Electronic Engineering

VOL. 39. No. 469.

COMMENTARY

THE National Electronics Research Council is to undergo a change of name and assume a new and wider function under the aegis of the Ministry of Technology. That is the main substance of a written reply by Mr. Anthony Wedgwood Benn, Minister of Technology, to a parliamentary question on 24 January last. The question was designed to elicit information on the relationship existing between the Ministry of Technology and the N.E.R.C. The full text of the reply was as follows:

"The National Electronics Research Council, an independent non-governmental organization, was set up in July 1964 under the chairmanship of Earl Mountbatten of Burma, in the belief that there was an urgent need to co-ordinate pure and applied research in electronics, and with the intention that it should indicate gaps in research, suggest priorities and prevent unnecessary duplication of effort.

"Since then, however, the Government has made new central arrangements for dealing with questions of science and technology. In particular the Ministry of Technology has become the sponsoring Department for the electronics industry. With the imminent transfer to it of the present functions of the Ministry of Aviation its research stations with an interest in electronics will be further augmented by the inclusion of the Royal Radar Establishment and the Royal Aircraft Establishment. Responsibility for civil scientific policy and questions relating to basic scientific research lies with the Department of Education and Science and the Science Research Council.

"These developments have profoundly affected the role of the National Electronics Research Council. In future its research will be continued under different arrangements. The Government has therefore proposed, with Lord Mountbatten's agreement, that the Council should accept a new and wider function under the aegis of the Ministry of Technology.

"The change would be reflected by an alteration in the title to the National Electronics Council (N.E.C.). The new Council would, in place of a more limited interest in research, assume responsibility that would embrace the impact of major developments in electronics on society. It will consider and advise the Government on the application of electronics to the national life. If it sees the need to promote research or encourage other specific action it will take the appropriate initiative either directly with the Ministry of Technology or through the various existing bodies operating in the electronics field. The new Council will continue to develop the valuable contacts which its predecessor had established with Commonwealth countries.

"There will be an appropriate widening of the membership of the governing body of the Council to reflect its new role. Also, as from the date at which this new arrangement takes effect, the Secretariat will be provided by the Ministry of Technology.

"I believe that the new Council has a very important job to do and am looking forward greatly to working with it."

During the first two years of its existence one of the major tasks of N.E.R.C. has been the formulation of plans and the carrying out of a feasibility study for a S.D.I. (Selective Dissemination of Information) system for scientists and engineers concerned with electronics. Immediately N.E.R.C. was established its Council, through various working parties, set out to obtain information on the existence and development of various 'current-awareness' information retrieval systems then in operation both in Great Britain and the U.S.A. As a result of this fact finding survey it was decided to operate a project basically of the type in which the documents received by the information centre are compared by computer with the interests of each individual taking part in the scheme, and details are sent to him only of those documents which sufficiently match his requirements.

A proposal on these lines was prepared and submitted to the Department of Scientific and Industrial Research, where it was warmly welcomed, in October 1964. As a result of this, Government support was obtained for the initial stages of the project: a Director of the S.D.I. Project was appointed and a five-month design study stage was carried out between October 1965 and February 1966. Proposals for the remainder of the project were then submitted to the Government but, while it considered the project well planned, technically feasible and of great potential, it also considered that a substantial proportion of the costs should be met from industrial sources.

However, despite the widespread interest in the S.D.I. Project, industrial support was slow in coming forward and, to safeguard the lead given by N.E.R.C. in this sector of the information field, it was decided that the best solution would be for the project to be taken over by the Institution of Electronic and Radio Engineers which, as a learned institution, would be eligible for 100 per cent financial support from Government sources.

Subsequent discussions between the I.E.R.E. and the I.E.E. have resulted in the S.D.I. Project being taken over by the latter, where it can be co-ordinated with the projected scheme for the computer production of *Science Abstracts*. The I.E.E. in its turn has invited the I.E.R.E. to be associated with it not only over the S.D.I. Project but also in connexion with *Science Abstracts*.

This seems a very sensible arrangement and one which should make a real contribution to dealing with the present 'information explosion' and in turn to the prosperity of the Country.

'Noise Margins' in Binary Electrological Systems

By E. L. Jones*, B.Sc.

The term 'noise margin' (or 'noise immunity'), as applied to a binary electrological system, has been defined in many different ways. Each definition involves a certain number of assumptions. Some of these have been recognized explicitly, but others have not. In many cases alternative assumptions would lead to basically different definitions of noise margin; but in other cases assumptions are introduced solely to simplify the problem, for ease of exposition or for ease of measurement, without affecting the basic considerations involved.

The article attempts to make explicit the assumptions involved, and to indicate which of them are basic and which are introduced solely for simplicity. Where alternative basic assumptions merit consideration, the reasons for preferring one rather than another are indicated.

Consideration is for the present limited to 'static' (very-low-frequency) noise margins, since it is felt that until these are thoroughly understood, and their definition and interpretation agreed, it is useless to try to consider 'dynamic' (transient) noise margins.

Graphical means are given for the determination and expression of the static noise margins of integrated logic units.

(Voir page 202 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 209)

Initial Assumptions of Convenience

THE following assumptions (which will be made throughout this article) do not in any way affect the basic principles of the definition of noise margin. The modifications necessary to accommodate relaxations of these 'assumptions of convenience' are in most cases fundamentally simple and straightforward, although they may be very difficult to implement in practice.

For the sake of definiteness, so as to avoid excessive abstractions and circumlocutions in the descriptions, *assume* that the system under construction is one in which the significant variable is (in the steady state) a steady positive voltage; let the more-positive state be H and the less-positive state be L, and let the inputs and outputs of all units concerned be nominally compatible.

For simplicity, *assume* that the system is nominally in a steady logical state, and that any disturbances change so slowly that effects dependent on their rate of change may be neglected.

For brevity, *assume* that an adequate set of limiting conditions has been defined, covering supply voltages, ambient temperature, input and output loadings, and so on.

For simplicity, *assume* that only one input terminal and one output terminal of each unit need be considered at any one time, all other terminals being maintained under defined conditions throughout.

For simplicity, *assume* that any disturbance to which the system is subjected may be adequately represented by a disturbing voltage introduced in series between the output of one unit and the input of the next unit in the logic chain.

For simplicity, *assume* that the output and input impedances of the units are such that there is negligible change in the actual output terminal voltage of a unit when a disturbing voltage is introduced between that terminal and the input terminal of the following unit.

To avoid unprofitable discussion of hypothetical cases of little practical importance, *assume* that the shapes of the various characteristics within the operating region are such that the only disturbances that need be considered are those that tend to shift the level of the H state negatively, or that of the L state positively; the opposite disturbances in either case having no significant effects.

Basic Assumptions

These assumptions are fundamental to the arguments that follow. Any change in them reflects a change in the basic approach to the subject.

Worst-Case Units

In any but the smallest digital systems, it is highly probable that some of the units will be found to have characteristics close to the extremes of the production spread, particularly if the units are not all from the same batch, and that some of the interconnexions will occur between extremely-unfavourable combinations of units.

Assume therefore that the noise margin must be so defined that it applies to any possible combination of unit characteristics within the overall production spread for the type or types involved. (The overall production spread may involve products from more than one manufacturer.)

Worst-Case Conditions

Different parts of a system may be at different ambient temperatures, and may be operated from separate power supplies; and individual units will have varying fan-in and fan-out, from unity up to the maximum permitted (or an equivalent in external loading). Some interconnexions will therefore represent extremely-unfavourable combinations of operating conditions.

Assume therefore that the noise margin must be so defined that it applies to any possible combination of operating conditions within the permitted ranges.

(In some cases the differences between the ambient temperatures or between the supply voltages of two interconnected units may be limited to something less than their overall permitted ranges. Such cases may be dealt with by dividing the overall ranges into appropriate segments and treating each of these separately.)

Single or Simultaneous Disturbances

There are two limiting cases to be considered with respect to possible definitions of noise margin. In one

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case only one single adverse disturbance in one single interconnexion is assumed to occur, all other interconnexions remaining undisturbed. In the other case an adverse disturbance is assumed to occur in every interconnexion, the disturbances in all L-state interconnexions being equal, and those in all H-state interconnexions being also equal (but not necessarily the same in magnitude as those in the L-state interconnexions).

The single-disturbance assumption appears to be over-optimistic, because very few sources of disturbance (internal or external to the system) will affect one interconnexion only.

The simultaneous-disturbance assumption is somewhat pessimistic, but probably not unduly so in general, because many sources of disturbances will undoubtedly affect a large number of interconnexions simultaneously.

Assume therefore that the noise margin must be so defined that it applies to the simultaneous-disturbance case.

PESSIMISM

Assume that the noise margin has to take into account the requirements of the three preceding sections simultaneously. Then the 'degrees of improbability' involved are:

(1) Each interconnexion joins a unit with a worst-case output characteristic to a unit with an adversely worst-case input characteristic.

(2) Each of these units independently is operating under the most adverse conditions permissible.

(3) This applies to every single interconnexion along the whole length of any logic chain.

(4) Every such interconnexion is subjected simultaneously to a maximum adverse disturbance.

This may be regarded as unduly pessimistic. Points that should be carefully considered, however, are:

(1) Disturbances (particularly those from outside the equipment) do tend to affect the interconnexions in groups rather than one at a time.

(2) The magnitudes of expected disturbances cannot be predicted with any great accuracy.

(3) The operating conditions to be encountered in different parts of the equipment throughout its life are not known exactly.

(4) The number of possible logic chains involved in an equipment is usually so very much higher than the number of units involved, that some chains at any rate are likely to come very close to being 'worst cases'.

(5) The problem has been simplified by considering only one input at a time to each unit.

(6) Changes may occur in the characteristics of the units during life.

All in all, it seems probable that the safety factor introduced by the pessimistic approach is just sufficient to cover the uncertainties in the estimation of the disturbances to be expected and the uncertainties introduced by the various simplifications of the problem.

Classification of Types of Unit

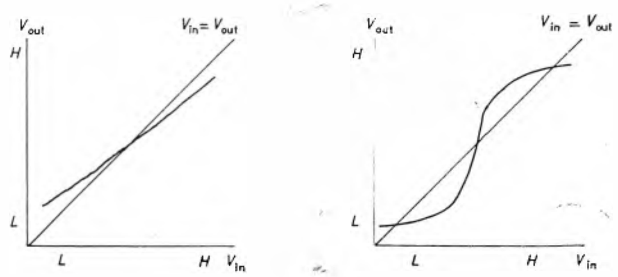
The basic units of a logic system may be classified into four groups; complex units such as shift-registers or counters may be regarded as built up from these basic units, and will not be considered further.

Figs. 1 to 4 show the basic transfer characteristics typical of the four groups of unit.

The four groups are formed by two binary classifications. The first is the classification into 'inverting' and 'non-inverting': these terms are used in their usual sense. The second is the classification into 'regenerating' and 'non-regenerating'. Broadly speaking, regenerating units have characteristics with a slope numerically greater than unity over the central part of the working range, and numerically less than unity at the ends of the working range; non-regenerating units have transfer characteristics with a slope numerically less than unity over the whole working range.

Transistorized invertors and buffers, involving amplification and limiting, provide examples of regenerating units, both inverting and non-inverting.

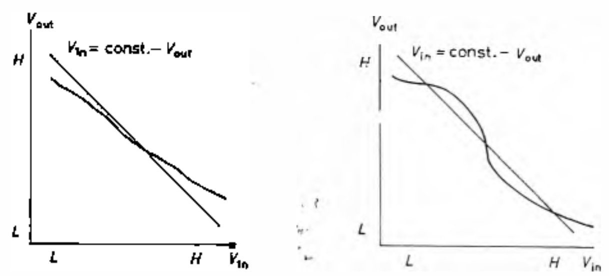
Diode gates, without amplification, provide examples of non-inverting non-regenerating units.



(1) *Non-inverting Non-regenerating*

(2) *Non-inverting Regenerating*

Figs. 1-4. Basic static voltage-transfer characteristics of the four classes of simple binary logic units



(3) *Inverting Non-regenerating*

(4) *Inverting Regenerating*

Examples of inverting non-regenerating units do not normally occur in practice in systems of the kind under discussion, and will not be considered further.

Regenerating units are responsible for the noise margins of any logic system; non-regenerating units can only decrease the available noise margins provided by the regenerating units.

Non-regenerating units cannot be used by themselves to form a logic system, and for present purposes they are most easily dealt with by considering them as sources of signal degradation inserted between the output of one regenerating unit and the input of another regenerating unit. In this way they can be treated in the same way as any other source of signal degradation or 'noise'. The next three sections therefore deal only with regenerating units, leaving the consideration of non-regenerating units to the last section.

Static Voltage Noise Margins of Non-inverting Regenerating Units

It is not normally possible to build an equipment of any complexity using non-inverting units only; but from the point of view of noise margins they are simpler to consider than inverting units, and will therefore be dealt with first.

The measured voltage transfer characteristic of any individual unit under given operating conditions may be plotted as in Fig. 2. If many units are measured, their characteristics may be superimposed on the same diagram. If these units form a sufficiently large and sufficiently representative sample, the envelope of all the characteristics thus drawn may be regarded as indicating the total production spread of the characteristics of the units, under the particular operating conditions chosen for the measurement. If the process is repeated for all significant sets of operating conditions, the overall envelope of all the characteristics so measured will indicate the total spread of characteristics for which the user must allow in his system design, as far as units of this one type are concerned. (The operating conditions to be varied include temperature, supplies, fan-in, fan-out, and so on.)

Suppose that such a 'grand spread' of characteristics has been produced, for a given type of unit, by a sufficiently-large number of measurements, and that it appears as in Fig. 5. (The smooth shape shown is convenient for illustrative purposes, although in practice there may be sharp corners on the envelope where one individual limit characteristic crosses over another.)

Logic Chains: the Noise-free Case

Now consider a long logic chain of such units. In the absence of all noise disturbances, the input to each unit is the same as the output from the preceding unit. Fig. 5 shows how an input H_1 to the first stage becomes, in the worst case, an output H_2, and how this output may be

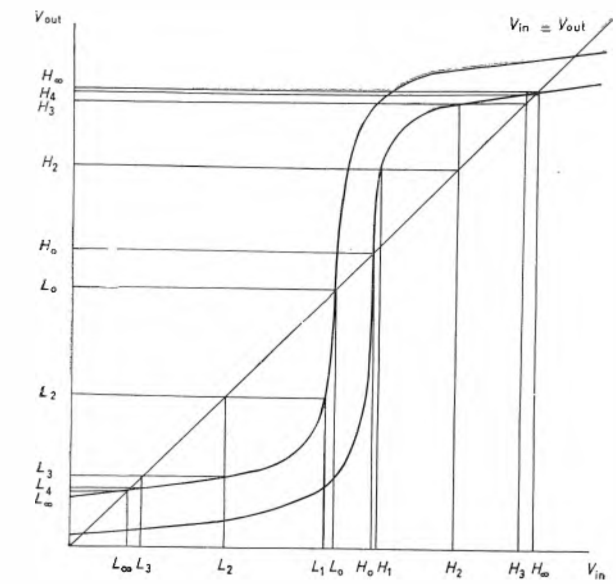


Fig. 5. Grand spread of characteristics for non-inverting regenerating units; worst-case signal-level relationships in the absence of noise disturbance

transferred graphically, by means of the line $V_{in} = V_{out}$, to become the input to the second stage. This stage has in the worst case an output H_3; and so on. (On the basis of the assumptions previously made, only the worst case need be considered here.)

Fig. 5 also shows two limiting values H_o and H_∞, defined by the intersections of the worst-case characteristic with the line $V_{in} = V_{out}$. Provided only that H_1 is greater than H_o, it is clear from the graphical construction that in the worst case all the subsequent values of H_n along the logic chain will lie between H_1 and H_∞. It follows that the logic chain can be used satisfactorily in any noise-free logic system wherein the defined minimum level H_h representing the H state lies anywhere between H_o and H_∞; because then if the input H_1 to the logic chain is greater than H_h, the output from any point of the logic chain will also be greater than H_h, even in the worst case.

Exactly similar considerations, with the signs of the inequalities reversed, apply to the L state; and the chain will be satisfactory in any noise-free system wherein the defined maximum level L_h representing the L state lies anywhere between L_o and L_∞.

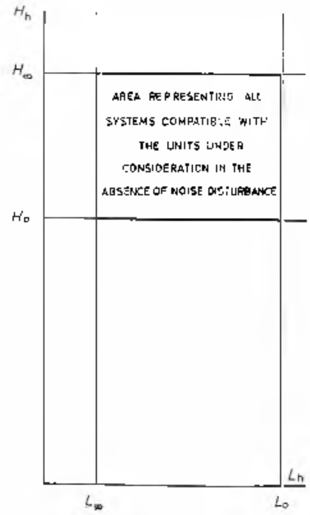


Fig. 6. Compatibility diagram for non-inverting regenerating units in the absence of noise disturbances

Representation: The Compatibility Diagram

It is convenient, for reasons that will become apparent later, to characterize any complete logic system (as distinct from the units of which it is made up) by its arbitrarily-defined values of H_h and L_h, and to represent the system graphically by the corresponding point in the (L_h, H_h) plane. In this plane, four straight lines may be drawn to represent H_o, H_∞, L_o, L_∞ for a given type of unit, as shown in Fig. 6. A chain of any number of logic units of the type under consideration will operate satisfactorily as a part of any noise-free logic system represented by a point within the rectangle bounded by the four lines (assuming of course that the current and impedance requirements are met). In such a case, the units are said to be 'compatible' with the system and the diagram may be referred to as the 'compatibility diagram' for that type of unit.

The Noisy Case

It is now necessary to consider the effects of noise disturbances. In accordance with the assumptions made previously, it is supposed that an adverse noise voltage N_H appears in series with every input connexion that is in the H state, and an adverse noise voltage N_L in series with every one in the L state. The simplest way of taking these into account is to consider them as inherent parts of the units in question, and to modify the transfer characteristics accordingly. The modification consists simply of a shift of the worst-case characteristic by an amount N_H or N_L, parallel to the axis of V_{in}, as shown in Fig. 7. When the compatibility diagram is constructed for these modified characteristics, the area of compatibility is smaller than it was in the noise-free case. This is shown in Fig. 8.

By repeating the process for different values of N_H and N_L, a complete set of noise-margin contours can be built up on the compatibility diagram. (There is an alternative

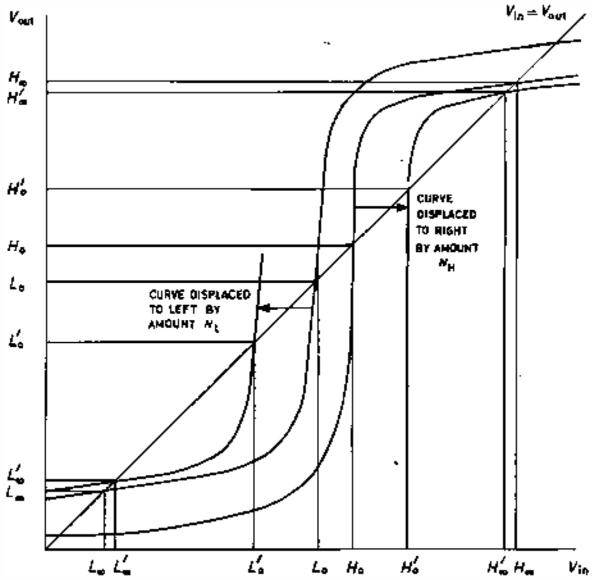


Fig. 7. Grand spread of characteristics for non-inverting regenerating units; modification (indicated by primed letters) to allow for noise disturbances N_H and N_L

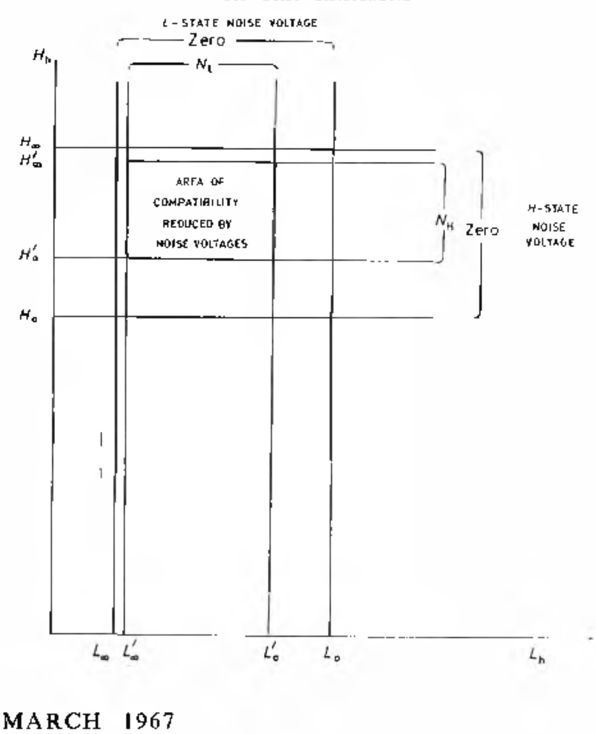
graphical construction that simplifies the work of plotting these, but it will not be discussed here.)

The complete compatibility contour diagram for non-inverting units appears as in Fig. 9.

Advantage of the Compatibility Diagram

The advantage of the compatibility diagram over other methods of presenting noise-margin information is that one single diagram can be drawn for any given type of unit under its limiting operating conditions, and that this one diagram can then be used to find the noise margins of any system into which the units may be incorporated. If such diagrams for several types of unit are superimposed, it is immediately obvious whether the units are compatible as far as signal voltage is concerned; and, if so, what noise margins can be achieved when they are used together.

Fig. 8. Compatibility diagram for non-inverting regenerating units, allowing for noise disturbances



Inverting Regenerating Units

In the foregoing discussion of non-inverting units, it was possible to discuss the noise margins for the two logic states completely independently. Their independence is shown in the compatibility diagram by the contours appearing as two sets of straight lines at right angles to one another. This is not the case when inverting units are considered; the two noise margins are not independent of one another, and the compatibility diagram shows two sets of curved contour lines. The basic procedure is, however, the same as for the non-inverting case, and the compatibility diagram has the same convenient properties.

The Noise-free Case

Fig. 10 shows the grand spread of characteristics for an

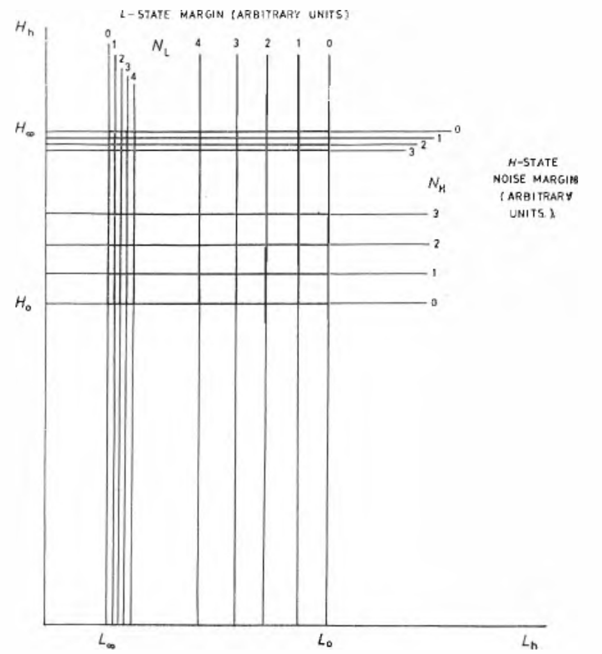


Fig 9. Compatibility contour diagram for non-inverting regenerating units

inverting unit. An input H_1 to the first stage of a chain of such units results in an output L_2, in the worst case. This becomes the input to the second stage and produces an output H_3, again in the worst case. As previously, the graphical construction can be repeated as often as necessary, to show the worst-case signal levels all along the logic chain. As with non-inverting units, the worst-case signal levels tend towards limiting values H_∞ and L_∞ in the later stages of the chain; but here these limiting values do not correspond to the intersections of the worst case characteristics with the line $V_{in} = V_{out}$. They correspond instead to the intersection of each worst-case characteristic with the mirror-image of the other, formed by reflection in the line $V_{in} = V_{out}$. This reflection is equivalent to an interchange of the axes of the graph. Because of the symmetry due to reflection, only one half of the diagram showing the intersections need be studied; and only the part of the reflection that is shown dotted in Fig. 10 need actually be constructed. The two intersections define the four limiting signal levels H_0, H_∞, L_0, L_∞, as shown. These limiting signal levels have exactly the same significance as in the non-inverting case, except that here they cannot be considered separately.

(It should be noted that the same 'reflection' technique may be applied to the non-inverting case. In that case all the intersections due to the reflection lie on the line $V_{in} =$

V_{out}, and the results are the same as were obtained previously.)

THE COMPATIBILITY DIAGRAM

The compatibility diagram for inverting units in a noise-free system shows an area of compatibility having exactly the same size, shape, and position as the area enclosed by the worst-case limiting characteristics (one 'real' and one 'reflected') of Fig. 10. This is shown in Fig. 11.

THE NOISY CASE

As before, the noise disturbance is represented by a voltage N_H or N_L in series with the input of every unit,

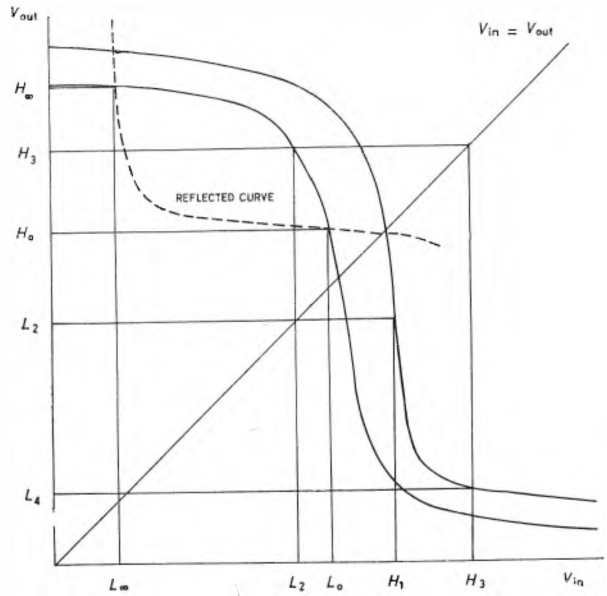


Fig. 10. Grand Spread of characteristics for inverting regenerating units, showing the construction (by the reflection in the line $V_{in} = V_{out}$) for finding H_0, H, L_0, L

and this results in a horizontal displacement of the worst-case characteristics; this becomes a vertical displacement when reflected. The contours in the compatibility diagram are therefore shifted bodily parallel to themselves; the upper curve is moved to the left by N_L, and the lower, because it results from a reflection, is moved upwards by N_H. The complete compatibility contour diagram therefore consists of two intersecting sets of evenly-spaced curves, as shown in Fig. 12.

Alternative Presentation

It is equally possible to produce corresponding contour plots where the axes represent limiting values of the input voltage rather than of the output voltage.

However, in the majority of practical cases the signal level is defined primarily by the output characteristics of a stage, rather than by the input characteristics of the next stage; and so for general use the system already described seems preferable to this alternative.

Non-regenerating Units (Non-inverting)

The introduction of a non-regenerating unit between two successive regenerating units in a logic chain has two distinct effects on the noise-resisting properties of the system. Firstly, the total available noise margin between the two regenerating units is reduced by the amount of the voltage difference (assumed adverse) between the input and output terminals of the non-regenerating unit. Secondly, what would otherwise be a single interconnexion is split into two separate successive interconnexions—an 'input' and an 'output' connexion to the non-regenerating unit—and the reduced noise margin must be shared between these two.

The proportions in which the noise margin is to be regarded as shared between these two interconnexions are not fixed by the characteristics of the units, but may be chosen arbitrarily. As far as the pure voltage noise margin is concerned, the optimum choice is to regard the available total noise margin as being shared equally between the two interconnexions. This is in line with the 'equal-disturbance' assumption used throughout the discussion of regenerating units. However, this 'equal-sharing' approach appears to be unrealistic, for two reasons.

Firstly, the input and output impedances of the non-regenerating units are likely to differ considerably from those of the associated regenerating units. In itself this difference of impedance has no effect on the voltage noise margin; but it is likely to have a marked effect on the relative probability of the two interconnexions picking up any given amplitude of interfering voltage in practice.

Secondly, the structure of a given logic system (existing or proposed) will often require that the non-regenerating units should always be physically very close to those regenerating units that immediately follow them in the logic chain, or (alternatively) close to those that immediately precede them; so that a particular one of the two connexions to the non-regenerating unit will always be very short, and will thus be exposed to considerably less interference than the other. (In the limit, of course, the non-regenerating unit and the adjacent immediately-following or immediately-preceding regenerating unit may be regarded as forming a single 'two-level-gating' regenerating unit, and as such may be treated adequately by the methods already described.)

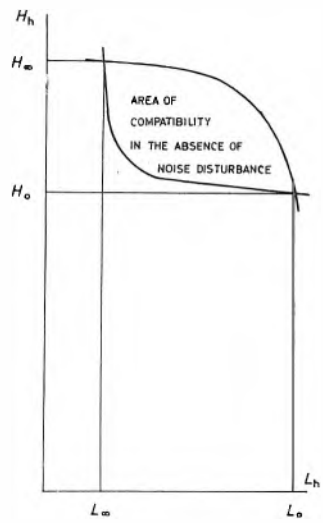


Fig. 11. Compatibility diagram for inverting regenerating units, in the absence of noise disturbance

In view of these considerations, therefore, the present general discussion will be restricted to the consideration of the total available noise margin between two successive regenerating units in the logic chain, irrespective of how this is to be apportioned between the input and output connexions of the intervening non-regenerating unit. A limiting value of this 'net noise margin' will be found that will guarantee safe operation of the system no matter how it is apportioned between the two interconnexions.

The extension of the discussion to the relatively-rare cases where two or more non-regenerating units may be introduced between two regenerating units is fairly straightforward.

ASSUMPTIONS OF CONVENIENCE

The following assumptions are made, in addition to the earlier ones, in order to provide a clear and specific basis for the ensuing discussion. They do not introduce any new basic ideas.

Assume that, in the system under consideration, regenerating units may be interconnected either directly or by way of just one non-regenerating unit.

Assume that the limiting signal levels for the system have been chosen, as previously described, so as to permit the direct interconnexion of regenerating units. Let the limiting signal levels at the output of a regenerating unit be H_h, L_h, and the corresponding worst-case input levels H_j, L_j, so that the noise margins associated with the direct interconnexion of regenerating units are $N_H = H_h - H_j$, $N_L = L_j - L_h$, as before.

Assume that the grand spread of the transfer characteristics of the non-regenerating units has been determined and is as shown in Fig. 13.

Basic Assumption

Assume that the quantity to be determined is the 'net noise margin', T_H or T_L, as defined above: this then represents a total noise voltage that cannot cause false operation of the system, however it may be divided between the input and output connexions of the non-regenerating unit.

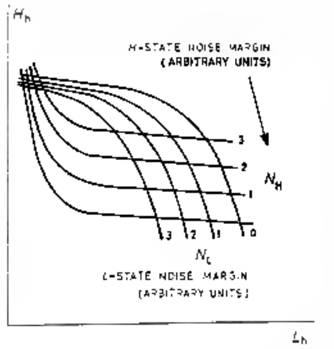


Fig. 12. Compatibility contour diagram for inverting regenerating units

Net Noise Margin

Consider first the L state.

The net noise margin T_L is less than the previously-defined 'regenerating-unit-only' noise margin N_L by the amount of the maximum worst-case adverse voltage difference between output and input terminals of the non-regenerating unit.

This voltage difference will of course vary somewhat with the voltage at the input terminal of the non-regenerating unit, and therefore with the noise voltage in the input connexion to that unit (the output terminal of the preceding regenerating unit remaining fixed at its worst-case level of L_h). What is required is the most adverse value that the voltage difference can reach within the operating region corresponding to the assumed worst-case conditions.

This operating region is bounded by the lines $V_{in} = L_h$ and $V_{out} = L_j$ shown in Fig. 13; and the required most adverse value of the voltage difference occurs at the point, on the worst-case characteristic, that is furthest from the line $V_{in} = V_{out}$, within the operating region. If a line is drawn parallel to the line $V_{in} = V_{out}$ and passing through this 'furthest' point, it will cut the axis of V_{out} in a point P_L representing the required value of the maximum worst-case adverse voltage difference between output and input terminals of the non-regenerating unit.

Then $T_L = N_L - P_L$.

Similar reasoning of course applies to the H state, the operating region corresponding to the assumed worst-case conditions being bounded by the lines $V_{in} = H_h$ and $V_{out} = H_j$ and the line through the 'furthest' point of the worst-case characteristic within that region cutting the axis of V_{in} in the point P_H; and $T_H = N_H - P_H$.

Simplification

In most practical cases the slope of the worst-case characteristic is everywhere less than unity within the worst-case operating region. (Fig. 13 shows this condition in the H state only.)

Where the slope is less than unity, the worst-case voltage difference always occurs where $V_{in} = L_h$ or H_h. This corresponds to the fact that, with a gain less than unity in the non-regenerating unit, noise disturbances occurring at the input of the unit are less harmful than similar disturbances occurring at its output.

Deliberate Offset

If the non-regenerating unit introduces a definite voltage offset between output and input, as happens with a simple uncompensated diode gate, it is possible for either P_L or P_H (but not both) to be negative. In this case the noise margin in the corresponding state is actually *increased* by the introduction of the non-regenerating unit; but that of the opposite state is always decreased by a somewhat greater amount.

As long as direct interconnexion of regenerating units is also permitted, however, this increase of one of the noise margins cannot be claimed for the system as a whole.

Conclusion

The foregoing discussion of static voltage noise margins has been primarily aimed at clearing up some of the confusion that surrounds the topic, in the hope that it will help towards eventual agreement as to what is being talked about when noise margins are mentioned.

It is obviously possible in principle to extend the compatibility contour diagram to the representation of the sensitivity of the system to low-frequency *current* noise, by dividing the voltage figure for each contour by the maximum circuit impedance occurring under the corresponding conditions; but the details of such a transformation will require very careful formulation if the results are to be of practical utility.

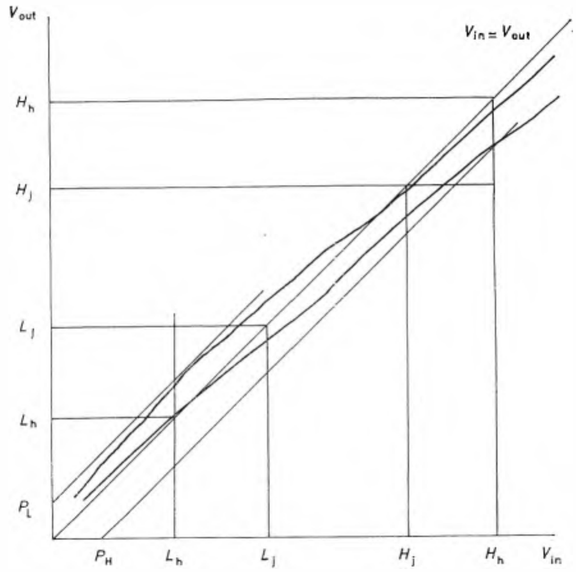


Fig. 13. Grand spread of characteristics for non-inverting non-regenerating units, showing construction for loss of noise margin when such units are introduced into a system

It is hoped that it will be possible later to extend some of the present concepts to the much more difficult problems associated with 'dynamic' or 'transient' noise margins.

Another possible direction of extension is towards the general questions of compatibility between separate electrological systems.

Acknowledgment

The author wishes to express his thanks for many helpful discussions with colleagues within S.T.C. Semiconductors Ltd and with various members of the BSI and VASCA integrated electronics committees.

A Current-Switching Phase-Sensitive Detector
and its Application to a Resolved Components Voltmeter

By J. G. Lacy*, B.E., C.Eng., M.I.E.E.

A phase-sensitive detector is described which uses transistors as current switches to obtain improved amplitude linearity and give a wide operating frequency range. The sinusoidal reference signal is converted to a square wave by a circuit using a tunnel diode as a hysteretic switch and employing a servo-loop to obtain unity mark-to-space ratio. A description is also given of a resolved components voltmeter using these detectors which displays simultaneously the in-phase and quadrature components of a sinusoidal a.c. signal with respect to a related reference signal. The instrument operates over the frequency range 20Hz to 150kHz.

(Voir page 202 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 209)

THE operation of an ideal p.s.d. (Phase-Sensitive Detector) or synchronous rectifier can be considered, mathematically, to be equivalent to that of multiplying the input waveform by a reference square wave of unity mark-to-space ratio. If the square wave is considered by Fourier analysis to contain a fundamental sinusoidal component plus a number of odd order harmonics, the output of the detector can be seen to contain a d.c. component which depends on the amplitudes and phase relationships of the various frequency components of the input waveform in the following way:

$$I_{DC}=K[v_1 \cos \theta+v_3 \cos 3\theta+v_5 \cos 5\theta+ + \text{- - - -} + v_{2n+1} \cos (2n+1)\theta]$$

Where K = constant of the detector

v_{2n+1} = peak amplitude of the component of the input signal of the same frequency as the $(2n+1)^{\text{th}}$ harmonic component of the square wave

$(2n+1)\theta$ = angular displacement of the various components of the input signal relative to the square wave.

The output will also contain a.c. components proportional to the amplitudes of other frequency components of the input signal, including those that are even-order harmonics of the fundamental frequency component of the square wave.

If the detector is used to drive a moving-coil instrument the a.c. components of the output signal will produce no deflexion and may be neglected. When the input waveform is a sine wave of the same frequency as the square wave the output reduces to $I_{DC} = Kv \cos \theta$.

If a second detector is fed with this same input signal but has its reference waveform phase shifted through 90° its output will be $I_{DC} = Kv \sin \theta$. An instrument using two phase-sensitive detectors in this manner could therefore be used to display simultaneously the in-phase and quadrature components of a sinusoidal a.c. signal with respect to a related reference waveform.

Choice of P.S.D.

When one comes to consider the use of p.s.d's described in the literature, in relation to this application, one finds that their performance falls short of ideal in some of the following respects:

(a) Amplitude linearity.

(b) Frequency response.

(c) Stability of gain and zero.

(d) Rejection of frequencies other than the reference and its odd-order harmonics.

A non-linear relationship between the d.c. output and the peak amplitude of the input signal can result from the switching action being dependent on the amplitudes of the input and reference signals, and the non-linear resistance of the switches when conducting at low levels.

If the switching is done on the output of a constant-current, rather than a constant-voltage source, then the non-linear resistance of a practical switch may be neglected, as the voltage developed across the load will be proportional to the current and independent of any included non-linear resistance.

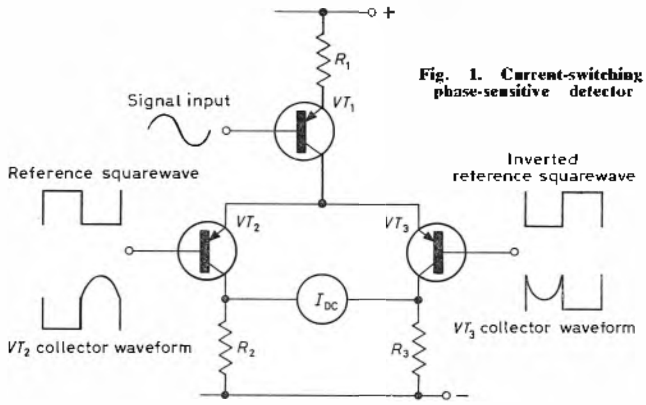


Fig. 1. Current-switching phase-sensitive detector

This 'current-switching' principle has been used by others[1,2,3] in detectors employing valves. When transistors are used as the switching elements, operation in this current-switching mode eliminates the effect of the non-linearity of the switch and in addition, since the transistors are unsaturated, the switching speed is a maximum for a given cut-off frequency.

