucune pièce mobile permettant ainsi d'employer des vitesses de fonctionnement maxima et d'éliminer les pertes de temps causées par l'atmosphère polluée ou la nécessité d'effectuer des réglages.

Le type 2C5A répond immédiatement à toute variation dans l'état normal d'enregistrement soit en avançant soit en retardant un moteur électrique auxiliaire ou tout autre dispositif devant être contrôlé, suivant la position des repères d'enregistrement.

L'autre élément est un système de correction unidirectionnel pouvant être incorporé à des machines où le matériau d'emballage a été délibérément tiré sur chaque cycle. Le système de correction accomplit la tâche de retarder l'alimentation lorsque le repère d'enregistrement obscurcit la cellule photoélectrique.

EE 85 758 *pour plus amples renseignements*

BLOC D'ALIMENTATION TRANSISTORISÉ

Fenlow Electronics Ltd, Springfield Lane, Weybridge, Surrey

(*Illustration à la page 622*)

Le bloc d'alimentation PU5 a été conçu pour l'utilisation sur courant alternatif de secteur de 200 à 250 V ou 100 à 150 V, 50 Hz ou 60 Hz, la consommation de courant maximum étant de 290 W. La sortie est réglable de 0 à 150 V c.c., stabilisés jusqu'à 1,5 A.

Ces blocs peuvent fonctionner sur deux modes: tension constante et courant constant, et ils sont équipés de circuits de permutation à "mode automatique" et d'un dispositif d'indication visuelle. Ils sont protégés contre les courts-circuits accidentels des bornes de sortie.

Dans le mode à tension, la tension de sortie est réglée par incréments de 1 V au moyen de commutateurs de décades étalonnés et le courant tiré par la charge appliquée est indiqué sur l'instrument de mesure du panneau. Dans le mode à courant, ce dernier est réglé par un bouton du panneau marqué pour pouvoir s'y référer rapidement et au moyen de l'instrument de mesure du courant. Le passage d'un mode à l'autre se produit automatiquement à mesure que l'on fait varier la charge extérieure à travers la valeur critique et deux lampes colorées signalent le changement de mode. On peut percevoir à distance la tension de manière à éviter les effets de l'impédance dans les câbles de connection. En outre, la tension peut être établie à distance et des éléments peuvent être reliés en série ou en parallèle pour augmenter la sortie de tension ou de courant.

L'appareil est fourni sous forme de "demi-bâti" permettant de monter une paire sur un bâti standard de 48,26 cm au moyen d'un assemblage de cadre.

EE 85 759 *pour plus amples renseignements*

INTERRUPTEURS À BOUTON-POUSSOIR DE CIRCUIT IMPRIMÉ

Mills & Rockleys (Electronics) Ltd, Swan Lane, Coventry

(*Illustration à la page 622*)

Cet interrupteur a été prévu spécialement pour pouvoir être monté sur des panneaux de circuits imprimés sous forme de composants. On évite ainsi la nécessité d'acheter un système interrupteur préassemblé de conception plus complète qu'il est presque impossible de détacher du circuit imprimé en cas de panne ou d'autres défauts communs à ce type d'interrupteur.

L'interrupteur est basé sur le dessin du circuit imprimé ayant un dispositif conducteur approprié pour coïncider avec les contacts portés par les boutons-poussoir pour l'enclenchement ou la rupture.

Les boutons-poussoirs, qui portent des contacts de liaison, sont reliés au circuit imprimé à l'intérieur d'une fente s'étendant sur le bouton-poussoir qui est parallèle à la surface plate du panneau de circuit imprimé.

Chaque bouton-poussoir est muni d'un ressort intérieur de sorte que lorsqu'on presse un autre bouton, n'importe quel bouton-poussoir déjà enclenché et qui n'est pas pressé en même temps retourne à sa position originale. Ceci est réalisé par le dessin de la forme du moulage du bouton-poussoir en liaison avec la barre de retenue qui est fixée au panneau de circuit imprimé.

L'assemblage des composants se réduit donc à la fixation des moulages des boutons-poussoir complets avec leur contacts et ressorts aux panneaux de circuits imprimés. Ils sont retenus par la longueur de barre appropriée, déterminée par le nombre de boutons-poussoir nécessaires. Ce montage permet également un espacement variable des boutons, si on préfère cette réalisation, sur n'importe quel appareil.

Les boutons-poussoirs deviendront des éléments standard et il se sera possible à l'utilisateur d'adopter n'importe quel nombre de ces boutons puisqu'il suffira de choisir la longueur voulue de barre et de réaliser le circuit imprimé en conséquence. Ce sera donc le rôle normal de la production Mills & Rockleys de circuits imprimés de fournir les outils appropriés aux circuits imprimés, de fabriquer les panneaux de circuit et de fournir ou d'assembler les composants des boutons-poussoirs pour les assortir aux panneaux.

EE 85 760 *pour plus amples renseignements*

VOLTMÈTRES NUMÉRIQUES

Roband Electronics Ltd, Lowfield Heath Road, Charlwood, Horley, Surrey

(*Illustration à la page 622*)

La société Roband Electronics Ltd a réalisé deux nouvelles gammes de voltmètres numériques compacts. Les deux instruments de base sont le modèle RDV 32 et RDV 42, livrables en coffrets pour montage sur bâti ou sur panneau frontal. Il existe également des éléments de lecture à distance, un élément de mesure en série et des voltmètres à gamme unique pour certaines fonctions particulières.

Les deux instruments sont du type à approximation successive, fonctionnant à des vitesses de conversion dépassant 30 par seconde; ils donnent des signaux de sortie décimaux ou décimaux codés binaires pour le contrôle ou l'impression.

Le modèle RDV 42 lit (sur la totalité de l'échelle) de 0,9999 à 999,9 V en quatre gammes avec un résolution de 100 μV. Le RDV 32 lit (sur totalité de l'échelle) de 1,099 à 999,0 V en quatre gammes avec une résolution de 1 mV. La précision minima des deux instruments est de 0,1 % $\pm$1 chiffre.

Ces instruments qui comprennent des circuits constitués de corps solides sont prévus pour répondre aux conditions industrielles les plus ardues et ils existe en outre des modèles scellés pour utilisation en atmosphère corrosive.

EE 85 761 *pour plus amples renseignements*

SONDE DE COMPTAGE

Racal Instruments Ltd, Crowthorne, Berkshire

(*Illustration à la page 622*)

La société Racal Instruments Ltd vient de créer un nouvel accessoire de comptage de points: la sonde active type SA.544. Cette sonde tout transistors a été prévue pour l'emploi avec les compteurs numériques Racal à lecture directe, types SA.540 et SA.550 pour la mesure de fréquence de 10 MHz et 100 MHz.

L'emploi de la sonde active améliore le comptage, portant l'impédance d'entrée de comptage à 10 MΩ en 5 gammes et la sensibilité à 5 mV, rendant ainsi possibles de nombreuses mesures dans la gamme de fréquence de 100 MHz qui n'étaient possibles jusqu'ici qu'à l'aide d'amplificateurs externes spéciaux et de dispositifs d'équilibrage d'impédance. Le compteur conserve sa précision inhérente élevée pour toutes les mesures. La nouvelle sonde, entièrement munie de semiconducteurs est alimentée directement par le compteur connexe. Elle est de construction robuste et en mesure de résister aux conditions de fonctionnement rigoureuses imposées par les applications militaires. Elle a reçu le code de référence de l'OTAN No. 6625-99519-1573.

EE 85 762 *pour plus amples renseignements*

TÊTE ENREGISTREUSE À SEPT PISTES

Gresham Lion Electronics Ltd, Lion Works, Hanworth Trading Estate, Feltham, Middlesex

(*Illustration à la page 623*)

La société Gresham Lion Electronics Ltd a réalisé une tête enregistreuse magnétique à sept pistes de lecture après écriture. Cette tête est prévue pour les applications numériques et effectue la vérification immédiate des données enregistrées.

Elle existe en deux versions: avec éléments enregistreurs à gauche, s'adaptant au matériel IBM, ou avec éléments enregistreurs à droite. La face frontale est profilée de manière à minimiser le bruit des grandes longueurs d'onde est elle est repliée au delà de la piste pour empêcher les cannelures provoquées par les bords de bandes.

La tête mesure 3,17 cm de hauteur $\times$ 2,61 cm de profondeur et 2,54 cm de largeur; l'azimuth est tenu entre $\pm$0,0002 radians et l'élévation entre 0,0005 radians. La dispersion d'entrefer est contenue dans une bande de 0,00008 pouces de largeur et les lignes d'entrefer sont parallèles à 0,0002 radians. Les pistes de lecture ont 0,031 pouce de largeur à intervalles de 0,070 pouce et les pistes d'écriture ont 0,046 pouce de largeur au même intervalle.

Le courant de saturation sur bande 3M T488 à 300 chiffres binaires par pouce et saturation de 99 % est de 36 mA + 15% de crête à crête. La sortie de lecture, avec courant d'écriture de 54 mA de crête à crête, est de 44 mV de crête à crête à une vitesse de bande de 120 pouces/sec.

L'alimentation durant l'écriture sur cette piste simultanément en phase avec un signal d'onde carrée de 100 kHz à temps de montée de 2 μsec et à valeur de courant de 60 mA de crête à crête est inférieure à 10 mV sur n'importe quelle piste de lecture. Ce chiffre peut être réduit d'un rapport d'au moins 2:1 par un écran couvrant la face frontale et placé immédiatement à l'arrière de la bande.

EE 85 763 *pour plus amples renseignements*

SOURCES FROIDES

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

(*Illustration à la page 623*)

La société A.P.T. Electronic Industries Ltd, constructeurs des châssis et bâtis Lektrokit, vient d'ajouter à sa gamme Lektrokit une série de sources froides. Ces dernières sont livrables en quatre modèles.

La source froide No. 1 (pièce No. LK.1111) a été conçue pour pouvoir être montée dans le transistor de type international TO-3, mais elle peut cependant être montée en plusieurs manières différentes, soit individuellement soit en groupe. La construction à fentes permet la circulation de l'air libre à travers l'ensemble de la source froide, permettant ainsi le refroidissement efficace à n'importe quel angle. Les ouvertures de connexion du transistor sont allongées, permettant de fixer deux sources froides dos à dos avec un transistor, augmentant ainsi dans une grande mesure le régime de dissipation.

La source froide No. 2 (pièce No. LK.1121) est de forme identique mais avec ouvertures de montage pour les grands transistors de puissance du type TO-36 ainsi que pour les diodes zéner de puissance et les redresseurs.

Les sources froides 1 et 2 sont construites en aluminium à haute conductivité de 1,22 mm, sous forme pliée et à fentes. Elles sont perforées pour montage sur transistors de puissance et peintes en noir pour faciliter la dissipation de chaleur. Leur résistance thermique à l'air libre est d'environ 3,5°C/W.