Circuit

The circuit of a full-wave current-switching p.s.d. is shown in Fig. 1. VT_1 operates as a constant-current source presenting a high output impedance to the emitters of the grounded-base switches VT_3 and VT_3. The bases of these transistors are driven by unity mark-to-space ratio anti-phase square waves of fixed amplitude and d.c. level, derived from the reference signal. If no input signal is applied to the base of VT_1 its constant collector current is switched alternately into R_3 and R_3 at reference frequency, producing anti-phase square waves at the collectors. The magnitude of this current is chosen so that VT_2 and VT_3 can never bottom. The deflexion of the moving-coil meter connected between the collectors of VT_2 and VT_3 will be proportional to the average value of the instantaneous difference of the collector currents. In the absence of an input signal this deflexion should of

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course, be zero assuming that VT_2 and VT_3 have been matched for current gain and leakage current and that $R_2 = R_3$. In practice a small preset potentiometer is added in series with R_2 and R_3 to balance out any residual inequalities in R_2 and R_3 and the collector leakage currents. It has not been found necessary to readjust this in use. The use of silicon transistors reduces any drift of zero with temperature to negligible proportions.

If a sinusoidal input signal at reference frequency is applied to the base of VT_1 the collector load waveforms will be as shown in Fig. 1. The meter deflexion may be shown to be proportional to $v \cos \theta$ where v is the peak value of the applied signal and θ the phase difference between the applied signal and the reference. Gain stability and a linear relationship between the collector current of VT_1 and the input signal is assured by the negative current feedback provided by the un-bypassed emitter resistor R_1. The considerable mismatch between the output impedance of VT_1 and the non-linear input impedance of either VT_2 or VT_3 completely swamps the effect of the latter on the input current to the switches. Measurements made by the author on the circuit showed no deviation from linearity greater than 1 per cent which was the limit of accuracy of the test equipment used. The true distortion figure is probably at least one order better.

The lower frequency limit of operation is set by the signal-input coupling capacitor and the time-constant of the indicating instrument used. In applications where the response time of the detector is unimportant the lower limit may be reduced by shunting either the meter or the collector loads with large capacitors. The high-frequency response is limited effectively by the switching speed of the transistors. Silicon planar epitaxial transistors with an F_t of 300MHz were used in the prototype with the idea of eventually raising the upper limit of operation above the design figure of 150kHz. No extensive tests have been made above this frequency but a quick check showed satisfactory operation at 1MHz.

Effect of Mark-to-Space Ratio

In order to realize the theoretical performance of the p.s.d. it is essential that the mark-to-space ratio of the reference square wave be unity and that the rise and fall times be a very small fraction of the cycle time. Any deviation from unity mark-to-space ratio will introduce a zero error proportional to ϕ (where $\pi \pm \phi$ is the mark and $\pi \mp \phi$ is the space of the square wave). With a non-unity mark-to-space ratio reference square wave the output for a sinusoidal input at reference frequency will be of the form $I_{DC} = \pm K' \phi + (K/2)\, v\, [\cos \theta - \cos (\pi \pm \phi - \theta)]$ i.e. for a constant amplitude input of variable phase with respect to the reference there is no longer a simple cosine relationship between output and phase angle. In addition there will no longer be complete suppression of any even-order harmonics present in the input signal.

Irrespective of the circuit configuration the switch or switches must operate at unity mark-to-space ratio to give correct operation. This point has been mentioned by Wright[4] who gives a circuit to produce a square wave from a sine wave reference signal.

For the proposed application the p.s.d. was required to operate over the frequency range 20Hz to 150kHz without adjustment. Considerable experimental work on various standard squaring circuits, e.g. overdriven amplifiers, Schmitt triggers, clippers etc., showed the following difficulties:

(a) The mark-to-space ratio of the square wave in many cases was dependent on the input signal amplitude.

(b) To set the mark-to-space ratio to unity usually involved the critical adjustment of some component values.

(c) Unity mark-to-space ratio was not preserved over the required frequency range.

(d) The mark-to-space ratio was temperature-dependent.

(e) The output square wave was not in phase with the input sine wave.

(f) Most circuits were critically dependent on power supply stability.

The circuit developed by the author used a tunnel-diode biased by a constant current as a hysteretic switch. If a sinusoidal current biased to the centre of the hysteresis range is applied to such a switch the output will be a square-wave voltage of unity mark-to-space ratio. The phase angle between input and output will be $\sin^{-1} (h/2I_p)$, where h is the difference between the input currents corresponding to the two states of the switch and I_p the peak value of the input current. This can be made negligible by ensuring that the hysteresis range or window is kept small compared with the peak signal amplitude.

Complete Circuit

The complete squaring circuit and associated p.s.d. is shown in Fig. 2. The operation of the circuit may be explained by initially assuming points A and B to be at a common fixed positive potential with respect to earth. The current in the tail resistor R_4 defined by this potential divides between VT_6 and VT_7. The constant collector current of VT_7 may be represented as a horizontal load line on the tunnel-diode characteristic Fig. 3. Two stable states a and d are defined by the intersections of this load line and the diode characteristic. If the diode current is caused to increase above point b or to decrease below point e by applying an alternating signal to the base of VT_7 the diode will switch rapidly from one state to the other along the locus $abcdefa$. The mark-to-space ratio of the voltage square-wave produced across the diode will be unity if the load line is set equidistant between the diode peak and valley currents. This waveform is amplified by the 'long-tailed-pair' current-switching amplifier VT_4 and VT_5, the collector voltage excursions being limited by the catching diodes to the potentials defined by the Zener diodes MRZ_1 and MRZ_2.

The two anti-phase collector square waves drive the bases of the transistor switches VT_2 and VT_3 of the p.s.d. As the maximum and minimum potentials of the driving square waves are fixed the average or d.c. values of the waveforms depend only on their mark-to-space ratio. The average values of the two waveforms are abstracted by the two RC filters R_5, R_6, C_1, C_2 and R_7, R_8, C_3, C_4 and compared through difference by VT_6 and VT_7. The effect of a difference is to change the value of the tunnel-diode current in the correct direction to reduce the difference to zero, which corresponds to unity mark-to-space ratio square waves at the collectors of VT_4 and VT_5. With the circuit values shown the mark-to-space ratio of the output square wave is held without adjustment to within 0·2 per cent of unity over the frequency range 20Hz to 150kHz when supplied with a sinusoidal input of 4 ± 2 V peak-to-peak. A 10 per cent change in the supply voltages produces less than 0·1 per cent change in the mark-to-space ratio. The input signal is pre-clipped by a two-stage diode clipper and applied to the base of VT_7 through the emitter-follower VT_8. A small capacitor is paralleled across R_9 to compensate for phase lag caused by hole storage in the diodes used. Many audio-signal generators have been found to have a small d.c. component in their output due to the use of an electrolytic capacitor to couple the anode of the output valve to the output attenuator. As the presence of this d.c. component would forward-bias two of the input clipping diodes and result in asymmetric clipping of the input waveform the output of a variable current source is added in at the junction of R_9 and R_{10} to cancel this effect.

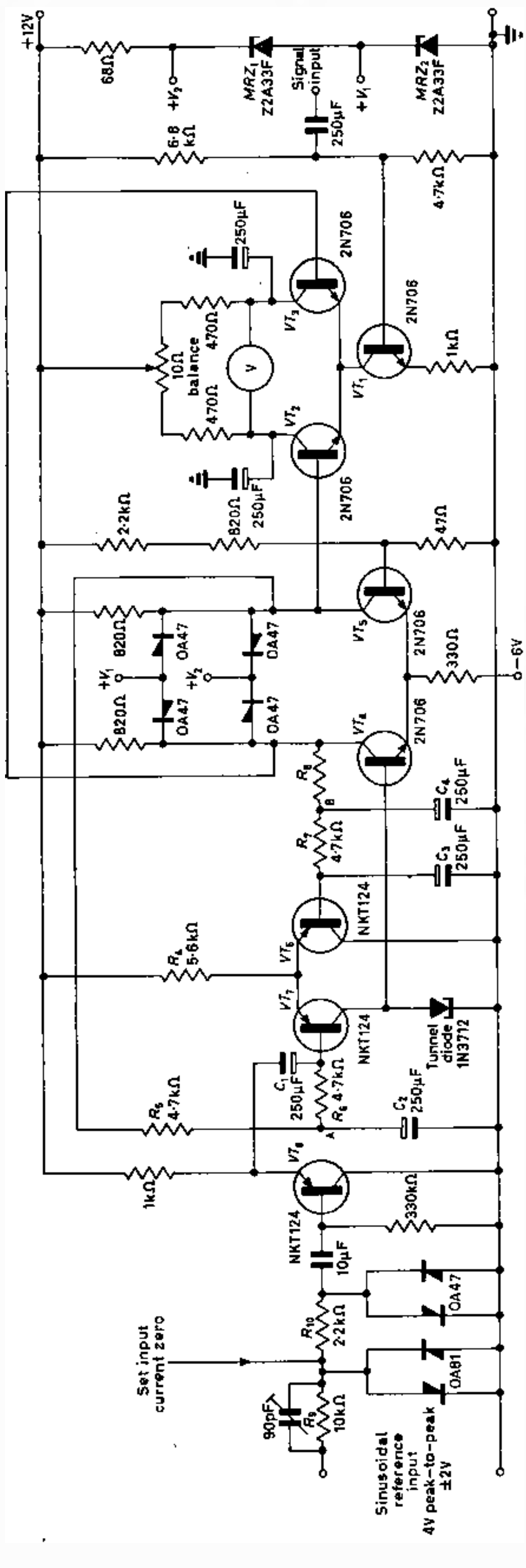

Fig. 2. The complete circuit

Resolved Components Voltmeter

The arrangement of the instrument is shown in Fig. 4. The input is applied through a compensated attenuator giving ranges of 0·5, 5 and 50V f.s.d. A field-effect transistor operating as a source follower is interposed between the attenuator and the paralleled inputs of the reference-phase and quadrature-phase p.s.d.'s, to give a high input

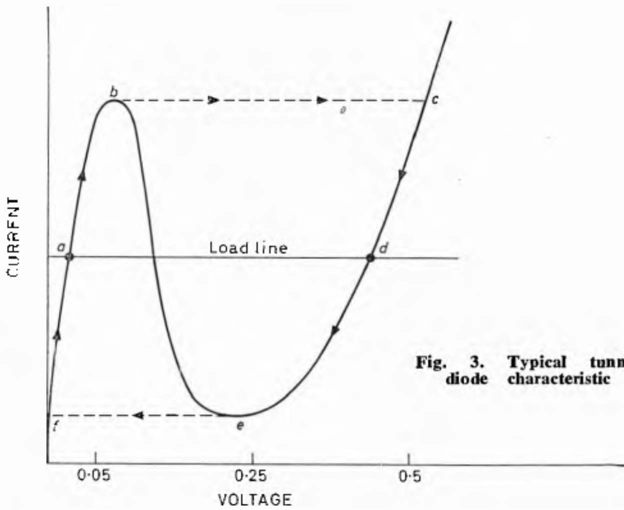

Fig. 3. Typical tunnel diode characteristic

impedance on all ranges. Each p.s.d. drives a centre-zero moving-coil meter giving direct reading of the cartesian co-ordinates of the tip of the phasor representing the applied voltage. The deflexion sense of the meters gives direct identification of the phasor quadrant, the reference phasor being taken to be positive along the x axis.

In order to render the use of the instrument independent of special-purpose signal generators providing a two-phase output, circuits are incorporated to derive a quadrature phase from a single-phase reference input. The reference signal is applied directly to the squaring circuit—which drives the reference-phase p.s.d. It is applied also to a voltage-controlled phase-shifter incorporating a Luxistor, the phase-shifted output being applied to the squaring circuit driving the quadrature phase p.s.d. This squaring circuit also drives a further p.s.d. that has as its signal input a fraction of the original sinusoidal reference signal. If the output of the squaring circuit and the reference signal are not in quadrature an error signal appears at the output of the p.s.d. This error signal, after filtering, is amplified by a d.c. amplifier and applied as a control signal to the phase shift circuit, its polarity being such as to tend to reduce the error signal to zero. The phase shift circuit is shown in Fig. 5. The sinusoidal reference signal is applied to VT_9 which operates as a phase splitter, the anti-phase signals at its collector and emitter being applied through the compound emitter-followers VT_{10}, V_{11}, VT_{12} and VT_{13} to the phase-shift network consisting of the switched capacitors and the resistive element of the Mullard ORP39 Luxistor. The Luxistor consists of a cadmium sulphide cell and a small tungsten filament lamp encapsulated in an epoxy resin block. The resistance of the CdS cell may be varied over four decades by controlling the lamp voltage from 0 to 12V, there being complete electrical isolation between the control voltage and the variable resistance. This resistance change would have been sufficient to hold the phase shift constant at

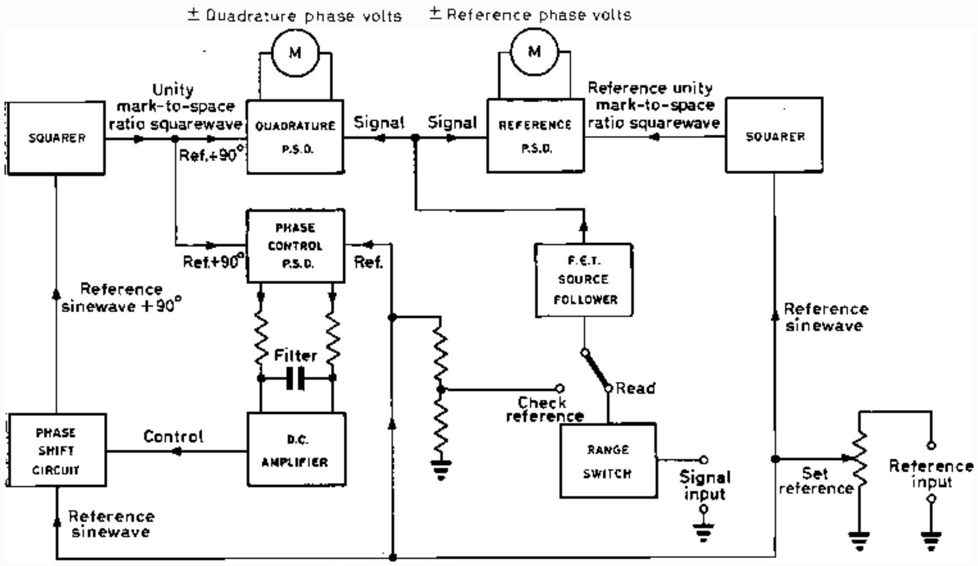

Fig. 4. Resolved components voltmeter

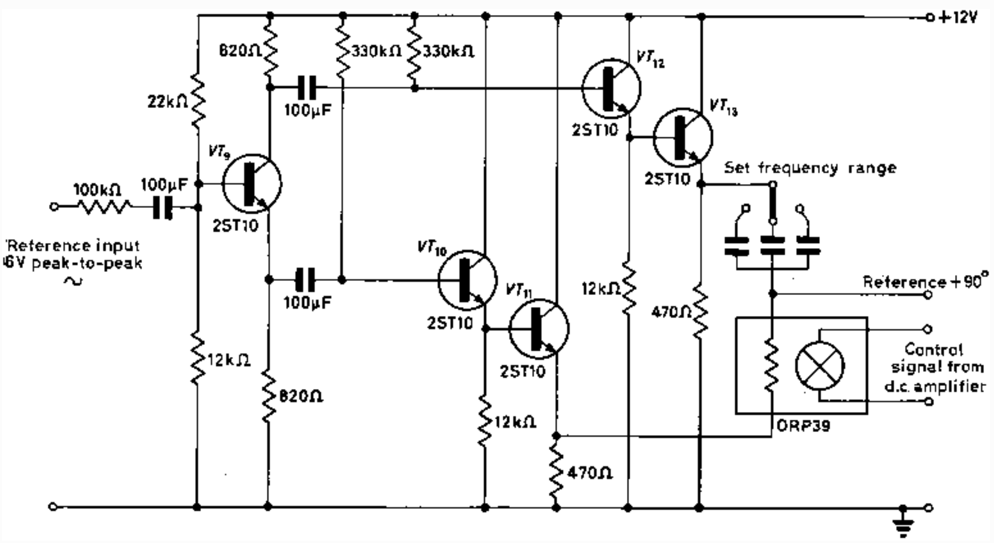

Fig. 5. Phase-shift circuit

90° over the design frequency range without changing the capacitor value but due to the greatly increased lag in the response at lamp voltages less than 4·5V it was found necessary to split the overall frequency range into three bands. The effect of the increased lag was to render the phase-control servo unstable and compensation for this lag resulted in the servo having a very sluggish response. The three bands are selected by switching in three different capacitors by a panel mounted switch. A 'press to test' switch on the front panel connects the signal input of the p.s.d.'s to a fraction of the reference voltage which gives a standard deflexion on the reference-phase meter for a nominal reference input of 4V peak to peak. The reference amplitude is adjusted to give this reading and at the same time it is noted that the quadrature-phase meter reads zero. This gives an indication that the correct phase-shift capacitor for the particular frequency is in circuit and that the instrument is functioning correctly. The phase-control servo holds the maximum error between the reference and quadrature phases to less than 1°. The overall accuracy of the instrument when measuring a sinusoidal voltage is ±3 per cent. Normally the instrument is used in the construction of polar plots of the response of feedback amplifiers or networks. In this application unless the device under test is very non-linear the error introduced by the odd harmonics generated will be negligible. In other applications, since the second and all higher even-order harmonics are rejected by the instrument, a simple tunable filter used in front of the instrument will be sufficient to eliminate any third or higher odd-order harmonics present in the input.

Conclusion

A phase-sensitive detector has been described suitable for use in a measuring instrument, having stable gain and zero, excellent amplitude linearity and a wide frequency response.

A description is also given of the design of a resolved components voltmeter using the circuit. The upper frequency limit of operation at present is 150kHz but it should be possible to extend it to enable measurements to be made on networks at video frequencies.

Acknowledgments

The author wishes to extend his thanks to Professor J. J. Morrissey, Head of the Electrical Engineering Department, at University College Dublin, and his colleagues, for their continued interest in this work, and to Mr. K. A. Comerford, B.E., who assisted in the development of the instrument.

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Decade Boxes with Binary Output

By D. P. Franklin*

Circuits are derived for decimal to binary conversion in which the requisite combinational logic is performed entirely passively. The AND *function is fulfilled through the series connexion of contacts of multi-pole ten position switches, while the* OR *function is through the employment of parallel paths, often via diodes.*

Two modes of operation are described in which the inputs to the logical units either have or have not first been put in binary form.

(Voir page 203 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 210)

WHEN numerical information has to be set into electronic apparatus containing binary logic, ergonomic considerations often favour the use of decimal numbers and ten-position rotary switches rather than binary numbers and a bank of simple on-off switches. The conversion of the decimal to binary form wherein relays perform the necessary logic functions has been described previously[1,2,3].

Towards the end of his note[3], de Visme asserts that, in principle, it is possible to construct a code convertor using only inter-connexion of the contacts of 10-position wafers (together with diodes to provide isolation where necessary), but that the complication is likely to be formidable. In the following, two forms of such a circuit, 'binary' and 'decimal', will be realized for the two and three decade cases so that their complexity will no longer be a matter of conjecture.

Principles

A primitive example is given in Fig. 1 of the familiar 'two-way switch' lighting circuit; in algebraic terms, a voltage will appear at the output terminal of B when:

$$A + B \pmod 2 = 1$$

This requires the logical condition $A\bar{B} + \bar{A}B$ (i.e. A on and B off OR A off and B on). It exemplifies the principle of first furnishing alternative functions (in this simple case, A or $\bar{A}$) and then selecting the appropriate one according to the state of the other switch (here simply one of two states; B or $\bar{B}$). It will be seen that there are two possible paths, $A\bar{B}$ and $\bar{A}B$. The conjunctive (AND) role is thus filled by the provision of contacts in series while disjunction (OR; +) is provided through parallel alternative paths.

In progressing to the general problem, one has the situation shown in Fig. 2, in which each stage receives numerical information, N, from the switch positions, and carry information, C, from the preceding stage. In response to these inputs, each stage gives the sum (modulo 2) to the output and a carry to the following higher stage. To achieve the objective, it is necessary for the carry input, C, at any one stage to be in electrical form and to be operated upon by the numerical information, N, through switch connexions only, so as to yield the S and C outputs.

Now, although the outputs, S_1, etc., must be in the required binary form, it is not necessary for the inputs, N_1 etc., to be expressed in the binary scale. In the circuit described by Nairn[1], however, these inputs were in fact binary, each component consisting of the appropriate digit in the binary expression of each decimal component set in by its particular switch. Each stage there performed in the scale of two.

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An alternative arrangement is one in which N_1, N_2, etc., are decimal (1 out of 10) inputs of which each output S_1, etc., is a binary function. Clearly, in this extreme case, N_1, N_2, etc., will all be identical, consisting at each stage of the units, tens and hundreds inputs. Further, the carries, C_1, etc., will all be zero, each stage being energized direct from the supply.

AND functions will, as in the example of Fig. 1, be furnished by series connexions of contacts in order to

Fig. 1. (right) Simple example of switch combination

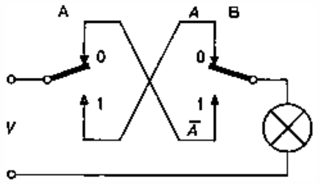
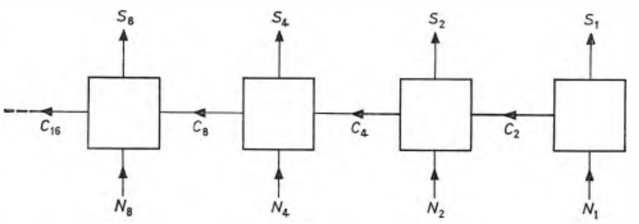


Fig. 2. (below) Block diagram of general convertor



obviate the use of alternate AND and OR logic gates or active NOR elements. For the same reason, any carries are conveyed in a 1 out of n code, with the value of n as necessary; if desired, each one of the n possible channels may be regarded as carrying a binary function derived in the same way as are S_1, etc.

Binary Operation

First, each input is expressed in the binary scale, and the results then summed.

TWO SWITCH CODE CONVERTOR (BINARY)

Logical Relationships

The circuit of Fig. 3 will be seen to be divided into seven stages, each corresponding to one of the blocks of Fig. 2, carries between stages being communicated in each case by alternative paths, X and Y. Here, the symbol X denotes 'no carry' and Y denotes 'carry 1'. Each stage receives binary inputs through T (tens) and U (units) switches whose state conforms to the value in the corresponding column in the appropriate binary expressions. The binary equivalents of the tens and units inputs, taken separately, are given in Tables 1 and 2 of the Nairn article[1].

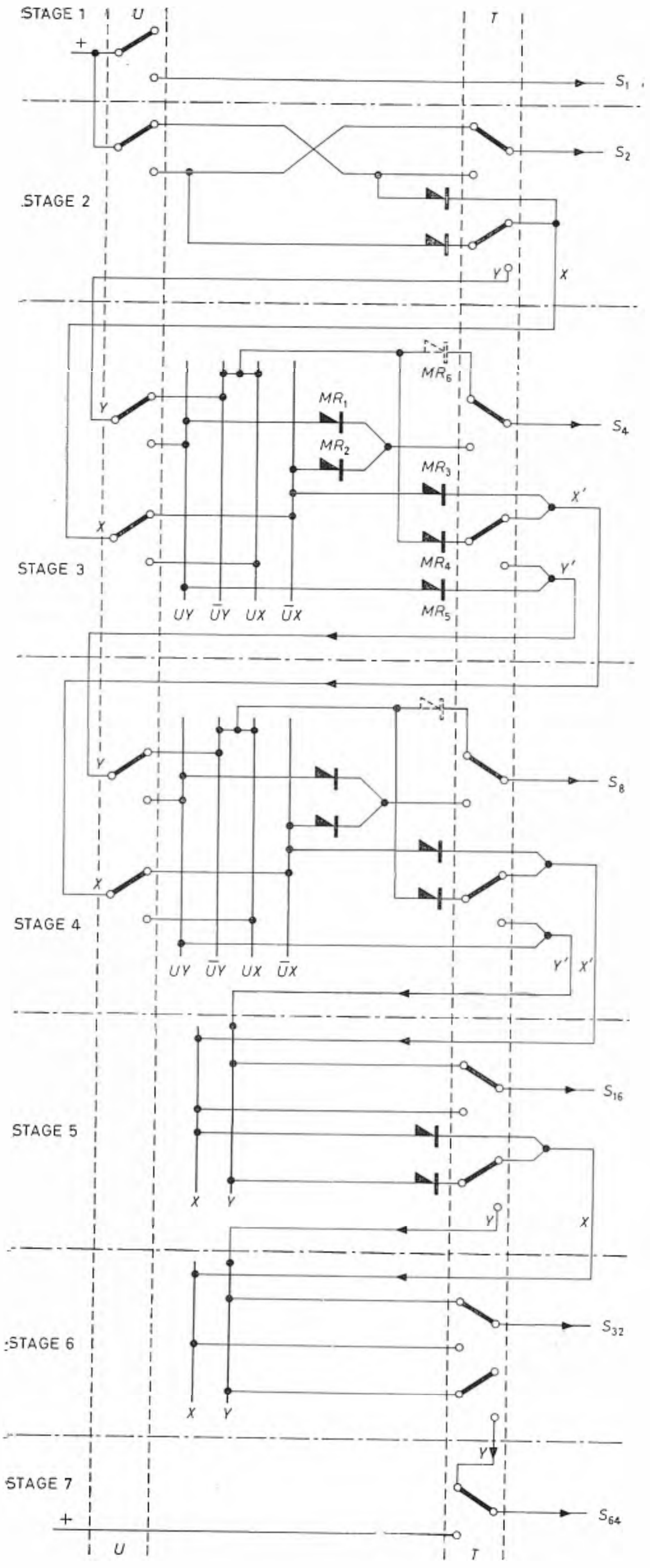


Fig. 3. Circuit for two-switch convertor (binary)

The third stage may be taken as illustrating the general case, and the possible combinations are listed in Table 1(a). At the intersection of any one row and any one column is given the corresponding total input. From this it is easy to construct logic maps for the outputs S, X' and Y'. For S, odd entries in Table 1(a) call for a 1 in

TABLE 1

(a) Total input

	$\bar{U}X$	UX	$\bar{U}Y$	UY
$\bar{T}$	0	1	1	2
T	1	2	2	3

(b) Sum (mod 2)

	$\bar{U}X$	UX	$\bar{U}Y$	UY
$\bar{T}$	0	1	1	0
T	1	0	0	1

(c) No carry, X'

	$\bar{U}X$	UX	$\bar{U}Y$	UY
$\bar{T}$	1	1	1	0
T	1	0	0	0

(d) Carry 1, Y'

	$\bar{U}X$	UX	$\bar{U}Y$	UY
$\bar{T}$	0	0	0	1
T	0	1	1	1

TABLE 2

(a) Total input

	$\bar{H}X$	HX	$\bar{H}Y$	HY	$\bar{H}Z$	HZ
$\bar{U}\bar{T}$	0	1	1	2	2	3
$U\bar{T}$	1	2	2	3	3	4
$\bar{U}T$	1	2	2	3	3	4
UT	2	3	3	4	4	5

(b) Sum (mod 2)

	$\bar{H}X$	HX	$\bar{H}Y$	HY	$\bar{H}Z$	HZ
$\bar{U}\bar{T}$	0	1	1	0	0	1
$U\bar{T}$	1	0	0	1	1	0
$\bar{U}T$	1	0	0	1	1	0
UT	0	1	1	0	0	1

(c) No carry, X'

	$\bar{H}X$	HX	$\bar{H}Y$	HY	$\bar{H}Z$	HZ
$\bar{U}\bar{T}$	1	1	1	0	0	0
$U\bar{T}$	1	0	0	0	0	0
$\bar{U}T$	1	0	0	0	0	0
UT	0	0	0	0	0	0

(d) Carry 1, Y'

	$\bar{H}X$	HX	$\bar{H}Y$	HY	$\bar{H}Z$	HZ
$\bar{U}\bar{T}$	0	0	0	1	1	1
$U\bar{T}$	0	1	1	1	1	0
$\bar{U}T$	0	1	1	1	1	0
UT	1	1	1	0	0	0

(e) Carry 2, Z'

	$\bar{H}X$	HX	$\bar{H}Y$	HY	$\bar{H}Z$	HZ
$\bar{U}\bar{T}$	0	0	0	0	0	0
$U\bar{T}$	0	0	0	0	0	1
$\bar{U}T$	0	0	0	0	0	1
UT	0	0	0	1	1	1

Table 1(b) and even entries a 0. For X', entries in 2(c) are 0 when 1(a) shows 2 or 3, and otherwise 1, while, for Y', Table 1(d) is the same as 1(c) with 0 and 1 interchanged.

Switch Connexions

According to which one of the four column headings agrees with the inputs, a voltage can be made to appear on one of four bus bars by connecting two changeover contacts operated by U, one to the X input wire and the other to the Y. Furthermore, on examining Tables 1(b), (c) and (d), it will be seen that UX and $\overline{U}Y$ always appear together in any one row. Thus, it is permissible for these two bars to be connected permanently. The $\overline{U}X + UY$ function of the lower row of Table 1(b) is derived via two diodes MR_1, MR_2 and this or the $UX + \overline{U}Y$ may be selected according to the appropriate state of T by means of a changeover contact to give the S output of the stage.

Turning to the X' function, from Table 1(c) it will be seen that the truth of $UX + \overline{U}\overline{Y}$ should generate an output when T is off and that $\overline{U}X$ should also do so whatever the state of T. This combination is provided via another T contact and diodes MR_3, MR_4. For Y', $UX + \overline{U}\overline{Y}$ is taken via the complementary T contact and combined with UY in diodes MR_4, MR_5. If desired, an extra diode, MR_6, may be introduced as shown dotted in order that each possible path from an input to an output includes one and only one series diode.

In the fifth and sixth stages, the U input is identically zero, so only those circuits carrying $\overline{U}$ functions need be used. MR_1 and MR_3 can therefore be eliminated, whereupon MR_2 also becomes redundant. Stage 2 receives no carry, so the circuit can be used, with $\overline{U}$ and U in place of X and Y respectively.

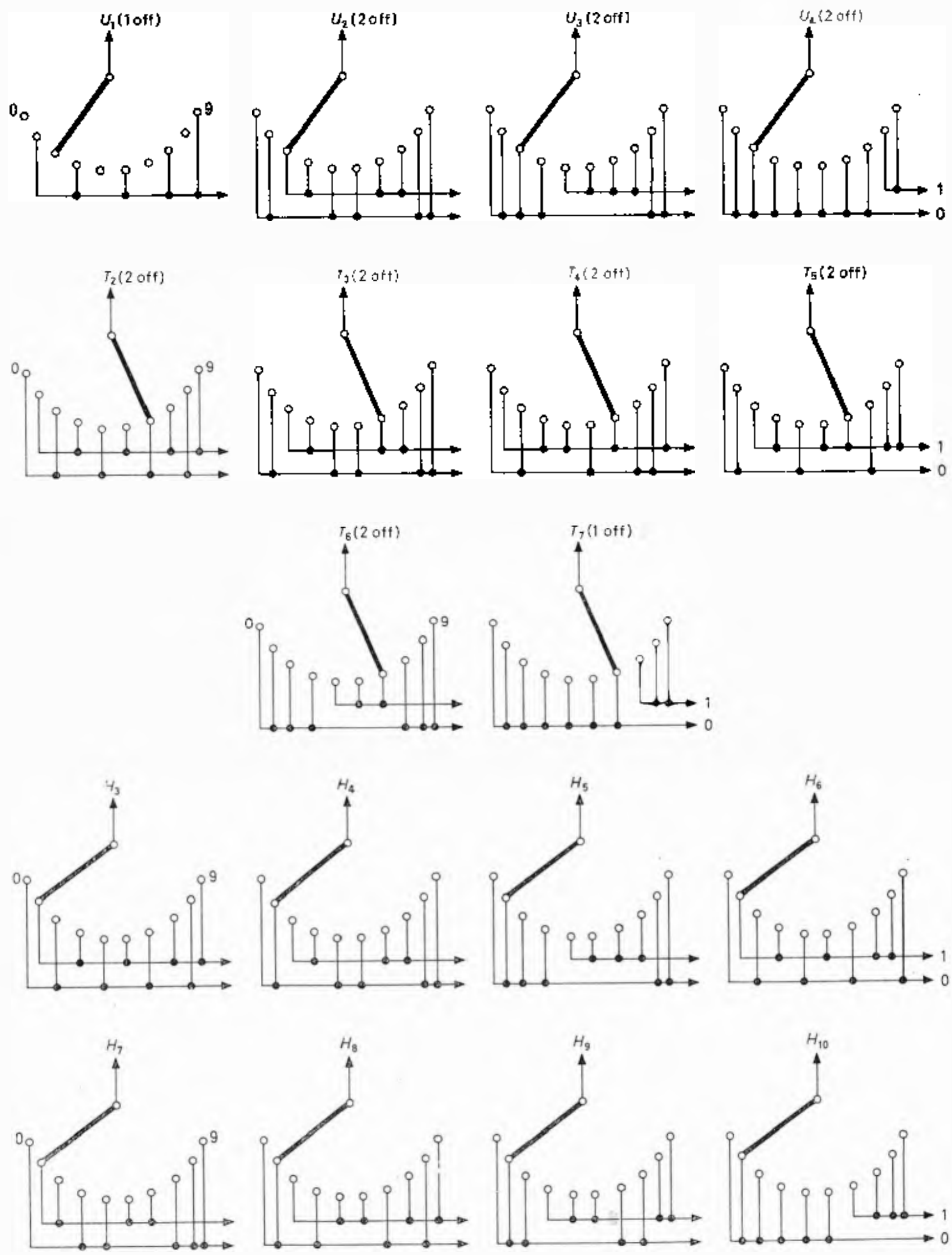
No special logic is needed for the stage 1 output since the T input is identically zero and there is no carry. It is, therefore, easily obtained from a single contact operated by U. For the seventh stage, one may take advantage of the largest number being 99, which is less than 127. A 1 output is induced by the tens input being at least 70 or there being a carry from the previous stage. The two conditions do not arise simultaneously, so forward carry circuits are unnecessary as is also an X input. A simple alternative connexion will therefore produce the required function, as shown in the figure.

Complete Circuit

The switches in Fig. 3 are all shown in the 0 state and each switch shown corresponds to a 10-way wafer whose contacts are wired to one pole or the other according to the appropriate binary values, as shown in Fig. 4. It will be seen that six wafers are required for the units switch, U, and eleven for the tens, T, together with fourteen diodes. Tracing a path from the + supply to, say, the S_{16} output, one includes a succession of diode OR configurations separated by series contacts in the closed state. Unlike conventional AND devices, these contacts impose

Fig. 4. Connexion of wafer switches

(Quantities quoted refer to Fig. 3)



no limit to the current available and therefore any restrictions upon 'fanning out' are only trivial. The available output voltage at the later stages will be reduced because of the number of series diodes each with, for silicon, a 0·6V drop. A supply of 10V or more is therefore advisable.

Three Switch Code Convertor (Binary)

Logical Relations

With the addition of a hundreds switch, the output will consists of ten bits*, and ten stages therefore appear in Fig. 5. The Z carry denotes 'carry 2', while the X and Y are as before. The most general stage is the fourth, which can have three inputs from H, T and U and up to two through the Z carry. If the possible combinations are examined, it emerges that Z carries are required at one time or another between the third and fourth, fourth and fifth, fifth and sixth and, for 768 and 769 only, the sixth and seventh stages. However, the properties of the decimal system are such that no Z carry is called for between the seventh and eighth stages. Also, the total input to the fourth stage never exceeds 4, but there proves to be no rewarding way of making use of this property.

A total input Table 2(a) may be compiled for all possible input conditions as before, and the logic tables 2(b), (c), (d) and (e) derived for S, X', Y' and Z' respectively.

* *See Nairn (Ref. 1); Table 4.*

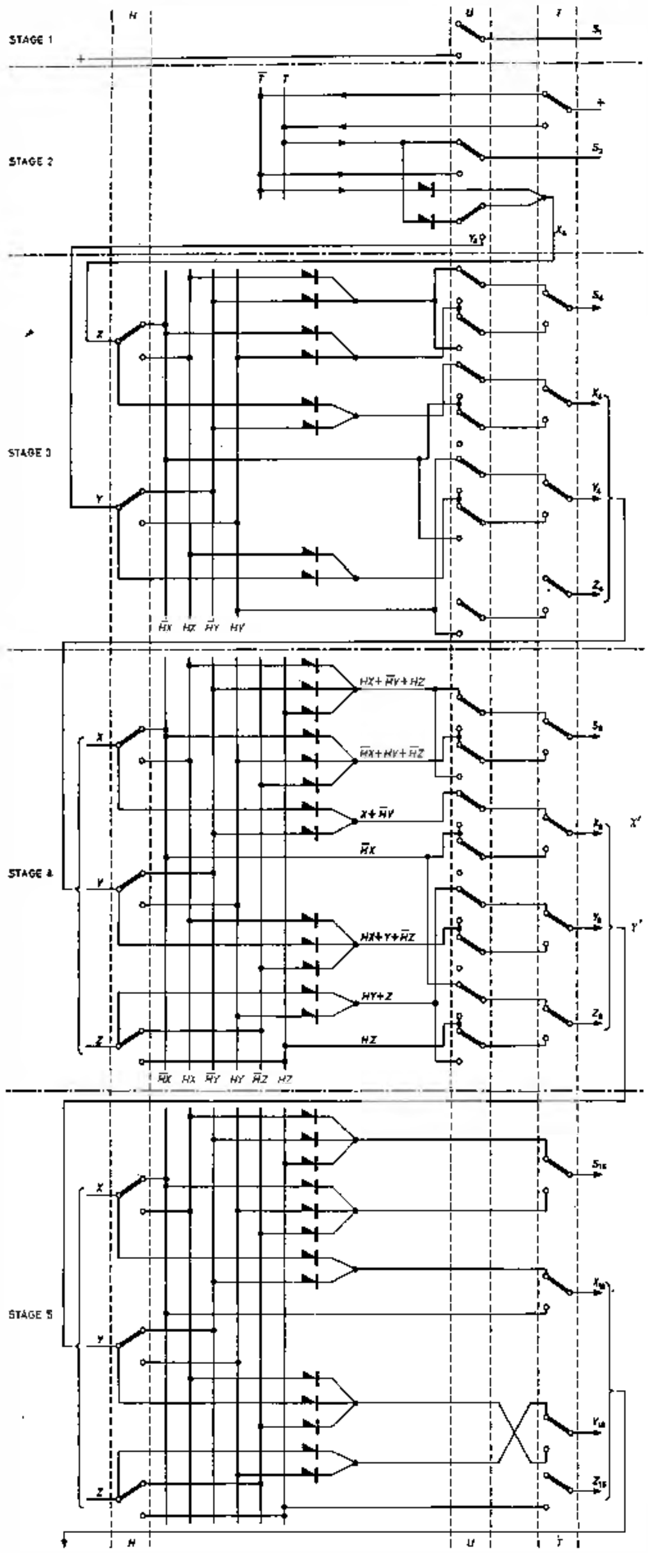
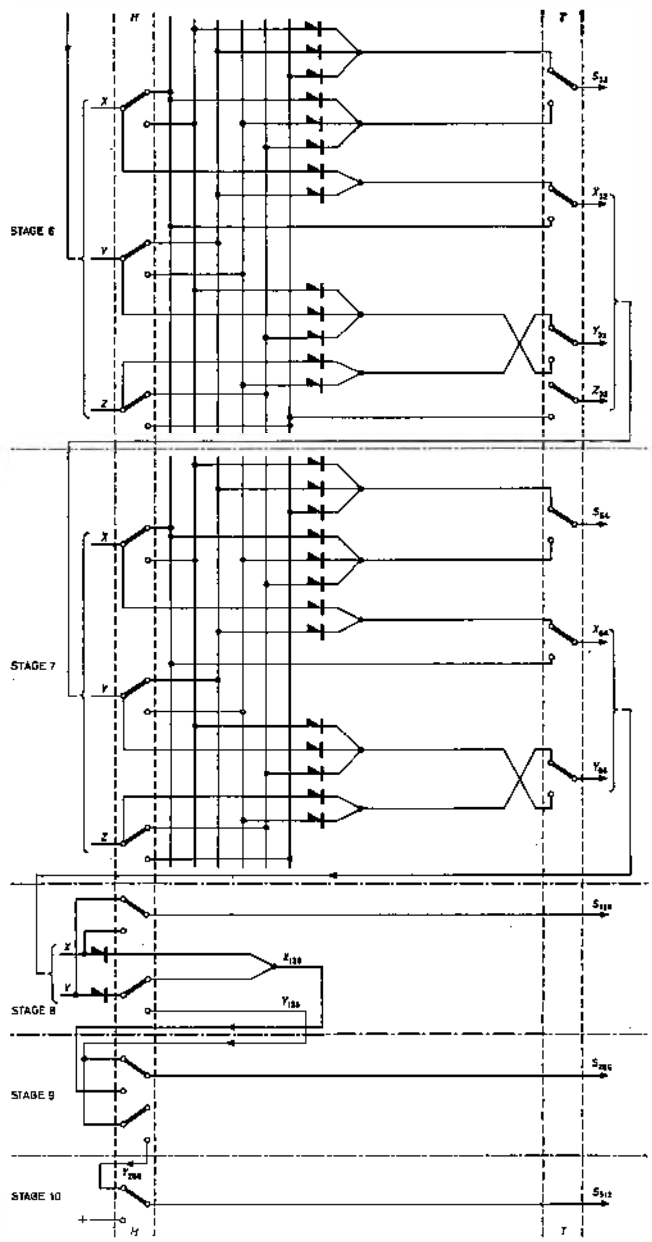


Fig. 5. Circuit for three-switch convertor (binary)



Switch Connexions

As previously, each carry line in Fig. 5 feeds a changeover contact, this time to produce a voltage on one bus bar out of six, corresponding to the column of Table 2. It will be seen that the same combination appears in more than one table, e.g. top row of 2(c) and bottom row of 2(d); each different such row function is accordingly assembled using diodes appropriately connected, noting that the relation $\overline{H}X + HX = X$, for example, can save a diode.

Starting, now, from each output of stage 4, a tree of contacts will provide four circuits, corresponding to the four rows of Tables 2. Connecting each circuit to the appropriate logic function, derived above, ensures that the outputs become correctly energized. Thus the truth of one of the six combinations of H with X, Y or Z at the left leads to the energization of the appropriate bus bar, and therefore of the diode circuits to give rise on seven wires to the functions shown. These are then routed to the appropriate output through the *T-U* contact trees at the right. It will be seen that, in the general stage, 3 + 8 + 4 changeover contacts are required.

The third stage has H, T and U inputs, but only X or Y carry. As a result, the Z' output occurs only for condition $UTHY$, and the T wafer associated with $\overline{U}$ is not required. This circuit is therefore eliminated, and only 2 + 7 + 4 wafers are needed.

Another specialization of the general circuit applies to stages five, six and seven which receive T and H inputs and X, Y and Z carry. Instead of constructing a complete *T-H* tree for each output, one may proceed as if, in the tables, the inputs were T and U, and H was always in the $\overline{H}$ state. The result is given in Fig. 5 using 3 + 4 wafers each time. (N.B. The Z' circuit is not needed for the seventh stage, so only 3 + 3 wafers are required in this case).

The remaining stages will, in general, follow the circuits used for the two switch case. Thus, stages two, eight, nine and ten correspond exactly to stages two, five, six and seven respectively of Fig. 3, apart from the switch identification.

Allocation of Wafers to Switches

The inputs provided by the U, T and H switches are each of unit value at the binary weight in question, so the labelling of each group of wafers in a given stage is arbitrary. So as to share the required number of wafers among the switches as evenly as possible, it is necessary to adopt the allocation shown. Thus there is a total requirement of 18 wafers on the units, 20 wafers on the tens, and 19 wafers on the hundreds switch, leading to the complete circuit of Fig. 5. The switches are again connected as shown in Fig. 4 and a total of 64 diodes is required.

Decimal Operation

Associated with each of the binary outputs is a circuit which accepts inputs in 1 out of 10 form, each such circuit being independent of the others.

Two Switch Code Convertor (Decimal)

Logical Relationships

The tens switch has no influence on the S_1 output and this may therefore be derived as before from a single wafer on the U switch. Other outputs are given as functions of the T and U inputs in Table 3 and, if all the functions of U corresponding to each of the rows are

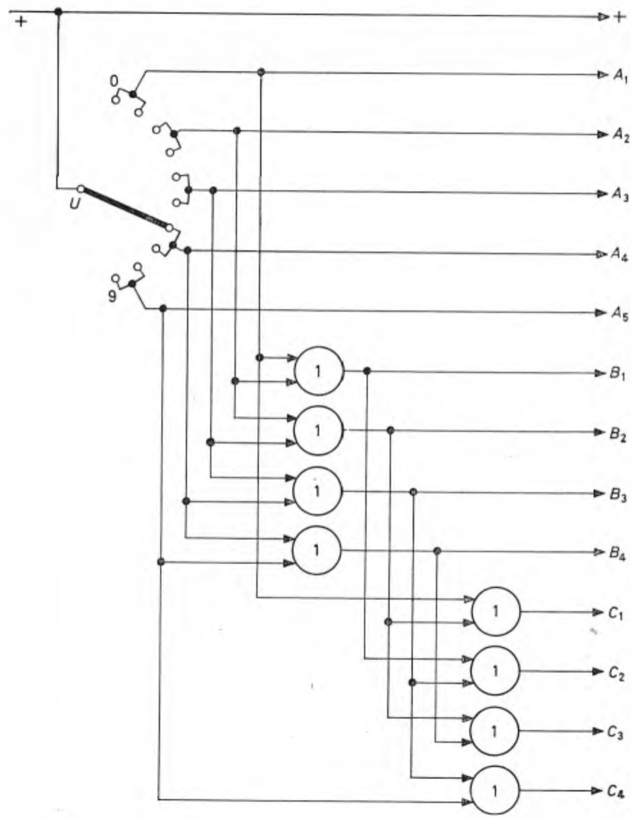


Fig. 6. Two-switch convertor (decimal). Provision of basic row functions

Fig. 7. Two-switch convertor (decimal). Selection of row functions

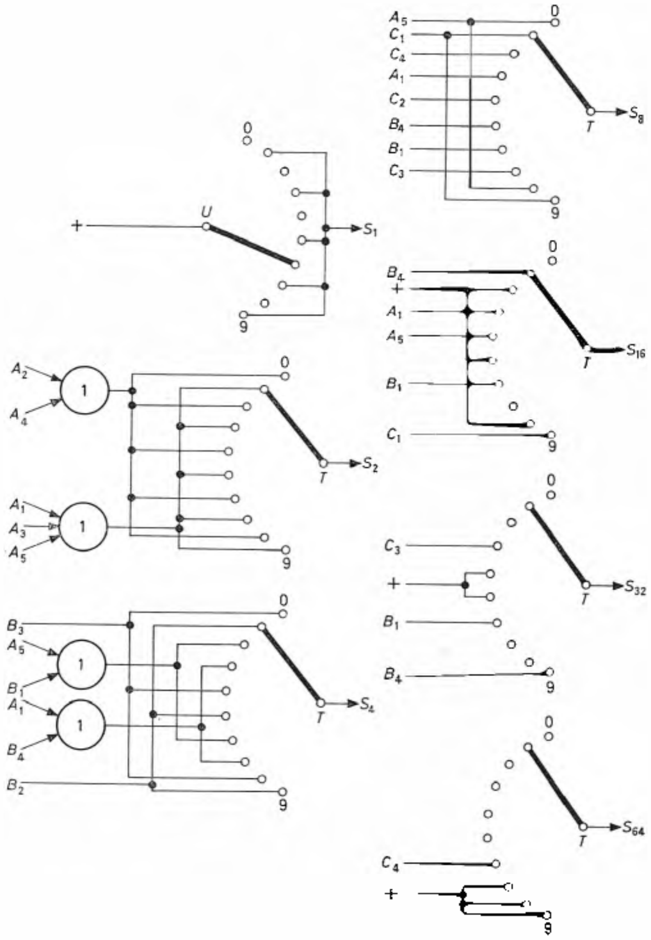


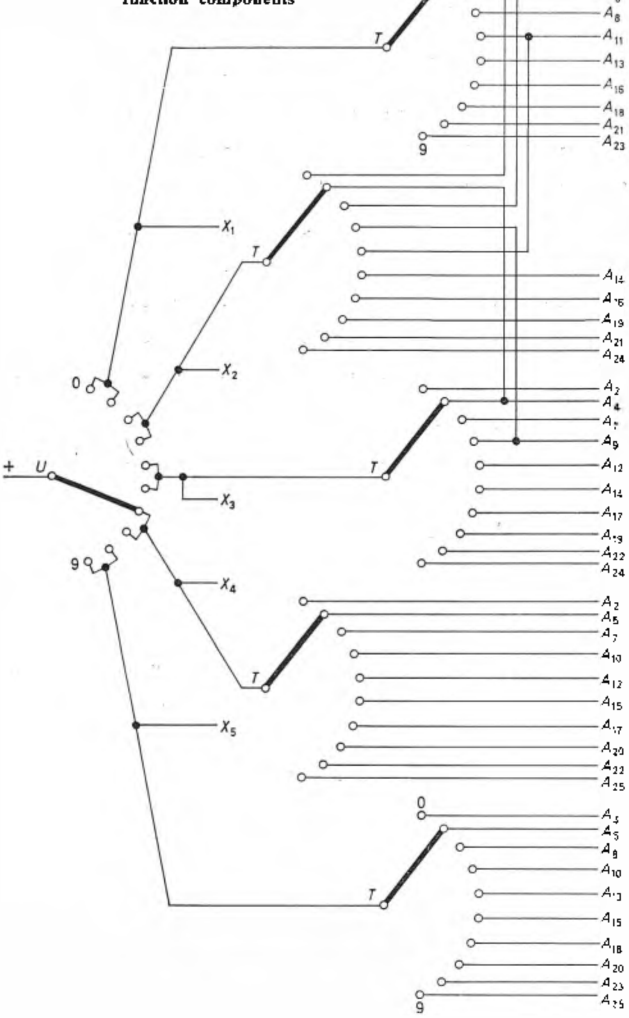
TABLE 3

Binary outputs as functions of two decimal inputs

BINARY OUTPUT	T INPUT	U INPUT 0,1	2,3	4,5	6,7	8,9	DECOMPOSITION OF ROW FUNCTION
S_2	Even 10's		*		*		A_2+A_4
	Odd 10's	*		*		*	$A_1+A_3+A_5$
S_4	00, 40, 80			*	*		B_3
	10, 50, 90		*	*			B_2
	20, 60	*	*			*	B_1+A_5
	30, 70	*			*	*	A_1+B_4
S_8	00, 80					*	A_5
	10, 90	*	*	*			C_1
	20			*	*	*	C_4
	30	*					A_1
	40	*	*	*	*		C_2
	50				*	*	B_4
	60	*	*				B_1
	70		*	*	*	*	C_3
S_{16}	00, 70						0
	10				*	*	B_4
	20, 50, 80	*	*	*	*	*	1
	30	*					A_1
	40					*	A_5
	60	*	*				B_1
	90	*	*	*			C_1
S_{32}	00, 10, 20, 70, 80						0
	30		*	*	*	*	C_3
	40, 50	*	*	*	*	*	1
	60	*	*				B_1
	90				*	*	B_4
S_{64}	00, 10, 20, 30, 40, 50						0
	60			*	*	*	C_4
	70, 80, 90	*	*	*	*	*	1

* *and a space represent a '1' and '0' respectively*

Fig. 8. Three-switch convertor (decimal). Provision of basic row function components



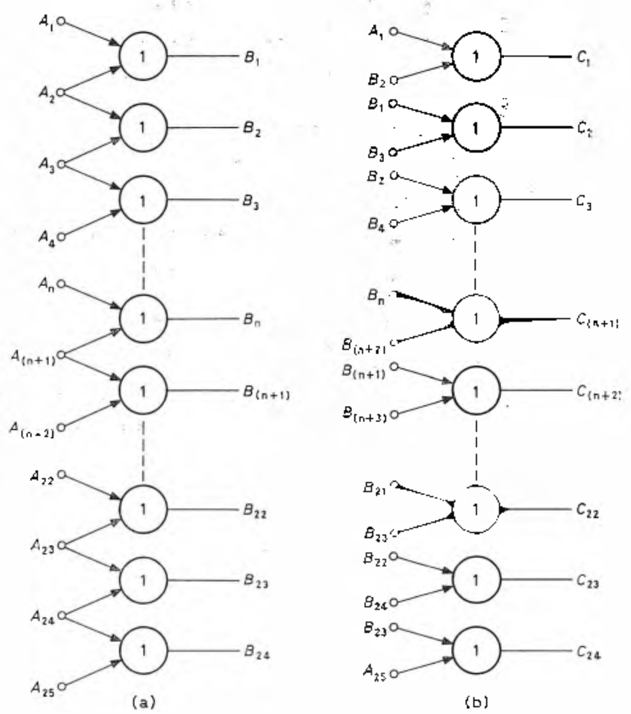


Fig. 9. Three-switch convertor (decimal). Grouping into larger row function components

Fig. 10. Three-switch convertor (decimal). Provision of 'twenties' functions

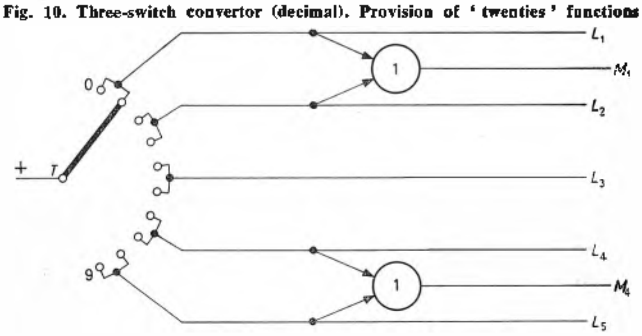


TABLE 4

Binary outputs as functions of three decimal inputs

BINARY OUTPUT	H INPUT	T, U COMBINATION																									DECOMPOSITION OF ROW FUNCTION
		00 to 03	04 07	08 11	12 15	16 19	20 23	24 27	28 31	32 35	36 39	40 43	44 47	48 to 51	52 55	56 59	60 63	64 67	68 71	72 75	76 79	80 83	84 87	88 91	92 to 95	96 99	
S_4	Even 100's		•		•		•		•		•		•		•		•		•		•		•		•		Even A's
	Odd 100's	•		•		•		•		•		•		•		•		•		•		•		•		•	Odd A's
S_8	000, 400, 800			•	•			•	•			•	•			•	•			•	•			•	•		$B_3+B_7+B_{11}+B_{15}+B_{19}+B_{23}$
	100, 500, 900		•	•			•	•			•	•			•	•			•	•			•	•			$B_2+B_6+B_{10}+B_{14}+B_{18}+B_{22}$
	200, 600	•	•			•	•			•	•			•	•			•	•			•	•			•	$B_1+B_5+B_9+B_{13}+B_{17}+B_{21}+B_{25}$
	300, 700	•			•	•			•	•			•	•			•	•			•	•			•	•	$A_1+B_4+B_8+B_{12}+B_{16}+B_{20}+B_{24}$
S_{16}	000 800					•	•	•	•						•	•	•	•					•	•	•	•		$C_6+C_{14}+C_{22}$
	100, 900				•	•	•	•						•	•	•	•					•	•	•	•			$C_5+C_{13}+C_{21}$
	200			•	•	•	•						•	•	•	•					•	•	•	•				$C_4+C_{12}+C_{20}$
	300		•	•	•	•						•	•	•	•					•	•	•	•					$C_3+C_{11}+C_{19}$
	400	•	•	•	•					•	•	•	•					•	•	•	•					•	$C_2+C_{10}+C_{18}+A_{25}$
	500	•	•	•					•	•	•	•					•	•	•	•					•	•	$C_1+C_9+C_{17}+B_{24}$
	600	•	•					•	•	•	•					•	•	•	•					•	•	•	$B_1+C_8+C_{16}+C_{24}$
	700	•					•	•	•	•					•	•	•	•					•	•	•	•	$A_1+C_7+C_{15}+C_{23}$
S_{32}	000									•	•	•	•	•	•	•	•									•	$C_{10}+C_{18}+A_{25}$
	100	•	•	•	•	•	•	•									•	•	•	•	•	•	•	•			$C_3+C_5+C_{17}+C_{21}$
	200							•	•	•	•	•	•	•	•									•	•	•	$C_6+C_{12}+C_{24}$
	300	•	•	•	•	•									•	•	•	•	•	•	•	•					$L_1+C_{15}+C_{19}$
	400					•	•	•	•	•	•	•	•									•	•	•	•	•	$C_6+C_{16}+L_5$
	500	•	•	•									•	•	•	•	•	•	•	•							$C_1+C_{13}+C_{17}$
	600			•	•	•	•	•	•	•	•									•	•	•	•	•	•	•	$C_4+C_8+C_{20}+C_{24}$
	700	•									•	•	•	•	•	•	•	•									$A_{11}+C_{11}+C_{15}$
	800	•	•	•	•	•	•	•	•									•	•	•	•	•	•	•	•		$C_2+C_6+C_{18}+C_{23}$
	900								•	•	•	•	•	•	•	•									•	•	$C_9+C_{13}+B_{24}$
S_{64}	000																	•	•	•	•	•	•	•	•	•	$C_{19}+L_5$
	100	•	•	•	•	•	•	•																	•	•	$L_1+B_8+B_{24}$
	200	•	•	•	•	•	•	•	•	•	•	•	•	•	•												M_1+C_{12}
	300						•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•					$L_2+L_3+L_4+A_{21}$
	400													•	•	•	•	•	•	•	•	•	•	•	•	•	$C_{14}+M_4$
	500	•	•	•																	•	•	•	•	•	•	$C_1+A_{16}+L_5$
	600	•	•	•	•	•	•	•	•	•	•																M_1
	700		•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•									$C_3+L_2+L_3+B_{19}$
	800									•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•		$B_9+L_3+L_4+C_{22}$
	900																•	•	•	•	•	•	•	•	•	•	M_4
S_{128}	000																										0
	100								•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	$C_9+(L_3+M_4)$
	200	•	•	•	•	•	•	•	•	•	•	•	•	•	•												(M_1+C_{12})
	300																						•	•	•	•	C_{23}
	400, 900	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	+
	500	•	•	•																							C_1
	600											•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	(L_3+M_4)
	700	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•									$(M_1+C_{12})+C_{11}$
	800																									•	A_{26}
S_{256}	000, 100, 600																										0
	200															•	•	•	•	•	•	•	•	•	•	•	$A_{15}+M_4$
	300, 400, 800, 900	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	+
	500	•	•	•																							C_1
	700																		•	•	•	•	•	•	•	•	$C_{10}+L_4$
S_{512}	000 to 400																										0
	500				•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	$C_4+\{C_6+(L_3+M_4)\}$
	600 to 900	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	•	+

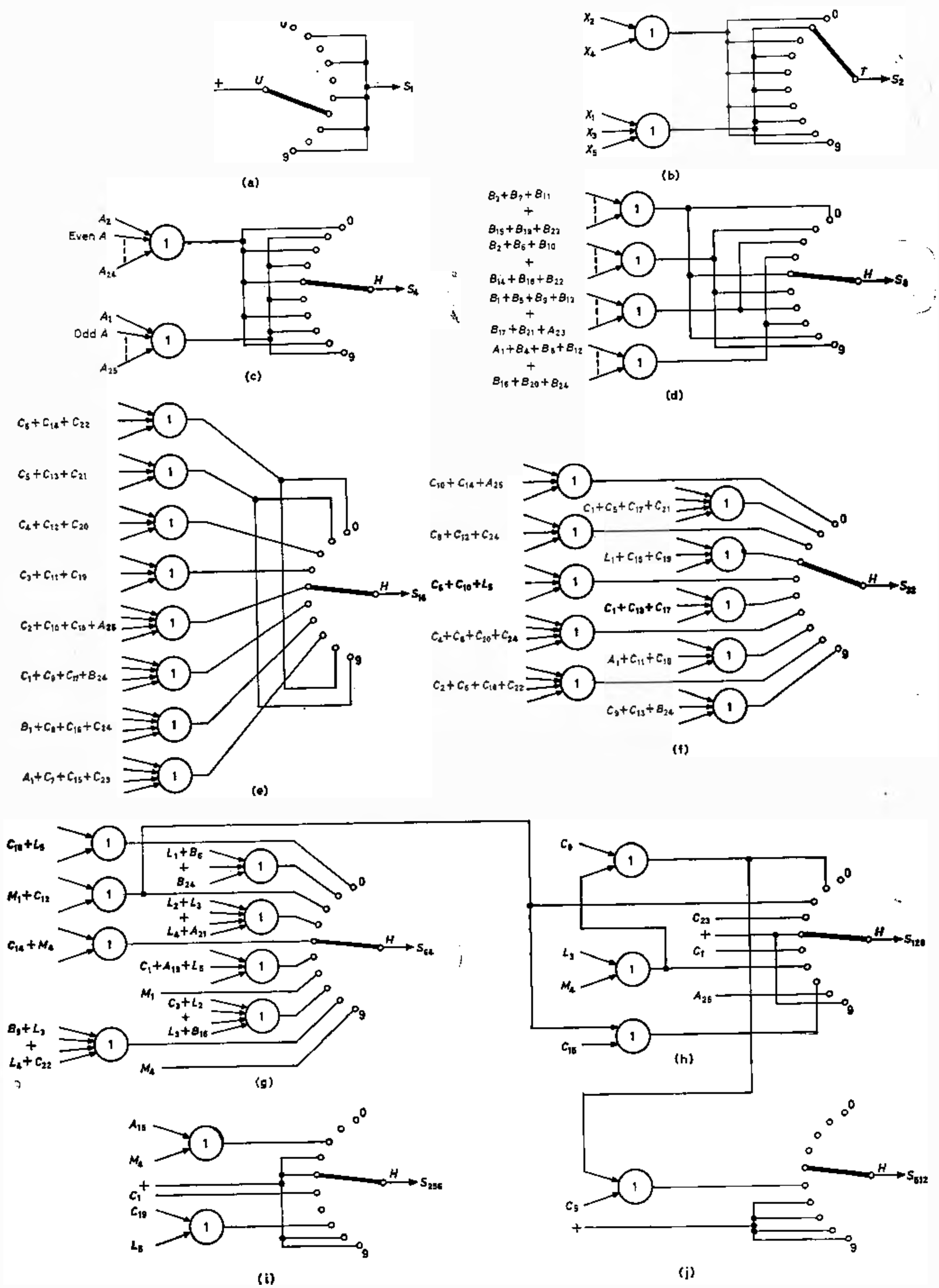


Fig. 11. Three-switch convertor (decimal). Derivation and selection of row functions

made electrically available (c.f. the functions in Tables 1), a T wafer allocated to each binary output can be connected so as to select the function appropriate to the T switch position. (c.f. the binary T switches of Fig. 5, stage 5 selecting the functions provided by the diodes).

Derivation of Functions

A U switch causes a voltage to appear (Fig. 6) on one of five terminals, A_1 to A_5, corresponding to the five columns of Table 3, and these are connected in pairs to diode OR gates to give four outputs, B_1 to B_4. Further combinations, C_1 to C_4, of adjacent threes or fours are also given by a further four OR gates. From these thirteen components and the voltage supply terminal, +, the required functions may be obtained or economically derived in a manner entered at the right of each row in Table 3.

SWITCH CONNEXIONS

Fig. 7 shows the connexion of a T switch bank for each binary output from S_2 onwards, together with OR gates as called for in Table 3. It will be seen that, altogether, 2 + 6 wafers suffice, but with 25 diodes in one three-input and eleven two-input gates.

THREE SWITCH CODE CONVERTOR (DECIMAL)

Logical Relationships

In this case, the S_1 output is influenced only by the U switch, and the S_2 only by the U and T switches. These two stages may therefore be identical to those just described above. As far as outputs S_4 and higher are concerned, one may consider the range 0 to 99, controlled by U and T, as being made up of twenty-five states, only one of which is energized at a time (c.f. the six bus bars of Fig. 5, stages 4, 5, 6 and 7).

Table 4 lists the various outputs for the possible combinations of these twenty-five states with the ten possible H states. Again, if all the functions of T, U given in the rows can be made available, the appropriate one can be selected by means of H switches.

Derivation of Functions

A U switch, strapped as in Fig. 6, has five output poles; if each such pole is allocated one ten way T wafer, and is connected as in Fig. 8, fifty outlets are obtained from the combination, only one of which is energized at a time. These may be paired off as shown to give twenty-five outputs, A_1 to A_{25}.

Further pairing through twenty-four two-input OR gates (Fig. 9(a)) gives paired outputs B_1 to B_{24} which, in turn, may be further paired (Fig. 9(b)) to give twenty-four more outputs C_1 to C_{24}. B outputs thus occur in adjacent pairs of A while C outputs involve adjacent fours (or threes at the two extremes). From these outputs it is straightforward to build up the functions required for S_4 to S_{32} outputs.

For higher order outputs, however, some economy in diode gates if afforded by bringing in five 'twenties' circuits from an extra T wafer. This is shown in Fig. 10, providing five outputs L_1 to L_5 and end pairs, M_1 and M_4.

The disjunction of the various row functions into A, B, C, L or M functions is given in the right-hand column of Table 4, for which is required:

2	inputs	gates;	Quantity	59 =	118	diodes
3	,,	,,	,,	13 =	39	,,
4	,,	,,	,,	10 =	40	,,
6	,,	,,	,,	2 =	12	,,
7	,,	,,	,,	2 =	14	,,
12	,,	,,	,,	1 =	12	,,
13	,,	,,	,,	1 =	13	,,
				Total	248	,,

Switch Connexions

Figs. 11(c) to (j) show the connexion of an H switch bank for each binary output from S_4 onwards, together with the requisite input OR gates. The total count is 2 + 7 + 8 wafers and 248 diodes.

At the expense of further wafers, it is possible to reduce the number of diodes a little; for example, remove the need for the 12 and 13 input gates from the S_4 circuit.

Conclusion

A method has been described for avoiding the use of active logic units in the conversion of a decimal switch input into a parallel binary output, and examples given for the two and three decade cases for direct decimal and binary coded switch inputs. In principle, larger numbers of decades could be tackled in the same way, but with the number of wafers or diodes involved receding further from the bounds of practical possibility.

Whether the three decade box is worth making depends upon a number of factors including:

(a) The ease of accommodating a suitable switch. There are now switches on the market which can carry ten wafers, each in turn carrying two 10-way contact sets. For some applications the space occupied may prove a liability.

(b) The practicability and cost of providing and accommodating several hundred diodes.

(c) The cost of wiring and soldering so many joints relative to the cost of 21 half-adders and 3 + 4 + 6 wafers quoted by Nairn[1].

(N.B. If the use of multi-contact relays is allowed, their substitution in Fig. 3 for U switch banks 3 and 4 and T switch banks 3, 4, 5 and 6, and their operation by switches as described by Nairn[1], may have advantage in some applications.)

It is not claimed that, for the decimal case, the most economical logical network was chosen for producing the necessary functions from twenty-five poles, but a substantial reduction in the total number of diodes or connexions seems unlikely.

It would be interesting to carry through a design based upon ternary inputs and to compare the total number of wafers and of diodes with those for the two cases described above. Intuition suggests the quantities would be intermediate.

Any adventure to the right of the decimal point is limited to the provision of an on/off switch adding in 0·5, equivalent to 0·1 in the binary scale. All other multiples of 1/10 convert to recurring binary numbers and are clearly inadmissible.

Acknowledgments

Thanks are due to Dr. E. L. C. White, Chief Scientist, E.M.I. Electronics Ltd., for permission to publish this article and to Mr. A. J. S. Udall for suggesting the decimal approach.

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A Mains Stabilized High Voltage Power Supply Using Controlled Avalanche Rectifiers

By R. W. J. Barker*, B.Sc.

A novel technique employing controlled avalanche rectifiers to produce a mains stabilized, low power, high voltage power supply is described.

A method of calculating the performance of such a power supply using the conventional voltage doubler configuration and specifying the components is derived, and graphs produced to assist in the evaluation of the derived equations.

The practical performance of a laboratory model power supply designed using the derived design procedure is compared with the theoretical performance.

(Voir page 203 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 210)

THERE exists a family of rectifying circuits[1] capable of giving a d.c. output, which is largely insensitive to supply voltage variations. These circuits are characterized by some or all of the rectifying elements working in the avalanche (or Zener) mode over a portion of each cycle.

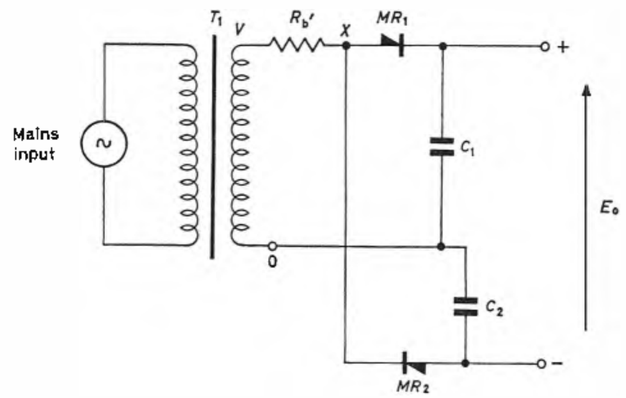


Fig. 1. Voltage doubler circuit

A modified form of the conventional voltage doubler configuration, see Fig. 1, will be considered in detail, although the same basic principles of operation and treatment may be extended to other members of the family.

Principle of Operation

With reference to the circuit diagram, Fig. 1, the mains transformer T_1 feeds avalanche diodes MR_1 and MR_2 through a resistor R_b'. MR_1 and MR_2 represent one or more avalanche diodes in series. Capacitors C_1 and C_2 are charged by MR_1 and MR_2 respectively. Items MR_1, MR_2, C_1 and C_2 comprise a conventional voltage doubler configuration.

Each diode (or set of diodes), MR_1 and MR_2, has an avalanche voltage of approximately the value required for the stabilized d.c. output voltage, E_o.

Under working conditions a potential of $E_o/2$ appears across C_1 and C_2. At the instant when MR_1 becomes forward biased, the voltage at point x with respect to the negative output terminal is limited by the avalanche effect in MR_2 to $+E_o$. Similarly when MR_2 is forward biased the potential at point x with respect to the positive output terminal is limited by the avalanche effect in MR_1 to $-E_o$. The a.c. potential developed between 0 and x is, therefore, a clipped sine wave of peak-to-peak amplitude E_o. R_b' is a ballast resistor which also limits the avalanche current flowing in the diodes.

The output voltage of the circuit is largely determined by the peak-to-peak amplitude of the a.c. potential between 0 and x. It is evident that if the mains input voltage rises or falls, the peak-to-peak amplitude of this potential will be substantially the same.

The output voltage of the supply will, therefore, be little affected by mains voltage variations.

Analysis of the Supply

Initial Considerations

It will normally be necessary to specify the circuit components to achieve a certain stabilization ratio dE_o/dv for a given d.c. output voltage E_o and current I_o.

The d.c. output voltage is determined principally by the value of the avalanche voltage ($2E_a$) of MR_1 and MR_2. MR_1 and MR_2 may each be one or more diodes in series. It is essential that the mismatch in $2E_a$ between each diode set is a minimum, as any mismatch in voltage will manifest

Symbols

v	=	Transformer secondary voltage on no load
i	=	Transformer secondary r.m.s. current
E_a	=	$\frac{\text{Total avalanche voltage of each diode set}}{2}$
ϕ	=	parameter = $\sin^{-1}(E_a/\sqrt{2}v)$
R_b	=	Resistance of the ballast resistor + transformer winding resistance referred to the secondary
R_a	=	total dynamic resistance of each diode set when operating in the avalanche mode
E_o	=	D.C. output voltage ($\simeq 2E_a$)
I_o	=	D.C. output current
θ_j/a	=	Thermal resistance of each diode from junction to ambient (°C/W)
T_a	=	Temperature coefficient of avalanche voltage of diode type (/°C)
$V_{Rb(av)}$	=	Average voltage across R_b
$I_{Rb(av)}$	=	Average current through R_b
$V_{Rb(RMS)}$	=	R.M.S. voltage across R_b
R_b'	=	Resistance of the ballast resistor
P_{Rb}'	=	R.M.S. voltage across R_b'
n	=	Number of avalanche diodes in each set

* Formerly Edwards High Vacuum International Ltd.

itself as a supplementary ripple voltage across C_1 and C_2. Large mismatches in $2E_a$ will also invalidate the design equations which assume perfectly balanced conditions.

In general output voltage changes due to mains voltage variations will be influenced by the dynamic resistance of the diodes during avalanche conduction $\left.\frac{\partial E_o}{\partial v}\right|_{T_a=o}$ and the changes in avalanche voltage due to junction temperature changes caused by a change in power dissipation, $\left.\frac{\partial E_o}{\partial v}\right|_{R_a=o}$.

$$\frac{dE_o}{dv} = \left.\frac{\partial E_o}{\partial v}\right|_{T_a=o} + \left.\frac{\partial E_o}{\partial v}\right|_{R_a=o} \quad \ldots (1)$$

Determination of $\left.\frac{\partial E_o}{\partial v}\right|_{T_a=o}$

It is initially desirable to formulate an expression between R_b and ϕ.

Where $\phi = \sin^{-1} E_a/\sqrt{2}v$ .... (2)

It is evident that if the value of v increases (due to an increase in mains voltage) that the lost charge in C_1 and C_2 will be replaced before the end of the conduction angle. Once the capacitor charge is replaced all current flowing through R_b will flow through the corresponding avalanche diodes causing E_a to increase due to the action of R_a, and giving rise to an increased output voltage, E_o.

To carry out an exact analysis based on the above discussion would lead to rather unwieldy expressions which would be of little practical use. To simplify matters the assumption is made that when a change in v occurs, *all* the current change through R_b flows through the appropriate diode set, which is conducting in the avalanche mode.

In making the above assumption one will, in practice, obtain lower values for $\left.\frac{\partial E_o}{\partial v}\right|_{T_a=o}$ than the calculated values.

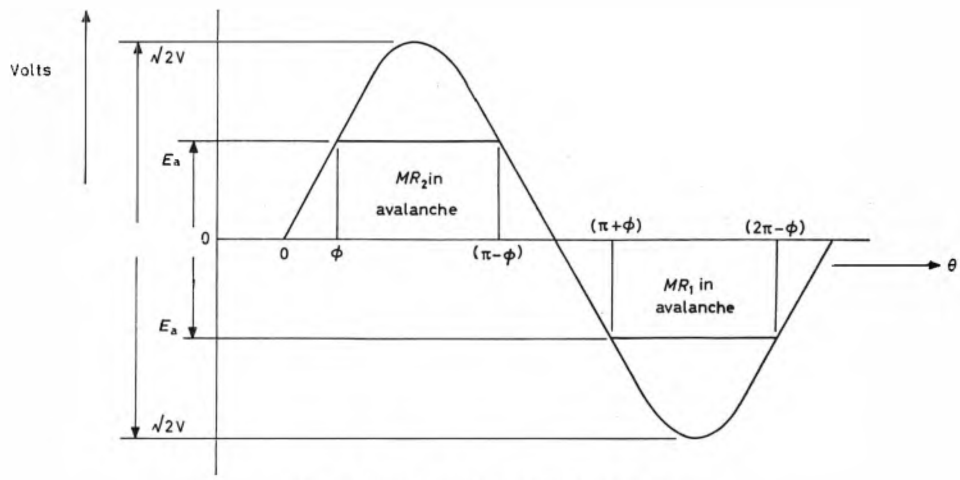


Fig. 2. Illustration of the conduction angle

If conduction in the avalanche mode occurs over an angular range ϕ to $(\pi - \phi)$, see Fig. 2, the average voltage across R_b is given by:

$$V_{Rb(av)} = \frac{1}{\pi - 2\phi}\int_{\phi}^{(\pi-\phi)} \sqrt{2}v\,(\sin\theta - \sin\phi)\,d\theta$$

$$\therefore\ V_{Rb(av)} = \frac{\sqrt{2}v}{\pi - 2\phi}\,[2\cos\theta - \sin\phi\,(\pi - 2\phi)] \quad \ldots (3)$$

If a load current I_o flows, the average current required to maintain the charge in the capacitors is given by:

$$I_{Rb(av)} = \frac{I_o\,2\pi}{\pi - 2\phi} \quad \ldots\ldots\ldots\ldots\ldots (4)$$

From equations (3) and (4):

$$R_b = \frac{V_{Rb(av)}}{I_{Rb(av)}} = \frac{\sqrt{2}v}{2\pi I_o}[2\cos\phi - \sin\phi\,(\pi - 2\phi)] \quad \ldots (5)$$

(assuming that the transformer reactances have a negligible affect).

Simplifying by substituting equation (2) in equation (5):

$$R_b = E_a/I_o\,\frac{2\cot\phi - (\pi - 2\phi)}{2\pi} \quad \ldots\ldots\ldots\ldots (6)$$

In equation (6) R_b gives the value of the total ballast resistance that is required to replace the lost charge in C_1 and C_2 during the conduction angle ϕ to $(\pi - \phi)$. This is calculated when ϕ is a maximum, i.e. v is at its minimum value.