Les sources froides 3 et 4 (pièces No. LK.1131/1141) sont construites à partir de profilés en aluminium solide, perforés pour montage sur transistors et sont également de couleur noire. La résistance thermique de ces deux instrument est d'environ 1,6°C/W, pour montage vertical, et de 1,75°C/W pour montage horizontal à l'air libre. Elles ont été étudiées pour avoir une capacité thermique et une dissipation de chaleur maxima. La source froide No. 3 est pour montage sur transistor de type international TO-3 et la source No. 4 est pour transistor de puissance du type TO-36 ainsi que pour les diodes zéner et les redresseurs de puissance.

EE 85 764 *pour plus amples renseignements*

TRANSDUCTEURS DE PRESSION

S.E. Laboratories (Engineering) Ltd, North Feltham Trading Estate, Feltham, Middlesex

(Illustration à la page 623)

Une gamme de transducteurs de pression type S.E. 37 est offerte par S.E.L. sous forme d'éléments autonomes préétalonnés fonctionnant sur tension continue. Ils ont été conçus essentiellement pour les systèmes de fusées supersoniques et d'aviation, ainsi que pour les conditions d'emploi où la plupart des autres transducteurs échoueraient par suite de leurs vibrations extrêmes.

Ils sont d'une très grande précision et fiabilité. Leur sortie de tension ou de courant est prévue pour branchement direct sur divers appareils tels que magnétophones à oscillographes, appareils de télémesure, oscillateurs contrôlés par tension de haut niveau, commutateurs, instruments de comptage numérique ou indicateurs à lecture directe.

Leur circuit soudé pleinement intégré assure une grande sécurité de fonctionnement et l'utilisation continue à des température de 150° C avec une erreur de température de moins de 0,015 %/°C. Pour un fonctionnement intermittent de 3 minutes à une température de 300°C l'erreur totale est inférieure à 0,35 %.

Les effets de bruit dus à des circuits de sol et à de grands courants circulaires provenant par exemple d'avions ou d'installations de contrôle au sol sont réduits au minimum au moyen de l'isolement total des lignes d'alimentation en tension continue et des circuits de sortie.

Le transducteur S.E. 37 fonctionnant sur courant continu convient aux applications de jauge différentielle ou de pression absolue avec gaz ou liquides corrosifs, avec des gammes de pression de 0 à 2 et de 0 à 4 000 livres/pouces carrés. La tension de sortie de —5 V à +5 V pour une pression d'échelle totale peut être réglée suivant la demande du client pour varier de 0 au maximum ou de la négative à la positive (par exemple de 0 à 1 000 livres/pouce carré, 0 à 5 V; ou —1 000 à +1 000 livres/pouce carré, 0 à ±5 V). La construction soudée et l'exclusion de tout scellement en caoutchouc ou en matière plastique rend ce transducteur particulièrement indiqué pour l'emploi avec tous les fluides compatibles avec l'acier inoxydable.

EE 85 765 *pour plus amples renseignements*

PONT R.F. DE PRÉCISION

The Wayne Kerr Laboratories Ltd, Sycamore Grove, New Malden, Surrey

(Illustration à la page 623)

Ce nouveau pont à gamme étendue assure une précision de 0,1 % pour les mesures de capacitance et de conductance entre 100 kHz et 1 MHz et il peut être utilisé avec une précision inférieure jusqu'à 5 MHz. Le pont B201 est entièrement portatif, y compris la batterie (ou le redresseur), la source à fiches, les détecteurs, les étalons et un instrument de mesure de zéro. La modulation interne à 1 kHz et un jack de sortie à détection permettent d'employer des écouteurs. On peut ajuster le niveau du signal et le gain du détecteur à l'aide de commandes sur le panneau frontal. La gamme de couverture totale va de 0,0001 pF à 0,1 μF et de 0,001 μmho à 1 mho (1 Ω à 1000Ω).

Les dessin de l'instrument est basé sur celui des transformateurs pour rapports précis de tension et de courant au moyen desquels l'inconnue est comparée à des étalons internes de capacitance et de conductance. Une borne neutre sur la barrette de connexions permet d'effectuer des mesures à trois bornes. Les fréquences pouvant être obtenues actuellement du détecteur et de la source à fiches sont de 100 kHz et 1 MHz mais des éléments feints permettent de brancher un matériel extérieur pour des mesures dans la gamme de 100 kHz à 5 MHz.

EE 85 766 *pour plus amples renseignements*

INSTRUMENT DE MESURE MULTIGAMMES

Salford Electrical Instruments Ltd, Peel Works, Silk Street, Salford, 3, Lancashire

(Ilustration à la page 624)

Le Selectest mark II est un modèle perfectionné de l'appareil original avec des raffinements et des modifications supplémentaires suggérées par les ingénieurs spécialisés.

La plaquette des bornes comprend maintenant des bornes avec prises pour câble à fiches. Les interrupteurs de changement de gammes ont été révisés pour assurer une plus grande légibilité et un choix de gammes simplifié. Ils peuvent être tournés dans l'une et l'autre direction à travers un angle de 360° et ils sont inter-verrouillés électriquement.

Le disjoncteur à maxima automatique et les composants internes sont montés sur une robuste plaquette de circuit imprimé enfermée dans un solide coffret en Mélomine. Il existe deux modèles: le Selectest Super 50 ayant un mouvement de base de 50 μA et 32 gammes internes pour la mesure du courant alternatif et du courant continu ainsi que de la résistance. Le Select Super K a un mouvement de 1 mA avec 39 gammes intérieures.

EE 85 767 *pour plus amples renseignements*

INDICATEUR NUMÉRIQUE MINIATURE

Counting Instruments Ltd, 5, Elstree Way, Boreham Wood, Hertfordshire

(Ilustration à la page 624)

La société Counting Instruments Ltd a réalisé un nouvel élément d'affichage numérique microminiature à projection arrière. Il mesure 1,90 cm × 1,27 cm de large × 5,08 cm et malgré ces dimensions fort réduites projette une image de 0,95 cm de hauteur.

Son fonctionnement est semblable aux éléments d'affichage à 12 lampes de 9,52 cm, 2,54 cm et 1,58 cm. Il comporte un assemblage de 11 lampes incandescentes miniature à l'arrière, une série de lentilles et un écran de vision frontal sur lequel on peut projecter individuellement ou en combinaison 11 messages. Etant donné ces messages sont sur pellicule, l'appareil peut afficher n'importe quels chiffres, lettres, mots, symboles ou couleurs pouvant être reproduits photographiquement.

L'appareil se trouve dans un coffret insérable. Ce dernier est branché dans un logement comprenant les connexions de lampes qui peuvent être bobinées de manière permanente. Le coffret insérable peut être retiré du logement par l'avant du panneau afin de faciliter le changement des lampes ou il peut être remplacé par un coffret à message différent. Il existe quatre types de coffrets d'insertion: le modèle à sous-panneau utilisable avec un écran continu en assemblages multiples; le modèle à écran individuel encastré dans le panneau, ou le modèle à écran individuel avec cadre à couvercle partiel ou total. Chacun des types peut être groupé pour former un assemblage multiple.

EE 85 768 *pour plus amples renseignements*

AIMANTS DE RÉFÉRENCE STANDARD

Preformations Ltd, Cheney Manor, Swindon Wiltshire

(Ilustration à la page 624)

La société Preformations Ltd, du groupe des composants Plessey, fournit maintenant une gamme d'aimants de référence standard entièrement blindée.

Conçus pour l'étalonnage des bobines de recherche, des fluxmètres, des gaussmètres d'effet Hall et d'autres instruments analogues, ces aimants avaient été prévus à l'origine pour l'utilisation avec une bobine de recherche de format standard fabriquée par la société W. G. Pye & Co. Ltd de Cambridge. Cependant ils peuvent être utilisés également pour l'étalonnage de n'importe quelle bobine de recherche ou de sonde d'effet Hall avec un diamètre maximum de 10 mm et une épaisseur maximum de 4,75 mm.

Les aimants de référence sont pleinement stabilisés, tant magnétiquement que par cyclage de température durant la production, afin de les protéger contre les pertes irréverssibles résultant des changements de température et de champs magnétiques extérieurs.

L'étalonnage s'effectue par référence à un magnétomètre nucléaire de résistance magnétique.

Il existe des aimants avec champs transversaux ou axiaux, avec les densités de flux suivantes:

Densité axiale 300 et 500 gauss.

Densité transversale 1000, 3000, 5000, 10 000 et 15 000 gauss.

La précision est de ±0,5 % de la densité de flux marqué à 20° C. La stabilité magnétique est telle que les aimants ne sont pas affectés de manière permanente par des champs de dispersion au-dessous de 150 oersteds et la stabilité à la température est de l'ordre de 0,2 %/° C entre −10° C et +90° C. La température de fonctionnement maxima est de 100° C.

EE 85 769 *pour plus amples renseignements*

DOUILLES DE CONTRÔLE DE CIRCUIT IMPRIMÉ

Sealectro Ltd, Walton Road, Farlington, Portsmouth, Hampshire

(Illustration à la page 624)

Le SKT-105 PC L1 est une nouvelle douille de contrôle en matière plastique isolante thermorésistante pouvant être insérée dans les châssis de circuit imprimé. L'extrémité de contact de ce nouveau composant est à angle droit par rapport à la plaquette de circuit imprimé et elle est prévue pour être placée plus haut que les composants conventionnels par rapport à la surface du châssis. Cette hauteur supplémentaire au-dessus du châssis permet d'intercaler des douilles telles que le modèle SKT-103 PC et d'augmenter le nombre d'éléments pouvant être montés sur le châssis. Le SKT-105 PC L1 comprend des capots protecteurs en matière plastique isolante thermorésistante et les contacts en cuivre de béryllium sont dorés sur argent.

EE 85 770 *pour plus amples renseignements*

MINUTERIES ÉLECTRONIQUES

Solid State Controls Ltd, 30-40 Dalling Road, London, W.6

(Illustration à la page 625)

Le "Variset" est une minuterie entièrement intégrale avec sortie à relais à fiches pouvant commuter des charges de puissance de 5 A. L'élément de base est un circuit de minutage entièrement transistorisé d'une durée de vie allant jusqu'à 100 millions d'opérations. Elle est parfaitement protégée contre les phénomènes transitoires de haute tension et elle n'est pas affectée par les variations étendues de tension et de température de régime. Elle est donc d'une grande fiabilité, même dans des conditions industrielles défavorables et sa précision de répétition caractéristique n'est que de 1 %. Un relais standard à fiches et à l'épreuve des poussières est donc la seule pièce pouvant être usée et elle est d'un accès facile.