Proceeding with the above assumption. If the mains voltage increases giving rise to an increase in v of Δv the change in E_o is given by:

$$\Delta E_o = \frac{\sqrt{2}\,\Delta v\,R_a}{R_a + R_b}$$

$$\therefore\ \left.\frac{\partial E_o}{\partial v}\right|_{T_a=o} = \frac{\sqrt{2}\,R_a}{R_a + R_b} \quad \ldots\ldots\ldots\ldots (7)$$

Combining equations (6) and (7):

$$\left.\frac{\partial E_o}{\partial v}\right|_{T_a=o} = \frac{2\sqrt{2}\pi\,R_a I_o}{2\pi R_a I_o + E_a\,[2\cot\phi - (\pi - 2\phi)]} \quad \ldots (8)$$

Determination of $\left.\frac{\partial E_o}{\partial v}\right|_{R_a=o}$

Examination of the power dissipation in the circuit of Fig. 1 shows that:

$$2P_d = vi - \frac{V_{Rb(RMS)}{}^2}{R_b} - 2\,E_a I_o \quad \ldots\ldots (9)$$

also:

$$i = \frac{V_{Rb(RMS)}}{R_b} \quad \ldots\ldots\ldots\ldots\ldots (10)$$

Combining equations (9) and (10):

$$2P_d = \frac{v\,V_{Rb(RMS)}}{R_b} - (V_{Rb}{}^2/R_b) - 2\,E_a I_o \quad \ldots\ldots (11)$$

From Fig. 2 it can be seen that:

$$V_{Rb(RMS)} = [2v^2/\pi \{ \int_{\phi}^{(\pi-\phi)} (\sin\theta - \sin\phi)^2 . d\theta \}]^{1/2} \quad \ldots . (12)$$

After integration and substitution equation (12) becomes:

$$V_{Rb(RMS)} = v/\pi^{1/2} [(\pi - 2\phi)(2 - \cos 2\phi) - 3\sin 2\phi]^{1/2} \quad \ldots\ldots\ldots\ldots (13)$$

From equations (2), (11) and (13):

$$2P_d = \frac{E_a^2}{2\pi^{1/2}R_b} . (\lambda^{1/2}/\sin^2\phi) - (E_a^2/2\pi R_b) . (\lambda^{1/2}/\sin^2\phi) - 2\, I_o\, E_a \quad \ldots\ldots\ldots\ldots (14)$$

where:

$$\lambda = (\pi - 2\phi)(2 - \cos 2\phi) - 3\sin 2\phi \quad \ldots\ldots (15)$$

$\frac{\partial E_o}{\partial v}\Big|_{R_a=0}$ may be found from the relationship:

$$\frac{\partial E_o}{\partial v}\Big|_{R_a=0} = (dE_o/dP_d) . (d\phi/dv) . (dP_d/d\phi) \quad \ldots\ldots (16)$$

dE_o/dP_d may be found by considering the result of a change in dissipation, in a diode set containing n diodes, on the combined avalanche voltage $2E_a$.

$$dE_o/dP_d = (dE_o/dT) . (dT/dP_d) \quad \ldots\ldots. (17)$$

where:

$$dE_o/dT = 2E_a\, T_a \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (18)$$

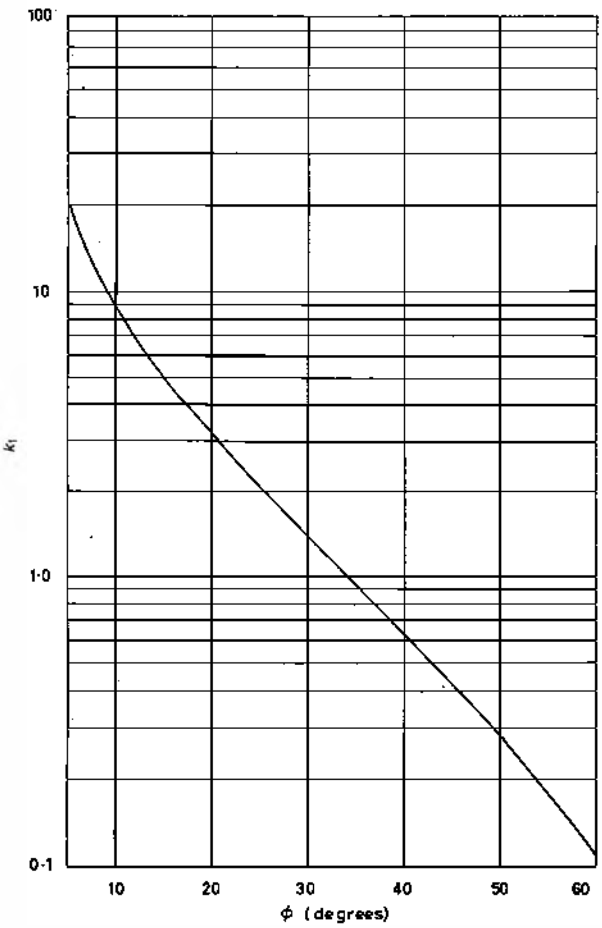
and:

$$dT/dP_d = \frac{\theta_j/a}{n} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (19)$$

Combining equations (17), (18) and (19):

$$dE_o/dP_d = \frac{2\, E_a\, T_a\, \theta_j/a}{n} \quad \ldots\ldots\ldots\ldots\ldots (20)$$

Fig. 3. ϕ against k_1



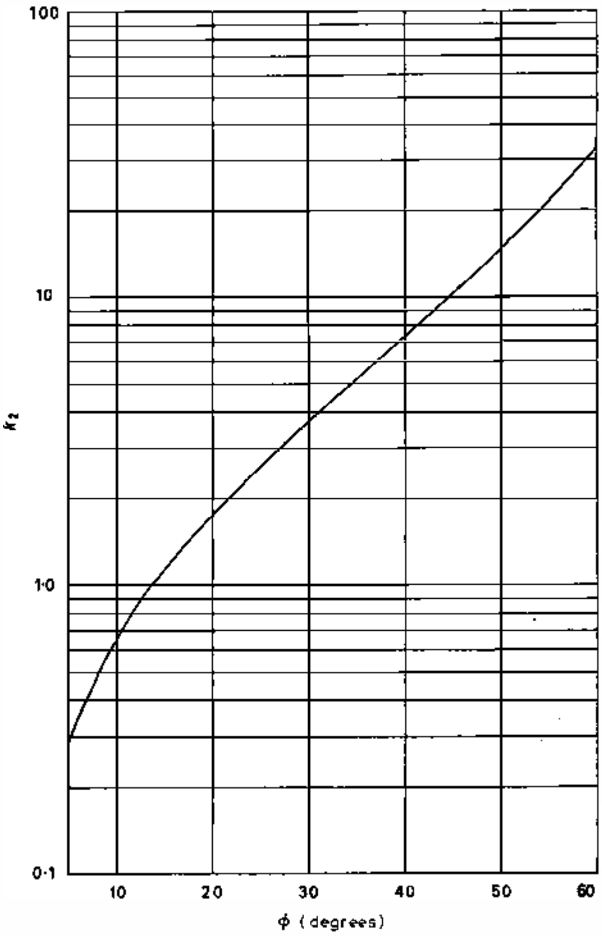


Fig. 4. ϕ against k_2

$d\phi/dv$ may be found from equation (2).

$$\phi = \sin^{-1}(E_a/\sqrt{2}v) \quad \ldots\ldots\ldots\ldots\ldots\ldots (2)$$

$$d\phi/dv = -(\sqrt{2}/E_a) . \frac{\sin^2\phi}{\cos\phi} \quad \ldots\ldots\ldots\ldots (21)$$

$dP_d/d\phi$ may be found from equation (14) and (15).

$$\frac{2dP_d}{d\phi} = \frac{E_a^2}{2\pi R_b} \cdot \frac{1}{\sin^3\phi} [\{(\pi\lambda)^{1/2} - \lambda\} \{(\sin\phi/2\lambda) . (d\lambda/d\phi) - 2\cos\phi\} - (1/2)\sin\phi\,(d\lambda/d\phi)] \quad \ldots\ldots\ldots\ldots (22)$$

where:

$$d\lambda/d\phi = 2(\pi - 2\phi)\sin 2\phi - 4(1 + \cos 2\phi) \quad \ldots . (23)$$

From equations (16), (20), (21) and (22):

$$\frac{\partial E_o}{\partial v}\Big|_{R_a=0} = \frac{\sqrt{2}\, T_a\, \theta_j/a\, E_a^2}{\pi\, R_b\, n \sin 2\phi} [(\sin\phi/2) . (d\lambda/d\phi) - \{(\sin\phi/2\lambda) . (d\lambda/d\phi) - 2\cos\phi\} \{(\pi\lambda)^{1/2} - \lambda\}] \quad \ldots\ldots\ldots\ldots (24)$$

Substituting for R_b from equation (6) in equation (24):

$$\frac{\partial E_o}{\partial v}\Big|_{R_a=0} = \frac{T_a\, \theta_j/a\, E_a\, I_o}{n} \left[\frac{2\sqrt{2}}{2\cot\phi - (\pi - 2\phi)}\right] \frac{1}{\sin 2\phi} [(\sin\phi/2) . (d\lambda/d\phi) - \{(\sin\phi/2\lambda) . (d\lambda/d\phi) - 2\cos\phi\} \{(\pi\lambda)^{1/2} - \lambda\}] \quad \ldots\ldots\ldots\ldots (25)$$

DETERMINATION OF v AND R_b FOR A GIVEN dE_o/dv

From equations (1), (8) and (25) an expression can be found for dE_o/dv in terms of the various constants and parameter ϕ.

$$dE_o/dv = \frac{2\sqrt{2}\pi\, R_a\, I_o}{2\pi\, R_a\, I_o + E_a\,(k_1)} + \frac{T_a\,\theta_j/a\, E_a\, I_o\,(k_2)}{n} \quad ..\ (26)$$

where:

$$k_1 = 2\cot\phi - (\pi - 2\phi) \quad ....\ (27)$$

$$k_2 = \left[\frac{2\sqrt{2}}{2\cot\phi - (\pi - 2\phi)}\right]\frac{1}{\sin 2\phi}$$
$$[(\sin\phi/2)\,.\,(d\lambda/d\phi) - \{(\sin\phi/2\lambda)\,.\,(d\lambda/d\phi)\ -2\cos\phi\}$$
$$\{(\pi\lambda)^{1/2} - \lambda\}] \quad ....\ (28)$$

Plots of ϕ against k_1 and k_2 are given in Figs 3 and 4. With the aid of equation (26) and Figs 3 and 4 parameter ϕ can be found for the required dE_o/dv by an iterative process.

From this value of ϕ, the transformer secondary voltage v can be determined from equation (2). This will correspond to the secondary voltage when the mains voltage is at its minimum value. R_b can be found from equation (6).

The expression dE_o/dv in equation (26) is used to determine ϕ when the mains voltage is at a minimum (i.e. ϕ is at a maximum). The value of dE_o/dv will be modified for higher mains voltages. It is evident that $\left.\frac{\partial E_o}{\partial v}\right|_{T_a = 0}$ will be unaffected as it is independent of ϕ (see equation (7)). In order to determine $\left.\frac{\partial E_o}{\partial v}\right|_{R_a = 0}$, for higher mains voltages equation (24) should be used. To assist in the evaluation of equation (24) a plot of ϕ against k_4 is given in Fig. 5 where:

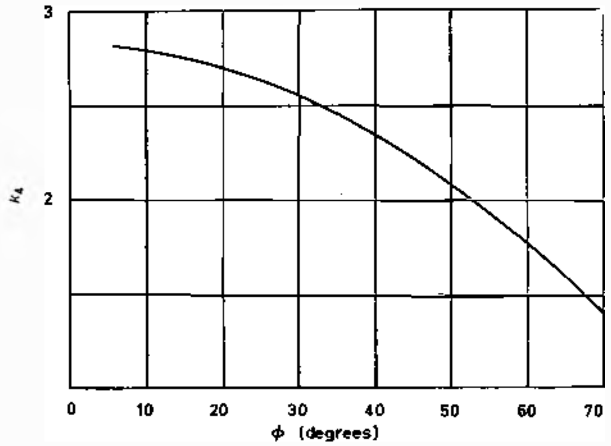


Fig. 5. ϕ against k_4

$$k_4 = \frac{\sqrt{2}}{\sin 2\phi}\ [(\sin\phi/2)\,.\,(d\lambda/d\phi) -$$
$$\{(\sin\phi/2\lambda)\,.\,(d\lambda/d\phi) - 2\cos\phi\}\ \{(\pi\lambda)^{1/2} - \lambda\}]$$
$$............\ (29)$$

Determination of i and P_{Rb}'

The transformer secondary current, i, and the power dissipation in R_b', P_{Rb}', will normally be evaluated when the mains is at its maximum value. A value of ϕ corresponding with maximum mains voltage can be found from equation (2). i may be found by using this value for ϕ in equation (13) and then substituting for $V_{Rb(RMS)}$ in equation (10). P_{Rb}' can be found knowing i and R_b'. Calculated in this manner, i and P_{Rb}' give the necessary ratings for the transformer secondary current and the ballast resistor.

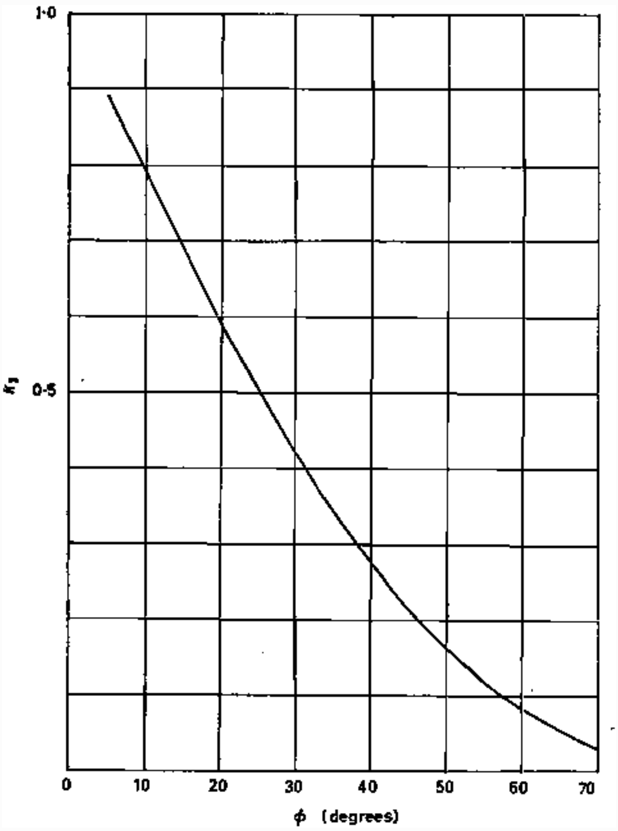


Fig. 6. ϕ against k_3

To assist in the evaluation of equation (13) a plot of ϕ against k_3 is given in Fig. 6 where:

$$k_3 = \left[\frac{(\pi - 2\phi)\,(2 - \cos 2\phi) - 3\sin 2\phi}{\pi}\right]^{\frac{1}{2}} \quad ..\ (30)$$

Determination of Diode Power Dissipation

The power dissipation in each diode set may be obtained by substituting in equation (11), from which the power dissipation of any individual diode can be assessed.

Design Procedure for a Laboratory Model Supply

Load Conditions

$I_{o(max)} = 1{\cdot}2$mA $\quad E_o \simeq 3\,000$V

Stabilization Ratio

$dE_o/dv \simeq 0{\cdot}1$ with mains variations of 90 to 110 per cent of the nominal value.

Diode Information

Type RAS 508.AF on 3in × 3in × 1/16in bright aluminium heatsink.

$\theta_j/a = 11{\cdot}5$°C/W (from published data)

$T_a = 0{\cdot}15\%/$°C (by sample measurement)

$E_a \simeq 1\,500$V (by selection)

$R_a = 2 \times 20$kΩ (sample measurement, average over the range 0 to 2mA avalanche current).

$n = 2$

Equation (26):

$$\frac{2\sqrt{2}\,\pi\, R_a\, I_o}{2\pi\, R_a\, I_o + E_a\,(k_1)} + \frac{T_a\,\theta_j/a\, E_a\, I_o\,(k_2)}{n} = dE_o/dv$$

let $\phi = 25°$ $\quad k_1 = 2{\cdot}05,\ k_2 = 2{\cdot}55$

$\therefore\ dE_o/dv = 0{\cdot}126 + 0{\cdot}040 = 0{\cdot}166$ (Too high)

let $\phi = 19°$ $\quad k_1 = 3{\cdot}4,\quad k_2 = 1{\cdot}6$

$\therefore\ dE_o/dv = 0{\cdot}0784 + 0{\cdot}0248 = 0{\cdot}103$

With $\phi = 19°$ From equation (2) $v = \frac{E_a}{\sqrt{2} \sin 19°} = 3\,260\text{V}$

This is the transformer voltage when the mains is 90 per cent of its nominal value, therefore the nominal value of $v = 3\,620\text{V}$.

PRACTICAL RESULTS

A power supply was tested using the values calculated above. The results obtained with $C_1 = C_2 = 0{\cdot}25\mu\text{F}$ are given in Table 1.

TABLE 1

		MAINS CHANGE 90%-100% OF NOMINAL			MAINS CHANGE 100%-110% OF NOMINAL		
		$\frac{\partial E_o}{\partial v}\big\vert_{T_a=0}$	$\frac{\partial E_o}{\partial v}\big\vert_{R_a=0}$	$\frac{dE_o}{dv}$	$\frac{\partial E_o}{\partial v}\big\vert_{T_a=0}$	$\frac{\partial E_o}{\partial v}\big\vert_{R_a=0}$	$\frac{dE_o}{dv}$
E_o=3000V	I_o=1·2mA	+0·078	+0·029	+0·107	+0·063	+0·027	+0·09
E_o=3116V	I_o=0mA	+0·030	+0·025	+0·055	+0·028	+0·027	+0·055

Effective output resistance = 116/1·2 = 97kΩ
Mismatch in diode sets = 10V

From equation (6):

$$R_b = E_a/I_o \frac{2 \cot \phi - (\pi - 2\phi)}{2\pi} = 660\text{k}\Omega$$

Total transformer resistance referred to the secondary = 48kΩ

Total transformer impedance referred to the secondary = 60kΩ

$\therefore$ Transformer reactance $= \sqrt{(60^2 - 48^2)} = 36\text{k}\Omega$ which may be neglected compared with 660kΩ in quadrature.

$\therefore$ Ballast resistor $R_b' = 660\text{k}\Omega - 48\text{k}\Omega = 612\text{k}\Omega$

When mains is at 110 per cent of nominal value

$$v = 3\,260 \times 1{\cdot}22 = 3\,990\text{V}$$

$\therefore \phi = 15{\cdot}5°$

From equations (13) and (30), $V_{Rb(RMS)} = v(k_3)$

From Fig. 6 when $\phi = 15{\cdot}5°$, $k_3 = 0{\cdot}68$.

$\therefore V_{Rb(RMS)} = 0{\cdot}68 \times 3\,990 = 2\,700\text{V}$.

From equation (10) $i = \frac{V_{Rb(RMS)}}{R_b} = 4{\cdot}1\text{mA}$

$$P_{Rb}' = i^2 R_b' = 10{\cdot}3\text{W}$$

When mains is at 110 per cent of nominal value $\frac{\partial E_o}{\partial v}\big|_{R_a=0}$ is given by equation (24)

$$\frac{\partial E_o}{\partial v}\bigg|_{R_a=0} = \frac{T_a\, \theta_j/a \,.\, E_a^2}{\pi R_b n}[k_4]$$

From Fig. 5 when $\phi = 15{\cdot}5°$ $k_4 = 2{\cdot}75$

$$\therefore \frac{\partial E_o}{\partial v}\bigg|_{R_a=0} = 0{\cdot}0258$$

$\therefore dE_o/dv = 0{\cdot}0784 + 0{\cdot}0258 = 0{\cdot}104$

Maximum diode dissipation is given by equation (9) when $I_o = 0$, i.e. no load conditions and the mains is at 110 per cent of nominal value.

$$2 P_d = vi - \frac{V_{Rb(RMS)}^2}{R_b} - 2 E_a I_o = 4{\cdot}3\text{W}$$

$\therefore P_d = 2{\cdot}15\text{W}$

Summarizing:

Transformer secondary rating 3 620V at 4·1mA

$R_b' = 612\text{k}\Omega$ 10·3W

Discussion of the Results

In the above results, values of $\frac{\partial E_o}{\partial v}\big|_{R_a=0}$ are within 17 per cent of the calculated values, the main source of error is expected to be in the value of θ_j/a used.

As anticipated, the values of $\frac{\partial E_o}{\partial v}\big|_{r_a=0}$ are, in general, lower than the predicted values due to the assumptions made in the analysis, and because the value of R_a assumed constant in the analysis, decreases at larger values of avalanche current.

The value of $\frac{\partial E_o}{\partial v}\big|_{T_a=0}$, obtained with $I_o = 1{\cdot}2\text{mA}$ over the mains range 90 to 100 per cent of nominal, is relatively high approaching the theoretical figure (0·0784). This is expected to be due to the fact that the supply is operating under marginal conditions where the criterion for the value of R_b (described by equations (3), (4) and (5)) is established. The value of R_a is also likely to be at its maximum under these conditions giving rise to a high $\frac{\partial E_o}{\partial v}\big|_{T_a=0}$. The situation could be improved by a small increase in v.

Conclusions

The circuit described is a potentially cheap method of stabilizing the d.c. output voltage of a low power supply against mains voltage changes. Suitable applications could include its use as a pre-stabilizer feeding a conventional feedback stabilizer, an e.h.t. supply for cathode-ray tubes and a cold cathode vacuum gauge power supply.

ADVANTAGES

(1) The basic number of components required is the same as for a non-stabilizing conventional voltage doubler except for the addition of the ballast resistor, R_b'.

(2) The capacitor rating need only be that required to cope with the d.c. output voltage.

(3) The inherent voltage ripple for a given capacitor value is lower than would be obtained with the corresponding non-stabilizing circuit.

DISADVANTAGES

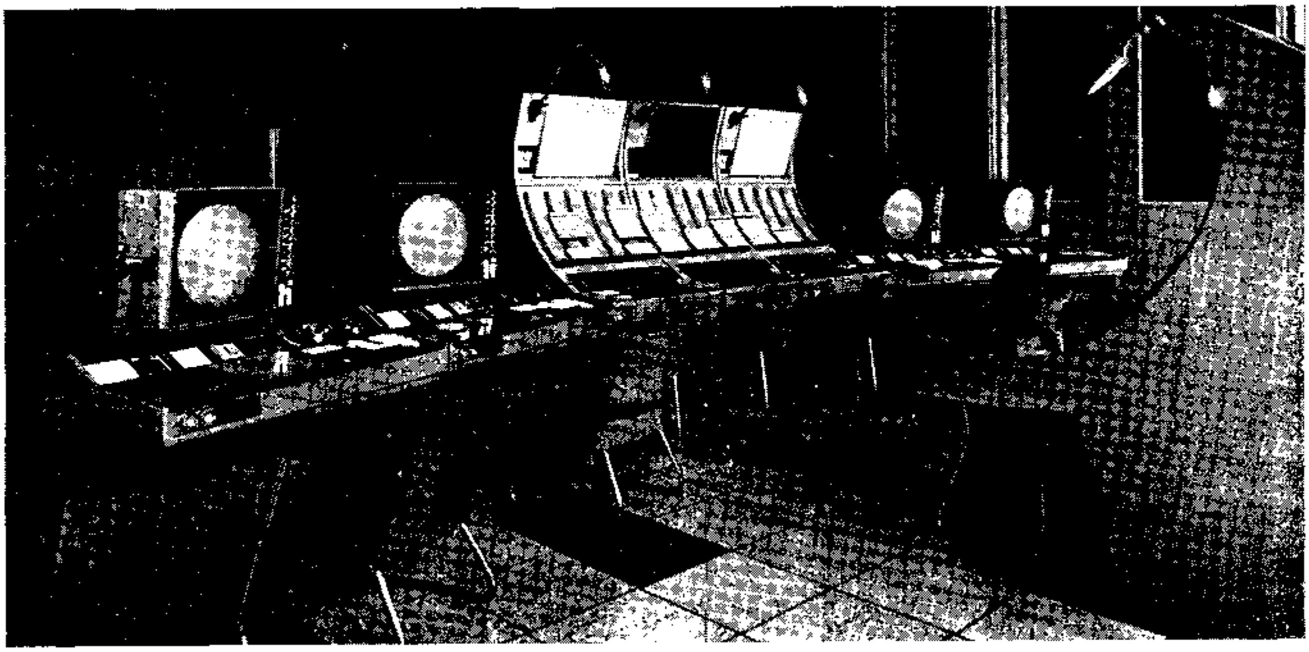
(1) The output voltage, being dependent on the avalanche voltage of the diodes is non-adjustable.

(2) Due to the current practice of manufacturers of specifying avalanche diode breakdown voltage characteristics with a very large tolerance, a selection procedure is necessary to obtain balanced sets of diodes which will give the necessary supply output voltage.

(3) The mismatch between the diode sets gives rise to a supplementary output voltage ripple.

REFERENCES

1. British Patent Appl. 43 268/65.



The Eurocontrol Experimental Centre

AN air traffic control simulator system has now been brought into operation at the EUROCONTROL Experimental Centre at Bretigny-sur-Orge, Essonne, France (40km south-west of Paris).

It will be recalled that the European Organization for the Safety of Air Navigation (EUROCONTROL) was set up in 1960 by seven European countries to strengthen the co-operation in matters of air navigation and in particular to provide a common organization of the air traffic services in the upper air space over the countries concerned (United Kingdom, Ireland, West Germany, Belgium, France, Holland and Luxembourg).

The first step is the creation of a EUROCONTROL operational upper air space control centre at Maastricht (Holland) covering the Belgium, Holland and Germany areas, due for completion by the early 1970s.

During the preliminary discussions on the creation of EUROCONTROL, it was realized that there would have to be an experimental centre to provide the means for investigating the problems of navigation, control and technology which would arise and for testing proposed solutions in order to deal with the complex air traffic control of today and that expected in the 1970s with the introduction of supersonic aircraft.

Land was made available by the French Government at the Centre d'Essais en Vol airfield and a building was erected to house the main experimental tool—a Dynamic Air Traffic Control Simulator—together with a staff of about 80. In 1964 EUROCONTROL awarded a contract valued at approximately \$5M for this simulator to a consortium of three European companies namely:

Compagnie Génerale de Télegraphie sans Fil (C.S.F.) France for the radar generation equipment (radar simulators, fixed echo and video mapping equipment, moving clutter generators), the procedural flight progress consoles, and the telecommunications system.

Plessey Radar Ltd, Britain, for the buffer unit, display drive unit, synthetic and tabular display drive, viewing units and the large screen display equipment, the radar consoles and the pilots' consoles.

Telefunken A.G., Germany, for the computing unit, together with its peripheral equipment and all the computer programmes.

In addition C.S.F. acted as the Project Manager for the Consortium and was responsible for the installation of the equipment.

The value of the contract is roughly the same for each Consortium member.

The main functions of the simulator are as follows:

(a) To calculate the position of between 60 and 300 aircraft within an extended airspace of 600 nautical mile radius taking into account the flight plans, performances, weather conditions, position and performance of navigational aids, and manoeuvres effected by the pilots.

(b) To simulate six primary radars with six associated secondary radars the geographical arrangement of which may be anywhere within the exercise area.

(c) To reconstruct a control scheme consisting of five sectors with one or several centres; each sector includes:
procedural control positions and
radar control positions,
both equipped with classical telecommunications means.

The above illustration shows a section of the control room with three C.S.F. procedural units (centre) and four Plessey vertical radar consoles

(d) To generate a synthetic video signal to be used by each control position.

(e) To collect the principal information, in a synthetic form, for the supervision position.

(f) To be able to replay and analyse an exercise.

The operation of this simulator can be on a mathematical basis whereby the main functions are represented by various formulae in which different numerical values for the principal variable parameters can be used to allow the main framework of a system, such as a route structure, to be tested in order to determine its optimum performance.

The simulator can also be operated on a dynamic basis so as to provide the environment in which the human factors of operation and the vital aspects of information display and interpretation, communications, and decision making can be studied in real-time, that is against the clock operating to normal time, or under speeded up conditions to compress the problems, thereby increasing the load on the operators and decreasing the time available for their reactions.

When operating dynamically the simulator can be used to:

Test new equipments for fitness for purpose.

Try out different configurations of airspace, control room layouts, a.t.c. organizations.

Investigate problems of traffic flow so as to obtain a measure of the significant parameters for examination and feedback into mathematical simulation.

Train operators and controllers in new procedures, techniques, and the use of new equipments.

Check plans and develop the means whereby they can be introduced smoothly into systems which are already operating.

Develop the computer programmes that modern control techniques using automatic data processing require.

Generally aid the operational and technical staff of EUROCONTROL to a better understanding of their problems and possible solutions so as to be ready to meet the demands being made on them by the increasing complexity and volume of air traffic.

The radar simulation sub-system by C.S.F. provides a network of six radar installation units each one containing an antenna rotation mechanism, an interface with the TR4 computer and a number of 'aircraft' circuits (15 for each of category I radars and 20 for the category **II** unit) each capable of generating the primary and secondary radar video corresponding to an air target. Circuits for the mixing and distribution of the generated radar videos and for the generation of internal clock and synchronization pulses is also incorporated. Furthermore, each radar simulation unit is equipped with a control panel by means of which the desired characteristics of the radar may be selected.

The following are some of the characteristics which can be selected for each simulated radar:

Antenna rotation speed	3-20rev/min
Pulse repetition frequency	200-500Hz
	400-1 000Hz
Pulse length	1·2, 2·5, 5 and 10μsec
Azimuth beamwidth (primary radar)	0·7, 1·4, 2·8, 5·6 and 11·2°
Azimuth beamwidth (S.S.R.)	1·4, 2·8, 5·6 and 11·2°

The telecommunication system supplied by the Belgian company S.A.I.T. Electronics as a sub-contractor to C.S.F. provides radiocommunication simulation between controllers and pilots on air ground equipment and consists of the following:

20 units each of four transmitters/receivers connected with each pilot,

5 units each of ten transmitters/receivers connected with control group,

20 telecontrol devices connected with each controller giving access to ten transmitters/receivers reunited in groups,

2 telecontrol devices connected with the two supervisors and giving access to all the 50 transmitters/receivers of the control groups.

A free choice of frequency, out of 50 frequencies available for the system assembly, may be allowed to the various transmission equipments mentioned above.

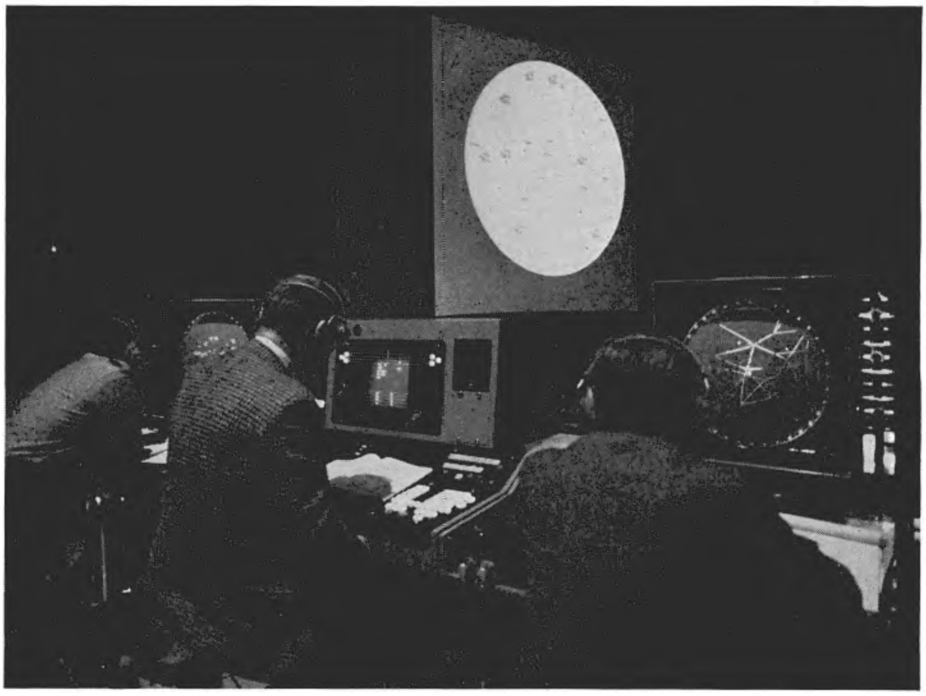
A further part of the telecommunication system consists of three separate ground to ground sub-assemblies providing intercommunication simulation between controllers, simulation exterior telephone circuits, and connexions between pilots and the pilot supervisor.

The Buffer Store unit, supplied by Plessey Radar Ltd, and based on a system of logic modules is fundamentally a stored instruction machine operating only on demand from an exterior source.

If more than one demand exists at any time then a queueing system is provided to give priority to the demands according to a pre-arranged order. A new demand cannot interrupt a demand that is already being dealt with. In general, action taken in response to a demand will be kept as short as possible, so that the waiting time of any demand is a minimum. This is particularly important with ancillaries having a low priority which would otherwise hold up those of a higher priority.

In this application the store is of 4 096 words, 25 bits with a read-write cycle time of 6μsec.

Horizontal radar display unit with video map, micro-tabular display, controls and keyboard



Supervision position—two radar consoles, one pilot's console, input keyboard and large screen display

The basic clock frequency is $1\frac{1}{3}$MHz allowing 8 microsteps within the 6μsec store cycle. Each operation that occurs within the buffer store is synchronized with these microsteps. The number of microsteps to answer a particular demand is dependent on the number of operations required.

The clock has two phases, X phase and Y phase, and each clock pulse is of 430nsec duration. There are three modes of clock operation, single step, single reduction and continuous.

The Plessey display equipment consists of:

(a) Eleven radar modules, eight with a 16in diameter p.p.i. display arranged vertically and mounted on a two-axis support to permit ergonomic and operational experiments, and three with a horizontally mounted 21in diameter conference display. In addition, five 16in monitor displays are provided.

(b) Ten control modules, each with a tabular keyboard for use with the radar modules.

Radar simulator control panel for selection of radar characteristics

(c) One large screen display providing a 6ft square picture, for use by the control supervisor, able to show by selection either any single simulated radar video or a synthetic plan display.

(d) Twenty 'pilots' modules each with a keyboard and Plessey Mk. 7 c.r.t. tabular display, suitable for operating up to 15 simulated aircraft from each position.

The 16in and 21in radar displays have video mixing facilities to show simulated primary and secondary radar video from any of the six radars, range markers, angle markers and video map presentation, with interscan markers for inter-console marking and range and bearing measurements; or if selected, a purely synthetic display with alpha-numeric labels generated by the computer.

Inter-console marking is achieved by means of a rolling ball position control and electronic circuit markers.

The computer supplied by Telefunken A.G. is the TRA, general-purpose programme controlled binary parallel machine of the one address type with a clock frequency of 2MHz. It handles binary numbers, decimal numbers and alpha-words. It has the following facilities:

A full set of 230 machine instructions.
A permanent store of 4 096 words of 52 bits.
An index store of 256 words of 16 bits.
A working store of 32 768 words of 52 bits.
One input/output channel for the control desk.
Six input/output channels for connexion with the peripheral units.

A control desk contains a typewriter and a control panel for the input and output of alpha-numeric control data, instructions and indications.

Six (with the possibility of eight) input/output channels can operate simultaneously and independently of each other. Up to eight input/output devices of different operating speeds may be linked to each one of the input/output channels, such as punched card readers and punchers, punched paper tape readers and punchers, high speed page printers, magnetic tape units.

An Electromechanical Running Averager

By K. C. Ng, Ph.D. and K. K. Murthy, M.E.

A new electromechanical running averager is described. It is inexpensive, simple in construction and reliable in operation. It is particularly useful when the running average of signals are required to be simulated on an analogue computer.

(Voir page 203 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 210)

IN certain analogue computer applications, the 'moving average' or the 'running average' of a functon of time is required. The running average $R(t)$ of a function of time $x(t)$ over an interval of time T is defined as:

$$R(t) = \int_0^T x\,(t-s)\,ds = \int_{t-T}^{t} x(s)\,ds \quad \ldots\ldots\ldots \quad (1)$$

where $x(t-s)$ is the delayed version of the function $x(t)$.

An example of a practical situation where the realization of equation (1) on the analogue computer is desired is the determination of the integral of cross-correlation function between two signals in a particular adaptive control system[1].

$$\text{Let } I = \int_0^T \phi_{xy}(s)ds \quad \ldots\ldots\ldots\ldots \quad (2)$$

where $\phi_{xy}(t)$ is the cross-correlation function between two signals $x(t)$ and $y(t)$.

Equation (2) can be written as:

$$I = \int_0^T \int_{-\infty}^{+\infty} y(t)\,x(t-s)dt\,ds \text{ (from the definition of } \phi_{xy}(t))$$

Changing the order of integration gives:

$$I = \int_{-\infty}^{+\infty} y(t)\,\{\int_0^T x(t-s)ds\}\,dt \quad \ldots\ldots\ldots\ldots \quad (3)$$

In equation (3), $\int_0^T x(t-s)ds$ is the running average of $x(t)$ over time T which is required to be simulated on the analogue computer for the determination of I.

This article explains a new method of obtaining the running average using a simple electromechanical system.

Equation (1) can be written as:

$$R(t) = \int_{-\infty}^{t} x(s)ds - \int_{-\infty}^{t} x(s-T)\,ds \quad \ldots\ldots\ldots \quad (4)$$

By the definition of equations (1) and (4), the running average $R(t)$ at time t of the signal $x(t)$ is the integral of the signal over the interval $(t-T)$ to t. That is, it is equal to the integral of $x(t)$ minus the integral of the delayed signal $x(t-T)$. The mechanization for the computation of $R(t)$ is shown schematically in Fig. 1.

From Fig. 1, the impulse response of the running averager can be expressed as:

$$h(t) = u(t) - u(t-T) \quad \ldots\ldots\ldots\ldots \quad (5)$$

where $u(t)$ is the unit step function and $u(t-T)$ is the delayed step function.

Thus, the impulse response of the running average has a box car waveform.

Methods of Simulation

The method of simulation of the running average as represented in Fig. 1 requires the delayed version of the time function. The accuracy with which the running average can be simulated largely depends upon the accuracy with which the delayed signal can be generated.

The delayed versions of time functions can be obtained by conventional techniques such as electronic delay circuits or magnetic tape delay lines. With these methods, precise unity gain setting for the delay block (Fig. 1) is difficult. If the gain is not exactly unity, the residual difference when integrated results in considerable deviations from the actual running average at the output of the integrator. This is a serious limitation on the accuracy obtainable by this method.

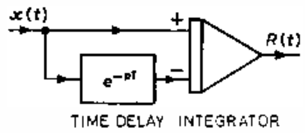


Fig. 1. Arrangement for averager

An alternative but rather expensive method is to convert the analogue signal into a digital signal. The delayed versions are then readily obtained and the computations of the running average is easily performed using reversible binary counters. A digital to analogue convertor is then used to obtain the analogue running average signal. The accuracy here is dependent only on the analogue to digital and digital to analogue convertors used.

The electromechanical running averager described in the next section has the advantages of low cost and simplicity. No complicated equipment or circuits are involved. It is particularly useful when analogue signals are involved.

Principle of Operation and Construction

The principle of operation of this is the capability of a capacitor drum to generate delayed versions of the time function. The performance of the capacitor drum as a delay unit is described first, before describing the running averager proper.

A number of capacitors are connected at one end each to a commutator segment on a rotating drum and at the other ends to a common point. Two brushes are provided on the commutator at diametrically opposite points. The input voltage (a function of time) is connected to one of the brushes, the other brush is taken to the input of a high input-impedance amplifier. The common point of the capacitors is earthed.

When a capacitor is in contact with the input brush through its connected commutator segment, the input voltage charges the capacitor to the instantaneous value of the input voltage. The capacitor moves on with its full charge until it comes in contact with the next brush when the charge on the capacitor causes the output voltage of the amplifier to rise to a value proportional to the stored charge. If the leakage is negligible, the output

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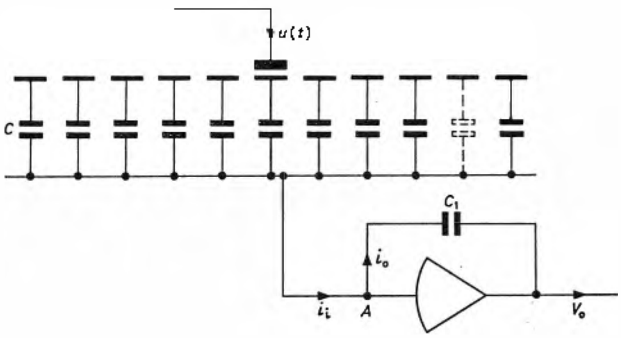


Fig. 2. Principle of the running averager

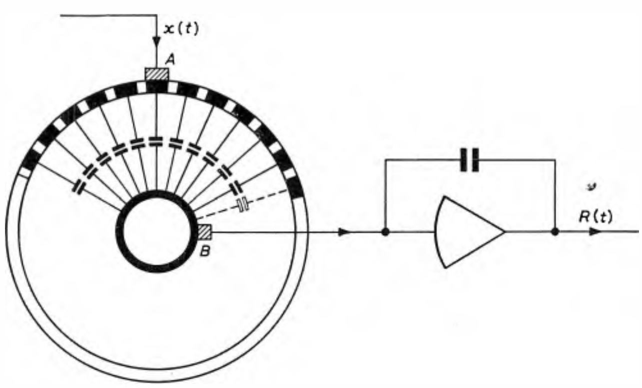


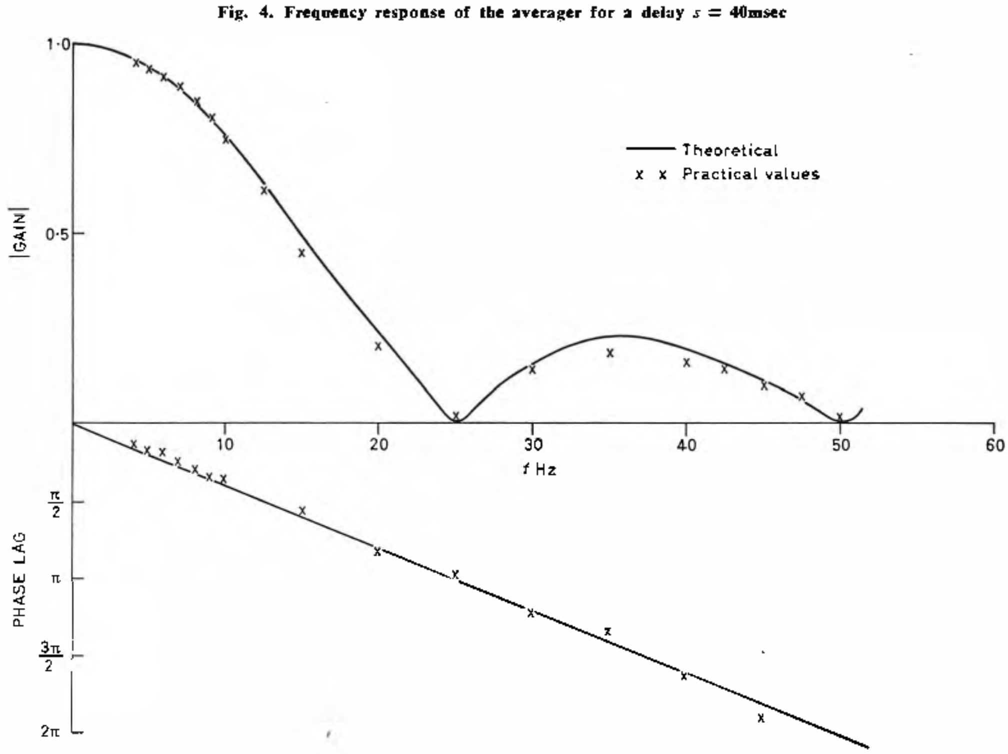
Fig. 3. Practical arrangement of running averager

voltage is the same as the instantaneous value of the input voltage, but delayed by a time interval equal to the time taken by the capacitor to move from one brush to the other. Similarly, successive capacitors delay the successive instantaneous values of the input voltage and the voltage at the output of the amplifier is thus a delayed version of the input voltage. The length of the delay is conveniently varied by varying the speed of rotation of the drum.