Le cadran de réglage est étalonné directement en secondes et des temps spécifiques peuvent donc être facilement réglés d'avance. Des périodes de retard de temps de 0,2 sec à 4 min sont assurées en trois gammes de minutage cependant que huit tensions de fonctionnement standard peuvent être obtenues dans la gamme de 12 à 110 V c.c. et de 48 à 415 c.a. Les minuteries Variset sont munies de deux contacts de permutation d'une puissance nominale de 5A, 250 c.a.; d'autres valeurs nominales peuvent être également prévues. Quatre modes de base de fonctionnement sont prévus: retard sur excitation; retard sur désamorçage; mise en marche par bouton-poussoir ou recyclage. En outre, les bonnes caractéristiques de rétablissement permettent un nouveau calage instantanné et précis à n'importe quel moment de la période de minutage ou au terme de cette période.

EE 85 771 *pour plus amples renseignements*

POTENTIOMÈTRE BOBINÉ MINIATURE

Miniature Electronic Components Ltd, Copse Road, Woking, Surrey

(Illustration à la page 625)

Conçus à l'origine pour l'utilisation sur circuit imprimé sur module de 0,1 pouce, le potentiomètre T20P est du format et du modèle du transistor TO-5, le boîtier mesurant 0,345 pouce de diamètre × 0.255 pouce de hauteur. Il convient aux applications aérospatiales et industrielles.

Le réglage vertical à un tour avec arrêts d'extrémité positifs réduit le temps de calage; l'axe est à autoverrouillage; le coffret métallique et l'élément de formage fournissent une dissipation maxima de chaleur; la faible inductance permet l'emploi du potentiomètre pour les applications r.f. à impulsion.

Valeurs de résistances standard: 50, 100, 200, 500 Ω; 1, 2, 5, 10 KΩ.

Puissance nominale: ½ watt à 70° C.

Gamme de température: −55° C à +150° C.

EE 85 772 *pour plus amples renseignements*

ANALYSEUR DE BRUIT D'IMPACT

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustration à la page 625)

L'analyseur de bruit d'impact, Dawe type 1412, est un instrument portatif actionné par batterie. Il mesure l'amplitude de pointe et analyse les caractéristiques du bruit d'impact. Utilisé en liaison avec la gamme Dawe de sonomètres, le type 1412 donne une lecture directe de l'intensité de pointe du bruit et des renseignements permettant d'évaluer la durée et la forme d'onde d'un seul impact. Un circuit de "quasi pointe" est utilisé pour les sons à répétition et des filtres de bande d'octave peuvent être reliés extérieurement pour analyser la composition de fréquence. En outre l'instrument peut être utilisé pour l'étude des vibration en l'alimentant à partir d'un accéléromètre approprié; il peut analyser une forme d'onde impulsive de n'importe quelle source pouvant fournir 1 à 10 V de pointe à travers une impédance d'entrée de 50 kΩ dans la gamme de fréquence de 5 Hz à 40 kHz.

Le type 1412 se compose d'un amplificateur à tube actionnant trois redresseurs de signaux et un voltmètre électronique à courant continu qui peut être commuté sur chacun des circuits redresseurs à tour de rôle. Ces circuits redresseurs fournissent les signaux de pointe, de quasi pointe et de moyenne de temps qui permettent de déterminer la forme et l'ampleur de la forme d'onde.

Le circuit de point a un temps de montée extrêmement court; pour une impulsion rectangulaire, l'indication monte à 1 dB près de la valeur définitive en moins de 50 µsec. La charge est maintenue dans les limites de 1 dB pendant au moins 10 sec. Le circuit de quasi pointe a un temps de montée rapide de moins de 1 msec, et un temps de déclin lent d'environ 0,6 sec. Les caractéristiques sont sensiblement les mêmes que celles décrites dans la norme BS727:1954 "caractéristique et comportement des appareils pour mesurer les interférences radiophoniques". Le circuit de moyenne de temps donne une indication du niveau maintenu pendant une période donnée, des constantes de temps de 2 à 200 msec étant fournies en changeant une résistance. Dans des conditions d'emploi normales, la charge est maintenue à 1 dB près pendant au moins une minute.

L'instrument est fourni dans un coffret métallique portatif avec couvercle amovible et courroie.

EE 85 773 *pour plus amples renseignements*

COMPTEUR-MINUTERIE

Advance Electronics Ltd, Roebuck Road, Hainault, Ilford, Essex

(Illustration à la page 625)

Le nouveau compteur-minuterie Advance TC4 peut effectuer des mesures de fréquence directe dépassant 5 MHz.

Quatre temps de porte et sept éléments de minutage de µsec à 1 sec permettent d'effectuer des mesures précises de fréquence, de période et de temps. L'instrument comporte un élément d'affichage en ligne à quatre chiffres avec une décimale automatiquement fixée pour la mesure de fréquence. Le dispositif à périodes (jusqu'à 10^5 périodes), permet d'effectuer la mesure précise de fréquence extrêmement basse.

Pour la mesure de temps et de fréquence, la précision de base du TC4 est régie par un oscillateur à cristal piloté par four et réglé à une partie dans 10^6 à la température ambiante. Au cas où une précision supérieure serait nécessaire, un étalon externe produisant des signaux sinusoïdaux ou à pulsations peut être utilisé également.

EE 85 774 *pour plus amples renseignements*

POTENTIOMÈTRE BOBINÉ MINIATURE

Darstan Ltd, 263 Church Road, Thundersley, Essex

(*Illustration à la page 625*)

La construction compacte de ce potentiomètre le rend propre à l'emploi dans des circuits imprimés ainsi que pour d'autres applications exigeant un espace fort réduit.

Ce potentiomètre bobiné avec coffret métallique de protection, comporte un capuchon à fente avec un indicateur de position pouvant être facilement réglé à l'aide d'un tourne-vis. Les câbles sont sur module standard de circuit imprimé de 0,1 pouce.

Les câbles permettent un montage facile sur plaquette de circuit imprimé, aucune autre forme de fixation étant nécessaire. Un bouton préréglé est prévu et le potentiomètre peut être muni d'un support avant pour montage sur panneau.

Les valeurs de résistance normale sont de 5, 10, 20, 50, 100, 200, 300, 500 Ω, 1, 2, 5 Ω. La tolérance de résistance normale est de ±10 % et la puissance nominale est de 1 W.

EE 85 775 *pour plus amples renseignements*

Résumés des Principaux Articles

Un quantificateur à diodes tunnel par R. V. L'Archeveque et B. McA. Sayers

Résumé de l'article aux pages 574 à 579

Cet article décrit un générateur en escalier à diodes tunnel et à plots rapides. Le taux rapide de progression de l'ordre de 10^7 plots par seconde permet d'utiliser l'appareil pour la quantification, la conversion analogique/numérique et d'autres opérations avec un degré élevé de précision et une grande vitesse potentielle.

Un système de contrôle du courant de loi quadratique par C. I. Sach

Résumé de l'article aux pages 580 à 583

Cet article décrit un appareil qui contrôle un courant I_2 dans une charge de manière a ce qu'il varie suivant le courant I_1 parcourant une charge identique de telle manière que la somme des carrés des deux courants demeure substantiellement constante. Lorsque ces charges sont des bobinages sur nu noyau commun, les tours d'ampères nets sont contrôlés tout en maintenant la dissipation constante.

Données de réalisation de générateurs de fonction de résistance dépendants de la tension par G. Brown et R. A. B. Bond

Résumé de l'article aux pages 584 à 588

Un circuit amplificateur opérationnel, comprenant un réseau d'entrée à résistance dépendant de la tension et une réaction à résistance linéaire, est analysé en tant que générateur de fonction de la loi de puissance. Des données tabulaires sont indiquées qui permettent de calculer rapidement les valeurs optima de résistance pour des exponents de loi de puissance dans la gamme de 1,5 à 4. Le critère d'optimisation est l'erreur carrée minima. La table indique également les erreurs du générateur de fonction qui en résultent. La compensation d'erreurs et de température est brièvement examinée.

Un inverseur redresseur piloté au silicium de 12 volts c.c à 230 volts c.a. par D. I. Ross et B. E. G. Goodger

Résumé de l'article aux pages 589 à 591

Il s'agit ici d'un redresseur piloté au silicium de 12 volts c.c à 230 volts c.a. La fréquence de l'inverseur est contrôlée par un oscillateur de 1 kHz pour produire une sortie stable de 50 Hz. Le circuit fonctionne efficacement avec des charges variant de 0 à 70 watts. Un circuit de sécurité est incorporé à l'inverseur pour le protéger contre toute surcharge accidentelle.

La réalisation de circuits logiques minimals NI/ET par D. Zissos et G. W. Copperwhite

Résumé de l'article aux pages 592 à 597

Les auteurs décrivent une méthode systématique de réalisation de circuits logiques NI/ET minimals en cas d'absence des compléments des signauz d'entrée.

Ils présentent une technique de minimisation pour les circuits NI/ET et font mention d'une méthode de classification des circuits logiques.

Un amplificateur différentiel à courant continu utilisant des transistors à effet de champ par J. C. S. Richards

Résumé de l'article aux pages 598 à 600

Des transistors à effet de champ polarisés pour donner une failbe dérive de température constituent l'étage d'entrée d'un amplificateur différential stable ayant une largeur de bande de 100 kHz, un rapport de rejet de plus de 5000 et un gain différentiel stabilisé à 100. La derive de zéro est inférieure à 100 µV/°C, et le courant d'entrée inférieur à 10^{-9}A. L'utilité de l'amplificateur est limitée par le bruit basse fréquence produit dans les transistors (environ 50 µV efficaces).

Un dispositif à retard et un interrupteur logique à courant alternatif dans la logique à transistor à couplage résistance-condensateur par G. Philokyprou

Résumé de l'article aux pages 601 à 603

Le système de logique à transistor à couplage résistance-condensateur de circuit logique, introduit par P. L. Cloot, possède de nombreuses caractéristiques intéressantes pour la logique à faible vitesse. Le circuit de base de ce système, lorsqu'il est relié dans le montage à courant alternatif (entrée par condensateur seulement) se comporte comme un dispositif à retard. D'aurte part, en tenant compte des états 0 et 1, le même circuit peut être utilisé comme interrupteur logique à courant alternatif. L'auteur explique la caractéristique à retard (monostable) du circuit de base et calcule le temps de retard pendent que les propriétés de commutation sont utilisées pour la construction de portes ET et OU. Les propriétés de commutation à retard et à courant alternatif rendent le système logique à transistor à couplage résistance-condensateur du circuit logique beaucoup plus puissant et souple.

La distorsion dans les oscillateurs à réaction par P. F. Ridler

Résumé de l'article aux pages 604 à 605

L'auteur analyse une théorie permettant de prédire la distorsion dans les oscillateurs à réaction et se rapportant à la distorsion dans la section d'amplification de l'oscillateur. Un circuit pratique ayant une distorsion de forme d'onde inférieure à 0.01 % est indique à l'appui de la théorie.

Transistors à effet de champ au silicium par R. Cullis

Résumé de l'article aux pages 606 à 608

La prolifération des types de transistors à effet de champ pouvant être obtenus actuellement à pu jeter la confusion dans l'esprit d'utilisateurs potentiels.