The capacitor drum described can be used to realize equation (4) and generate the running average.

This is best illustrated by considering the arrangement shown in Fig. 2. The voltage whose running average is required is connected to a brush on the drum. The common point of the capacitors is taken to the virtual earth of an electronic integrator. The capacitors are charged to the instantaneous value of the input voltage $u(t)$ as they pass under the brush in turn. After a complete revolution, when the first capacitor again comes under the brush, it is charged to the new value of the input voltage at that instant of time and thus the previous charge is automatically erased.

Fig. 4. Frequency response of the averager for a delay s = 40msec



If $u(t)$ is an impulse $A_1 u_0(t_0)$ occurring at $t = t_0$, the capacitor which is in contact with the brush at $t = t_0$ gets charged to the value:

$$q_i = A_1 C \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (7)$$

where C is the value of the capacitance.

At point A, the virtual earth of the amplifier.

$$i_i = i_o$$

$$\therefore\ q_i = q_o$$

Hence the output voltage V_o is given by:

$$V_o = \frac{-A_1 C}{C_1} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (8)$$

After one full revolution, when the same capacitor passes through the brush again the original charge is erased and hence the output voltage falls to zero. Thus the impulse response is a constant of magnitude $-A_1C/C_1$ from t_0 to $t_0 + n\Delta t$ and zero for all other values of t, where $n\Delta t$ is the time required for one complete revolution.

Thus, the successive values of the input signal is remembered over the time required for one revolution. The output of the integrator is the average value of the input signal over the last period, i.e. it is the running average of the input signal over this period.

Fig. 3 shows the practical set-up of the running averager. The capacitors are connected between two copper laminate disks on which the desired number of conducting segments (commutator segments) has been etched. Low leakage, high stability polystyrene capacitors are used. In the unit built, 127 such capacitors are employed. The capacitor drum is driven by a servo-motor. Since the time over which the running average is taken is the time required for one complete revolution of the drum, the running average of the signal over a wide range of time intervals can be obtained by driving the drum at the appropriate speed.

Frequency Response of the Running Averager

By transforming equation (5), the frequency response function $H(j\omega)$ of the running averager is obtained as

$$H(j\omega) = \frac{T \sin(\omega T/2)}{\omega T/2} \angle -(\omega T/2) \quad \ldots\ldots \quad (9)$$

This is plotted in Fig. 4 where the practical values obtained are also shown. It is seen that the experimental and theoretical values agree quite closely.

Conclusions

The running averager described is simple to construct and reliable in operation. The accuracy is limited by the number of capacitors used and in the experimental unit 127 discrete elements perform satisfactorily. The accuracy also depends upon the leakage in the capacitors and brush ripple. For the experimental unit, the error is within 2·5 per cent in the low frequency range. Six of these units are now under construction for use in a multi-channel adaptive computer financed by a Ministry of Technology Research Contract.

Acknowledgment

One of the authors (K. K. Murthy) wishes to thank the Commonwealth Scholarship Commission in the U.K. for the Scholarship Award for higher studies in this country.

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The Belfast Flight Simulator

A flight simulator, designed and manufactured by Miles Electronics Ltd of Shoreham, Sussex, has now been put into operation at the RAF Transport Command Squadron Station at Brize Norton for flight training with the Short S.C.5 Belfast aircraft.

This transport aircraft is a 4-turboprop machine having a maximum payload of over 35 tons with a speed of 306 knots and a range of 5300 miles. As a trooper aircraft it can carry 250 men.

The simulator (Fig. 1) is equipped with a cockpit motion system providing sensations of movement to the crew of three which are representative of the movements of the aircraft during flight. The system is free to move in roll, pitch and heave.

All aspects of the aircraft's operation are faithfully simulated. The noise system provides the sounds of engines, propellers, wind noises, alternators, the noise of tyres on the run-way and of the wheels running over the joints in the concrete taxi-tracks. Even such things as the thud of the undercarriage going down when it is lowered and the difference in noise when the flaps are operated, are simulated.

In addition to full simulation of all the aircraft systems, the Belfast is fitted with the Smiths Autoland system, similar to the one used on Trident aircraft operated by British European Airways. This simulator is the first to be fully equipped for practising automatic landings so that pilots can become familiar with the operation of the aircraft while on automatic flight.

Simulated aircraft failures of the various systems, including the Autoland system, can be introduced at will by the instructor. The aircrew can then practise the emergency procedures necessary to maintain control under all combinations of these failures.

To provide practice in world wide flying the Belfast simulator is provided with a radio navigation simulation system which is capable of producing synthetic signals representing 60 ground radio stations simultaneously. When the aircrew tune from one or other airborne receivers to a ground radio beacon they will receive the proper identification signal and information automatically from the simulator computer. The computer calculates the position of the radio station and of the aircraft and varies the signal strength as the aircraft gets further away from the radio station.

This vital facility is provided by a special purpose digital computer of Miles design and manufacture. To ensure that the coverage is really world-wide a facility is provided for changing the programme of radio stations by the simple insertion of a punched programming tape, thus an exercise which has been concerned with flying in Europe in the morning can be replaced with a North American flight in the afternoon.

The instructors' positions (Fig. 2) are completely comprehensive and in addition to the radio aids and faults selectors, contain all the controls and instruments necessary to efficiently conduct and predefine the exercises. These positions are ergonomically designed and contain automatic devices to record the situations existing at crucial points in the exercise such as landing.

Within the next few months the simulator will be equipped with a synthetic vision system to allow pilots to practise day and night landings at an airfield in visual conditions under reduced visibility. The display to the pilots will be provided by a 3-barrel colour projection television system with projectors mounted under the front of the fuselage and the image back projected on to a screen mounted in front of the pilots' windscreen. An ultra light weight mirror is placed in front of the projectors to reflect the picture on to the back of the screen.

Fig. 1. The Belfast Three-axis Simulator weighing approximately 8000lb

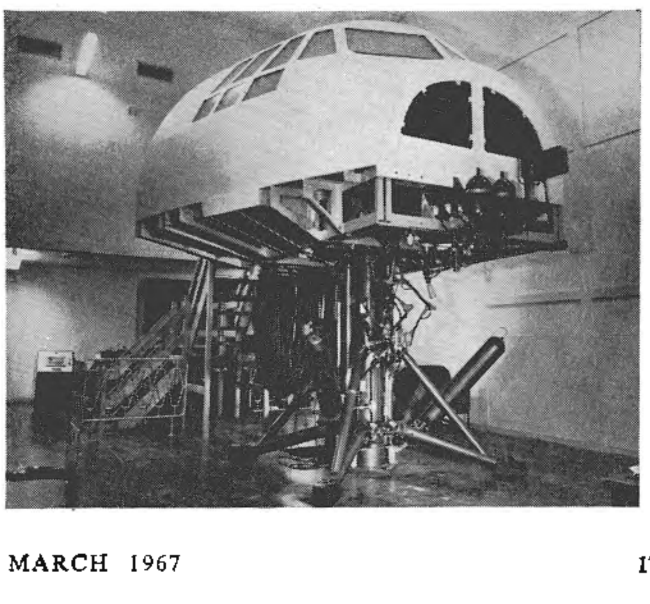
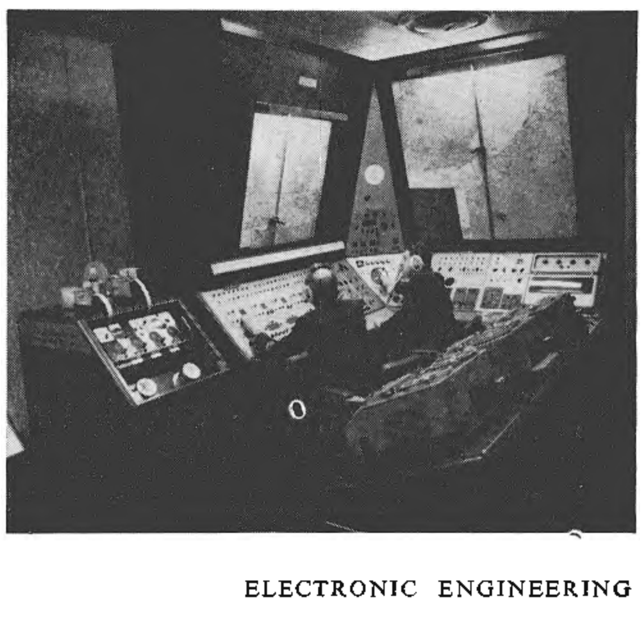


Fig. 2. The Instructors' Control Desk showing the several tape recorders and the maps of the routes of the simulated flight



Determination of Non-linearity by Analogue Simulation

By G. W. Holbrook*, B.Sc., M.Sc., Ph.D., and J. S. Collins†, B.Eng., M.Eng.

The majority of non-linear devices can be described by a power series relating output and input currents or voltages. The coefficients of the earlier and significant terms may be obtained by synthesizing the output waveform of the device by means of a sequence of integrating circuits, using a ramp input signal. The output of each integrator circuit may be used to provide a direct plot of the corresponding coefficient over a range of bias voltages.

(Voir page 203 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 210)

NON-LINEAR devices which exhibit a continuous and single valued relationship between input voltage and output current can generally be described by an infinite power series of the type:

$$I = a_0 + a_1V + a_2V^2 + a_3V^3 + \text{etc.} \quad \ldots\ldots (1)$$

While a number of methods exist for determining this describing power series, the majority of these require a large number of measurements which are usually followed by a tedious calculation. Addtionally, any of these methods which depend upon a d.c. measurement often require that the device be held, over the period of the measurement, in a steady condition which is beyond its normal continuous power rating. This is of particular consequence in semiconductor devices which are susceptible to even small temperature changes above the normal.

Methods[1] which depend upon the measurement of the harmonics, generated with a single sine wave input signal, require a knowledge of the relative phase as well as amplitude of each harmonic with respect to the fundamental. While the amplitude of harmonics may easily be determined, phase measurements of higher harmonics represent a distinct problem.

The following paragraphs describe a method of obtaining the first few significant coefficients of the describing series, directly and under dynamic operating conditions. It will be appreciated that the degree of significance of the coefficients will depend upon the degree of nonlinearity of the device and also on the amplitude of the input voltage which is contemplated. Additionally, the method described below is limited in use to non-linear devices which are resistive in nature and thus exhibit no 'hysteresis loop' in their characteristics.

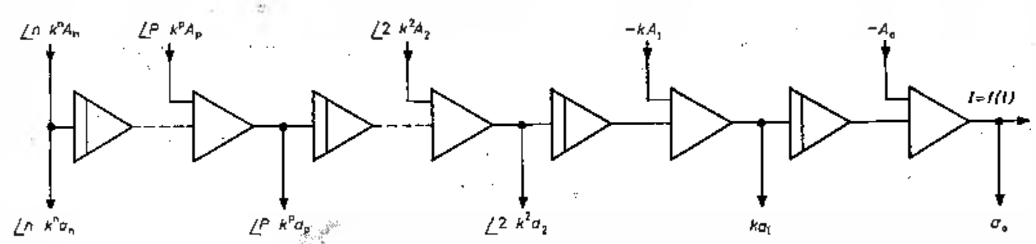


Fig. 1. Arrangement of integrators

Waveform Synthesis

This method depends essentially on synthesizing a waveform which will match the output waveform of the device in question. Synthesis is achieved by a number of successive integrations performed by electronic integrating circuits whose initial conditions can be adjusted to provide an optimum match. In order to integrate voltages and currents, with respect to time, a linear sawtooth signal is used as the input to the non-linear device allowing the input signal to be written as:

$$V = V_0 + kt$$

where V_0 is the input steady-state bias and k is a scaling factor. If equation (1) is rewritten as a Taylor's series:

$$I = \sum_{p=0}^{n} A_p (V - V_0)^p \quad \ldots\ldots\ldots\ldots\ldots (2)$$

where:

$$A_p = 1/P!\, \frac{d^pI}{dV^p}\bigg|_{V=V_0}$$

then:

$$I(t) = \sum_{p=0}^{n} k^pA_pt^p \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (3)$$

where:

$$d^pI/dt^p = P!\; k^pA_p$$

Such a series can be generated using a number of operational amplifiers, connected alternately as adders and integrators as indicated in Fig. 1. At any time, t_1,

$$I(t) = \sum_{p=0}^{n} k^pa_p\,(t - t_1)^p$$

Thus coefficients of equation (1) can be written as:

$$\left.\begin{aligned} a_0 &= A_0 + kA_1t_1 + k^2A_2t_1^2 - - - - k^nA_nt_1^n \\ ka_1 &= kA_1 + 2k^2A_2t_1 + 3k^3A_3t_1^2 - - - - nk^nA_nt_1^{n-1} \\ 2!\; k^2a_2 &= k^2A_2 + 6k^3A_3t_1 - - - - n(n-1)\,k^nA_nt^{n-2} \\ n!\; k^na_n &= n!\; k^n A_n \end{aligned}\right\} \quad \ldots . (4)$$

A comparison of equation (4) with Fig. 1 indicates that the total input to the integrator establishes the value of:

$$P!\; k^pa_p \text{ for } t = t_1$$

Thus, providing that the output of the integrating circuits matches the output of the nonlinear device, when it is driven with a ramp signal, the coefficients of equation (1) may be read off, for any point in the operating region, as the input to the corresponding integrator.

Typical Test Circuit

In order to establish a match between the output waveform of the non-linear device and the waveform as generated by the integrators, a visual comparison of the differences between these two signals may be made on a c.r.o. This is achieved as shown in Fig. 2.

The control device, on which depends the basic timing of the whole system, provides the energizing and reset current of all of the s.p.d.t. relay contacts in the system. Additionally, it provides a triggering pulse for the c.r.o. Amplifier No. 1, connected as an integrator, produces a ramp signal of kt to the non-linear device, via the biasing voltage V_0 and also drives the X axis of the XY plotter

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† *Northern Electric Co. Ltd, Canada.*

directly. Operational amplifier No. 3 compares the output of the non-linear device with the waveform generated by the main integrator chain (amplifiers 4, 5, 6 etc.) and drives the Y axis of the c.r.o.

The outputs of amplifiers 4, 6, 8 etc. are carried to a manual selector which connects each of these, as desired, through potentiometer P_y and amplifier No. 2 to the Y axis of the XY plotter.

With this arrangement simulation is best achieved by adjusting potentiometer P_x until the desired sweep amplitude is achieved at the output of the non-linear device. Potentiometers P_0, P_1 etc. are then adjusted in turn to minimize the vertical deflexion presented on the c.r.o. Successive adjustment of the potentiometers P_0, P_1 etc., with increasing sensitivity of the c.r.o., will allow a precise null to be achieved. The coefficients a_0, a_1, a_2 etc. can then be determined by selecting the appropriate output on the manual selector and observing the plot of the coefficient, with apropriate scaling in potentiometer P_y, as a function of the input voltage, about the bias voltage V_0, on the XY plotter.

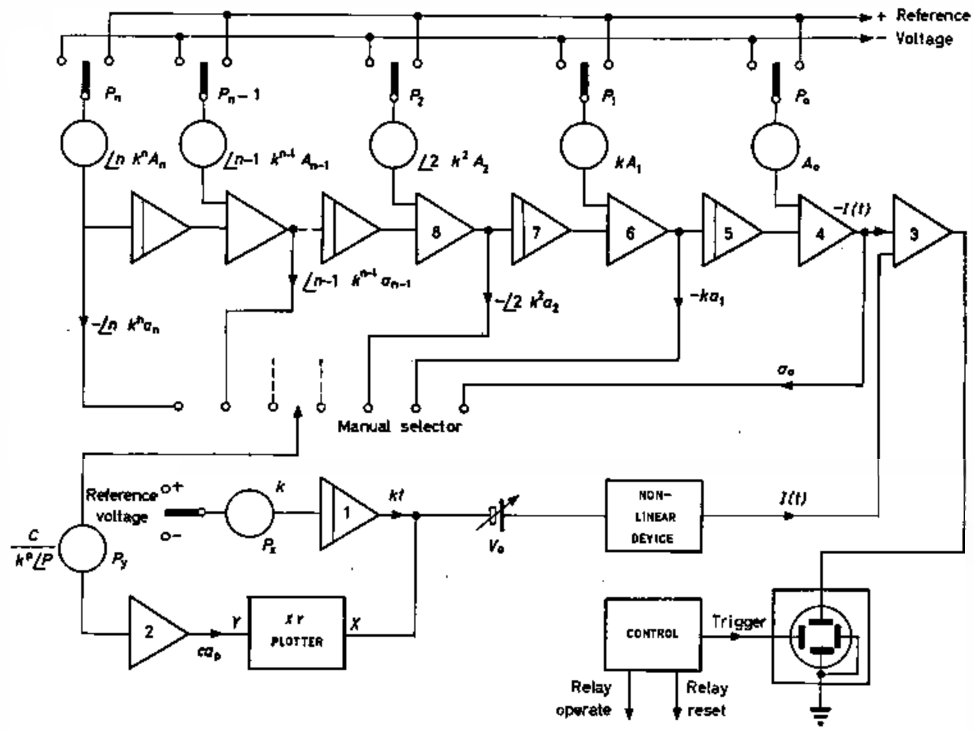
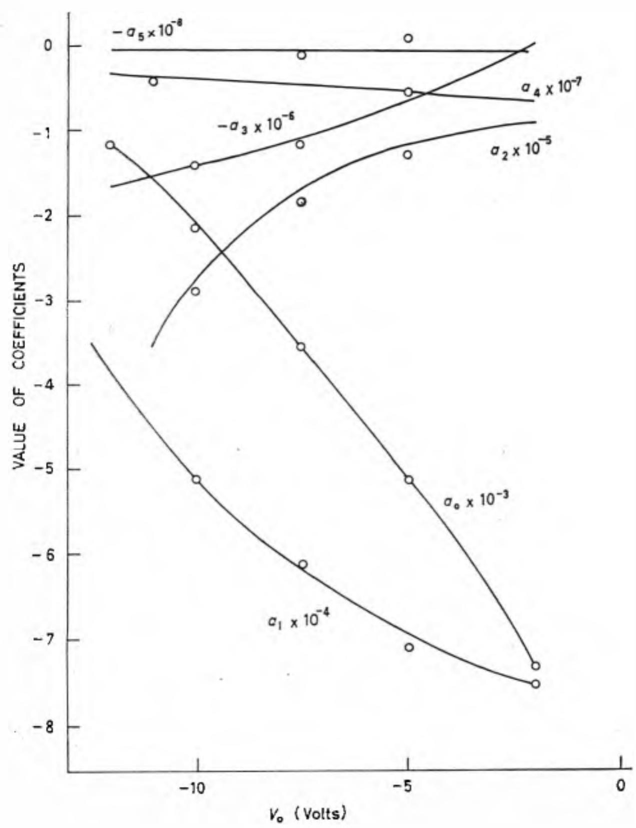


Fig. 2. Complete test circuit

Fig. 3. Typical results for triode amplifier



Practical Results

Fig. 3 illustrates a set of results obtained for a triode voltage amplifier stage using a 6J5 tube with an anode load of 20kΩ. The lines in this figure represent the replotting of the traces of the XY recorder while the marked points illustrate the values of coefficients obtained for the same device using a somewhat exhaustive method of harmonic analysis[1]. In this case, it will be noted that six integrating circuits were used, providing a_0 and the next five coefficients. It should be noted that the dimensions of the coefficients depend upon the order of coefficient and are shown in terms of amps. volts^{-m}, where m represents the numerical order of the coefficient. Additionally, each set of coefficients has its own individual scaling factor.

In this particular case, a time of one second is used for each recurring sweep, thus the value of $k = 10$ volt sec^{-1} and the input signal to the non-linear device is given by V_t volts. In order to prevent an overload of amplifier No. 3, a potential divider with a d.c. bucking bias is employed at the output of the 6J5 tube.

Conclusion

The results illustrated in Fig. 3 indicate that coefficients up to the fifth order can be plotted with an accuracy which is at least equal to that of measuring harmonics. The method is rapid and once the equipment has been established a large number of samples can be tested in a short period of time. It is evident that a similar method, depending upon successive differentiation of the output of the non-linear device, with a ramp input signal, can be devised. However, the inherent noise produced by a succession of differentiating circuits makes this alternative somewhat impractical.

Acknowledgment

The work described in this article was supported by a grant-in-aid of research from the National Research Council of Canada.

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Combination Leading and Trailing Edge Variable Slope Pulse Modulation

By O. E. Kruse*, D. L. Lewis*, and J. K. Jackson†

This article discusses a system of modulation wherein two channels of information are carried by the same train of triangularly shaped pulses; one channel is carried by the slopes of the leading edges and the other by the slopes of the trailing edges.

(Voir page 203 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 210)

CIRCUITS for and analyses of leading edge variable slope pulse modulation and of trailing edge variable slope pulse modulation have been developed and described[1-6]. In leading edge variable slope pulse modulation the intelligence is sampled periodically and is carried in the slopes of the leading edges of a train of triangularly shaped pulses. In trailing edge variable slope pulse modulation the intelligence is carried in the trailing edges of a train of triangularly shaped pulses. This article discusses a system of modulation wherein two channels of information are carried by the same train of pulses, one channel being carried by the slopes of the leading edges and the other channel by the slopes of the trailing edges. The system is illustrated in Fig. 1.

Circuits and Results

The train of trapezoidal pulses carrying the two channels of intelligence is produced by the controlled charging and discharging of a capacitor. This process is illustrated in block diagram form in Fig. 2.

The circuit employed in the modulator is shown in Fig. 3, and is an adaptation of a circuit previously described for the purpose of generating pulses with adjustable linear rise and fall times[7]. The details of the operation of the circuit are given in the reference cited and will not be presented here, although a few words concerning the operation of the circuit are in order. The slope of the leading edge of a pulse is determined by the current into the capacitor C, and this charging current is determined by the collector current in the transistor VT_2. The trailing edge slope of the pulse is determined by the collector current in VT_4. The leading edge of the trapezoidal pulse begins at the time of arrival of the leading edge of the input rectangular pulse. The beginning of the trailing edge pulse is coincident with the trailing edge of the input rectangular pulse.

In the original description of the circuit, the leading and trailing edge slopes were adjusted by changing the resistances above and below A and B, respectively. For the purpose of producing a train of trapezoidal pulses with one channel of information on the leading edges of the pulses and a second channel of information carried on the trailing edges, the circuit is modified above A and below B as indicated in Fig. 3. Capacitor C is charged during the conduction of VT_2 with VT_4 not conducting. The amount of charging current is the emitter current in VT_6. This current is controlled by the 10kΩ potentiometer setting and by the leading edge modulating signal applied between point E and ground. Thus the slope of the leading edge of the voltage pulse across C is linearly related to the voltage of the leading edge modulating signal at the time of sampling. VT_5, operating in an emitter-follower configuration, provides a modulator output preserving the linear relationship between the leading edge slope of a trapezoidal pulse and the leading edge modulating signal voltage at the time of sampling. The sampling time duration is negligible compared to the period of the modulating signal and so the current does not change significantly during the time of rise of the pulse. At the next time of sampling the modulating signal

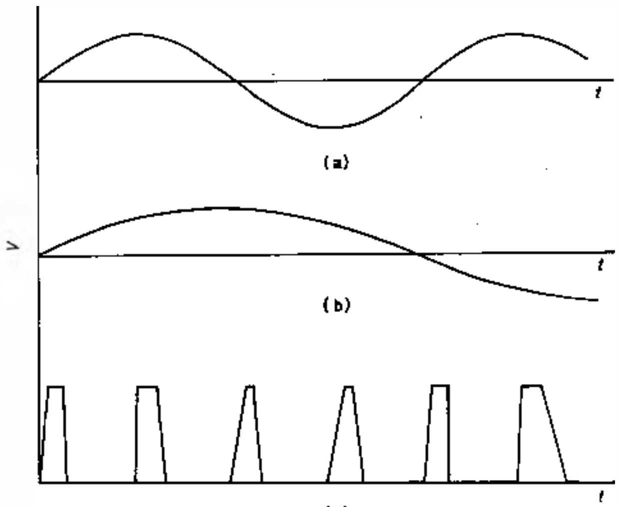

Fig. 1. (a) Leading edge modulating signal
(b) Trailing edge modulating signal
(c) Leading and trailing edge slope modulated trapezoidal pulse train

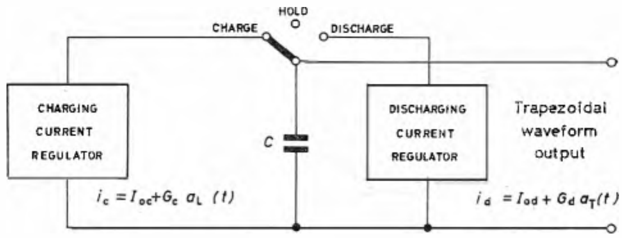

Fig. 2. Modulator block diagram

will have changed appreciably, and so the charging current, and thus the slope of the next leading edge, will be different. Changes in the leading edge slope from the unmodulated value of the leading edge slope will be directly proportional to the modulating signal voltage at the time of sampling. The unmodulated leading edge slope is determined by the quiescent emitter current during conduction of VT_6 and is controlled by the setting of the 10kΩ resistor.

The discharge of the capacitor C begins after MR_4 has clamped the voltage across the capacitor at +3V

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and when the trailing edge of the input rectangular pulse causes VT_4 to conduct and VT_2 to cease conducting. The capacitor discharge current is determined by the setting of the 10kΩ potentiometer in the circuit built around the transistor VT_7 and by the trailing edge modulating signal applied between point F and ground. The discharge current, and consequently the trailing edge slope, remain essentially constant during the discharge of the capacitor because the discharge time duration is negligible compared to the period of the modulating signal. By the next time of sampling the modulating signal will have changed; thus, the next trailing edge slope will be different and will, of course, be linearly related to the modulating signal voltage at the time of sampling. The voltage determined by the setting of the 10kΩ resistor will fix the unmodulated trailing edge pulse slope.

The output of the modulator is, then, a train of trapezoidal pulses with one channel of intelligence carried in the slopes of the leading edges of the pulses and with a second channel of intelligence carried in the slopes of the trailing edges of the train of pulses. The desired pulse repetition frequency is determined by the p.r.f. of the rectangular pulse train at the modulator input and may be altered if desired.

The demodulation of the train of trapezoidal pulses requires first the separation of the leading edge channel from the trailing edge channel and then the appropriate detection of each individual channel. The required steps in the demodulation process are illustrated in Fig. 4. The stretcher and filter have been described[5] and will not be presented here. The only difference between the detectors for the two channels is the use of a simple phase invertor that is required at the input to the trailing edge stretcher and filter for the purpose of inverting the train of negative pulses coming out of the separator.

The separator circuit employed is shown in Fig. 5, and is a simple differentiating circuit. The diodes MR_1 and MR_2 are, ideally, perfect switching devices with zero forward direction resistance and infinite reverse direction resistance. Under these conditions, the separator circuit

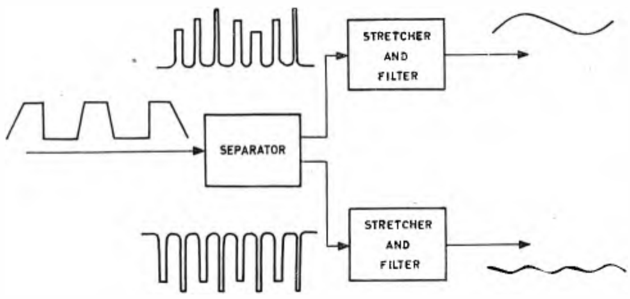


Fig. 4. Demodulation of a leading and trailing edge slope modulated trapezoidal pulse train

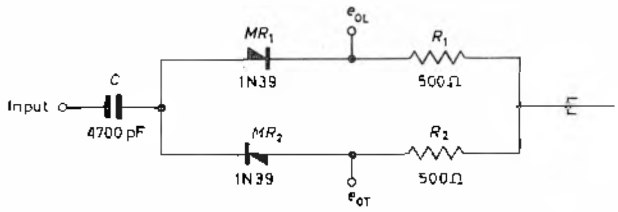
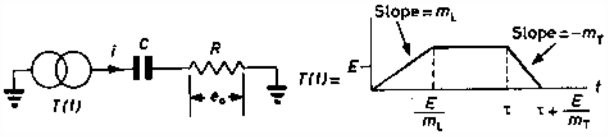


Fig. 5. Separator



$$T = \frac{1}{C}\int_0^t i\,dt + Ri$$

Fig. 6. Equivalent circuit for separator

may be represented as in Fig. 6. Then, by use of Laplace transform methods, it is easy to show that:

$$e = \begin{cases} RCm_L\,[1 - e^{-(t/RC)}] & \text{for } 0 \leqslant t \leqslant E/m_L \\ RCm_L\,e^{-(t/RC)}\,[\exp(E/m_LRC) - 1] & \text{for } E/m_L < t < \tau \\ RCm_L\,e^{-(t/RC)}\,[\exp(E/m_LRC) - 1] + \\ \qquad RCm_T\,[1 - e^{-(t-\tau)/RC}] & \text{for } \tau \leqslant t \leqslant \tau + (E/m_T) \end{cases}$$

By choosing $RC \ll (E/m_L)$ and $RC \ll (E/m_T)$

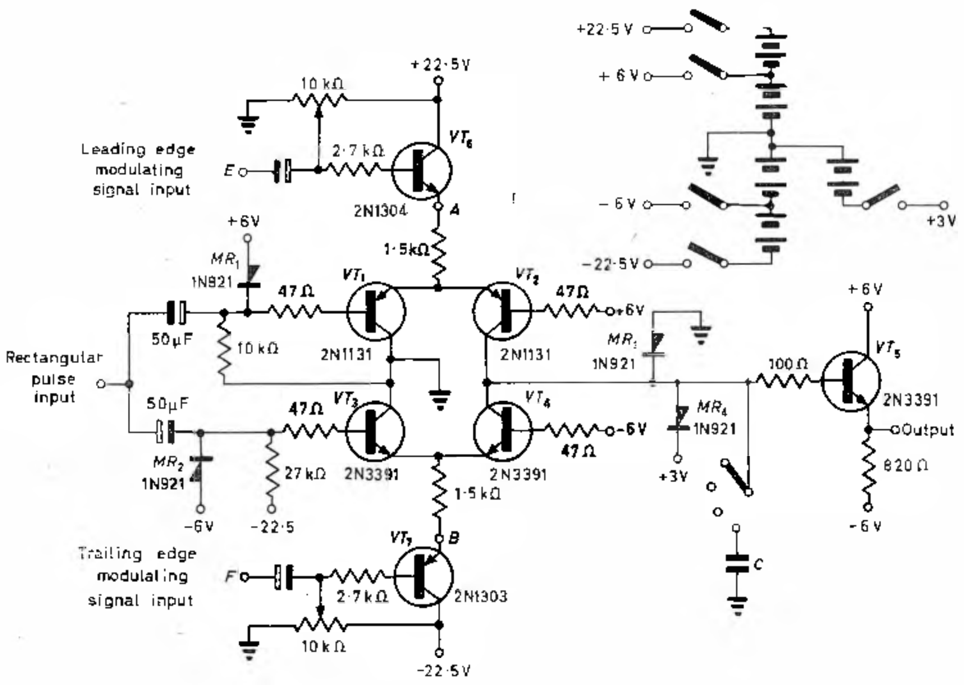
$$e_{oL} \simeq \begin{cases} RCm_L & \text{for } 0 \leqslant t \leqslant E/m_L \\ 0 & \text{otherwise} \end{cases}$$

$$e_{oT} \simeq \begin{cases} RCm_T & \text{for } \tau \leqslant t \leqslant \tau + E/m_T \\ 0 & \text{otherwise} \end{cases}$$

where e_{oL} is the peak value of the leading edge pulse across R and e_{oT} is the peak value of the trailing edge pulse across R.

Thus, the two channels of intelligence put into the leading and trailing edge slopes of the trapezoidal pulse train are converted into two variable amplitude trains, completely separated, and having amplitudes directly proportional to the slopes in the trapezoidal pulse train, and so linearly related to the modulating signal voltages at

Fig. 3. Modulator circuit



the times of samplings. As mentioned, the two variable amplitude pulse trains are then applied to the inputs of two stretcher and filter circuits already described in the literature. The outputs of the stretcher and filter circuits are reproductions of the two original intelligence signals. Fig. 7 includes c.r.o. displays of a leading and trailing slope modulated trapezoidal pulse and waveforms associated with the demodulation processes.

input to channel 1, when the input signals to both channels are of equal magnitude, an intermodulation of considerably less than 1 per cent is observed so long as the train of pulses is kept trapezoidal (not allowed to become triangular). This result indicates an excellent degree of isolation between the two channels, a characteristic of significant importance in possible applications of the system.

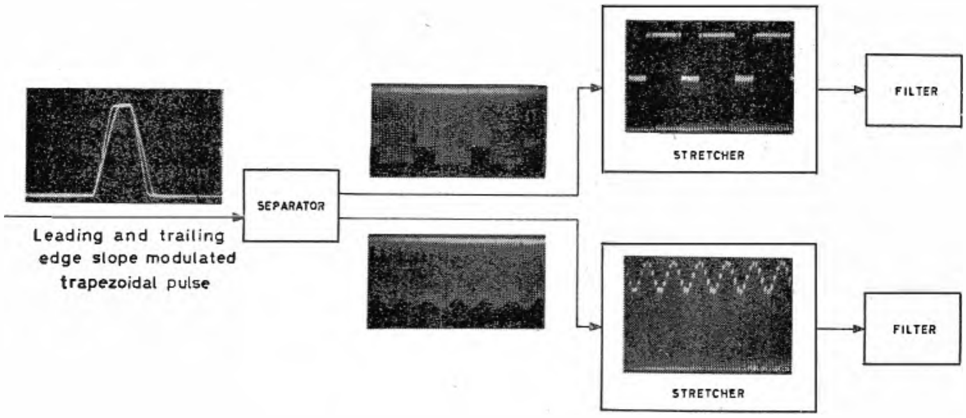


Fig. 7. C.R.O. displays

Conclusion

The system described thus carries two channels of intelligence on a single train of pulses. Response of the system is as good as for the individual leading edge variable slope modulation system and the trailing edge system previously reported in the literature. An important consideration for a system transmitting two channels on the same pulse is that of a crosstalk or intermodulation. Employing a definition of intermodulation as the output in channel 1 identifiable with the input to channel 2 divided by the output in channel 1 identifiable with the

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A Flight Simulator for America

GPS, General Precision Systems Ltd, of Aylesbury, is supplying the first digital simulator with integral visual system ever built as a complete 'package' unit, to American Airlines Inc. for installation at the airline's Fort Worth, Texas, training centre.

The order, worth over £1M to the British manufacturers, includes a BAC One-Eleven simulator, two colour visual flight attachments, and a GP4 digital computer. The second visual will be linked with a Boeing 727 simulator, manufactured by GPS's associated company in the United States.

The whole complex, which is scheduled to enter service early this year, is to be used for transitional and conversion training, as well as research into low weather minima operations. In addition to raising safety standards, it will allow considerable financial savings to be made by reducing in-flight training time.

The One-Eleven simulator incorporates a replica of the airliner's fuselage and cockpit, with all controls, instruments, radio and navigational aids functioning as in the complete aircraft under operational conditions. A three-axis hydraulic motion platform, on which the simulator is mounted, provides representative movement in pitch, roll and heave.

Both colour visual flight attachments incorporate three-dimensional models of Dulles Airport, Washington D.C., and its surrounding terrain. One will operate in conjunction with the 727, and the other with the One-Eleven.

The GPS Colour Visual Flight Attachment (VFA) has been developed to provide added realism for air crew training in flight simulators. It supplies a realistic visual simulation of the selected terrain, to reinforce the sound effects, movements and instrument readings in the simulator. The coverage of the picture and its perspective respond automatically to the movement of the pilot's controls in the cockpit of the associated simulator as he approaches the airport—under a variety of weather conditions—and lands on the runway or overshoots and joins the traffic circuit.

The system consists basically of a moving terrain model, a closed-circuit colour television system, a computer attachment, and controls incorporated within the simulator.

A Marconi colour camera, with an optical system specially developed by GPS, views the terrain model from a position representing that of the pilot's eye. The resulting picture is displayed to the crew under training on a screen situated some three yards in front of the windshield, by means of television projectors mounted above the flight deck.

The terrain model, which is to a scale of 1:2000 representing an area 14·3 miles long by 3·8 miles wide, includes the airfield complete in every detail with buildings, approaches and runway lighting, constructed on an endless belt running on rollers in a vertical gantry. Major features of this system are its special camera mounting and optics which avoid the need to move the camera over the full length of the terrain model; illumination close to the model surface, reducing intensity and power consumption, and eliminating heat dissipation problems; and continuous viewing, without the need for camera resetting. Floor area is also reduced by vertical configuration of the unit.

A Data Logger Scaler and Alarm Limit Comparator

By R. J. S. Ruckman*

The article describes a method of scaling using counting techniques, the scaling taking into account zero offsets, and negative as well as positive values. Also alarm limits are compared with the scaled values at the same time as the scaling is being carried out, and the information as to whether a particular variable is in 'alarm' state or 'no alarm' state can be obtained.

(Voir page 203 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 210)

MANY small system data loggers, whether for telemetery, marine or any other application, require small, flexible scalers. As most typewriters use binary coded decimal (b.c.d.) or 'one out of ten' input codes, less decoding logic would be required for typewriter inputs if the output of the scaler was in b.c.d. At the same time more efficient use is made of telemetery systems if the data is received in binary form, as against b.c.d., and scanning time is saved.

Although calculations using counting techniques usually take longer than other systems, data loggers do not employ very high speed printers or typewriters. On average the print or type rate is 10 characters/sec, and high speed logic can carry out a scaling operation of an accuracy of 0·1 per cent in 2msec.

Zero offset constants are taken into account in the calculation. These may be positive or negative offsets, and are generated by wired in logic, or if they need to be revised, by matrix plugboard or other storage media. The same may be said for scale factors.

The counting technique used in the scaler enables b.c.d. alarm limits to be compared with the variables, by less logic than would be required normally.

All four calculations are carried out in one counting operation, i.e.:

(a) Scaling

(b) Adding or subtracting zero offset constant

(c) Binary to b.c.d. conversion

(d) Alarm limit comparison

The Techniques of Counting Used for the Scaler

The main technique of the system used is to count in two counters, one a pure binary counter and the other a b.c.d. counter, with the same clock and at the same time. When the input binary counter has counted the number of clock pulses which is equivalent to the incoming binary number, the b.c.d. counter is stopped, and the number which is left in this counter is the b.c.d. equivalent of the binary input number.