Cet article décrit la fabrication de variantes courantes du transistor à effet de champ au silicium et indique les propriétés des différentes structures dans le but d'éclairer certains points obscurs se rapportant au nouveau composant. Il indique enfin certains des développements futurs qui se produiront vraisemblablement dans cette nouvelle branche de la technique des transistors.

Fréquence de gain et de coupure d'un étage amplificateur à transistor par S. Turk

Résumé de l'article aux pages 609 à 610

On a déterminé l'effet de l'impédance interne de la source de signaux sur le gain et sur les caractéristiques de fréquence d'un étage amplificateur à transistor. Il a été constanté que la fréquence de coupure obtenue est toujours supérieure à ^{W}B dans un montage à émetteur commun et toujours inférieure à ^{W}a dans un montage à base commu^{n}e.

Un convertisseur c.c./c.a. à fréquence stable par G. Smiljanic

Résumé de l'article aux pages 610 à 611

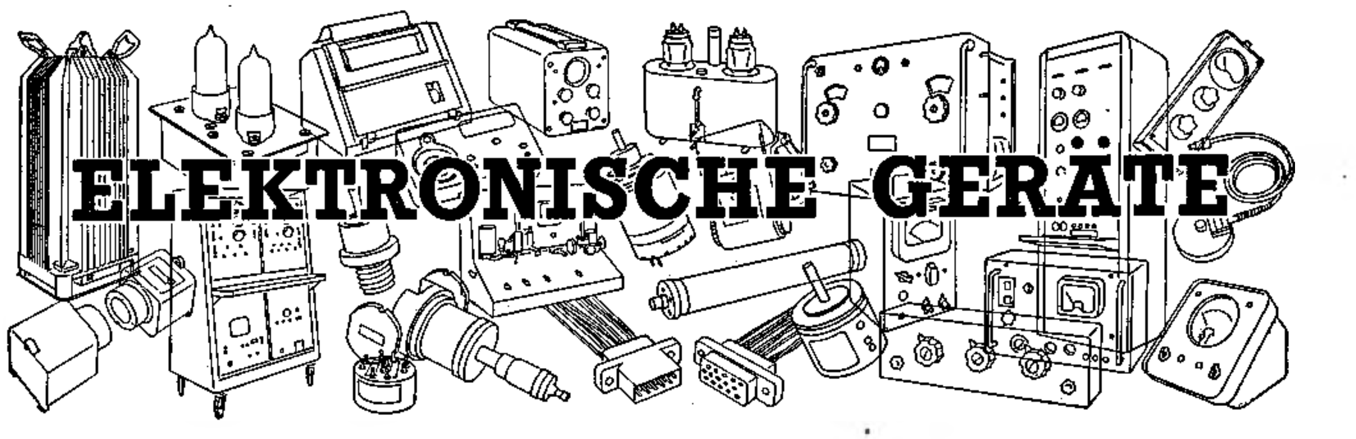
La fréquence d'un convertisseur c.c./c.a. peut être rendue indépendante des changements de tension d'alimentation en utilisant des diodes Zéner pour stabiliser l'amplitude de tension d'un transformateur d'entraînement à base de transistor qui détermine la fréquence. Les variations de fréquences obtenues avec le circuit en question sont de 1% ou de moins de 1% pour des changements de tension d'alimentation de 20%.

La solution automatique de la réponse de fréquence de réseaux par T. Fleetwood

Résumé de l'article aux pages 612 à 614

L'auteur décrit une méthode courante pour résoudre la réponse de réseaux électriques. Il décrit en outre un programme de calcul permettant de mettre en oeuvre cette méthode. A l'aide de ce type de programme, on peut obtenir aisément des résultats pour n'importe quel montage de réseaux.

ELEKTRONISCHE GERÄTE



Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 620 bis 625

Frequenznormal

The Marconi Co. Ltd, Chelmsford, Essex

(Abbildung Seite 620)

Die Marconi Co Ltd hat einen äusserst konstanten Oszillator entwickelt, dessen Genauigkeit mit der eines modernen Atom-Frequenznormals vergleichbar ist. Dieses Gerät hat eine Frequenzschwankung von nur 30×10^{-6} einer Periode bei 100 kHz und ist eins der konstantesten und genausten Frequenznormale in der Welt, die aus der Fertigung lieferbar sind.

Der Oszillator hat eine Kurzzeitkonstanz von besser als 3×10^{-10}, eine Alterungskonstanz von 1×10^{-9} je Monat und erzeugt drei gleichzeitige Normalfrequenzen von 100 kHz, 1 MHz und 2,5 MHz.

Ein im Werk Hackbridge der Firma in Süd-London entwickeltes und gefertigtes, hochpoliertes Quarzkristall mit A-T-Schnitt schwingt mit der fünften Harmonischen bei 2,5 MHz. Das Polierverfahren geht in mehreren Stufen über das in der Fertigung üblicher Kristalle angewandte hinaus. Innerhalb des Kristalls, das einen Gütefaktor von über 6×10^6 hat, gibt es acht perfekte Grenzflächen. Für die Aussenanschlüsse werden auf zwei Aussenflächen des Kristalls Goldkontakte thermisch aufgedampft.

Die ausserordentliche Temperaturstabilität des Kristalls und Oszillators wird durch Einbau in den inneren von zwei Thermostaten gewährleistet, die beide proportional geregelt werden und schon auf Temperaturschwankungen von einem Tausendstel Grad ansprechen. Ausserdem arbeitet das Kristall mit dem extrem niedrigen Leistungspegel von 0,5 μW.

Das 2,5-MHz-Signal des Quarzes wird in die Frequenzteilerstufe gespeist, wo ein "Treppenteiler" die 1-MHz- und 100-kHz-Signale mit hoher Phasenkonstanz erzeugt. Diese Schaltungen sind zusammen mit den Puffer-Endstufen als volltransistorisierte Druckschaltungsplatten ausgeführt.

Das ganze Gerät ist mit seinen Stromversorgungen für Betrieb an normalen Netzen an eine 19"-Frontplatte angebaut. Bei Ausfall der Netzversorgung wird der Oszillator mit derselben Frequenzkonstanz aus den eingebauten Notbatterien betrieben.

Wegen seiner Genaugkeit und Konstanz ist dieses Gerät für Navigations-Leitsysteme, Rechner, Satellit-Antennensteuerung, militärische und Verteidigungszwecke sowie viele andere Anwendungsmöglichkeiten in Systemen hoher Präzision ideal geeignet.

EE 85 751 *für weitere Einzelheiten*

Digitalanzeiger

K.G.M. Electronics Ltd, Bardolf Road, Richmond, Surrey

(Abbildung Seite 620)

Der zierliche, schlanke Anzeiger wurde speziell für unkomplizierten Anbau an Druckschaltungsplatten entwickelt, wo er gleichzeitig als Handgriff zum Herausziehen der Platte und als Digitalanzeige dient.

Der mit K.G.M.-Anzeiger M23 bezeichnete Baustein kann bis zu 10 Zeichen darstellen, ist mit den für Streitkräfte zugelassenen Zwerglampen mit Nennspannungen von 6 ... 28 V bestückt und hat die isolierte gemeinsame Anschlusseinrichtung.

EE 85 752 *für weitere Einzelheiten*

Vorwahlzähler-Zeitgeber

Advance Controls Ltd, Cheltenham, Gloucestershire

(Abbildung Seite 620)

Advance Controls Ltd gibt die Ergänzung ihres Fertigungsprogrammes durch Einführung des neuen Zeitgeber-Vorwahlzählers TB4 bekannt. Dieses vielseitige Gerät hat eine Zählgeschwindigkeit von 5000 Hz; für Zuverlässigkeit und Wirtschaftlichkeit ist es als Kombination transistorisierter Schaltungen und Kaltkathoden-Zählrohre ausgelegt.

Das Modell TB4 hat vier Dekaden und zählt jede beliebige Quantität bis zu 9999 ab; man kann es für Programmsteuerung oder als Zeitgeber für zyklische Vorgänge mit jeder beliebigen Zeitdauer bis zu 99,99 s in 0,1-s-Schritten einsetzen. Ein einstellbares Vorsignal für Abzählen oder Zeitdauer gibt ein Alarmsignal oder löst eine Steuerfunktion aus.

Auf der Rückwand ist ein Mehrfachprogrammstecker angeordnet, und Eingang-, Ausgang- sowie Rückstellanordnungen sind so einstellbar, dass sie fast allen Anforderungen angepasst werden können. Standard- oder Sonderzweck-Programmstecker sind entweder einsatzbereit oder zur Verdrahtung beim Kunden nach Anweisungen im mitgelieferten Handbuch lieferbar.

EE 85 753 *für weitere Einzelheiten*

Steckleisten

Plessey-UK Ltd, Wiring & Connectors Division, Ilford, Essex

(Abbildung Seite 621)

Der Geschäftsbereich Verdrahtung und Steckverbindungen der Plessey-UK Ltd hat die Fertigung einer Serie neuer Steckleisten für gedruckte Schaltungen in Streifen aufgenommen, die mit bis zu 80 Kontakten bestückt werden können.

Die Steckleisten haben die Baumusterbezeichnung 603 und können genau nach Kundenwünschen abgeschnitten werden; es gibt auch verschiedene Kombinationen, die den Anforderungen der einzelnen gedruckten Schaltungen gerecht werden.

Das Baumuster 603 wird mit den

Normen entsprechenden Befestigungsabmessungen geliefert. Es gibt auch Befestigungsfüsse, die die Streifen an beiden Enden abschliessen. Sind jedoch offene Presslinge erwünscht, so können Löcher für selbstschneidende Befestigungsschrauben in die Presslinge gebohrt werden.

Die neuen Steckleisten bestehen aus Gehäusen, die direkt mit 1,6 mm starken Leiterplatten zusammenpassen. Die hohe Qualität teurer Steckverbindungen wird durch Bestückung mit einfach geformten freitragenden Kontakten und optimale Ausnutzung von Werkstoff und Raum erzielt.

Die Kontakte sind speziell für gute Kontaktgabe und Kontakt-Kraft-Kenndaten geformt und reduzieren Abnutzung und Kontaktwiderstand trotz Schwankungen in der Plattenstärke auf ein Minimum. Niedriger Polwiderstand wird durch Verwendung von Messingkontakten mit Lötanschluss gewährleistet. Für die Oberflächenbehandlung stehen Hartgold, Schnellvergoldung auf Silber oder Spezialvergoldung zur Wahl.

Die Streifen sind aus glasverstärktem Polycarbonat gepresst, haben ausgezeichnete mechanische und elektrische Eigenschaften und sind feuerhemmend.

Die für 750 V_{eff} bei 5 A Dauerstrom bemessenen Steckleisten sind für Betriebstemperaturen von $-40°C...+85°C$ bestimmt und haben dann einen Kontaktwiderstand von weniger als 15 mΩ.