This is what happens if the scale is 1:1 and no zero offset constant exists. However, if a scale factor is generated, say 1:2, the b.c.d. counter needs to be counted on every second clock pulse, whereas the binary counter is counted on every clock pulse. This will leave the b.c.d. counter holding the equivalent of half the input binary number when the binary counter stops the input to the b.c.d. counter. It is possible for any ratio of full numbers to be translated to scale the output b.c.d. number up or down, although in the circuit used for this article the output b.c.d. is scaled down.

In order that the analogue to digital convertors used in data loggers can be used for most if not all of the analogue inputs, the output scale is standard. In the case of a 10-bit binary analogue to digital convertor (a.d.c.), the usual scale is 0-1023 (0-1111111111 binary). Since it would be most convenient to have a maximum scale output of 0-999 in b.c.d., it was thought necessary for the a.d.c. to produce a 0-999 scale in binary. Although two different clocks could be used; one for the binary counter and one for the b.c.d. counter, unnecessary inaccuracies would result. The output scale of the 10-bit a.d.c. can be modified to 0-999 by increasing the reference supply voltage to the analogue input digital to analogue convertor (d.a.c.) gates by 1023-999/10·23 per cent. The a.d.c. can also be specifically designed to give a binary output of 0-999. In order to avoid the b.c.d. counter counting from 999 back to 000 the counter input is blocked when 999 is reached. Therefore if the a.d.c. gives out a number over 999 the maximum output of the scaler will be shown, assuming that the scale is 1:1. When the output scale is decreased and the binary input to the scaler is more than 1111100111 (999) the output b.c.d. number is proportionally more, according to the scale being used. Normally this should not occur.

If a positive offset zero constant has to be taken into account the b.c.d. counter has to be counted up to this constant value before scaling begins, and before the binary counter starts to count. If a negative offset zero constant exists the binary counter is counted up to the value of the constant before the b.c.d. counter starts to count, so that the final output b.c.d. number is reduced by the value of the constant. The number of pulses which any offset zero constant represents is determined by another counter, and this counter produces the term that starts the counter which is not counting the constant. Whether or not an offset zero constant is used at any time, and if so, whether it is negative or positive, is decided by which scale is being used. Also the values of the constants are generated from the scale terms.

The scale terms themselves are generated from decodes of the addresses for the data logger. Only a few scales may be needed for many addresses.

Operation of Logic Circuits

The main rule for using NOR logic is that *any* inputs of a NOR gate being in a '1' state negative will make the output of the NOR gate a '0' state (0V). If *all* of the inputs to a NOR gate are in the '0' state the output of the gate will be in a '1' state. Hence NOR gates are used, with invertors, as AND logic and OR logic[1]. The counters count up on the counter input going from a '1' state to a '0' state, and also have the normal set and reset inputs. Stores only have the set and reset inputs. (See Fig. 1.)

On inspection of the scaler logic circuits (Fig. 2) it will

* *Formerly Serck Controls Ltd.*

be seen that there are two small counter chains, one at the input to the binary counter, and the other at the input to the b.c.d. counter. These two counters determine, from the scale factor being used, what proportion of the full count is put into the relevant counter. For example, if the scale S_0 is used the input scaling counter will count

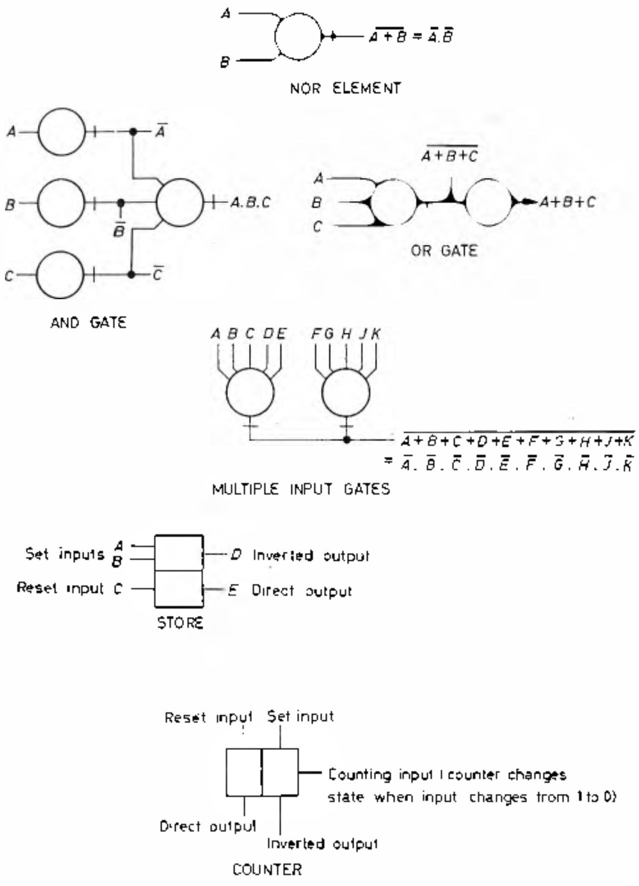

Fig. 1. Logic elements used

up to '001' and be reset by the term $\overline{\text{COUNT 1}}$, while counting up the main input binary counter by one. If the scale S_1 is being used the input scaling counter will count up to '010', and from the setting of the COUNT 1 store from the decode '4$\bar{2}$1' will count up the main counter by one pulse, and reset on $\overline{\text{CLOCK}}$. Therefore, in the first case every input clock pulse was passed on to the main input binary counter, while in the second case every other clock pulse was passed on.

The output scaling counter operates in exactly the same way, but affects the output b.c.d. counter. If one in every two clock pulses are passed to the output b.c.d. counter, and every clock pulse is passed to the input binary counter the effect would be to divide the input binary number by two. In the S_0 scale this is the case, the decodes making the scale factor $001/010 = 1/2$. Assuming that the maximum output scale is $0 - 1000$ b.c.d., the output scale in this case would be $0 - 1000/2 = 0 - 500$ b.c.d. In the case of S_1, it will be seen that the input binary counter decode is '4$\bar{2}$1' and the output b.c.d. counter decode is $\bar{4}2\bar{1}$, making an output scale of $-(2/5 \times 1000) = 0 - 400$. As the input binary counter will be counting at half the rate that it would if the input scaler counter decode were '42$\bar{1}$', it would take twice as long for the main counter to count the number of pulses for the same input binary number. However, assuming a clock frequency of 0·5MHz, even if an awkward scale factor like 4/5 is required, the scaling would take 8msec which is a short time compared with the 100msec required for the typing of one character.

The input binary complement is inserted into the input binary counter by the output a.d.c. register. The scaling process is then started, and the input binary counter is counted up until all of the counters are set in the '1' state, making '$A1$-$A512$' all '1's' and '$\overline{A1\text{-}A512}$' all '0's'. The fact that '$\overline{A1\text{-}A512}$' are all '0', resets the SCALE store, which was first set when the START SCALING term was received. The term SCALE then blocks both the input binary and output b.c.d. counter input gates, thereby inhibiting further counting. The output b.c.d. counter now holds the b.c.d. value of the binary input number, scaled down according to the scale factor which was used. After the b.c.d. output number has been transferred to the typewriter circuits the RESET SCALER term is generated, and the two main counters reset ready for the next SET SCALER and START SCALING signals.

Zero Offset Logic Operation

Zero offset constants relative to the output scale are generated in binary form from wired in logic, plug-board or other forms of number storage. They are generated in binary form so that the binary complement may be inserted into the zero offset counter, and the value counted up in the same way as with the input binary counter.

If a positive offset constant is required for a scale the output b.c.d. counter has to count up to the value of that constant before the input binary counter starts to count. Therefore the scaling itself commences after the b.c.d. counter has counted to the value of the positive offset zero constant. The logic circuit carries out this operation by blocking the input of the binary counter until the zero offset counter, as well as the b.c.d. counter, have counted up the value of the constant.

The sequence of the logical operations carried out for a positive zero constant is as follows, the appropriate logic diagrams being Figs. 3 and 4.

(1) The SCALE store is set when CLOCK is in a '0' state. If a positive zero constant is required, POS. ZERO is generated by the scale factor being used, and the BLOCK store is set. $\overline{\text{POS. ZERO}}$ and BLOCK generate the term BLOCK INPUT, which blocks the binary input counter input gate before any counts have been received from the input scaling counter.

(2) The term $\overline{\text{BLOCK}}$ has opened the zero offset counter input gate, and this counter therefore counts up the pulses from the output scaling counter, the initial counter configuration being the complement of the positive zero offset constant required. This constant is applicable to the final output scale. At the same time as the zero offset counting is proceeding, the same input pulses are counting up the b.c.d. counter.

(3) When zero offset counter holds a full house (all counters set), this indicates that the number of pulses which have been counted is equivalent to the positive offset constant value. Therefore the input binary counter block must be removed, and the scaling proceed as normal. The full house of the zero offset counter sets the store END ZERO COUNT, and this term resets the BLOCK store, therefore removing the term BLOCK INPUT, and permitting the input binary counter to start counting on the

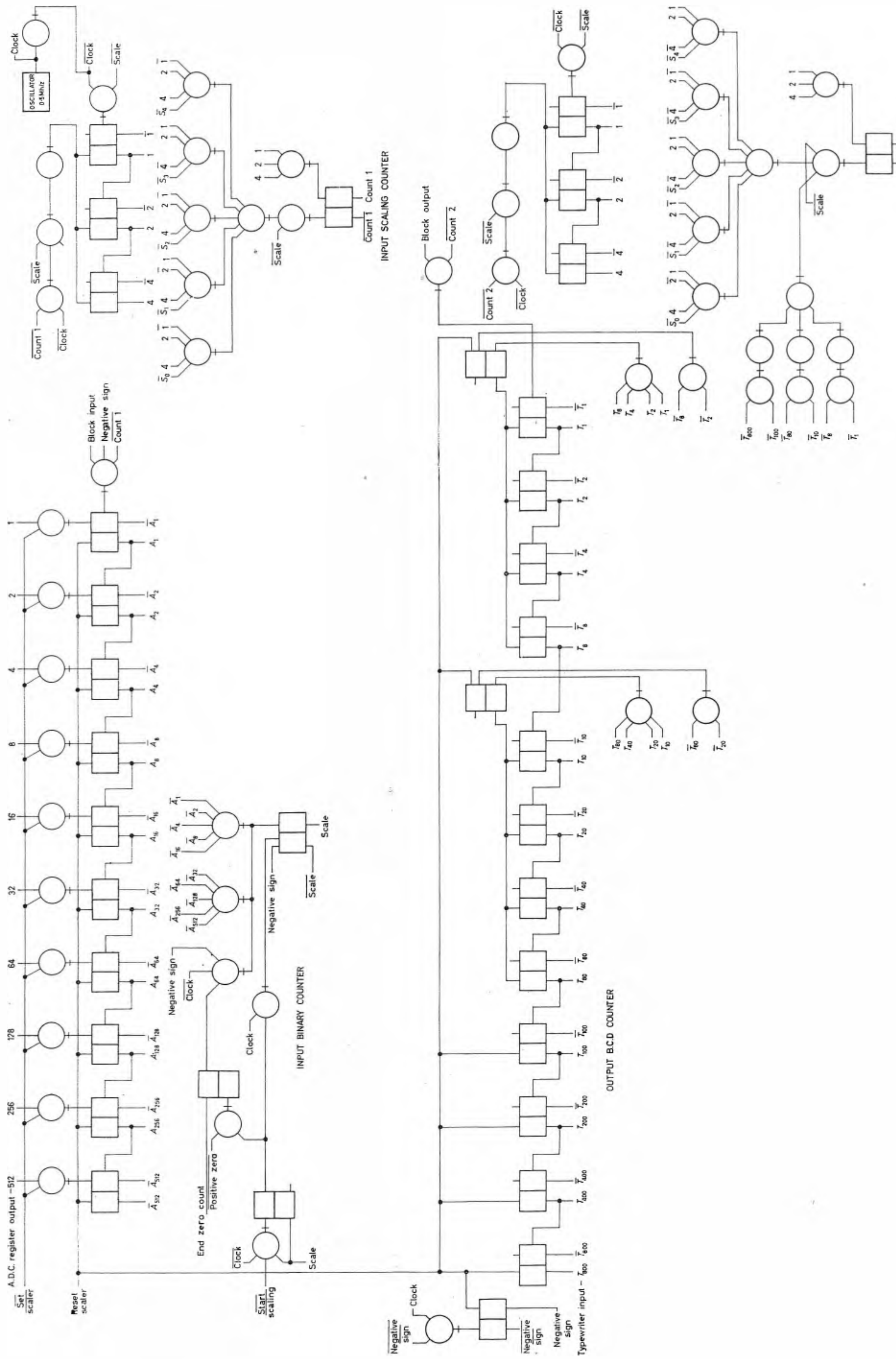

Fig. 2. Scaler counters logic

next pulse from the input scaling counter. The BLOCK term blocks the input gate to the zero offset counter, and resets the same counter. When all the counters have been reset, and CLOCK is in the '0' state, the END ZERO COUNT store is reset.

When a negative zero offset constant is required the output b.c.d. counter has to be held blocked until the case of the above example with a negative zero constant of 100, the SCALE store would be reset before the zero count had ended if the incoming binary number was less than 200. When this happens the output b.c.d. counter block is removed, and this counter and the zero offset counter continue to count until the END ZERO COUNT is generated. Using an incoming binary value of 150, and

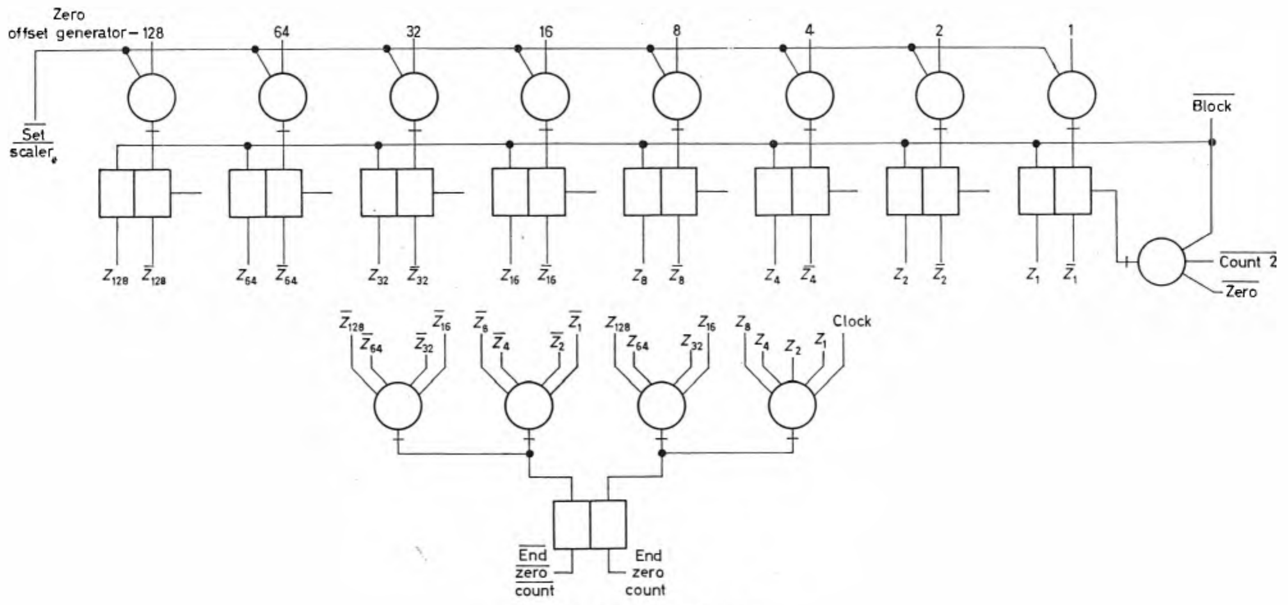

Fig. 3. Zero offset counters

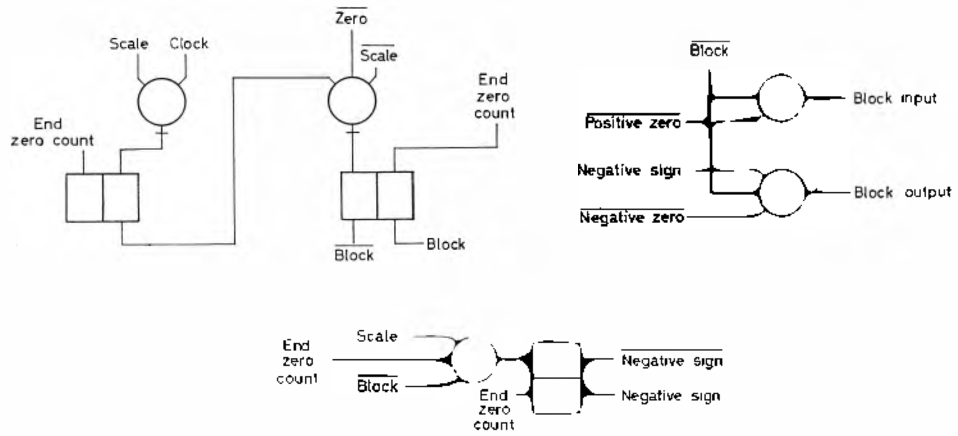

Fig. 4. Zero offset

number of pulses equivalent to the constant has been counted up on the zero offset counter. At the same time the input binary counter must count up the equivalent of the negative constant, relative to its own scale. For example, if a negative output zero constant of (−100) in an output scale of (−100 to + 400) is required, the input binary counter would have to count up (100 × 1000/500) pulses before scaling begins, as the input binary scale is always 0 to 1000. The input of the input binary counter is taken from the input scaling counter which would give the appropriate number of pulses for the input scale, while the zero offset counter input is taken from the output scaling counter.

If, however, the input binary number is a negative number of the output scale, the SCALE store would be reset before the END ZERO COUNT term is generated. In the same output scale as before, an example follows of the logical sequence used when a negative output number is required on the output scale.

(1) Input binary counter set to complement of: 0010010110 (150) i.e. 1101101001. Zero offset counter set to complement of: 01100100 (100) i.e.: 10011011.

(2) The SCALE store is set and NEGATIVE ZERO generated by the scale factor. The BLOCK store is set, and the BLOCK OUTPUT term generated by NEGATIVE zero and BLOCK rule. This term blocks the output b.c.d. counter before the counting starts.

(3) The zero offset and the input binary counters are counted up until 150 pulses have been counted up on the input binary counter, when the SCALE

	Input Binary Counter										Zero Offset Counter								Output B.C.D. Counter											
	512	256	128	64	32	16	8	4	2	1	128	64	32	16	8	4	2	1	800	400	200	100	80	40	20	10	8	4	2	1
Set Scaler	1	1	0	1	1	0	1	0	0	1	1	0	0	1	1	0	1	1	0	0	0	0	0	0	0	0	0	0	0	0
Number of Pulses Counted	0	0	1	0	0	1	0	1	1	0	0	1	0	0	1	0	1	1	0	0	0	0	0	0	0	0	0	0	0	0
SCALE Store Reset	1	1	1	1	1	1	1	1	1	1	1	1	1	0	0	1	1	0	0	0	0	0	0	0	0	0	0	0	0	0
Number of Pulses Counted	0	0	0	0	0	0	0	0	0	0	0	0	0	1	1	0	0	1	0	0	0	0	0	0	1	0	0	1	0	1
END ZERO COUNT Store Set	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1	0	0	0	0	0	0	1	0	0	1	0	1

Fig. 5. Negative numbers sequence diagram

store is reset. In the meantime 75 pulses have been received at the zero offset counter input, as the output scale span is half of the input scale span. The BLOCK store has been set and not reset since setting, and also the SCALE store has just been reset so that the NEGATIVE SIGN store is then set. This store sets the SCALE store, blocks the input binary counter, and the BLOCK OUTPUT term goes to a '0' state so that the output b.c.d. counter can start counting. The term NEGATIVE SIGN also sets the NEGATIVE store when CLOCK goes to a '0' state.

(4) After a further 25 pulses have been counted up on the zero offset counter, this counter will then hold a full house which will set the END ZERO COUNT store. This store then resets the BLOCK store and also the NEGATIVE SIGN store, thereby making the SCALE store reset. Therefore all input counter gates are blocked, leaving the b.c.d. counter holding the number 25 which was counted up at the same time, and from the same scaling counter, as the zero offset counter. The END ZERO COUNT store is reset when all the counters of that chain are reset, and CLOCK is in the '0' state.

For the above example the contents of the three main counters over the operation period can be seen in the NEGATIVE NUMBERS SEQUENCE DIAGRAM (Fig. 5).

If a negative zero offset is required, but the input binary counter holds a number which would require the output b.c.d. number to be positive, then the SCALE store would not be reset before the BLOCK store, and therefore the NEGATIVE SIGN store would not be set. The output b.c.d. counter would start to count as soon as the END ZERO COUNT store had been set, and blocked when the SCALE store is reset.

If the input binary counter holds a number which would require the output b.c.d. number to be 000, and a negative zero offset is present, then the NEGATIVE SIGN store would not be set and no count up of the b.c.d. counter would start when the END ZERO COUNT store was set.

In the event of the input binary value being zero, the complement of zero will be inserted into the input binary register. Normally, without a zero offset being present, the SCALE store would be set from $\overline{\text{START SCALING}}$ and then reset straight away as $\overline{A1}$-$\overline{A512}$ would all be in the '0' state. When a zero constant is present, however, some counting has to take place in spite of the input being zero.

In the case of a positive zero constant the b.c.d. counter has to count up the constant before the SCALE store is reset. As can be seen from the logic circuit of the input binary counter the term $\overline{\text{POSITIVE ZERO}}$ is gated with the START SCALING signal to set a store, which forces a '0' state to the reset side of the SCALE store until END ZERO COUNT is generated. If the input binary number is more than zero the SCALE store will not reset immediately after zero count has finished.

In the case of a negative zero constant the BLOCK store is set in the usual way, but when the SCALE store is reset by the input binary number being zero the NEGATIVE SIGN store is immediately set and forces the SCALE store to set again while blocking the input binary counter counting input. Therefore the negative zero constant will be counted up on the b.c.d. counter and the negative sign store set, and the SCALE store reset with the NEGATIVE SIGN store being reset by END ZERO COUNT.

Alarm Limit Comparison

When this binary to b.c.d. scaler is used in a data logger, and the output b.c.d. values are to be inspected for being in alarm condition (i.e. outside certain safety ranges), then coincidence gates can be used to compare the b.c.d. alarm limit values with these output b.c.d. values. The principle used is for a store (COINCIDENCE store) to be set if and when the output b.c.d. value is the same as the alarm limit value. This inspection is carried out at the same time as the scaling is being carried out, and while the output b.c.d. counter is being counted up. Therefore the COINCIDENCE store shows, at the end of the scaling of a variable, whether or not that variable value is above or below the appropriate alarm limit, and so helps to decide whether that value is in an alarm state or not (see Fig. 6).

Other factors need to be considered before the final result, i.e.

(a) Whether the alarm limit is high or low

(b) The alarm limit polarity (negative or positive)

(c) The variable polarity (negative or positive).

For example, if the variable is found to be positive, and the alarm limit high and positive, then if coincidence occurs ($\overline{\text{COINCIDENCE}}$ in '0' state) there will be an alarm state, since the variable concerned has passed the value of the high limit and is above the safety range. If, however, the alarm limit was low the COINCIDENCE store being set would indicate a no-alarm state, as the variable value was found to be above the minimum safety limit. The eight combinations of the states of the alarm limits and variable polarity are given in the 'table of alarm conditions,' (Fig. 7). The first four combinations show which state the COINCIDENCE store should be in for an alarm to occur for those combinations. The second four combinations will be seen to have differing polarities, and it can be deduced that in each of these combinations the state of the COINCIDENCE store has no effect on the alarm condition. It will be seen that the first six combinations in the table give alarm states, and it is therefore these six combinations which are decoded to set the ALARM store, should

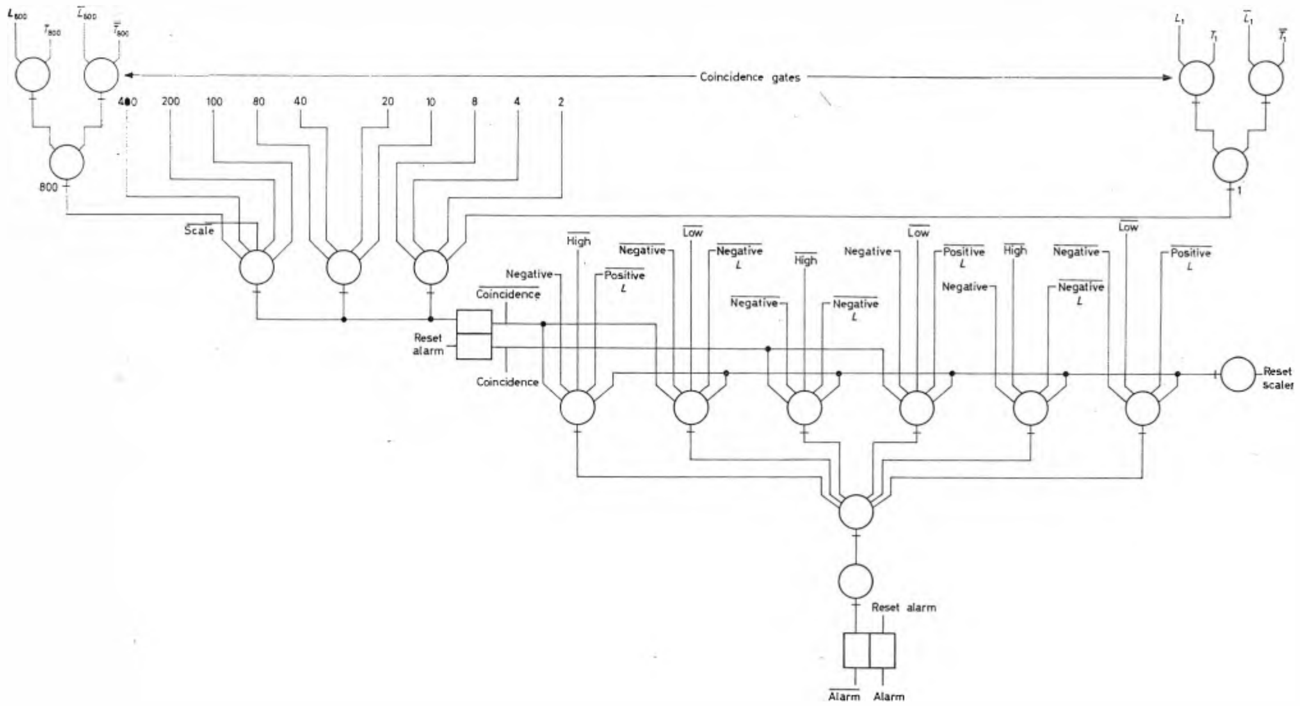

Fig. 6. Alarm limit comparator

HIGH OR LOW ALARM LIMIT	ALARM LIMIT POLARITY	VARIABLE POLARITY	STATE OF COINC. STORE	ALARM OR NO ALARM
High	Positive	Positive	Set	Alarm
Low	Negative	Negative	Set	Alarm
High	Negative	Negative	Reset	Alarm
Low	Positive	Positive	Reset	Alarm
High	Negative	Positive	—	Alarm
Low	Positive	Negative	—	Alarm
High	Positive	Negative	—	No Alarm
Low	Negative	Positive	—	No Alarm

Fig. 7. Table of alarm conditions

any of them be present. The time at which any one of the six combinations being present can set the ALARM store, is after the scaling has been finished, and when the scaler is reset. Before this has happened the COINCIDENCE store has detected a coincidence if there has been one, and is in the correct state ready for inspection, should one of the first four combinations be present.

The alarm limit terms $L1 - L800$ and $\overline{L1} - \overline{L800}$ are usually generated by b.c.d. coding logic, this coding logic being driven by plug-boards arranged into three sets of 10 columns, one for the hundreds, one for the tens and one for the units. When a particular row of the plug-board is driven, the alarm limit is generated by a diode plug being:

(a) In one of ten holes (0 — 9) for the units

(b) In one of ten holes (00 — 90) for the tens

(c) In one of ten holes (000 — 900) for the hundreds.

Other columns will be available for indications as to whether the alarm limit is:

(a) High or low

(b) Positive or negative.

Other storage media can be used for alarm limits, but it has to be of such a nature as to permit quick revision.

The alarm limit b.c.d. configuration is compared with the b.c.d. output counter configuration by 12 pairs of coincidence gates, one of every pair inspecting the direct outputs, and the other one of every pair inspecting the inverted outputs. If at any time during scaling either one of every pair of coincidence gate outputs is in a '1' state, then coincidence has occurred, and the COINCIDENCE store is set, ready for consideration in the final alarm state.

Conclusion

Although few scale factors were used in the logic circuit shown the range can be extended by increasing the number of input or output scaling counters. Increasing the number of input scaling counters means that if all the counters are used in a scale, the time taken for the scaling will also increase. If it is decided that the time limit for one scaling operation is the time taken for one character to be typed, say 100msec, then the maximum pulses that can be counted up on the input scaling counter is 50, assuming a clock speed of 0·5MHz. The b.c.d. output of the scaler can be not only a fraction but also a multiple of the binary input number, but this facility would require extra counters for the b.c.d. counter chain.

This scaler could be used for applications other than data logging, and some of these might require inputs other than binary and outputs other than b.c.d. If a b.c.d. input counter is used, the binary complement system is, of course, not effective and coincidence gates would have to be used for the input counter and the offset zero counter.

If two alarm limits are required to be considered for one variable (i.e. high and low), the scaling operation could be carried out twice for the same variable, and using another row on the alarm limit plugboard. If time is the important factor the alarm limit comparator logic circuit can be duplicated, one circuit being used for the high limits and the other for the low limits. In this case the alarm plugboard would have to be enlarged to take two alarm limits and their polarities on each row for each variable requiring two limits.

A data logging system employing the alarm comparator would usually have continuous alarm scanning facilities and the scaler would be in use all the time, as well as the alarm comparator. When actual data logging was in progress the alarm comparison programme would have to be interlaced with a data logging sequence, unless completely separate scalers were used for both facilities, and storage facilities provided for the data logging.

Acknowledgments

The author wishes to thank the Directors of Serck Controls Ltd for their permission for the article to be published, and to his colleagues for their help, in particular to Mr. John Bishop on whose original principle the article is based.

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Domain Originated Functional Integrated Circuits

Work has reached an important stage at Standard Telecommunication Laboratories Ltd, Harlow, on a completely new concept of electronic circuit function.

In current practice, circuit requirements are met by complicated networks of resistors, capacitors, semiconductors and associated interconnexions. Even when these are reduced to the form of microcircuits, they remain complex, costly and have inherent reliability problems due to interconnexions.

Work at STL by a team headed by C. P. Sandbank has produced a method of synthesizing complex electronic functions in a single solid state bulk effect device having only a few operational connexions.

DOFICS (Domain Originated Functional Integrated Circuits) make use of basic bulk effects in semiconductor material —in particular that it is possible to produce stable high electric field "domains" moving over a distance that is long compared with the width of the domain. These domains are launched when an applied voltage exceeding a certain threshold value is applied across samples of certain materials. Domains may be formed by virtue of various phenomena, including field dependent trapping, electron-phonon coupling, and intervalley transfer of hot electrons from lower to higher effective mass states. At STL, most of the work has been in the last two categories, in which domains travel at velocities up to 10^7 cm/sec.

If, when the domain moves through the material, it encounters changes in conductivity due to doping or changes in cross-sectional area, the current through the sample changes also. Thus, by specifying the conduction path in these or other ways, an output current waveform of almost any shape may be produced. Such conductivity profile shaping thus constitutes a static characteristic for the device.

Perhaps of even greater significance is the fact that dynamic control of the device is possible (i.e. while a domain is in motion) by varying the instantaneous bias. Thus, a domain can be arrested at any point along the drift path. Furthermore, the point along the path at which the domain is removed can be made, for example, proportional to the applied bias, by introducing an overall slope or taper to the profile.

Using these features, a simple analogue-to-digital convertor has been fabricated, which demonstrates the principle, i.e. a device in which the number of output current pulses is proportional to the applied bias. Although a 64-level, 6-digit coder using a 2mm long drift path in GaAs seems quite feasible, this work is still in the research phase and much remains to be done to develop the necessary materials and device-fabricating techniques.

The techniques described above realize the output waveform by monitoring the total current through the device. In addition, however, the high fields due to the domain can readily be picked up by electrodes placed sufficiently near to the surface of the semiconductor. Thus, one may have a completely separate electrode system bearing any additional code profile, etc., electrostatically coupled to the domain. Then, the sweeping domain will produce output potentials as it moves past the electrode, the latter being shaped to produce any desired time spacing of the output pulses.

SHORT NEWS ITEMS

The Ministry of Technology has placed a contract valued at £195 000 with the University of Edinburgh for the development in the Regional Computing Centre of master programmes for a large scale conversational computer system for a wide variety of users.

The University has ordered a System 4-75 computer from English Electric-Leo-Marconi Computers Ltd who are to provide technical support for the project. Programming staff of the company have joined the University software team.

The total cost of the project, with English Electric-Leo-Marconi's contribution, will be £350 000. Of the amount provided by the Ministry's contract, £125 000 is for the software development while £70 000 will cover the cost of additional hardware.

Initially, the aim is the establishment of a computing network primarily to serve University and Research Council activities in the Edinburgh area, but industry will be invited to use this facility to the extent of perhaps 20 per cent of capacity, and as the system develops access will be extended to more remote users. Research at the centre will include investigations into programming techniques and the study of complex computing systems, and may be expected to lead to important commercial and industrial applications.

The Paul Instrument Fund Committee, composed of representatives of the Royal Society, the Institute of Physics and the Physical Society and the Institution of Electrical Engineers has made grants as follows:

£4 132 to Dr. R. D. Davies, senior lecturer in radioastronomy, University of Manchester, to enable him to extend from 120 to 192 the number of frequency channels on the spectrometer for radio astronomy employing digital auto-correlation techniques, for the construction of which he received a grant of £7 800 in December 1964.

£500 to Dr. W. J. Jones, of the University Chemical Laboratory, Cambridge, and Fellow of Trinity College, Cambridge, for the purchase of equipment to enable him to increase the efficiency of the laser apparatus for the excitation of Raman Spectra, for the construction of which he received a grant of £2 500 in April 1965.

The British Standards Institution has published a supplement to B.S. 204 *Glossary of terms used in telecommunications (including radio) and electronics* bringing up to date the section dealing with *Colour television terms.* It includes many internationally agreed terms and definitions which were not available when B.S. 204 was issued in 1960. The supplement, which supercedes the Sub-section 742 of B.S. 204, was prepared before the colour television system for the U.K. was chosen. As a result it contains a few terms which are no longer used and some new terms necessary in the operation of the chosen system are omitted. However, the appropriate revision will be made in due course. It was considered preferable to make the existing terms available as early as possible rather than delay publication of the supplement.

Copies of Supplement 3 to B.S. 204 may be obtained from B.S.I. Sales Department, Newton House, 101/113 Pentonville Road, London, N.1. Price 6/- each (postage 9d extra will be charged to non-subscribers).

The Science Research Council has made a grant of £45 000 to Manchester University towards the cost of a design study for a new radio telescope. This is an essential preliminary to consideration of one of the principal recommendations made in the Report of the Radio Astronomy Planning Committee (The Fleck Report 1965) which recommended the provision of a new large radio telescope for the use of the Manchester University radio astronomers who are headed by Professor Sir Bernard Lovell, F.R.S.

The telescope, referred to as the Mk. V, will be a large fully steerable parabolic reflector. It is the latest concept of the instrument recommended in the Fleck Report and represents an agreed development from the Mk. IV and the Mk. V telescopes which were discussed at that time. Alternative design concepts will be explored before a particular design is chosen for a detailed study.

The Science Research Council has invited the United Kingdom Atomic Energy Authority, and the Authority has agreed, to act as agents for the design study. The Authority will be responsible for collecting the data required for evaluation of the possible design concepts, for commissioning and assisting in the detailed design study and for the preparation of cost estimates.

The Mk. V will be used individually and in association with the existing telescopes at Jodrell Bank in researches which should clarify understanding of the early history and evolution of the universe, of the stars, galaxies and planetary system.

A series of thyristors with a turn-off time of less than 3μsec has been introduced by the International Rectifier Company at Oxted, Surrey.

These reverse-blocking triode thyristors (types 8RCU and 10RCU) have been designed to cover seven voltage grades from 50 to 600 peak reverse voltage/peak forward voltage.

Current ratings are 8A full cycle average for the 8RCU and 10A for the 10RCU.

Prior to the introduction of these new devices, the development of high frequency invertors and heaters, ultrasonic equipment and a.c. to d.c. convertors has been restricted by the dynamic characteristics of the earlier thyristors.

With this new thyristor it is expected that the design will be simplified by the reduced number of components and by the less complicated circuits.

The Ministry of Aviation has placed a contract, valued at £5 000 per year, with the Department of Electrical and Electronic Engineering at the University of Leeds for a multi-discipline research effort into air traffic control systems.

A Pegasus computer and an orthodox airfield control radar system were installed in the University in 1965 and interface equipment is now complete.

With this equipment it is hoped to automate aircraft detection by radar as far as possible, leaving control tower staff to supervise the whole traffic picture.

Soon the radar will be tracking the bulk of scheduled flights in and out of the Leeds-Bradford Airport to collect statistical details about the accuracy of present-day air traffic guidance. The objective here is to determine more accurate touch-down times as soon as an aircraft appears on the radar screen. The differing behaviour of each type of aircraft will be recorded.

Elliott-Automation has been appointed by the European Launcher Development Organization (ELDO) as the prime contractor for the development of the inertial guidance system for 'Europa,' Europe's first international satellite launching rocket. Co-operating with Elliotts as sub-contractors are Ferranti Inertial Systems Department in Scotland, N.V. Van der Heem Electronics in the Netherlands, Fabbrica Italiana Apparechi Radio (FIAR) in Italy, Fluggerätewerk Bodensee GmbH (FGW) in Germany and the National Aerospace Laboratories (NLR) in the Netherlands. The Royal Aircraft Establishment at Farnborough is responsible for system accuracy and guidance laws and will therefore be collaborating closely with Elliott-Automation.

The necessary development and flight systems initially involve the supply of seven Elliott MCS 920M microminiature digital computers.

Each system will incorporate an Elliott MCS 920M microminiature digital computer, which will communicate with the platform, autopilots and the telemetry through an interface unit provided by Van der Heem Electronics. It will process attitude and acceleration information generated by a Ferranti lightweight inertial platform, a development of a system recently successfully launched at Woomera. The combined equipment will provide steering information to the vehicle autopilots. FIAR will provide the power supply, FGW certain ground equipment and the system test facility will be provided by NLR

The inertial system will provide self-contained guidance for Europa launch vehicles without the need for ground control during flight, and will first be flown in the ELDO F9 and F10 vehicles to be launched from Woomera during 1969.

Elliott-Automation has also been awarded two further contracts for electronic equipment for 360-ton C-5A heavy logistics transport aircraft being developed by the Lockheed Georgia Company for the United States Air Force Systems Command. The two latest contracts bring the total value of the work being done by Elliotts on this aircraft to over £1½M ($4·5M). The orders are for the development and production of equipment for the first 58 C-5As ordered for the USAF.

The first of the two new contracts is for an electronic control system to swivel the four main six-wheel bogies and the nosewheel of the giant transport aircraft to allow it to land without correcting for drift caused by cross-winds, and to swivel the bogies to facilitate tight turns on the ground. The system must never allow the bogies to become misaligned, even when turning tightly on rough ground.

The second new contract is for a computer to tell the crew at any time during flight how far they can fly, at what height and speed they will achieve the best range or endurance and what effect changes of height or speed will have on range capability. The systems are called Landing Gear Cross-wind Positioning, Controlled Castering and Indication System, and Energy Management Analog Computer (EMAC).