EE 85 754 *für weitere Einzelheiten*

Netzspannungs-Kontrollgerät

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(Abbildung Seite 621)

Kent Precision Electronics Ltd hat neuerdings eine Festkörperschaltung zur Überwachung von Standard-Drehstromversorgungen entwickelt, die wirksame, genaue und zuverlässige Kontrolle von Spannungsabweichungen gestattet und Schutz oder Warnung gibt. Das Gerät hat breite Anwendungsmöglichkeiten in Notstromversorgungen, der Spannungsregelung und in der Energieüberwachung für Datenverarbeitungsanlagen.

Im Prinzip besteht das Gerät aus einer Logikschaltung mit sechs Eingängen, die über einen Ausgangstrigger ein elektromagnetisches Standardrelais betätigt, wenn die Drehstrom-Eingangsspannungen entweder über oder unter dem vorgewählten Pegel liegen. Die Geräte lassen sich für jede beliebige Nennnetzspannung mit vorgegebener Prozentabweichung einstellen. Die Ausgangsrelaiskontakte stehen daher im Alarmzustand, wenn irgendeine der Drehstromspannungen von dem vorgewählten Wert abweicht, wodurch absoluter Unter- und Überspannungsschutz für jede Ausrüstung gegeben wird, die an das kontrollierte Netz angeschlossen ist.

Die Geräte sind entweder in Standardinstrumentgehäusen für Schalttafeleinbau oder in der neuen KPE-Modularkonstruktion wie abgebildet lieferbar.

EE 85 755 *für weitere Einzelheiten*

Leitfähigkeitsmesser

Electronic Switchgear (London) Ltd, 58 Wilbury Way, Hitchin, Hertfordshire

(Abbildung Seite 621)

Das Fertigungsprogramm der Electronic Switchgear (London) Ltd für Leitfähigkeitsinstrumente wurde neuerdings durch Einführung des tragbaren Leitfähigkeitsmessers MC3 ergänzt.

Ein Merkmal dieses Gerätes ist die praktisch unzerstörbare Messzelle, die direkt in die Seite des Instrumentgehäuses eingesteckt wird und ausser gelegentlichem Säubern mit einer mitgelieferten Bürste keinerlei Wartung erfordert. Die Zelle ist röhrenförmig und hat bei 10 cm^3 Kapazität einen unbeschränkten Innendurchmesser mit ringförmigen Elektroden, die in gleicher Weise Leitfähigkeitsmessungen von Flüssigkeiten, Pasten und Schlämmen gestatten.

Der Messumfang ist 1 ... 100 000 μS/cm in drei mittels Schalter einstellbaren Messbereichen; eine vierte Schaltposition gestattet Prüfen der effektiven Arbeitsweise und Messgenauigkeit des Instrumentes.

Die transistorisierten gedruckten Schaltungen werden aus einer kleinen, leicht auswechselbaren Langlebensdauerbatterie gespeist, die in einem Spritzgussgehäuse von 241 × 190 × 89 mm untergebracht ist.

EE 85 756 *für weitere Einzelheiten*

Vielfachmessgerät

Vertrieb: Croydon Precision Instrument Co., Hampton Road, Croydon, Surrey

(Abbildung Seite 621)

Dieses von Norma in Wien hergestellte Vielfachmesgerät hat 40 Messbereiche für Gleich- und Wechselspannungen, Gleich- und Wechselströme, Widerstand und Ausgangsleistung. Ausserdem hat es einen 12-mV-Gleichspannungsbereich mit 40 kΩ/V Empfindlichkeit und eine in Celsius geeichte Temperaturskala für Anschluss an Eisen-Konstantan-Thermoelemente.

Für Gleichspannungsmessungen ist die Empfindlichkeit 20 kΩ/V, für Wechselspannungen 4 kΩ/V. Die Messunsicherheit ist bei Gleichstrom ±2,5%, bei Wechselstrom ±3,5% bis zu 500 Hz und ±7,5% bis 30 kHz.

Das Gerät wiegt 340 g, hat ein Messwerk mit Spannbandaufhängung und kann Überlastungen bis zu 1 000 Prozent aushalten.

EE 85 757 *für weitere Einzelheiten*

Registereinrichtungen

Benson-Lehner Ltd, West Quay Road, Southampton

(Abbildung Seite 621)

Zwei von Benson-Lehner Ltd eingeführte neue Fotoregistereinrichtungen sind in erster Linie für Verpackungs-, Einwickel- und Beutelfüllmaschinen gedacht.

Beide Geräte sind volltransistorisiert und ohne Relais, Kontaktgeber oder bewegliche Teile konstruiert, so dass höchste Betriebsgeschwindigkeiten gefahren werden können und ein grosser Teil der in der schmutzigen Atmosphäre und für die Nachjustierung erforderlichen Ausfallzeit eliminiert wird.

Je nach Position der Registermarken spricht Typ 2C5A sofort durch Beschleunigung oder Verzögerung eines elektrischen Hilfsmotors oder einer anderen Regeleinrichtung auf jede Änderung des normalen Registers an.

Das andere Gerät ist eine nur in einer Richtung wirkende Regeleinrichtung für Einbau in Maschinen, in denen das Verpackungsmaterial absichtlich in jedem Arbeitstakt zu weit abgezogen wird. Das Korrektursystem verlangsamt den Vorschub, sowie die Registermarke die fotoelektrische Zelle deckt.

EE 85 758 *für weitere Einzelheiten*

Transistorisierte Stromversorgung

Fenlow Electronics Ltd, Springfield Lane, Weybridge, Surrey

(Abbildung Seite 622)

Die Stromversorgung PU5 ist für Anschluss an Wechselstromnetze von 200 ... 250 V oder 100 ... 150 V, 50 Hz oder 60 Hz ausgelegt und hat eine Höchstleistungsaufnahme von 290 W. Die Ausgangsspannung ist zwischen 0 und 150 V— regelbar und bis zu 1,5 A konstant.

Die Geräte können mit zwei Betriebsarten arbeiten, und zwar Konstantspannung und Konstantstrom; sie sind mit Betriebsartsautomatik - Übergangsschaltungen und visueller Anzeige der Betriebsart ausgerüstet und gegen zufälligen Kurzschluss an den Ausgangsklemmen geschützt.

In der Spannungsbetriebsart wird die Ausgangsspannung mittels geeichter Dekadenschalter in 1-V-Stufen eingestellt und der durch die angelegte Last entnommene Strom auf einem Schalttafel-Messgerät angezeigt. In der Strombetriebsart wird der Strom mittels eines für schnelles Einstellen markierten Bedienknopfes und eines Strommessers gewählt. Wenn die externe Last bei Regelung durch den kritischen Wert geht, wird automatisch umgeschaltet und dieser Vorgang durch zwei farbige Signallampen angezeigt. Spannungsfernmessung ist vorgesehen, um den Einfluss der Impedanz von Verbindungskabeln zu vermeiden. Ausserdem kann die Spannung fernprogrammiert werden; weiterhin können Geräte zur Erhöhung von Ausgangsspannung oder -strom in Serie oder parallel geschaltet werden.

Die Ausrüstung wird in halben Gestellgehäusen geliefert, so dass sich ein Paar mittels eines Zwischenchassis in ein 19"-Gestell einbauen lässt.

EE 85 759 *für weitere Einzelheiten*

Drucktastenschalter für gedruckte Schaltungen

Mills & Rockleys (Electronics) Ltd, Swan Lane, Coventry

(Abbildung Seite 622)

Der Schalter wurde besonders für Lieferung in Bausatzform für Zusammenbau (mit Druckschaltungsplatten entworfen. Man ist daher nicht auf die viel komplexeren, bereits zusammengebauten Schalterbaugruppen angewiesen, die man bei Ausfall oder ähnlichen diesen Typen anhaftenden Fehlern kaum aus der Druckschaltung ausbauen kann.

Die Konstruktion des Schalters setzt voraus, dass die gedruckte Schaltung geeignete Leiteranordnungen hat, die mit den öffnenden oder schliessenden Kontakten des Schalters zusammenfallen.

Die Kontaktbrücken tragenden Drucktasten arbeiten mit den Leitern der Plattine innerhalb eines Schlitzes in der Drucktaste, der parallel zur Oberfläche der Druckschaltungsplatte verläuft, zusammen.

Jede Drucktaste ist mit einer internen Feder bestückt, so dass bei Drücken der Taste jede andere eingerastete Taste, die nicht gleichzeitig gedrückt wird, in die Originalstellung zurückspringt. Das wird durch die Formgebung des Tastenpresskörpers und eine auf die Druckschaltung montierte Halteleiste erreicht.

Das Zusammenbauen der Bauteile ist daher sehr einfach, da die Tastenpresskörper komplett mit Kontakten und Federn lediglich auf die Druckschaltungsplatten aufgeschoben und durch eine der gewünschten Tastenzahl entsprechende Halteleistenlänge festgehalten werden. Die Konstruktion erlaubt auch unterschiedlichen Abstand zwischen Tasten, falls das für irgendwelche Geräte wünschenswert erscheint.

Die Tasten werden daher in Einheitsform hergestellt, und es bleibt dem Benutzer überlassen, sie in jeder beliebigen Anzahl zusammenzubauen, da er nur die geeignete Halteleistenlänge zu wählen und die gedruckte Schaltung entsprechend zu entwerfen braucht. Mills & Rockleys werden daher in ihrem Werk für gedruckte Schaltungen als Teil der Standardfertigung die entsprechenden Werkzeuge erstellen, die Druckschaltungsplatten fertigen und die für die Platten erforderlichen Drucktasten - Bauteile liefern und montieren.

EE 85 760 *für weitere Einzelheiten*

Digitalvoltmeter

Roband Electronics Ltd, Lowfield Heath Road, Charlwood, Horley, Surrey

(Abbildung Seite 622)

Roband Electronics Ltd hat zwei neue Serien kompakter Digitalvoltmeter eingeführt. Die beiden Grundmodelle haben die Bezeichnung RDV 32 und RDV 42; sie sind beide entweder im Gehäuse, als Gestelleinschub oder als Einbauinstrument lieferbar. Fernanzeigeeinrichtungen, ein Umsetzer für Serienübertragung und Einbereichvoltmeter für Sonderzwecke sind auch lieferbar.

Beide Geräte arbeiten nach dem Prinzip der aufeinanderfolgenden Näherungen mit einer Messhäufigkeit von 30 Hz; sie geben Dezimalsignale oder binärkodierte Dezimalsignale an Regelschaltungen oder Druckwerke ab.

Modell RDV 42 hat Bereichsendwerte von 0,9999 bis zu 999,9 V in vier Bereichen mit einem Auflösungsvermögen von 100 μV. Modell RDV 32 hat Bereichsendwerte von 1,099 bis zu 999,0 V in vier Bereichen mit 1 mV Auflösungsvermögen. Beide Geräte haben eine Mindestmessgenauigkeit von 0,1 Prozent $\pm$ Ziffer.