Miles Electronics Ltd of Shoreham, Sussex, has added colour television equipment to its flight simulators to give a more realistic visual aid for flight training.

The colour television equipment, manufactured by Peto Scott Ltd, uses the industrial Plumbicon colour television camera which is mounted on a mobile gantry above model countryside encompassing 9 by 5 nautical miles, and scaled down to 1200:1. An optical unit fitted under the camera focuses the image on the face plates of the television camera, and at the same time controls the various flight attitudes, such as height, roll, pitch and heading.

A local viewing colour monitor screen, mounted on the fuselage by the co-pilot's or pilot's window, and a colour projector mounted on top of the simulator fuselage, are both manufactured by Peto Scott.

The projector throws its image on to an angled screen 6ft by 8ft in size.

The fuselage, colour projector and screen are all mounted on a hydraulically operated base giving three axis fuselage motion and this, with flight noises fed into the cockpit via loudspeakers, simulates complete visual and oral flight conditions.

Varying weather conditions, as well as day and night flying conditions are also simulated.

Standard Telephones and Cables Ltd has received an order valued at £210 000 from the McDonnell Aircraft Corporation for the supply of the STR 70-P radio altimeter to be installed in the McDonnell Phantom F-4K and F-4M aircraft for the Royal Navy and the Royal Air Force.

These altimeters replace the American manufactured types which are fitted as standard equipment in the Phantom.

It is expected that this will be the first of a number of orders to be received from McDonnell for the STR 70-P which is already scheduled for numerous military aircraft including the Lockheed Hercules C.130K, HS.801 Maritime Comet, Sikorsky SH-3D Sea King helicopter, Andover, Wessex and General Dynamics F.111.

The STR 70-P, the world's first fully solid state microminiature radio altimeter has been wholly developed by STC's Radio Division. It is frequency modulated equipment employing monolithic integrated circuits to ensure maximum reliability. Designed for low level operation its capability includes terrain following, low level strike, low approach, automatic landing and auto-hover.

The Department of Geodesy and Geophysics at Cambridge University is carrying out experiments using new equipment to discover the different formations of the earth's crust under the sea in the English Channel.

Two prototype battery operated tape recorders together with gain level control units have been supplied by Royston Instruments, Byfleet, for this underwater research.

The equipment is enclosed in a water-tight buoyant sphere and dropped to the sea bed to a maximum depth of 100 fathoms by means of a heavy weight.

Previous experiments were carried out with a recorder suspended by a surface buoy, but wind and current affected the accuracy of readings.

In the present experiments, shock-waves produced by minor explosions fired by the team, are picked up by transducers mounted in the sphere and recorded on four track magnetic tape produced by Scotch Instrumentation (Type 499).

At the end of the experiment the sphere is released from its anchor weight by chemical action, set in motion by a pre-set clock which eats through a copper wire fixing the sphere to the weight.

On recovery, the tape is taken back to the Department and played at a higher speed and information extracted.

Underwater the recording speed is a quarter of an inch per second, and the 500ft of tape records for about six hours.

International Computers and Tabulators Ltd has endowed a Fellowship in the Group of Traffic Studies at University College London, in order to study the possibility of formulating a mathematical model for the area control of road traffic by digital computer.

At the moment, most computer control systems are designed to operate at a more sophisticated, but basically similar, level to that of electro-mechanical devices, such as are used to link traffic signals for green-wave progressions. Simple examples have been in use for some years at such places as Slough and Oxford Street, London, where the progression is along a single street.

A more extensive system exists in Cologne, West Germany, where a hierarchy of electro-mechanical controllers maintain green waves over whole areas. Different types of progression are introduced at different times of the day by switch control from a display room at the Headquarters building.

The use of a single central computer is fast becoming standard practice in area control, one of the first being the installation at Toronto, Canada. This type of centralization offers the advantage that the control procedures are easily amended and kept up to date.

LETTERS TO THE EDITOR

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

A Binary to Decimal Convertor

Dear Sir,—I was extremely interested in the binary to decimal convertor described by Mr. Dean in your October issue. Its modus operandi, however, was very difficult to appreciate, and so a few trial examples were worked. It then appeared that not all numbers were successfully evaluated by the proposed logic scheme. For example:

							1	1	1	0	1	1	
						1	1	1	0	1	1		
					1	1	1	0	1	1			
				1	1	1	0	1	1				
			1	1	1	0	1	1					greater than nine, ∴ add six
		1	0	1	0	0	1	1					
	1	0	1	0	0	1	1						
1	0	1	0	0	1	1							i.e. 53 in b.c.d.

which is different from the original binary number, 59.

The reason for this is as follows: Consider a b.c.d. number, of which only one cell (i.e. one decimal digit) is shown.

—	—	A	0	0	0	—	—
	$2^0.10^m$	$2^3.10^{m-1}$	$2^2.10^{m-1}$	$2^1.10^{m-1}$	$2^0.10^{m-1}$	$2^3.10^{m-2}$	where A equals 0 or 1

Let us assume that this cell represents the m^{th} decimal digit, in which case the weights to be ascribed to the various positions are as shown. If this number is now multiplied by two, corresponding to a shift to the left of one place in pure binary, the value of the bit A changes from $A.2^3.10^{m-1}$ to $A.2^0.10^m$, and not to $A.2^3.10^{m-1}.2$ as is required. This represents a deficiency of $A.2^3.10^{m-1}.2-A.2^0.10^m = 6.10^{m-1}.A$.

It is therefore necessary to add six into the cell every time that a '1' in the fourth location crosses the cell boundary during a shifting operation. This is, of course, in addition to the necessity for adding six every time that the numbers in a cell exceeds nine, as described by Mr. Dean.

Implementing this rule in the previous example now gives:

							1	1	1	0	1	1	
			1	1	1	0	1	1					add six because greater than nine
		1	0	1	0	0	1	1					
	1	0	1	0	0	1	1						
1	0	1	0	0	1	1							add six because cell boundary crossed
1	0	1	1	0	0	1							i.e. 59 in b.c.d.

which is now correct.

It seems unlikely that with this additional logic incorporated into the circuit that this is still a practicable method of radix conversion.

Yours faithfully,

F. K. Hanna,
Associated Electrical Industries Ltd.

The author replies:

Dear Sir,—I agree with Mr. Hanna's initial comments that the method of adding six does not give the correct result in every case. My article was introduced by using six because many readers would be familiar with the use of six as a filler. However, I think he has missed the whole point of the article. The shortcomings to which he draws attention are overcome by using the method of adding three. It was for this reason that apart from the introduction the whole of my article discussed the design of a convertor in which the addition of three rather than six was used. The example he quotes is then worked as follows:

		111011	shift 1 place left
	1	11011	shift 1 place left
	11	1011	shift 1 place left
	111	011	Add 3 and shift
1	0100	11	shift 1 place left
10	1001	1	Add 3 and shift
101	1001		stop
5	9		

I am grateful that Mr. Hanna's letter has enabled me to draw the attention of readers to the advantage of the method of adding three, where this problem of cell boundaries does not occur. Hence Mr. Hanna's conclusion that there is doubt about the practicability of the system is no longer valid, and it is my contention that this system using three can in many cases offer distinct advantages over other methods of binary-decimal conversion.

Yours faithfully,

K. J. Dean,
College of Technology,
Letchworth.

Null Networks and Frequency Stability of Null Network Oscillators

Dear Sir,—Three terminal null networks such as the *LCR* bridged-T, *RC* twin-T and four terminal balanced bridges—the Wien bridge, the Meacham bridge etc. have been widely used in low distortion oscillators in which the effective feedback can be arranged to be degenerative at practically all but the null frequency with consequent improvement in waveform. The purpose of this letter is to discuss the phase characteristics of these null networks and to derive a general expression for the frequency stability of oscillators incorporating them.

Consider any null network or balanced bridge whose voltage transfer function can be expressed as[1]

$$\beta_N = E_o/E_i = \frac{K}{1 - j\dfrac{b}{x-(1/x)}} \quad \ldots (1)$$

where ω = angular frequency of the input signal, ω_n is the null frequency and $x = (\omega/\omega_n)$. K and b are parameters characterizing a particular network. The phase shift of the network changes discontinuously at the null frequency and the phase slope is not defined at this point in the classical sense, although some authors[2,3] have expressed views to the contrary.

Writing $u = x - (1/x) = \omega/\omega_n - \omega_n/\omega$, the phase shift θ of the null network is, from equation (1)

$$\theta = \angle \beta_N = \tan^{-1}(b/u) \quad \ldots\ldots (2)$$

i.e. $\theta \to -(\pi/2)$ as $u \to 0_-$ (from the left) and $\theta \to +(\pi/2)$ as $u \to 0_+$ i.e. from the right. Differentiating equation (2)

$$\mathrm{Sec}^2\theta d\theta = -(b/u^2).du$$

Since

$$\mathrm{Sec}^2\theta = 1 + \tan^2\theta = 1 + (b^2/u^2)$$

and

$$du = dx(1+(1/x^2))$$

then

$$d\theta/dx = -(b/u^2)\frac{u^2}{b^2+u^2}(1 + (1/x^2))$$

Obviously $d\theta/dx$ does not exist at $u=0$ or $x=1$. However $d\theta/dx$ tends to the value $-2/b$ as $u \to 0$ and $x \to 1$ from both sides. By invoking the concept of the Dirac's δ-function the phase slope at $x = 1$ can be stated to be an impulse of area π.

Next consider the frequency stability of an oscillator incorporating a null network. Such an oscillator can always be viewed as a 'two path' oscillator because a true null network, in isolation, cannot provide the finite feedback voltage necessary to maintain oscillations. This can easily be arranged by providing a second, usually frequency-independent, feedback path of gain ν or by unbalancing the null network by an amount which depends on the available amplifier gain A. Thus the overall feedback factor β_T of the oscillator is

$$\beta_T = \nu + \beta_N = \nu + \frac{K}{1 - j(b/u)}$$

If the amplifier phase shift is negligible, the phase angle of the loop gain is

$$\phi = \angle\, \beta_T = \tan^{-1}(b/u) - \tan^{-1}\frac{b\nu}{(\nu + K)u}$$

or

$$\tan\phi = \frac{buK}{u^2(\nu+K)+b^2\nu} = P/Q \text{ say.}$$

$$\therefore \quad \mathrm{Sec}^2\phi d\phi = \frac{QdP - PdQ}{Q^2}$$

at $u = 0$, $\phi = 0$, $P = 0$ and $\mathrm{Sec}^2\phi = 1$.

$$\therefore \quad d\,\phi = dP/Q = \frac{bK}{u^2(\nu+K)+b^2\nu}\cdot du$$

Putting,

$$du = dx\,(1+(1/x^2))$$

We have

$$d\phi/dx = \frac{bK}{u^2(\nu+K) + b^2\nu}\,(1 + (1/x^2))$$

Thus the normalized phase slope of the loop gain at the frequency of oscillation $\omega = \omega_0$ would be

$$\left.\frac{d\phi}{dx}\right|_{x=1} = 2K/b\nu$$

Since $A\beta_T = 1$ at $x = 1$ and $\beta_T = \nu$ at $x = 1$, $A = 1/\nu$

$$\left.\frac{d\phi}{dx}\right|_{x=1} = 2K/b\,.\,A$$

Yours faithfully,

U. S. Ganguly,
Jadavpur University,
Calcutta.

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A Simple Solid-State Multiplier

Dear Sir,—The problem of multiplying two time-varying voltages, especially in the course of analogue computation, has received much attention over recent years. A number of methods have been developed with varying bandwidths and degrees of accuracy.

The gain of a voltage-controlled variable-gain device may be made accurately proportional to the gain-setting voltage if a feedback configuration is employed. The multiplier to be described uses this principle to linearize the characteristic of a Luxistor, type ORP 39, and the circuit developed is a solid-state analogue of the servo-multiplier. The Luxistor is a device consisting of a small lamp and a cadmium sulphide photoresistor, electrically insulated, in an opaque epoxy-resin encapsulation.

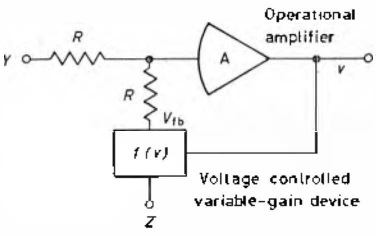


Fig. 1. Basic feedback loop

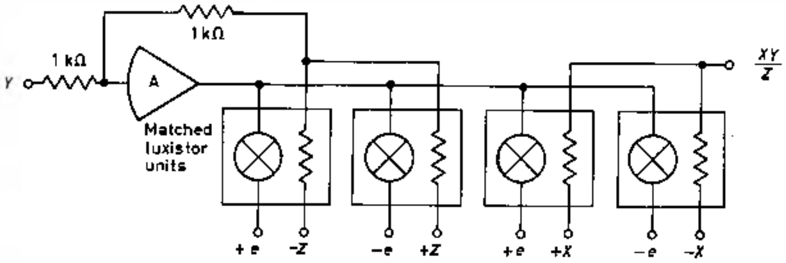


Fig. 2. Practical multiplier circuit

Referring to Fig. 1, it may be shown that

$$\nu = A(Y + V_{fb})$$

and

$$V_{fb} = Zf(\nu)$$

$$\therefore$$

$$AY + AZf(\nu) = \nu$$

where A = open-loop gain of amplifier

$f(\nu)$ = gain of non-linear device, assumed to vary monotonically.

Z = constant bias voltage

If $A \to \infty$, then

$$f(\nu) = -Y/Z \quad \ldots\ldots\ldots\ldots \quad (1)$$

The gain-setting voltage, ν, is used to control a second variable-gain device, assumed to be identical to the first. The input voltage, X, to this second unit will thus be operated upon by the function $f(\nu)$, which has been reduced to $-Y/Z$. Hence the quotient XY/Z is formed.

In the practical circuit shown in Fig. 2, the voltage e is a d.c. bias voltage of not greater than 6V, limited by the maximum lamp voltage. Inputs X, Y and Z may be time-varying voltages but, for normal multiplication, Z is a constant bias. Simultaneous multiplication and division is therefore possible.

The maximum input voltage on the Y-channel is limited to ± 5V, governed by the output capabilities of the operational amplifier. Channels Y and Z have input levels limited ideally by the photoresistor breakdown voltage of 70V. In a practical circuit this may be limited by the output voltage of an operational amplifier if inversion of these voltages is necessary.

The response of the multiplier to a varying Y-input is limited by the nature of the Luxistors. An error of less than 2 per cent can be realized with a sinusoidal input of 10Hz. The speed of response of the other channels is very high, being of the order of 2MHz for the same accuracy.

The input resistance of the Y-channel is 1kΩ and the X and Z channels have a common-mode input resistance of 150Ω.

This multiplier has the advantages of simplicity (especially if both polarities of one of the signals are available), simultaneous multiplication and division, and possible 'slave-channel' operation. Disadvantages are that the accuracy depends on the degree of matching of two pairs of Luxistors, and the response time of one channel is relatively long. The former could be rectified somewhat by employing a time-sharing scheme. The latter disadvantage has reduced importance in the application of the circuit as a slow-speed modulator or phase-sensitive demodulator.

Yours faithfully,

F. G. McIntosh,
N. D. Deans,
Robert Gordon's Institute of Technology,
Aberdeen.

Giro Handling Equipment

Dear Sir,—I am writing a sequel to *Giro Credit Transfer Systems*, my book that was instrumental in persuading Parliament to legislate for the highly-automated, computer-centred National Giro Service which will come into operation in 1968.

In view of the many requests for information I have received both from foreign governments and from firms anxious to budget now for giro form office handling equipment, I shall include a section on this and would like firms who are making—or proposing to make—equipment specially for this service to communicate with me.

Yours, etc.,

F. P. Thomson,
39 Church Road,
Watford, Herts.

NEW BOOKS

System Analysis by Digital Computer

Edited by Franklin F. Kuo and James F. Kaiser. 438 pp. Med. 8vo. John Wiley & Sons Ltd. 1966. Price 68s.

THE advent of computers, whether digital or analogue, has had a profound influence on man's approach to system analysis and design. By minimizing the time and effort spent on the necessarily large amount of routine calculation involved, the designer can make more efficient use of the faculties of judgement and experience. Analogue techniques were the first to be used in this way, but are now being replaced to a large extent by the greater flexibility and accuracy of digital methods.

This book, the first of its kind to be written, deals with the application of digital computers to the design of electrical networks. It is based on contributions made almost entirely by members of the Bell Telephone Laboratories U.S.A. to a summer conference on computer science held at Princetown University in 1966. Four important aspects of design are covered, namely: analysis, synthesis simulation, and non-numerical applications. Each chapter is stocked with a large number of references which, in themselves are sufficient to make the work a valuable acquisition to workers in this field. In general, the treatment is linear although some reference is made to advances in non-linear techniques.

The first four chapters are devoted to analysis in both the time and frequency domains. Of particular interest is the paper on state space techniques which enable many of the results of linear and matrix algebra to be applied directly to network analysis.

Design is covered in the next two chapters which describe programs for filter synthesis and optimization techniques respectively. Optimization is treated mainly from the decision aspect and a description is given of the SUPROX programme which has been written at Bell Telephone Laboratories.

Three chapters then follow on simulation which deal with digital filters, a sampled data simulation language, and the use of hybrid techniques. In this context hybrid techniques are those using combined analogue and digital computers for simulation of systems. No mention is made, however, of the application of the digital differential analyser or incremental methods which are admirably suited to both linear and non-linear simulation. For this purpose incremental techniques combine the desirable features of analogue and digital computers and avoid the interface problems which must be solved in the hybrid computer. It is felt that the book would have benefited considerably by the inclusion of a chapter on this topic.

With the present day emphasis on the importance of software, high level languages play an increasing part in any digital computer application. To what extent a special language will prove to be necessary in network design applications is open to question. Indeed, most of the programs referred to in this book have been written in FORTRAN. Two special launguages are nevertheless described, namely: BLODI, a sampled data system language, and ALTRAN for symbolic algebra. ALTRAN is perhaps more interesting and clearly covers a wider field than network design.

The last two chapters will be of interest to those people concerned with man-machine interfaces. 'Computer Produced Movies' deals with the hardware and software techniques which enable a computer to drive an automatic microfilm recorder. A first impression is that this is no more than moving photography coupled with character generation. However, as the subject is developed it becomes clear that the author is proposing dynamic animation of complex forms, in effect visual simulation of physical systems.

One example, after Zajac, deals with the characteristics of a communications satellite in an earth orbit. Instead of outputting the result conventionally in tabular form, they are converted into signals which cause a series of perspective drawing to appear on a c.r.t., the face of which is filmed. When subsequently projected the film presents a dynamic picture of events.

Another film, after Sinden, animates Newton's law $f = m \cdot a$ and photographs, reproduced in the book, made from various frames show clear, complex shapes conforming to mathematical equations which would be extremely difficult, and expensive, to produce by manual drafting techniques.

The last chapter 'Man—Computer Graphical Communication' deals with methods of manual input and visual output based on such items as c.r.t. displays, a fibre optics light pen and a RAND tablet which generates 10 bit by 10 bit co-ordinate information with a resolution of 100 lines/inch over an area of about 100in^2. Information is given on various system configurations of the above equipments with associated programming techniques.

References at the ends of both chapters show that a considerable amount of work has been done on this fascinating branch of digital computer technology. There is no doubt that it is going to play a part in digital systems of the future.

T. R. H. Sizer

P. L. Owen

Electromagnetic Shock Waves

By I. G. Katayev. 164 pp. Demy 8vo. Iliffe Books Ltd. 1966. Price 35s.

MUCH of the literature on electromagnetic wave propagation assumes linearity with respect to the medium. In general, however, non-linearities occur which can (somewhat arbitrarily) be divided into two classes. The first class results from the wave itself changing the thermodynamic state (e.g. the temperature) of the medium. Such non-linearities may be termed large amplitude, or large signal. The second arises from one, or more of the state parameters of the medium, namely the permittivity, permeability and conductivity, being functions of the wave field intensities. This second class of non-linearity can lead to 'shock' waves in much the same way as the acoustic analogue, and is the subject of the book under review.

After an interesting introduction the book opens with a chapter describing how shock waves are found. In this chapter the differential forms of permeability and permittivity are introduced leading to the quasi-linear Maxwell equations. Similarly the introduction of the differential inductance and capacitance leads to the corresponding quasi-linear transmission line equations. These equations are basic to the theory but are intractable in general, and it is for this reason that the author develops a systematic treatment of particular solutions. A comparison between gas dynamics and electrodynamics is drawn and it is shown that a wider definition of a shock wave than that usually used in the former case is desirable owing to the nature of the solutions in the latter case (i.e. in some cases there are no continuous solutions to the system describing electromagnetic waves, unlike acoustic waves). Thus it is asserted that two clearly distinguishable mechanisms give rise to the formation of shock waves: the first causes the advance of the wave crest owing to the permeability and/or the permittivity of the medium being non-linear in the sense described above. Such a mechanism will apply in gas dynamics also. The second mechanism, however, does not. This is the dissipation of energy at the wave front. The author substantiates his assertion by means of examples, and in the second chapter more detailed solutions are considered in the case of transmission lines containing ferrites with rectangular magnetization curves.

The analysis so far only applies to certain elementary waves (simple or stationary shock waves), and more general solutions are not forthcoming. However, the analysis is extended considerably by developing a system with given initial conditions in the form of a finite number of simple waves. Since the system is non-linear the superposition principle does not apply but there will be interaction between such waves. Several interesting cases are analysed.

The third and final chapter is devoted to the design of pulse shaping networks for short pulse generators using shock waves in lines containing ferrite. Since this field is fairly active it can hardly be expected that such a chapter will be fully up to date, nevertheless there are several useful hints and worked examples that makes it worth reading even for those well versed in the subject.

Taking the book as a whole the reviewer feels that it is an important addition to the literature of the subject more particularly because, as the author mentions in the introduction, much of the material is appearing here for the first time. It is well translated and reads well, although here and there the reviewer feels that the wrong word has been used (e.g. *multivalence* in place of *multiplicity*). The text is clear for this class of printing apart from very minor blemishes, but occasionally the captions on the figures in the reviewer's copy were rather blurred.

R. H. C. NEWTON

The Challenge of the Computer Utility
By D. F. Parkhill. 207 pp. Crown 4to. Addison Wesley Publishing Co. Inc. 1966. Price 45s.

This book stresses the close relationship between technological change and economic, political and social evolution under four main headings: (1) Principal historical and technological developments leading to computer utility, (2) Technology of computer utilities, (3) Economic and legal considerations, (4) Future possibilities.

Ionization Avalanches and Breakdown
By F. Llewellyn-Jones. 103 pp. Crown 8vo. Methuen and Co. Ltd. 1967. Price 25s.

This monograph explains the present position regarding the analysis and understanding of the physics of the electrical breakdown of gases. It also describes applications of modern time-resolving techniques in the investigation of the atomic, ionic and electrical collisional processes.

Automatic Control Systems Explained
By R. J. G. Pitman. 165 pp. Demy 8vo. Macmillan & Co. Ltd. 1966. Price 25s.

Intended as an introduction to the more specialized works, this book gives an elementary treatment of automatic control systems without resorting to differential equations and operational methods.

A basic knowledge of the fundamental electrical principles is assumed but a brief treatment is given of the required mathematics together with a complete analysis of an automatic position control system.

Semiconductor Devices
By M. J. O. Strutt. 313 pp. Med. 8vo. Academic Press Inc. 1966. Price $12.50

This book is the English version of 'Halbleiterbauelemente. I. Halbleiter und Halbleiterdioden' originally published in Switzerland in 1962.

The main concepts of semiconductor theory and the properties of semiconductor devices are presented and the fundamentals of semiconductor physics are dealt with, so as to afford an understanding of transport and storage phenomena in diodes and transistors.

Electrical Conduction Mechanisms in Thin Insulating Films
By A. I. Markov. 350 pp. Demy 8vo. Methuen and Co. Ltd. 1967. Price 25s.

This monograph first establishes that the ordinary energy band structure as applied to crystals can also be applied to amorphous and crystalline materials and then proceeds to show that the various conduction mechanisms can be postulated to explain this phenomena.

Fluctuations of Stationary and Non-Stationary Electron Currents
By C. S. Bull. 217 pp. Demy 8vo. Butterworths & Co. (Publishers) Ltd. 1966. Price 68s.

The theory of electricity is discussed in the first three chapters of this book with the object of supporting the hypothesis that the electric field is discontinuous and examining in detail the consequences of this hypothesis.

The book is based on the author's work on the research on the noise of thermionic valves.

Transduktorschaltungen (Transductor Circuits)
By W. Hartel and H. Dietz. 273 pp. Med. 8vo. Springer-Verlag, Berlin. 1966. DM 46,50

The book deals first with the basic characteristics of transductor chokes, particularly the static and dynamic magnetization, core characteristics and design. The analogies between controlled rectifiers and transconductor chokes follow, permitting the transfer of many of the characteristics of the static convertor circuits to the voltage-controlled transductor circuits operating under ideal conditions. Introduction of certain core constants permits presentation of a uniform theory of power amplification, time constants and magnification factor applicable to most of the more usual transductor circuits. Observations made originally under idealized conditions are more suitably adapted to practical conditions by taking into account feedback and dynamic magnetization. Due to their special properties, current-controlled transductor circuits are covered in a separate chapter. Knowledge of electrical engineering at the level of an introductory lecture and an introduction into the application of the calculus are presupposed.

A new chapter has been added to this English version dealing with noise in semiconductors.

NEW EQUIPMENT

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

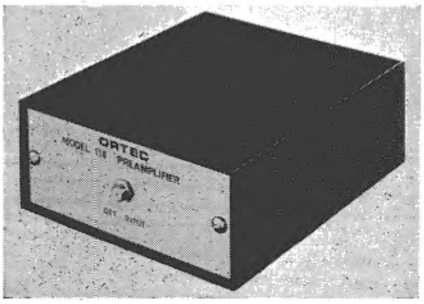
(Voir page 197 pour la traduction en français; Deutsche Übersetzung Seite 204)

HIGH RESOLUTION PRE-AMPLIFIER

High Volt Linear Ltd, 67 Dudley Street, Luton, Bedfordshire

(Illustrated below)

An extension to the ORTEC range of charge sensitive pre-amplifiers is the model 118, f.e.t. pre-amplifier. Primarily for high resolution gamma spectrometry this pre-amplifier is intended for use with cooled, low capacitance, low leakage semiconductor detectors such as the ORTEC 8000 range, or other similar lithium drifted germanium detectors. The charge gain, which is switchable either $\times 1$ or $\times 10$, is 365mV/MeV (Gc) in the $\times 10$ position, and the resolution is better than 1·25keV (0 capacitance, F.W.H.M. Ge). The model 118 pre-amplifier is compatible with the ORTEC 400 series of modular instrumentation and may be powered directly from the model 410 linear amplifier, or from a separate power unit model 115.



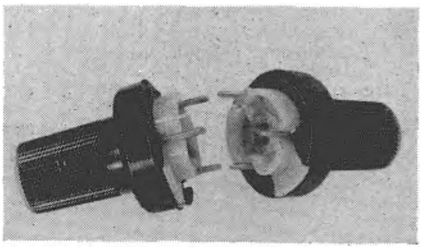
The overall dimensions of the model 118 pre-amplifier are $1{\cdot}8 \times 4 \times 6$in, and when it is used with the ORTEC range of detectors it can be mechanically attached to the cryostat such that capacitance on the pre-amplifier input may be kept to an absolute minimum.

EE 04 751 *for further details*

MINIATURE POTENTIOMETERS

Davall Electronics Ltd., Rothersthorpe Avenue, Northampton

(Illustrated above right)



The Davall type 72 miniature moulded track potentiometer is claimed to offer a substantially longer life than conventional sprayed track potentiometers now in common use in the commercial electronics field.

The type 72 has the insulating base, solid carbon track and terminals moulded into a single solid integrated unit. This type of construction ensures high stability under all conditions of humidity, temperature and vibration. Noise levels throughout the life of the component are very much lower than those achieved with sprayed track potentiometers and decrease with use. The type 72 has a power rating at 55°C of 0·5W (linear) and 0·25W (non-linear) and a resistance range of between 50Ω and 2MΩ (linear). Standard resistance values between 50Ω and 2MΩ are available to order. The maximum working voltage is 500V d.c.

Mounting is direct on to printed circuit boards with tripod tags on a 0·1in grid and the type 72 is compact enough to be contained within a cylinder 0·75in in diameter and 1in deep (excluding tags).

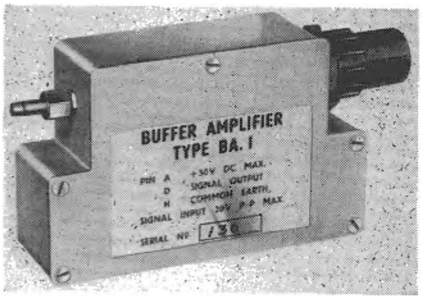
EE 04 752 *for further details*

BUFFER AMPLIFIER

Ether Ltd, Caxton Way, Stevenage, Hertfordshire

(Illustrated below)

The buffer amplifier type BA1 has been designed primarily for connexion between piezo-electric acceleration transducers and their associated measuring and recording devices, where an input impedance of the order of 1000MΩ is required to minimize loading effects. It can also be used in any signal detection application where negligible loading effect is necessary, provided the input signal amplitude is within the specified limits.



The unit is a two-stage all solid-state amplifier with a gain of unity (nominal) and with an exceptionally high impedance conversion ratio, 40×10^6 typical. One hundred per cent voltage negative feedback applied over both stages increases input impedance and reduces output impedance. The first stage is a high input impedance field-effect transistor; the second stage is a silicon alloy junction transistor which provides the necessary output current.

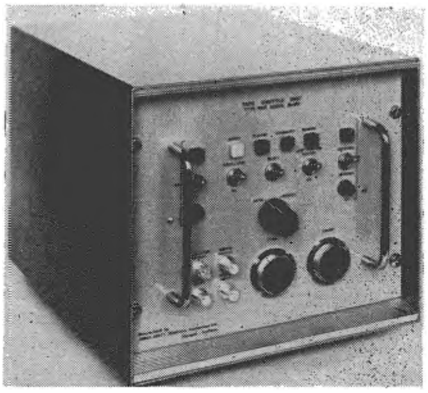
EE 04 753 *for further details*

TAPE SHUTTLE UNIT

James Scott (Electronic Engineering) Ltd, Carntyne Industrial Estate, Glasgow, E.2

(Illustrated below)

When recording experimental data or other information on multi-channel instrumentation tape recorders the information recorded may be spread over a long length of tape and it may be found that only a small portion of this length



contains information of particular interest to the user. The problem then arises that the user may wish to play and replay this portion a considerable number of times.

This tape shuttle unit utilizes the remote control socket on the tape recorder and, after the selected portion of the recording has been 'marked' electronically by the operator, will play this selected portion, stop the recorder at the end of the desired portion, set the recorder to rewind back to the start of the selection, stop the recorder and start it replaying automatically. This process will continue indefinitely without the necessity of the operator having to touch any of the recorder controls. At the end of the experiment it is only necessary to switch the shuttling unit from 'automatic' to 'manual' then the manual controls of the tape recorder will become effective and the recorder can be stopped manually.

The tape shuttle unit accomplishes the 'marking' of the selected portion of the tape by using one channel of the recorder and imposes a fixed duration pulse on to this channel when the operator depresses a switch on the unit. It is preferable to record these pulses at the same recorder speed as the desired playback speed.

The tape shuttle unit is self-contained having its own mains derived power supplies but it can also be run from an external 28V d.c. supply.

Shuttling can be accomplished at speeds from 1⅞in/sec to 60in/sec. The unit measures 11¼in wide × 11in high × 13¼in deep.

EE 04 754 *for further details*

HIGH INTENSITY MICROPHONES

D. J. Birchall, 5 Leys Avenue, Letchworth, Hertfordshire

(Illustrated below)

These piezo-electric microphone capsules are available in sizes varying from ⅜in to 1¼in diameter, and in charge sensitivity from 40pC per lb/in² to 2000pC per lb/in². The devices are suitable for operation in environments of high

ambient pressure, have a 250°C temperature capability, and are insensitive to vibration. Case material is stainless steel, diaphragms are welded. Illustrated is a 1⅜in diameter flange mounted device.

EE 04 755 *for further details*

MECHANICAL FILTERS

Distributed by: Impectron Ltd, Impectron House, 125 Gunnersbury Lane, London, W.3

(Illustrated above right)

Miniature mechanical filters and matching transformers made by Toko Inc., Tokyo, Japan, are now available from Impectron Ltd. They are designed mainly for use with valve or transistorized high grade radio receivers, selective radio paging systems, radio telephones and similar equipment. They have extremely good selectivity characteristics and are claimed to be far superior to the conventional *LC* variety. Their dimensions are 1·65in long by 0·42in wide by 0·4in deep. Connecting pins project 0·157in.

Five types are available, three with a 455± 1kHz frequency and two with a

455± 1·5kHz frequency. Bandwidths for the ±1kHz version at −6dB are 5± 1kHz, 6± 2kHz and 12± 1 kHz and at 40dB, 16kHz, 18kHz and 28kHz. For the ±1·5kHz types bandwidths at −6dB are 16+3·0kHz and 22+3·0kHz and at −40dB, 36kHz and 42kHz. Ripple characteristics are within 3dB and frequency shift is less than 150 in 10⁶/°C. Signal input to the transducer is 5V maximum at 455kHz, the operating temperature range is −10 to +70°C, and the weight is approximately 9g.

Basic elements are a sectionalized metal rod resonator and two transducers. No welded parts are used in coupling the resonant sections, thus keeping losses to a minimum. Drive and pick-up transducers are polarized high efficiency ceramic disks securely fixed to the resonator ends.

Special impedance matching transformers for use at the filter input and output are also available. These serve to secure maximum energy transfer and minimum ripple and may be obtained for either valve or transistor circuits. Transformers at the filter input will accept up to 30kΩ and at the output end have an impedance of 800Ω. They are small in size to match the filter.

EE 04 756 *for further details*

ROTARY WAFER SWITCHES

A.B. Metal Products Ltd, North West House, 119/127 Marylebone Road, London, N.W.1

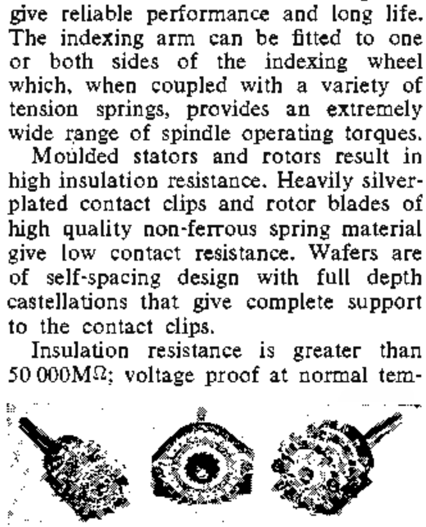
(Illustrated below)

A.B. Metal Products Ltd announce a new range of high quality rotary wafer switches—MINI-12 (12 position), MINI-24 (24 position) and PY (22 clips/12 position)—built for lasting reliability, and suitable for professional and military applications.

They feature a newly designed index mechanism employing a sintered iron index wheel and hardened detent pin to give reliable performance and long life. The indexing arm can be fitted to one or both sides of the indexing wheel which, when coupled with a variety of tension springs, provides an extremely wide range of spindle operating torques.

Moulded stators and rotors result in high insulation resistance. Heavily silver-plated contact clips and rotor blades of high quality non-ferrous spring material give low contact resistance. Wafers are of self-spacing design with full depth castellations that give complete support to the contact clips.

Insulation resistance is greater than 50 000MΩ; voltage proof at normal tem-



perature and pressure 2kV between electrical contacts and normally earthed components, 1kV between electrical contacts intended to be insulated from each other; contact resistance 4 to 6mΩ; switch breaking capacity (resistive loads) 50mA at 300V a.c./d.c. or 500mA at 30V a.c./d.c.; current carrying capacity 2A.

The MINI-24 and PY types are available with concentric spindles.

Electrostatic screens, bearing straps and mains switches are standard accessories.

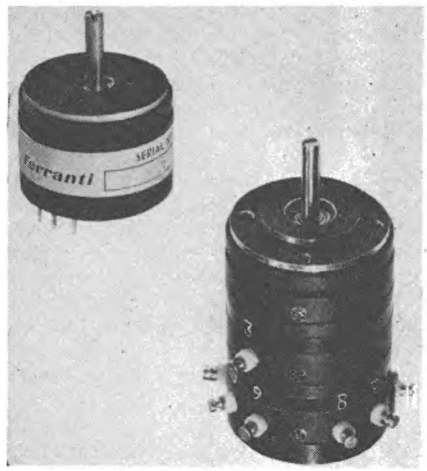
EE 04 757 *for further details*

WIRE WOUND POTENTIOMETERS

Ferranti Ltd, Thornybank, Dalkeith, Midlothian, Scotland

(Illustrated below)

Further developments of the recently introduced type 11 HL precision wire wound potentiometer has enabled Ferranti Ltd to provide sine/cosine functions on individual gangs. Single and multi-ganged units up to 6 gangs are available. Torque is only 3g.cm per gang and law conformity is better than ±1 per cent peak-to-peak.



Resistance values are from 50Ω to 50kΩ per quadrant and the operating temperature range is −65° to +150°C. Precious metal alloys are used for all wipers, slip-rings and most windings for maximum resolution and low noise, and taps are welded to single turns.

Using their cam winding method, Ferranti Ltd can also readily produce other trigonometrical or empirical laws such as loading correction.

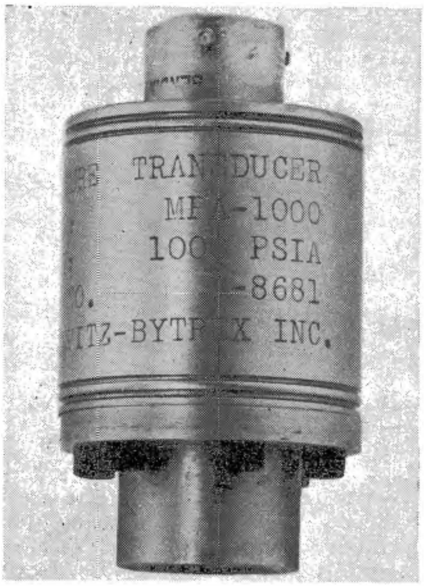
EE 04 758 *for further details*

ABSOLUTE PRESSURE TRANSDUCERS

Electro Mechanisms Ltd, 218-221 Bedford Avenue, Slough, Buckinghamshire

(Illustrated on page 192)

Electro Mechanisms Ltd announce the introduction of a new range of absolute pressure transducers only 1⅛in in diameter. These devices, fabricated entirely from stainless steel are available for most pressure ranges from 50lb/in² to 5 000lb/in². Internal mechanical overload stops protect the transducer



and permit overloads to five times rated pressure with less than 0·2 per cent zero shift. Units can be supplied to operate over the wide temperature range of −320°F to +250°F retaining a very high accuracy, less than 0·01 per cent error over the temperature compensated range. The sensing elements used are semiconductor strain gauges which provide a high electrical output of 150mV with low resistance. Meters and recorders can be driven directly from the transducer, thus dispensing with amplifiers which could introduce errors into the system. The combined non-linearity and hysteresis of 0·15 per cent permit use in high accuracy systems.

These pressure transducers are manufactured by Schaevitz Bytrex, Inc. of Massachusetts.

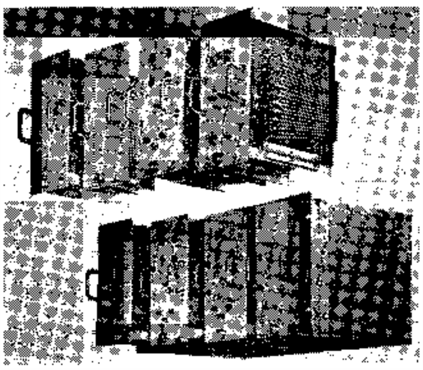
EE 04 759 *for further details*

VIBRATION MONITORING SYSTEM

Dawe Instruments Ltd, Western Avenue, Acton, London. W.3

(Illustrated below)

This system (type 1435B) is intended to monitor continuously and automatically the operational vibration of the widest range of plant and machinery, and to give warning or to fail safe if vibration reaches a preset value. To accommodate the many different kinds of operational requirements, three basic instrument modules are available. The



modules are vibration monitor (type 1435/1B), meter (type 1435/2) and a range of seven filters (types 1435/3B to 1435/9B). These can be combined by plugging into a pre-wired chassis (type 1435/11B) in any desired arrangement to make up systems of whatever operating mode is preferred.