Es finden durchgehend Festkörperschaltungen Verwendung; diese Geräte sind für die schwierigsten industriellen Betriebsbedingungen konstruiert, und für Einsatz in korrodierender Atmosphäre gibt es eine hermetisch dichte Ausführung.

EE 85 761 *für weitere Einzelheiten*

Zähler-Tastkopf

Racal Instruments Ltd, Crowthorne, Berkshire

(Abbildung Seite 622)

Racal Instruments Ltd kündigt an, dass der neue Aktive Tastkopf SA.544 nunmehr als Zählerzubehör lieferbar ist. Dieser volltransistorisierte Taster ist für Einsatz mit den direktanzeigenden Racal-Digitalzählern SA.540 und SA.550 für Frequenzmessungen von 10 MHz und 100 MHz gedacht.

Anwendung des aktiven Tasters erhöht die Zählerleistung, was durch Erhöhung der Eingangsimpedanz des Zählers in fünf Bereichen auf 10 MΩ und Erhöhung der Empfindlichkeit auf 5 mV charakterisiert ist; dadurch werden viele Messungen im 100-MHz-Frequenzbereich möglich, die bisher nur mit externen Spezialverstärkern und Impedanzanpassung durchführbar waren. Die dem Zähler eigene hohe Messgenauigkeit wird für alle Messfunktionen bewahrt.

Der ausschliesslich mit Halbleitern arbeitende Tastkopf wird direkt von dem Zähler gespeist, an den er angeschlossen ist. Er ist robust gebaut und für Betrieb unter den extremen Umgebungsbedingungen militärischen Einsatzes konstruiert. Dem Tastkopf wurde die NATO-Listennummer 6625-99-519-1573 zugeteilt.

EE 85 762 *für weitere Einzelheiten*

Siebenspurmagnetkopf

Gresham Lion Electronics Ltd, Lion Works, Hanworth Trading Estate, Feltham, Middlesex

(Abbildung Seite 623)

Gresham Lion Electronics Ltd hat einen Magnetkopf mit sieben Spuren und für Lesen nach Schreiben entwickelt. Der Kopf wurde für digitale Anwendungszwecke entwickelt und gestattet sofortige Verifizierung der aufgezeichneten Information.

Es gibt zwei Ausführungen, von denen die eine mit dem Schreibelement links mit IBM-Einrichtungen kompatibel ist, die andere das Schreibelement rechts hat. Die Vorderfläche ist zur Verringerung des Langwellenrauschens profiliert und zur Vermeidung von Rillen durch Magnetbandkanten ausserhalb der Spuren ausgespart.

Der Kopf ist 31,8 mm hoch, 26,2 mm tief und 25,4 mm breit; der Azimuth wird innerhalb $\pm$0,0002 RAD, die Erhebung innerhalb 0,0005 RAD eingehalten. Die Spaltstreuung fällt in ein 0,002 mm breites Band, und die Spaltlinien sind innerhalb 0,0002 Rad parallel. Die Lesespuren sind bei 1,78 mm Abstand 0,79 mm, die Schreibspuren bei demselben Abstand 1,2 mm breit.

Auf 2M-Magnetband T488 ist der Sättigungsstrom bei 300 Bit/Zoll und 99% Sättigung 36 mA + 15% (Spitze-Spitze). Bei einem Schreibstrom von 54 mA_{SS} und einer Bandgeschwindigkeit von 305 cm/s ist die Lesespannung 44 mV_{SS}.

Bei gleichzeitigem, gleichphasigen Schreiben auf sieben Spuren mit einem 100-kHz-Rechteckwellensignal, das eine Anstiegszeit von 2 μs und einen Stromwert von 60 mA_{SS} hat, ist der Durchbruch zu einer beliebigen Lesespur unter 10mV. Dieser Wert kann durch eine die Kopfvorderseite abdeckende, unmittelbar hinter dem Band montierte Abschirmung im Verhältnis 2:1 reduziert werden.

EE 85 763 *für weitere Einzelheiten*

Kühlkörper

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

(Abbildung Seite 623)

Die Hersteller des Gestell- und Chassis-Systems "Lektrokit", A.P.T. Electronic Industries Ltd, haben vor kurzem das Lektrokit-Programm durch eine Serie von Kühlkörpern erweitert, die vier Modelle umfasst.

Kühlkörper Nr. 1 (Teil-Nr. LK.1111) wurde für Transistoren im international genormten Gehäuse TO-3 entworfen und ist so geformt, dass die verschiedensten Aufbaumöglichkeiten—einzeln oder in Gruppen—bestehen. Das Schlitz-Muster gestattet freien Luftumlauf durch den Kühlkörper, so dass bei jedem beliebigen Winkel wirksame Kühlung gewährleistet ist. Die Löcher für die Transistorverbindungen sind als Langlöcher ausgeführt, so dass zwei Kühlkörper zur wesentlichen Erhöhung der Nennwärmeabfuhr Rückseite an Rückseite mit einem Transistor zusammenmontiert werden können.

Kühlkörper Nr. 2 (Teil-Nr. LK.1121) hat dieselben Abmessungen, jedoch Befestigungslöcher für die grösseren Leistungstransistorgehäuse bis zu TO-36 sowie für Leistungs-Zenerdioden und -Gleichrichter.

Kühlkörper 1 und 2 werden aus 1,22 mm starkem Aluminiumblech hoher Leitfähigkeit in gefalteter und geschlitzter Form hergestellt. Sie werden in der Fertigung für die Befestigung von Leistungstransistoren gelocht und zur Unterstützung der Wärmezerstreuung in schwarz ausgeführt. In stiller Luft haben sie einen Wärmewiderstand von 3,5°C/Watt.

Kühlkörper 3 und 4 (Teil-Nr. LK.1131 bzw. 1141) werden aus Vollaluminium in geripptem Strangpressprofil hergestellt, für die Transistorbefestigung gebohrt und ebenfalls in schwarz ausgeführt. Der Wärmewiderstand dieser Bauelemente in stiller Luft ist etwa 1,6°C/Watt wenn senkrecht montiert und 1,75°C/Watt wenn horizontal eingebaut. Die Konstruktion wurde auf maximale thermische Kapazität und Wärmezerstreuung ausgerichtet. Kühlkörper Nr. 3 ist für das international genormte Transistorgehäuse TO-3 ausgelegt, und Nr. 4 kann die grossen Leistungstransistoren im Gehäuse TO-36 sowie Leistungs-Zenerdioden und -Gleichrichter aufnehmen.

EE 85 764 *für weitere Einzelheiten*

Druckmesswandler

S.E. Laboratories (Engineering) Ltd, North Feltham Trading Estate, Feltham, Middlesex

(*Abbildung Seite 623*)

In sich geschlossene, gleichstromgespeiste, vorgeeichte Druckmesswandler des Types S.E.37 werden nunmehr von S.E.L. Ltd angeboten. Anwendungsmöglichkeiten bestehen in erster Linie in Raketen, Überschall- und anderen Flugzeugsystemen sowie in Umgebungsbedingungen, unter denen die meisten anderen Messwandler auf Grund der extremen Schwingungen ausfallen.

Messgenauigkeit und Zuverlässigkeit dieser Geber sind daher sehr hoch; mit der Ausgangsspannung oder dem Ausgangsstrom kann man direkt die verschiedenen Geräte wie z.B. Magnetbandgerät, Oszillograf, Telemetrieausrüstung, mittels hoher Spannungspegel gesteuerte Oszillatoren, Kommutatoren, Digitaldarsteller oder direktablesbare Anzeiger ansteuern.

Vollintegrierte geschweisste Schaltungen gewährleisten hohe Betriebssicherheit und Dauerbetrieb bei Temperaturen von 150° C mit einem Temperaturbeiwert von weniger als 0,015 Prozent/°C; bei unterbrochenem Betrieb kann man für drei Minuten mit einem Gesamtfehler von weniger als 0,35 Prozent bis zu 300° C gehen.

Auf Erdschleifen und hohe umlaufende Ströme—z.B. in Flugzeug- und Bodentestanlagen — zurückzuführende Rauscheffekte werden durch völlige Trennung von Gleichstrom-Speiseleitungen von den Ausgangsschaltungen auf ein Mindestmass reduziert.

Der gleichstromgespeiste Messwandler S.E.37 ist für Differenz- und Absolutdruckmessungen mit korrodierenden Gasen und Flüssigkeiten bei Messbereichen von 0...0,14 bis 0...280 kp/cm² geeignet. Die Ausgangsspannung von —5...+5V für Skalenendwertdruck kann je nach Kundenwünschen so vorgewählt werden, dass sie entweder zwischen Null und dem Maximum oder zwischen negativ und positiv pendelt (z.B. 0...70 kp/cm², 0...5 V oder -70 kp/cm²...$+70$ kp/cm², 0 ... ±5 V). Die geschweisste Bauform und Eliminierung aller Gummi- und Kunststoffdichtungen gestattet Einsatz mit allen Flüssigkeiten, die mit rostfreiem Stahl vereinbar sind.

EE 85 765 *für weitere Einzelheiten*

Präzisions-HF-Messbrücke

The Wayne Kerr Laboratories Ltd, Sycamore Grove, New Malden, Surrey

(*Abbildung Seite 623*)

Diese neue Grossbereich-Messbrücke arbeitet bei Kapazitäts- und Leitwertmessungen zwischen 100 kHz und 1 MHz mit 0,1 Prozent, bis zu 5 MHz mit verringerter Messgenauigkeit. Das Modell B201 ist einschliesslich Batterien (oder Gleichrichter), Einsteckquelle und Detektor, Normalen und einem Nullanzeiger völlig tragbar. Interne Modulation mit 1 kHz und eine Detektorausgangsklinke gestatten Benutzung von Kopfhörern; Bedienelemente auf der Frontplatte ermöglichen Regelung des Signalpegels und der Detektorverstärkung. Der Messumfang ist 0,0001 pF...0,1 μF und 0,001 μS...1 S (1 Ω...1 GΩ).

Das Konzept des Instrumentes beruht auf Transformatoren für die genauen Spannungs- und Stromverhältnisse, mittels derer der Prüfling mit internen Kapazitäts- und Leitwertnormalen verglichen wird. Die Spezialkonstruktion des Anschlussblocks hat eine neutrale Klemme, die Durchführung dreipoliger Messungen ermöglicht. Einsteckquellen und Detektoren sind zur Zeit für 100 kHz und 1 MHz lieferbar, jedoch gestattet Blindzubehör Anschluss externer Geräte zum Messen über den Bereich 100 kHz...5 MHz.

EE 85 766 *für weitere Einzelheiten*

Vielfachmessgerät

Salford Electrical Instruments Ltd, Peel Works, Silk Street, Salford, 3, Lancashire

(*Abbildung Seite 624*)

Selectest Baumuster II ist eine Entwicklung der Originalkonstruktion mit zusätzlichen Verfeinerungen und Modifikationen, die von Benutzern angeregt wurden.

Am Anschlussblock gibt es jetzt Buchsen zum Einstöpseln der Zuleitungen. Die Bereichsschalter wurden neu entworfen und geben sogar noch bessere Lesbarkeit sowie vereinfachte Bereichswahl. Sie können jetzt in beiden Richtungen durch 360° gedreht werden und sind elektrisch verriegelt.

Die automatische Überlastungsabschaltung und die internen Bauelemente sind auf robuste Druckschaltungskarten montiert, die in ein widerstandsfähiges Melamingehäuse eingebaut sind. Es gibt zwei Modelle: das 'Selectest Super 50' mit einem 50-μA-Messwerk und 32 internen Bereichen zum Messen von Gleich- und Wechselstrom (Spannung, Strom und Widerstand) und das 'Selectest Super K' mit einem 1-mA-Messwerk und 39 internen Bereichen.

EE 85 767 *für weitere Einzelheiten*

Miniatur-Ziffernanzeige

Counting Instruments Ltd, 5, Elstree Way, Boreham Wood, Hertfordshire

(*Abbildung Seite 624*)

Eine neue Mikrominiatur-Ziffernanzeige mit Projektion von hinten wurde von Counting Instruments Ltd angekündigt. Trotz der kleinen Abmessungen von 19 mm × 12,7 mm breit × 51 mm ist die projizierte Ziffer 9,5 mm hoch.

Der neue Baustein ist in der Arbeitsweise den 95-mm-, 25,4-mm- und 16-mm-Anzeigen der Firma mit 12 Lampen ähnlich. Er enthält am hinteren Ende elf Miniaturglühlampen, eine Reihe von Linsen und einen von vorn zu betrachtenden Schirm, auf den elf Meldungen entweder einzeln oder in Kombination projiziert werden können. Da diese Meldungen auf Film hergestellt werden, kann die Anzeige beliebige Ziffern, Buchstaben, Worte, Symbole oder Farben darstellen, die sich fotografisch wiedergeben lassen.

Die Anzeige ist eine Einsteckeinheit für ein Gehäuse, in dem die Lampenverbindungen untergebracht sind und das fest verdrahtet werden kann. Zum leichteren Auswechseln von Lampen oder Änderung der Meldungen ist die Einsteckeinheit nach vorn durch die Frontplatte herauszuziehen.

Es gibt vier verschiedene Typen der Einsteckeinheit: auf einer Zwischenfrontplatte mit durchgehendem Schirm für Mehrfachzusammenbau; mit Einzelschirm für bündigen Einbau in die Frontplatte; oder mit Einzelschirm und mit Voll- oder Teildeckrahmen. Alle Typen können zu Mehrfachanzeigen zusammengebaut werden.

EE 85 768 *für weitere Einzelheiten*

Normalbezugsmagnet

Preformations Ltd, Cheney Manor, Swindon Wiltshire

(*Abbildung Seite 624*)

Preformations Ltd, eine Firma der Plessey-Baulemente-Gruppe, liefert nunmehr ein Sortiment völlig abgeschirmter Normalbezugsmagnete.

Diese für die Eichung von Tastspulen, Flussmessen, Halleffekt- und ähnlichen Gaussmetern konstruierten Magnete waren ursprünglich für Einsatz mit der Tastspulen-Standardgrösse der W. G.

Pye & Co. Ltd in Cambridge bestimmt. Sie eignen sich jedoch auch für die Eichung beliebiger Tastspulen oder Halleffektsonden mit 10 mm Höchstdurchmesser, die nicht dicker als 4,75 mm sind.

Die Bezugsmagnete werden während der Fertigung sowohl magnetisch als auch gegen periodische Temperaturänderungen vollstabilisiert, so dass sie gegen durch Temperaturschwankungen oder externe Magnetfelder hervorgerufene nichtumkehrbare Verluste geschützt sind.

Für Eichzwecke wird ein magnetisches Kernresonanz-Magnetometer herangezogen.

Magnete sind mit Quer- oder Längsfeldern und folgenden Kraftliniendichten lieferbar:

Längs 300 und 500 Gauss
Quer 1 000, 3 000, 5 000, 10 000 und 15 000 Gauss

Die Genauigkeit liegt bei 20° C innerhalb ±0,05 Prozent der markierten Kraftliniendichte. Streufelder unter 150 Oersted haben durch die magnetische Stabilität keinen nachhaltenden Einfluss auf die Magnete; die Temperaturstabilität ist im Bereich −10° C . . . . . +90 C in der Grössenordnung von 0,2%/°C. Die Höchstbetriebstemperatur ist 100° C.

EE 85 769 *für weitere Einzelheiten*

Messbuchse für gedruckte Schaltungen

Sealectro Ltd, Walton Road, Farlington, Portsmouth, Hampshire

(*Abbildung Seite 624*)

Die neue PTFE-Messpunktbuchse SKT-105 PC L1 wurde für Einführung in Druckschaltungsplatten-Chassis ausgelegt. Das Kontaktende dieses neuen Bauelementes, das weiter aus dem Chassis vorsteht als die herkömmlicheren Typen, steht rechtwinklig zur Platte. Die zusätzliche Höhe über dem Chassis ermöglicht Zwischenstellen von Messpunktbuchsen wie z.B. SKT-103 PC und erhöht damit die Anzahl der Einheiten, die in einem Chassis untergebracht werden können. Die Einführungstülle des SKT-105 PC L1 besteht aus reinem Polytetrafluoräthylen, und die Beryllium-Kupferkontakte sind dünn vergoldet oder versilbert.

EE 85 770 *für weitere Einzelheiten*

Elektronischer Zeitschalter

Solid State Controls Ltd, 30-40 Dalling Road, London, W.6

(*Abbildung Seite 625*)

Der "Variset" ist ein völlig integraler Zeitschalter, für dessen Ausgang ein Steckrelais zum Schalten von Starkstrom bis zu 5 A als Standard mitgeliefert wird. Das Grundgerät besteht aus einem durchweg mit Transistoren bestückten Zeitnetzwerk mit einer Lebensdauer von bis zu 100 Millionen Schaltspielen. Es ist ausreichend gegen Hochspannungseinschwingvorgänge geschütz und wird durch weite Spannungs- und Temperaturschwankungen nicht beeinflusst. Das Gerät arbeitet daher selbst unter ungünstigen industriellen Bedingungen sehr zuverlässig, und kennzeichnend ist die wiederholbare Genauigkeit von 1 Prozent. Das staubdichte Standard-Steckrelais hat die einzigen abnutzbaren Teile und ist leicht zugänglich.

Die Skala ist direkt in Sekunden geeicht, und bestimmte Zeiten lassen sich daher sehr leicht einstellen. Es gibt Verzögerungsperioden von 0,2 s...4 min in drei Zeitbereichen; in den Spannungsbereichen 12...120 V− und 48...415 V∼ gibt es acht Standardbetriebsspannungen.

Die Variset-Serie ist mit zwei für 5 A 250 V∼ bemessenen Umschaltkontakten bestückt, kann aber auch für andere Belastungen geliefert werden. Es gibt vier Grundbetriebsarten: Verzögerung der Erregung, Verzögerung des Abfalls, Drucktasten-Start, oder mit Rückführung. Ausserdem gestatten die guten Erholungseigenschaften sofortiges und genaues Wiedereinstellen zu jeder beliebigen Zeit während oder nach Ablauf der Zeitperiode.

EE 85 771 *für weitere Einzelheiten*

Miniatur-Drahtpotentiometer

Miniature Electronic Components Ltd, Copse Road, Woking, Surrey

(*Abbildung Seite 625*)

Der hauptsächlich für gedruckte Schaltungen mit 2,5 mm Rastergrundmass entwickelte Typ T20P entspricht in Ausführung und Abmessungen dem Transistor TO-5 mit 8,76 mm Gehäusedurchmesser und 6,5 mm Höhe. Das Bauelement ist für Weltraum- und industrielle Anwendung geeignet.

Durch Senkrechteinstellung innerhalb einer Umdrehung und feste Endanschläge wird die Einstellzeit klein gehalten, die Welle ist selbsthemmend, das Metallgehäuse und der Wickelkörper geben maximale Wärmezerstreuung, und die niedrige Induktivität gestattet Verwendung in Impuls- und HF-Schaltungen.

Listenmässige Widerstandswerte 50, 100, 200, 500 Ω, 1, 2, 5, 10 kΩ.

Nennbelastbarkeit ½ Watt bei 70° C.

Temperaturbereich −55°C ... +150°C.

EE 85 772 *für weitere Einzelheiten*

Aufprallgeräuschanalysator

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(*Abbildung Seite 625*)

Der Dawe-Aufprallgeräuschanalysator 1412 ist ein portables, batteriebetriebenes Instrument zum Messen der Spitzenamplitude und Analysieren der Kenndaten von Aufprallgeräuschen. Bei Einsatz mit einem der Dawe-Schallpegelmesser gibt das Modell 1412 eine Direktanzeige der Spitzenstärke des Geräusches und Information, mit deren Hilfe sich die Dauer und Wellenform eines Einzelaufschlages schätzen lassen. Für sich wiederholenden Schall wird eine "Quasispitzenschaltung" benutzt; zur Analyse des Frequenzgemisches können extern Oktavfilter angeschlossen werden. Ausserdem kann man das Instrument für Untersuchungen mechanischer Schwingungen einsetzen, indem man es mit dem Ausgang eines geeigneten Beschleunigungsmessers ansteuert; es kann aperiodische Wellenformen von jeder beliebigen Quelle, die 1...10 V_S an eine Eingangsimpedanz von 50 kΩ im Frequenzbereich 5 Hz...40 kHz abgeben kann, analysieren.

Das Modell 1412 besteht aus einem Röhrenverstärker, der drei Signalgleichrichter treibt, und einem Gleichspannungs-Röhrenvoltmeter, das nacheinander in jeden der Gleichrichterkreise geschaltet werden kann. Diese Gleichrichterschaltungen geben die Spitzen-, Quasispitzen- und Zeitmittelsignale ab, mittels derer man Form und Grösse der Wellenform bestimmen kann.

Die Spitzenschaltung hat eine extrem kurze Anstiegszeit; für einen Rechteckimpuls steigt die Anzeige in unter 50 μs auf innerhalb 1 dB des Endwertes. Die Ladung wird für mindestens 10 s innerhalb 1 dB aufrechterhalten. Die Quasispitzenschaltung hat eine schnelle Anstiegszeit von 1 ms und langsame Abfallzeit von etwa 0,6 s. Diese Kenndaten entsprechen im wesentlichen der britischen Norm BS 727:1954 "Eigenschaften und Leistung von Geräten zum Messen von Funkstörungen." Die Zeitmittelschaltung gibt eine Anzeige für den über eine vorbestimmte Periode aufrechterhaltenen Pegel; durch Auswechseln eines Widerstandes können Zeitkonstanten zwischen 2 und 200 ms eingestellt werden. Unter normalen Zuständen wird die Ladung für mindestens eine Minute innerhalb 1 dB gehalten.

Das Gerät wird in einem portablen Metallgehäuse mit abnehmbarem Deckel und Tragriemen geliefert.

EE 85 773 *für weitere Einzelheiten*

Zähler-Zeitgeber

Advance Electronics Ltd, Roebuck Road, Hainault, Ilford, Essex

(*Abbildung Seite 625*)

Nach dem Pflichtenblatt des neuen Advance-Zähler-Zeitgebers TC4 können mit diesem Gerät selbst Frequenzen über 5 MHz direkt gemessen werden.

Vier Messzeiten und sieben Zeitgeber von 1 μs bis zu 1 s gestatten genaues Messen von Frequenz, Perioden und Zeit. Das Gerät hat eine vierstellige, einzeilige Anzeige mit Kommaautomatik für Frequenzmessungen. Die Periodeneinrichtung (bis zu 10^5 Perioden) erlaubt genaues Messen extrem niedriger Frequenzen.

Die Grundgenauigkeit des TC4 für Zeit- und Frequenzmessungen wird durch einen thermostatgeregelten Quarzoszillator bestimmt, der bei Raumtemperatur auf 1×10^{-6} eingestellt wird. Wenn höhere Genauigkeit erforderlich ist, können externe Normale, die Sinus- oder Impulssignale abgeben, eingesetzt werden.

EE 85 774 *für weitere Einzelheiten*

Miniaturdrahtpotentiometer
Darstan Ltd, 263 Church Road. Thundersley, Essex
(Abbildung Seite 625)

Die kompakte Bauart dieses Potentiometers eignet sich hauptsächlich für gedruckte Schaltungen und andere Anwendungsgebiete, bei denen Raumeinsparung eine ausschlaggebende Rolle spielt.

Dieses drahtgewickelte Bauelement hat ein Metall-Abschirmgehäuse und eine geschlitzte Kappe mit Positionsanzeige, die sich ohne Schwierigkeiten mit Hilfe eines Schraubenziehers einstellen lässt. Die Anschlussdrähte sind entsprechend dem Rastergrundmass von 2,5 mm für gedruckte Schaltungen angeordnet.

Das Potentiometer kann mittels der Anschlussdrähte auf die Druckschaltungsplatten montiert werden und benötigt keine anderen Befestigungsmittel. Ein Voreinstellknopf ist lieferbar, und das Potentiometer kann mit einem Frontmontagewinkel ausgerüstet werden.

Standardwiderstandswerte sind 5, 10, 20, 50, 100, 200, 300 und 500 Ω, 1, 2, 5 kΩ bei 1 W Belastbarkeit. Die Standardtoleranz der Widerstandswerte ist ±10 Prozent.

EE 85 775 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Ein Tunneldiodenzuordner von R. V. L'Archeveque und B. McA. Sayers

Zusammenfassung des Beitrages auf Seite 574-579

Ein schnellspringender, in Dekaden betriebener Tunneldioden-Treppengenerator wird beschrieben. Eine Sprungrate in der Grössenordnung von 10^7 Sprüngen je Sekunde gestattet Anwendung des Gerätes für Quantelung, Analog-Digitalwandlung und andere Arbeitsweisen mit annehmbarer Präzision und der Möglichkeit hoher Geschwindigkeiten.

Ein Stromregelungssystem mit quadratischer Kennlinie von C. I. Sach

Zusammenfassung des Beitrages auf Seite 580-583

In diesem Beitrag wird ein Gerät beschrieben, das einen durch eine Last fliessenden Strom I_2 so regelt, dass er mit dem durch eine identische Last fliessenden Strom I_1 so schwankt, dass die Summe der Quadrate der beiden Ströme im wesentlichen konstant bleibt. Wenn diese Lasten Wicklungen auf einem gemeinsamen Kern sind, werden die Netto-Amperewindungen bei Aufrechterhaltung konstanter Verlustleistung geregelt.

Entwurfsdaten für einen Funktionsgenerator mit spannungsabhängigem Widerstand von G. Brown und R. A. B. Bond

Zusammenfassung des Beitrages auf Seite 584-588

Eine Funktionsverstärkerschaltung, die aus einem Eingangsnetzwerk mit verlängertem spannungsabhängigen Widerstand und einer linearen Widerstandsrückkopplung besteht, wird als Potenzgesetz-Funktionsgenerator analysiert. Tabellierte Daten gestatten die schnelle Berechnung der optimalen Verlängerungswiderstandswerte für Exponenten des Potenzgesetzes zwischen 1,5 und 4. Das Kriterion für Optimisierung ist der kleinste quadratische Fehler. Die Tabelle enthält auch die sich ergebenden Fehler des Funktionsgenerators. Fehler- und Temperaturkompensation werden kurz besprochen.

Ein Wechselrichter mit steuerbaren Siliziumzellen für 12 V—/230 V~ von D. I. Ross und B. E. G. Goodger

Zusammenfassung des Beitrages auf Seite 589-591

Ein mit Thyristoren bestückter Wechselrichter für 12 V—/230 V~ wird beschrieben. Die Frequenz des Wechselrichters wird durch einen 1-kHz-Stimmgabeloszillator gesteuert, der eine konstante Frequenz von 50 Hz abgibt.

Die Schaltung arbeitet leistungsfähig unter Belastungen, die zwischen Null und 70 W schwanken. Eine eingebaute Schaltung schützt die Thyristoren gegen zufällige Überlastung.

Der Entwurf minimaler NOR/NAND-Logikschaltungen von D. Zissos und G. W. Copperwhite

Zusammenfassung des Beitrages auf Seite 592-597

Ein systematisches Verfahren für den Entwurf minimaler NOR/NAND-Logikschaltungen wird beschrieben für den Fall, dass die Komplemente der Eingangssignale nicht zur Verfügung stehen. Eine Technik zur Verringerung der NOR/NAND-Schaltungen auf ein Mindestmass wird eingeführt und eine Methode zur Klassierung logischer Schaltungen erwähnt.

Ein Gleichstrom-Differentialverstärker mit Feldeffekt-Transistoren von J. C. S. Richards

Zusammenfassung des Beitrages auf Seite 598-600

Auf Erzielung einer niedrigen Temperaturdrift vorgespannte Feldeffekt-Transistoren bilden die Eingangsstufe eines stabilen Differentialverstärkers, der eine Bandbreite von 100 kHz, einen Unterdrückungsfaktor von über 5000 und eine bei 100 stabilisierte Differentialverstärkung hat. Die Zerodrift liegt unter 100 $\mu V/°C$ und der Eingangsstrom unter $10^{-9}A$. Die Nützlichkeit des Verstärkers wird durch das in den Transistoren erzeugte niederfrequente Rauschen (etwa 50 μV_{eff}) beschränkt.

Die RC-gekoppelte Transistorlogik (R.C.C.T.L.) als Verzögerungseinrichtung und wechselstromgekoppelter logischer Schalter von G. Philokyprou, Ph.D.

Zusammenfassung des Beitrages auf Seite 601-603

Das RC-gekoppelte Transistorlogik-System der Schaltungslogik wurde von P. L. Cloot (1,2) eingeführt und hat bei niedriger Geschwindigkeit viele ansprechende Eigenschaften. In wechselstromgekoppelter Verknüpfung (Eingang nur über einen Kondensator) verhält sich die Grundschaltung dieses Systems wie eine Verzögerungseinrichtung. Dieselbe Schaltung kann gleichzeitig als wechselstromgekoppelter logischer Schalter benutzt werden, wenn man die Zustände 0 und 1 festlegt. Nachfolgend wird der monostabile Verzögerungscharakter der Grundschaltung erklärt und die damit verbundene Verzögerungszeit berechnet; die Schalteigenschaften werden für die Konstruktion von UND- und ODER-Gattern ausgenutzt. Durch die erläuterten Verzögerungs- und Wechselstromschalteigenschaften wird das RC-gekoppelte Transistorlogik-System viel leistungs- und anpassungsfähiger.

Verzerrungen in Rückkopplungsgeneratoren von P. F. Ridler

Zusammenfassung des Beitrages auf Seite 604-605

Eine Theorie für die Vorhersage der Verzerrung in einem Rückkopplungsgenerator, die die Verzerrung der Ausgangswellenform auf die Verzerrung im Verstärkerteil des Oszillators bezieht, wird entwickelt. Zur Unterstützung der Theorie wird eine praktische Schaltung gegeben, deren Wellenformverzerrung unter 0,01 Prozent liegt.

Silizium-Feldeffekt-Transistoren von R. Cullis

Zusammenfassung des Beitrages auf Seite 606-608

Die grosse Anzahl der nunmehr lieferbaren Feldeffekt-Transistortypen (FET) hat bei denen, die ihre Anwendung in Betracht ziehen, eine gewisse Verwirrung hervorgerufen. In diesem Beitrag wird die Herstellung der Varianten des Silizium-FET beschrieben, denen man im allgemeinen begegnet; die durch die verschiedenen Strukturen auftretenden Eigenschaften werden umrissen in der Hoffnung, dadurch einige der mit diesem neuen Bauelement zusammenhängenden Unklarkeiten zu beseitigen. Abschliessend werden zukünftige Entwicklungstrends in diesem neuen Sektor der Transistortechnik angedeutet.

Verstärkung und Grenzfrequenz einer Transistorverstärkerstufe von S. Turk

Zusammenfassung des Beitrages auf Seite 609-610

Der Einfluss der internen Impedanz der Signalquelle auf die Verstärkung und Frequenzkennlinien einer Verstärkerstufe mit Transistor wird bestimmt. Es wird gezeigt, dass die erreichbare Grenzfrequenz in Emitterschaltung immer höher als ω_B und in Basisschaltung immer niedriger als ω_α ist

Ein Wechselrichter mit konstanter Frequenz von G. Smiljanic

Zusammenfassung des Beitrages auf Seite 610-611

Die Frequenz eines Wechselrichters wird von Schwankungen der Speisespannung unabhängig, wenn die Spannungsamplitude eines den Transformator in Basisschaltung treibenden Transistors, die die Frequenz bestimmt, mittels Zenerdioden konstant gehalten wird. Die in der vorgeschlagenen Schaltung bei Speisespannungsschwankungen von 20 Prozent auftretenden Frequenzschwankungen sind 1 Prozent oder niedriger.

Auflösungsautomatik für den Frequenzgang von Netzwerken von T. Fleetwood

Zusammenfassung des Beitrages auf Seite 612-614

Eine Routinemethode zur Auflösung des Frequenzganges elektrischer Netzwerke und die Anwendung eines Elektronenrechnerprogrammes zur Durchführung der Methode werden beschrieben. Mit einem Programm dieser Art können die Ergebnisse für beliebige in Betracht gezogene Netzwerkanordnungen sehr schnell erlangt werden.