The monitor module is normally connected by wiring to velocity-sensing transducers mounted at a selected point on the machine (or structure) under surveillance. However, it is possible to derive a signal from accelerometers to be fed into the monitor, a 27V d.c. supply being provided for the excitation of emitter or source followers. The monitor provides a continuous electrical output proportional to the peak displacement of the machine within the frequency range from 10Hz to 800Hz (equivalent to 600 to 48 000rev/min). Two relays in each monitor may be preset to operate at any desired output level; they can be normally open to actuate an alarm, or normally closed to give 'fail safe' facilities. Since relay and amplifier circuits are separate, monitoring continues after the first relay is tripped. A gain change of 10:1 after the first warning can cause the second warning (or cut-out) to operate at up to 10 times the displacement level of the first.

To enable specific components, such as a bearing or gearwheel to be monitored, the input can be passed through a filter to give a specified frequency response of the monitor. Where maintenance logging or vibration limits is to be carried out (for instance to indicate varying loads on process plant or machine tools) the output of the monitor can be fed to a meter. A standard rack-mounting chassis (type 1435/11B) 19 × 8¾ × 10½in deep is available in which up to four modules can be mounted, each module measuring 8¼ × 4½ × 8in deep. External connexions for transducers, power supplies, relay contacts, external meters or recorders are provided at the rear of the chassis.

The system can operate with up to four monitors fed from four transducers, or with any desired combination of monitors, meters and filters up to a total of four. The meter has three channels (seven if required), the filter has two channels. For configurations using only three or fewer modules, blanking panels of module-size are available.

The monitor amplifier enables nominal gain settings from 0·001in peak-to-peak to 0·020in peak-to-peak to be obtained. The normal frequency response from 10Hz to 800Hz can be extended if required. It is suitable for use with velocity transducers having sensitivities from 10mV in/sec to 150mV in/sec. The signal is integrated internally to give voltage proportional to displacement. Dawe also supply moving-coil transducer (type 1403/1) for use at temperatures up to 70°C.

EE 04 760 *for further details*

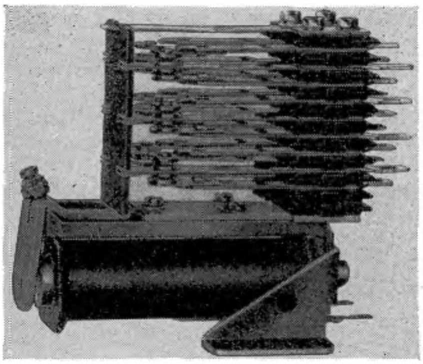
HEAVY DUTY RELAY

Magnetic Devices Ltd, Exning Road, Newmarket, Suffolk

(Illustrated below)

This heavy-duty relay comprising the coil and yoke assembly of the standard P.O. 3000 type relay but fitted with a heavy duty blade and contact assembly to provide a multi-contact relay of greater sensitivity than is normal for the switching capacity available, has been awarded C.E.G.B. approval (Certificate No. 173). The series C.350 (D.C.) relay conforms to Category 1 and light current category, covering generation, transmission and nuclear.

The series C.350 incorporates several interesting features, including specially designed contact blades, keeping bounce to a minimum and extending working life of the relay, while generous clearances between contacts increases the



maximum safe inter-contact potential that can be applied.

Contact arrangement is from 2 to 6 poles, with normally open, normally closed or changeover contacts, and typical operate speed is 70msec, release 15msec. Ambient temperature range is up to 35°C, adaptable to higher temperatures if specified.

Contact rating, with fine silver, suitable for resistive loads up to 10A at 30V d.c. or 250 V a.c., and with Elkonite D54X, suitable for resistive loads up to 15A at 30V d.c. or 250V a.c. Insulation is 1 500V r.m.s. coil to earth; 2 000V r.m.s. contact to contact and 2 500V r.m.s. contacts to earth.

Contact blade tags are suitable for soldered connexions or 'quick-connect' 250 series receptacles. Coil tags are suitable for soldered connexions only, but can be linked to two spare tags provided in the contact stack, if the use of quick connect receptacles is desired.

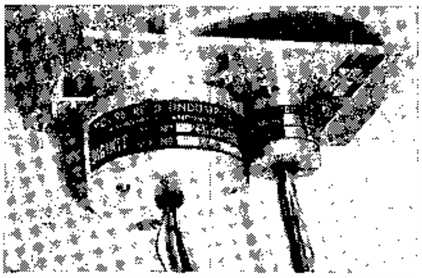
EE 04 761 *for further details*

DECIMAL DIGITAL SCALES

Moore Reed (Industries) Ltd, Walworth Industrial Estate, Andover, Hampshire

(Illustrated on page 193)

Electrical scales of different lengths and resolutions can be constructed from the 'Digikit' modules, manufactured by Moore Reed & Co. Ltd. The scale readings can be displayed directly on digital displays or recorded on tape, etc.



Selected positions can be determined for programmed control purposes.

By interposing Digishift switches between the scale and the display, the scale can be electrically slid or zero suppressed at will to any desired position, without having to mechanically uncouple the scale from the moving part under control. The scale can also be electrically reversed at any time or used for offset zero readings.

The illustration is of a typical 'Digikit' assembly, giving a scale length of 1 000 divisions in 10 turns of the input shaft. Resolutions of 1 000 in one turn and scale lengths in excess of 1 000 000 divisions are producible by selecting the appropriate combinations from the ranges of components available.

EE 04 762 *for further details*

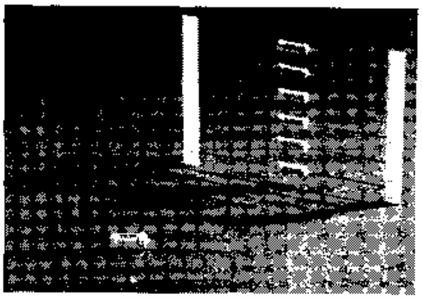
CARBON-FILM RESISTORS

Distributed by: Standard Telephones & Cables Ltd Rectifier Division, Edinburgh Way, Harlow, Essex

(Illustrated below)

To meet the demand for precision resistors having high stability and high reliability at an economic cost, the Rectifier Division of Standard Telephones and Cables Ltd has augmented its standard range of Beyschlag (West German) carbon-film resistors with a range of premium quality types claimed to offer the best performance: price ratio available.

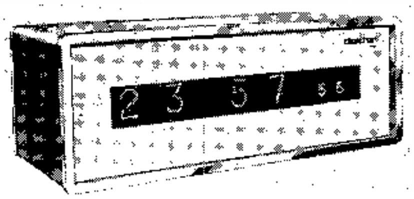
The new range now enables the Division to supply ± 1 per cent tolerance high stability carbon-film resistors, rated at $\frac{1}{4}$W and $\frac{1}{2}$W, in the IEC E24 series of preferred values from 51Ω to 1MΩ. Identically rated versions having ± 2 per cent tolerance are available from 10Ω up to 1MΩ and up to 2·2MΩ in the $\frac{1}{2}$W size. In addition, an $\frac{1}{8}$W series having ± 1 per cent tolerance can be supplied in E24 values from 51Ω to 68kΩ; ± 2 per cent versions are available from 10Ω up to 240kΩ. All these components are rated for operation at 70°C.



As the sole distributor in the UK, STC Rectifier Division holds large stocks of Beyschlag carbon-film resistors and can supply all types in pilot quantities of up to 200 in most values 'off the shelf'.

To assist the user in building his equipment, Beyschlag resistors are supplied on a continuous tape called an 'Autoband' so they can be fed directly from their packaging into automatic lead-cutting and bending machines employed to prepare components for printed circuit boards.

EE 04 763 *for further details*



DIGITAL CLOCK

Darang Electronics Ltd, Restmor Way, Hackbridge Road, Hackbridge, Surrey

(Illustrated above)

The model six-seventy digital electronic clock is the latest in the 'Digicron' range from Darang Electronics Ltd.

Intended for general time-keeping on either the 12 or 24 hour systems, the 'Six-Seventy' has a larger size neon digital display for reading at a distance in excess of 40ft even under bright ambient light conditions. It is presented in a contemporary styled case, which is available in a choice of four colours.

The timing source is derived from the 50Hz mains, or from a crystal source which can be fitted at extra cost. In the event of a mains failure the clock will not restart until it is reset. A single control is used to set the time and a single switch programmes the clock for either the 12 or 24 hour systems.

EE 04 764 *for further details*

POSITIVE TEMPERATURE COEFFICIENT RESISTORS

Thorn Parsons Co. Ltd, 40 Broadway, London, S.W.1

(Illustrated below)

A new range of 'Posistors', resistors with greatly increased positive tempera-

ture coefficients, is announced by Thorn Parsons Co. Ltd.

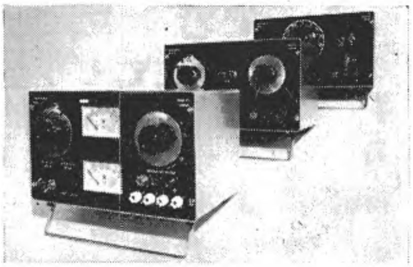
A feature of the Posistor is that its resistance remains almost constant up to a threshold temperature but at higher temperatures increases very rapidly. This characteristic and the ability to adjust the threshold temperature by varying the chemical composition of the device make the Posistor suitable for many safety and monitoring systems, especially for electric motor protection.

Modern electric motors are being designed to operate at temperatures closer to the point at which damage occurs than has hitherto been the practice. For the thermal protection of these motors the Posistor offers a high degree of reliability. Its small size (in special cases down to one-sixteenth of an inch diameter) allows it to be located in the motor windings, right at the 'hot-spot', so giving quicker responses to danger conditions than the bi-metal switches previously used. Its thermal time-constant of about 3sec and the absence of moving parts bring additional advantages.

For motor applications the devices are usually epoxy resin insulated. Most demands are for Posistors and controllers to work at switching temperatures of 120°C (BS 2757, Class E insulation) and 130°C (Class B). Other types of insulation can be used and Posistors are available for operating temperatures in the range 95°C to above 165°C.

Besides monitoring temperatures, Posistors can be used to limit currents or to sense changes in the environment of the device, for example liquid/vapour changes, rate of flow variations or flame blow-back.

EE 04 765 *for further details*



POWER SUPPLIES

Irwin & Partners Ltd, Kibworth Street, Dorset Road, London, S.W.8

(Illustrated above)

This new range of power supply units has been developed with the object of providing reasonably priced, universal, self-contained bench power-packs for use in laboratories, schools, etc. The output voltage is infinitely adjustable within the range chosen by a rotary transformer and on all low ranges the output is isolated from the mains supply by a seperate transformer. Seven models are manufactured, varying in the combination of voltage ranges and refinements.

On all units, with the exception of models A.1. and B.1., the selection of a particular voltage range is by means of a rotary switch, this switch being mechanically interlocked with the output switch to avoid the possibility of changing to a higher voltage range with the load still in circuit.

The rectification of the d.c. ranges is by means of silicon diodes. The smoothing uses a low-resistance choke to maintain reasonable regulation.

All meters are moving-coil, multi-range instruments and are automatically switched to the correct range by the output range switch. Where no meters are fitted, sockets are provided to facilitate the connexion of external meters. In the case of the ammeter a bridging switch is also fitted.

The units at present available are as follows:

Model A1—Outputs 0-250V a.c. at 1A
0-250V d.c. at 1A
—Unsmoothed
No meters fitted.

Model B1—Outputs 0-250V a.c. at 1A
0-250V d.c. at ½A
—Smoothed
No meters fitted.

Model C1—Outputs 0-250V a.c. at 1A
0-250V d.c. at ½A
—Smoothed
Fitted with voltmeter and ammeter.

Model D1—Outputs 0- 25V d.c. at ½A
—Smoothed
0-250V d.c. at ½A
—Smoothed
0- 25V a.c. at 1A
0-250V a.c. at 1A
Fitted with voltmeter and ammeter.

Model E1—Outputs 0-250V a.c. at 1A
5- 25V d.c. at ½A
—Stabilized
0-250V d.c. at ½A
—Smoothed
No meters fitted.

Model F1—Outputs 0- 25V a.c. at 1A
0-250V a.c. at 1A
5- 25V d.c. at ½A
—Stabilized
0-250V d.c. at 1A
—Smoothed
Fitted with voltmeter and ammeter.

Model H1—Outputs 0- 12V d.c. at 1A
—Smoothed
0-500V d.c. at 100mA
—Smoothed
5V a.c. at 3A
6·3V a.c. at 3A
Fitted with voltmeter and ammeter.

EE 04 766 *for further details*

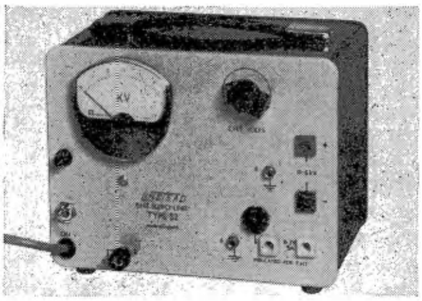
E.H.T. SUPPLY

Linstead Electronics Ltd, 35C Newington Green, London, N.16

(Illustrated below)

This instrument provides 0 to 6kV continuously variable, indicated on a meter. It provides maximum currents of approximately 1mA at 5kV, 1·3mA at 4kV, 2 to 3mA at lower voltages. A 3A 6·3V heater supply provides power for a wide range of c.r.t.'s or other hot-cathode tubes. Either side of the e.h.t. may be earthed, and the heater supply insulated for use at e.h.t. if required. The e.h.t. is obtained from an r.f. oscillator, needing small capacitance smoothing, and giving low short-circuit current. Thus it cannot cause sustained and dangerous shock.

EE 04 767 *for further details*



POTENTIOMETERS

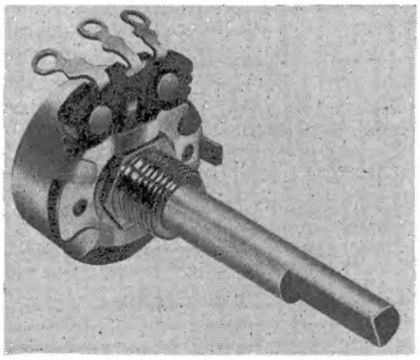
A.B. Metal Products Ltd, North West House, 119-127 Marylebone Road, London, N.W.1

(Illustrated below)

A.B. Metal Products Ltd announce a new range of carbon composition potentiometers—the 15/16in diameter, ½W rating series 45.

New mechanical design and the introduction of the most modern production techniques with mechanized assembly, enable controls of consistent high quality to be manufactured and marketed at a very economical price.

The standard resistance range is 500Ω to 10MΩ (linear, log, semi-log, reverse log and other laws); available



with single or double taps; standard bush, plug-in or twist tab mounting; standard lug, printed circuit or wire-wrap terminals; dual ganged or concentric controls.

The mechanical versatility of the series 45 permits almost unlimited combinations, thus offering immense scope to designers. Moulded spindles are standard. Controls can be supplied with integral plastic front-plates for simple plug-in mounting, and terminals for wire-wrap connexions—features that facilitate component installation and replacement, which could result in appreciable savings in time and cost.

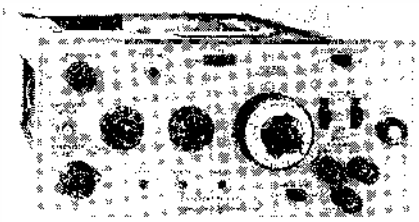
The controls can be fitted with a mains switch suitable for operation under the high surge current conditions usually associated with silicon rectifier power circuits.

EE 04 768 *for further details*

OSCILLOSCOPE DIFFERENTIAL AMPLIFIER

Hewlett-Packard Ltd, 224 Bath Road, Slough, Buckinghamshire

(Illustrated above right)



A new d.c.-coupled vertical plug-in oscilloscope amplifier from Hewlett-Packard measures microvolt-level a.c. signals in the presence of large d.c. levels, with no drift whatever. Both a.c. and d.c. voltage measurements may be made with 0·4 per cent accuracy. The plug-in has maximum sensitivity of 50μV/cm. With Hewlett-Packard's new zero-drift feature, and high-resolution calibrated d.c. offset, it easily makes some oscilloscope measurements which previously were impossible, and makes others far more conveniently.

The new plug-in, the HP Model 1406A differential amplifier, matches HP model 140A oscilloscope mainframes, and is even more useful with the recently-introduced HP model 141A variable-persistence/storage oscilloscope.

The calibrated offset feature of the 1406A turns it into an accurate voltmeter. Read-out is to four significant figures, and front-panel lights indicate the decimal point. The amplifier supplies up to 10V offset on the maximum sensitivity range (50μV/cm). This increases to 1 000V on less sensitive ranges.

The input of the 1406A amplifier may be used differentially, without offset. In differential operation, spurious common-mode signals, such as 50Hz hum, are rejected by 85dB or more.

Maximum bandwidth of the differential amplifier is 400kHz. Where high-frequency signal components are unwanted, built-in filters reduce bandwidth to 100, 25, or 5kHz by front-panel control.

Built-in, too, is an isolating amplifier to drive external equipment, such as an *X-Y* recorder or magnetic tape recorder.

The amplifier's ground connexion may either be floated to the measured circuit, or connected to the oscilloscope chassis, as desired, with a convenient front-panel switch. This helps to eliminate ground-loops, for clear pictures and high accuracy.

EE 04 769 *for further details*

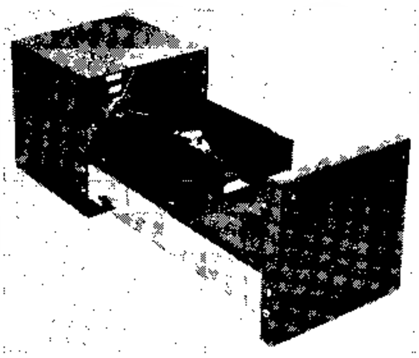
COOLING UNIT

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

(Illustrated on page 195)

The Lektrokit Division of A.P.T. Electronic Industries Ltd announces an addition to its range of components available in Lektrokit Rack and Chassis construction systems.

Designated LKU.511, the unit is primarily designed as a cooling unit for electronic equipment, especially for improving the efficiency of transistor heat sinks and for lowering the tem-



perature level inside assemblies as an aid to reliable operation and stability generally. Although the unit is dimensioned to fit into both Chassis and Rack System constructions, it can be incorporated into the design of any electronic equipment.

The housing measures 5in × 5in across the opening and it is 4·5in deep. Fixing holes are provided to enable the cooling unit to be used as a replacement for the normal side plate of any Lektrokit assembly with the Chassis rails bolted to it.

In operation the unit provides an air displacement of approximately 2 000ft³/h under conditions of reasonably free air flow. The fan blades are driven by a fractional h.p. 200/250V a.c. motor. Tapped bushes (two × 2BA) are fitted on the air intake face for mounting the unit on Lektrokit ventilation panels numbers 2 and 3.

The unit fits into the same space as a normal Lektrokit Chassis Assembly by cutting rails LK.201 to three quarter lengths and fitting the cooling unit at one side. In this way it may be fitted into a standard 19in rack, using normal Lektrokit mounting brackets (LK.601) for flush or proud mounting.

EE 04 770 *for further details*

VACUUM GAUGE

Edwards High Vacuum Ltd, Manor Royal, Crawley, Sussex

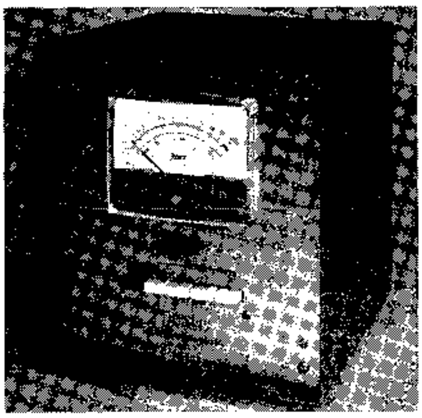
(Illustrated above right)

Edwards High Vacuum Ltd has introduced a new Pirani vacuum gauge, known as the model 9, which covers the pressure range 500 to 10^{-3}torr. The scale has three ranges and features automatic switching from one range to the next, coloured lamps on the front panel indicating which range is in operation. The gauge has been developed specifically for low pressure measurement in a variety of processes in the chemical, metallurgical and allied industries which, while not carried out at the highest vacua, still require an electrical gauge with facilities for recorder operation and external switching. The new instrument should also prove useful in a wide range of laboratory applications where a multi-head vacuum gauge is required.

A Wheatstone bridge voltage regulator and differential amplifier form the basic control circuit, a feature of which is the incorporation of a temperature dependent compensating resistor wound on the gauge heads. This circuit substantially reduces the effects of ambient temperature variation on gauge calibration. The inclusion of the voltage regulator also eliminates the need for manual voltage adjustment. Push-buttons on the front panel permit selection of any one of four series 9 gauge heads which can be connected to the control unit. The series 9 gauge heads are available in versions suitable for connexion to metal or glass vacuum systems and are electrically interchangeable.

Terminals on the rear panel provide: facilities for operating a moving coil or potentiometric recorder; a low impedance 19V d.c. switched supply for operating an auxiliary recorder pen for indicating range change points during recording, and heavy duty contacts for process control. Modular construction and transistor circuits throughout provide a compact instrument which is easily serviced.

EE 04 771 *for further details*



MULTIPLE DIODE ASSEMBLIES

SGS-Fairchild Ltd, Planar House, Walton Street, Aylesbury, Buckinghamshire

These new multiple diode assemblies contain up to 16 chips in one TO-5 transistor can and include many combinations in diode bridges, matrices, arrays and ring modulators. The Company also offers a special service to meet demands for additional combinations for individual circuit requirements.

The availability of so wide a range of standard assemblies is possible because of the characteristics of the silicon planar process employed by SGS-Fairchild. It is capable of producing high-conductance, high-speed diodes of exceptional uniformity and reliability, unaffected by long usage in the severest operating or environmental conditions.

These SGS-Fairchild multiple diodes are encapsulated in TO-18 and TO-5 cans, giving power dissipations as high as 400mW for each element. Operating and storage temperatures up to 200°C are permissible and, since all elements are contained in the same envelope, thermal conductivity is better than with discrete diodes and less affected by external factors.

These devices have also the advantage of reduced size, up to 16 diodes being accommodated in a single TO-5 can. This means considerably lower assembly costs since many interconnexions are already made—the 16-chip device needs only 10 connexions instead of 32 with discrete diodes.

Types BAX45 to BAX51 are silicon planar epitaxial diode matrices for high speed core driver applications in computer memory systems. They range from two-diode units with three leads to the 16-diode, 10-lead BAX51.

The BAX52 and BAX53 are diode bridges for high speed applications, contained in TO-5 and TO-18 cans respectively. The diode chips are not matched but are hermetically sealed in their encapsulations.

The BAX54 and BAX55 are silicon planar epitaxial diode ring modulators designed for applications in high speed multichannel equipment. They are particularly suitable for use in electronic telephone exchanges.

Finally, there is a choice of 18 SGS-Fairchild silicon planar epitaxial diode arrays, types BAX 56-BAX73. These are designed for high speed core driver applications, and are offered as common anode or common cathode arrays in combinations of two to eight diodes.

EE 04 772 *for further details*

Meetings THIS Month

THE INSTITUTION OF ELECTRICAL ENGINEERS

All evening meetings will be held at Savoy Place, commencing at 5.30 p.m. (tea at 5 p.m.), unless otherwise stated.

Date: 1 March.
Colloquium: Colour Cameras. (Joint meeting with the I.E.R.E., the I.E.E.E. and the Royal Television Society at 9.30 a.m.).
Date: 2 March.
Discussion: Control: A Unifying Force in Engineering Education.
(Joint meeting with the Institution of Mechanical Engineers).
Opened by: J. C. West and R. J. A. Paul.
Date: 6 March.
Lecture: U.H.F. Tunnel-Diode Amplifier.
By: M. K. McPhun.
Lecture: Short-Hop Radio-Relay Systems Using Tunnel-Diode Repeaters.
By: D. L. Hedderly, J. Hooper and M. K. McPhun.
Conference: Making the Bosworth Report Work. (Joint meeting sponsored by the Ministry of Technology and the Council of Engineering Institutions at 10 a.m. All wishing to attend must register by letter enclosing the Registration Fee of 15s.).
Date: 10 March.
Lecture: Broadband Transmission by Radio and Cable.
By: R. W. White.
Discussion: The Future Education of Electronic Engineers.
Opened by: G. D. Sims and B. H. Venning.
Date: 13-17 March.
Conference: Air Traffic Control.
Date: 15 March.
Lecture: Ergonomics in Electronic Equipment and Systems Design.
By: B. Shackel. (At 6 p.m., tea at 5.30 p.m.).
Date: 16 March.
Discussion: Automatic Processing in Nucleonic Isotope Studies.
(Joint meeting with the I.E.R.E. Medical and Biological Group (at 2.30 p.m.)).
Date: 17 March.
Discussion: The Present State of Stability Theory.
Opened by: A. G. Dewey, G. D. S. MacLellan, P. C. Parks and C. Storey.
(Joint meeting with the Automatic Control Group of the I.Mech.E.).
Lecture: Early Days in Radio Research.
By: R. L. Smith-Rose.
(Joint meeting of the I.E.E. History of Technology and the Newcoment Society).
Date: 20 March.
Colloquium: Automated Cartigraphy; Scientific Needs and Engineering Possibilities.
(Joint meeting with the Royal Society and the I.E.R.E. Computer Group at 2.30 p.m.).
Date: 21 March.
Discussion: Rapid Fault Diagnosis of the Future.
Opened by: L. E. Land and A. J. Cope. (At 6 p.m., tea at 5.30 p.m.).
Date: 22 March.
Lecture: The Theory of Oscillators.
By: R. Spence and P. J. Baxandall.

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

London Meetings

Tickets are required for meetings held in the Institution's Lecture Room, 9 Bedford Square, London, W.C.1 and for other meetings as indicated.

Date: 1 March.
Time: 9.30 a.m.
Held at: Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Symposium: Colour Cameras.
(Joint I.E.E./I.E.R.E. Television Groups and Royal Television Society.)
Tickets will be necessary: no charge to members.
Date: 13-17 March.
Held at: Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Conference: Air Traffic Control Systems Engineering and Design.
(Joint I.E.E./I.E.R.E.)
Advance registration will be necessary.
Further information from the I.E.R.E., 8-9 Bedford Square, London, W.C.1.
Date: 16 March.
Time: 2.30 p.m.
Held at: Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Colloquium: Automatic Processing in Isotope Studies.
(Joint I.E.E./I.E.R.E. Medical and Biological Electronics Group.)
Tickets will be necessary.
Date: 20 March.
Time: 2.30 p.m.
Held at: Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Colloquium: Automation in Cartography.
(Joint meeting with the Royal Society and the I.E.E.)
Tickets will be necessary.
Date: 22 March.
Time: 6 p.m.
Held at: Institution of Electronic and Radio Engineers, 8-9 Bedford Square, London. W.C.1.
Lecture: Dynamic Properties of Audio Frequency Compressors.
By: K. O. Bäder.
(Joint meeting with Electro-Acoustics Group.)

South Western Division

Held at: Bath Technical College.
Date: 8 March.
Time: 6 p.m.
Lecture: Communications and their Effect on Railway Operation.
By: Professor Banwell and H. H. Ogilvy.
(Joint meeting with the Western Centre of the I.E.E., Electronics Section).
Held at: Bristol University, Small Lecture Theatre.
Date: 21 March.
Time: 7.30 p.m.
Lecture: An Introduction to Microcircuitry.
By: R. A. Hyman.
(Joint meeting with the British Computer Society).
Held at: Cornwall Technical College, Camborne.
Date: 14 March.
Time: 7 p.m.
Lecture: Advances in Laser Technology.
By: R. C. Smith.
(Joint meeting with the Western Centre of the I.E.E., Electronics Division).

West Midland Section

Held at: Birmingham University, Department of Electronic and Electrical Engineering.
Date: 2 March.
Time: 7.15 p.m.
Lecture: Solid State Devices.
By: G. T. Wright.
Held at: Stafford College of Further Education, Tenterbanks, Stafford.
Date: 14 March.
Time: 7.15 p.m.
Lecture: Mathematics for the Engineer.
By: R. Woolridge.

East Anglian Section

Held at: The University Engineering Department, Trumpington Street, Cambridge.
Date: 2 March.
Time: 8 p.m.
Lecture: Electronically Assisted Acoustics.
By: J. Moir.
(Joint meeting with the Cambridge Electronics Section of the I.E.E.).
Held at: Chelmsford Technical High School, Patching Hall Lane, Broomfield.
Date: 8 March.
Time: 6.30 p.m.
Lecture: Electronics in Radio Astronomy.
By: D. M. A. Wilson.

South Wales Section

Held at: Welsh College of Advanced Technology, Cardiff.
Date: 8 March.
Time: 6 p.m.
Lecture: Satellite Control.
By: E. G. C. Burt.
(Joint meeting with the I.E.E.).

South Midland Section

Held at: North Gloucestershire Technical College, Cheltenham.
Date: 17 March.
Time: 7 p.m.
Lecture: Autoland Systems.
By: R. I. Bishop.

Scottish Section

Held at: Department of Natural Philosophy, Edinburgh University.
Date: 8 March.
Time: 7 p.m.
Lecture: Static Relay Techniques in Protection.
By: M. Legg.
(Joint meeting with the I.E.E.).
Held at: Department of Medical Physics, Royal Infirmary, Edinburgh.
Date: 9 March.
Time: 10 a.m.-6 p.m.
Meeting: Telemetry in Medical and Biological Research.
Joint I.E.R.E.-I.E.E. Scottish Medical Electronics Group Symposium.
Registration is necessary—apply to P. M. Elliott Esq., C.Eng., A.M.I.E.R.E., 21 Craigmount Loan, Corstorphine, Edinburgh 12.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow, C.2.
Date: 9 March.
Time: 7 p.m.
Lecture: Static Relay Techniques in Protection.
By: M. Legg.

Southern Section

Held at: Farnborough Technical College.
Date: 9 March.
Time: 7.30 p.m.
Lecture: Made to Measure Integrated Circuits.
By: P. Cooke.
(Joint meeting with the I.E.E. Southern Electronics and Control Section).
Held at: The Lanchester Theatre, University of Southampton.
Date: 21 March.
Time: 6.30 p.m.
Lecture: Electronic Equipment Design.
By: T. G. Clark.
(Annual General Meeting of the Section will precede this lecture.)

East Midland Section

Held at: Leicester University.
Date: 14 March.
Time: 6.30 p.m.
Lecture: Single-Sideband Transceivers.
By: E. T. Wilson.

Merseyside Section

Held at: Liverpool College of Technology, Byrom Street.
Date: 15 March.
Time: 7 p.m.
Lecture: The Early Days of Ships' Radio.
By: G. R. M. Garratt.

North Eastern Section

Held at: Institute of Mining and Mechanical Engineers, Neville Hall, Westgate Road.
Date: 8 March.
Time: 6 p.m.
Lecture: Time Sharing Computer Systems.
By: N. E. Heywood.

Thames Valley Section

Held at: J. J. Thomson Physical Laboratory, University of Reading.
Date: 16 March.
Time: 7.30 p.m.
Lecture: The Design of Linear Amplifiers.
By: E. A. Faulkner.

THE ROYAL TELEVISION SOCIETY

All meetings, unless otherwise stated are held in the I.T.A. Conference Suite at 70 Brompton Road, London, S.W.3, commencing at 7 p.m. Visitors' tickets and further information may be obtained from The Television Society office at 166 Shaftesbury Avenue, London, W.C.2. Telephone: TEM 3330.

Date: 9 March.
Lecture: Problems and Solutions of the Shadow Mask Tube.
By: R. R. Bathelt (Philips, Eindhoven).
Date: 31 March.
Lecture: A History of Television Equipment.
By: G. R. M. Garratt (The Science Museum).
(To be held at The Science Museum, South Kensington, London, S.W.7.)
Admission by ticket only.

NOUVELLES Réalisations

Traduction des pages 190 à 195

Une description basée sur des renseignements fournis par les fabricants de nouveaux composants, accessoires et instruments d'essai

PRÉAMPLIFICATEUR À HAUTE RÉSOLUTION

High Volt Linear Ltd, 67 Dudley Street, Luton, Bedfordshire

(Illustration à la page 190)

Le préamplificateur à transistors à effet de champ, modèle 118, constitue une extension de la gamme ORTEC de préamplificateurs sensibles aux charges. Initialement prévu pour la spectrométrie gamma à haute résolution, ce préamplificateur a été étudié pour l'emploi avec des détecteurs à semi-conducteurs à faible perte, de faible capacité et à refroidissement, tels que ceux de la gamme ORTEC 8000 ou d'autres détecteurs analogues au germanium et à dérive de lithium. Le gain de charge, qui peut être commuté soit sur X^1 ou X^{10}, est de 365 mV/MeV (Ge) dans la position de X^{10} et la résolution est supérieure à 1,25 keV (capacité 0). Le préamplificateur modèle 118 est compatible avec les instruments modulaires de la série ORTEC 400 et il peut être entraîné directement par l'amplificateur linéaire modèle 410 ou par un bloc d'alimentation à part modèle 115.

Les dimensions hors-tout du préamplificateur modèle 118 sont de 4,54 cm × 10,16 cm × 15,24 cm et lorsqu'il est utilisé avec les détecteurs de la gamme ORTEC il peut être fixé mécaniquement au cryostat de manière à ce que la capacitance dans l'entrée du préamplificateur puisse être maintenue à un minimum absolu.

EE 04 751 *pour plus amples renseignements*

POTENTIOMÈTRE MINIATURE

Davall Electronics Ltd., Rothersthorpe Avenue, Northampton

(Illustration à la page 190)

Le potentiomètre miniature Davall type 72 à piste moulée serait d'une durée de vie considérablement plus étendue que les potentiomètres à piste imprimée communément utilisés dans le domaine de l'électronique commerciale.

Le potentiomètre type 72 a la base d'isolement, une piste au carbone solide et des bornes moulées en un seul élément solide intégré. Ce type de construction assure une haute stabilité dans toutes les conditions d'humidité, de température et de vibration. Les niveaux de bruit durant la vie du composant sont beaucoup plus bas que ceux obtenus avec les potentiomètre à piste imprimée et ils décroissent en cours d'emploi. Le type 72 a une puissance nominale à 55° C de 0,5 W (linéaire) et de 0.25 W (non-linéaire) et une gamme de résistance de 50 Ω à 2 MΩ (linéaire). Des valeurs de résistance standard de 50 Ω à 2 MΩ sont prévus sur commande. La tension de régime maxima est de 500 V c.c.

Le montage se fait directement sur plaquette de circuit imprimé avec cosse reliée à un réseau de 0,1 pouce et le type 72 est suffisamment compact pour pouvoir être enfermé dans un cylindre de 1,90 cm de diamètre × 2,54 cm de profondeur (cosses exclues).

EE 04 752 *pour plus amples renseignements*

AMPLIFICATEUR INTERMÉDIAIRE

Ether Ltd. Caxton Way, Stevenage, Hertfordshire

(Illustration à la page 190)

L'amplificateur intermédiaire, type BA1 a été principalement conçu pour être inséré entre des transducteurs d'accélération piézoélectriques et les dispositifs d'enregistrement et de mesure connexes où une impédance d'entrée de l'ordre de 1000 MΩ est nécessaire pour minimiser les effets de charge. Il peut également être utilisé dans n'importe quelle application de détection de signaux où un effet de charge négligeable est nécessaire, à condition que l'amplitude du signal d'entrée soit à l'intérieur des limites spécifiées.

L'appareil est un amplificateur constitué de corps solides à deux étages avec gain unitaire (nominal) et avec un rapport de conversion d'impédance exceptionnellement élevé, soit 40 × 10^6 typiquement. Une contre réaction de tension de cent pour cent appliquée sur les deux étages augmente l'impédance d'entrée et réduit l'impédance de sortie. Le premier étage est un transistor à effet de champ d'une impédance à entrée élevée, le second étage est un transistor à jonction par alliage au silicium qui fournit le courant de sortie nécessaire.

EE 04 753 *pour plus amples renseignements*

NAVETTE À BANDE MAGNÉTIQUE

James Scott (Electronic Engineering) Ltd. Carntyne Industrial Estate, Glasgow, E.2

(Illustration à la page 190)

Lorsqu'on enregistre des données expérimentales ou d'autres informations sur magnétophone à multivoies, les informations enregistrées peuvent s'étendre sur une grande longueur de bande et on peut constater par la suite que seule une partie réduite de cette bande contient des informations d'un intérêt particulier pour l'utilisateur. Le problème qui se pose alors est celui de permettre à l'utilisateur de pouvoir reproduire à plusieurs reprises cette partie de la bande.

Le nouvel élément à "navette" utilise la prise de télécommande du magnétophone et après que la partie choisie de l'enregistrement ait été "marquée" électroniquement par l'opérateur, reproduit la partie choisie, arrête l'enregistreur au terme de la partie choisie, fait retourner la bande au début de la partie choisie, arrête le magnétophone et le remet en marche automatiquement. Ce processus peut continuer indéfiniment sans qu'il soit nécessaire pour l'opérateur de toucher l'une quelconques des commandes de l'appareil. Au terme de l'expérience il suffit de commuter la navette du fonctionnement automatique au fonctionnement manuel et les commandes manuelles du magnétophone deviendront alors effectives et l'appareil pourra être arrêté à la main.

La navette effectue le "marquage" de la partie choisie de la bande au moyen d'une des voies du magnétophone et impose une impulsion de durée fixe à cette voie lorsque l'opérateur presse un commutateur sur l'appareil. Il est préférable d'enregistrer ces impulsions à la même vitesse d'enregistrement que la vitesse voulue de réenregistrement.

La navette est un appareil autonome ayant son propre bloc d'alimentation secteur mais il peut également être alimenté d'une source extérieure de 28 V c.c.

La navette peut s'effectuer à une vitesse allant de 4,78 cm/sec à 152,4 cm/sec. L'appareil mesure 49,21 cm de large × 27,94 cm de hauteur × 34,29 cm de profondeur.

EE 04 754 *pour plus amples renseignements*

MICROPHONE À HAUTE INTENSITÉ

D. J. Birchall, 5 Leys Avenue, Letchworth, Hertfordshire

(Illustration à la page 191)

Ces capsules microphoniques piézoélectriques sont livrables en formats variant de 0,95 cm à 3,49 cm de diamètre et leur sensibilité de charge peut varier de 40 pC par livre/pouce² à 2000 pC par livre/pouce². Ces dispositifs peuvent être utilisés dans des environnements de haute pression ambiante ayant une capacité de température de 250° C et ils sont insensibles aux vibrations. Le boîtier est en acier inoxydable et les diaphragmes sont soudés. On voit dans notre illustration un dispositif à montage normal de 3,49 cm de diamètre.

EE 04 755 *pour plus amples renseignements*

FILTRES MÉCANIQUES

Distributeurs: Impectron Ltd, Impectron House, 125 Gunnersbury Lane, London, W.3

(Illustration à la page 191)

La société Impectron Ltd offre maintenant des filtres mécaniques miniature et des transformateurs d'adaptation fabriqués par la société Toko Inc., Tokio, Japon. Ces filtres sont essentiellement destinés à l'emploi avec des récepteurs de radio de haute qualité, à lampes ou transistorisés, ainsi qu'avec des systèmes de téléappel sélectif, des radiotéléphones et un matériel analogue. Ils ont d'excellentes caratéristiques de sélectivité et sont nettement supérieurs à la variété classique LC. Ils mesurent 1,65 pouce de long × 0,42 pouce de large × 0,4 pouce de profondeur. Les broches de connexion dépassent de 0,157 pouce.

Cinq types sont prévus, dont trois à 455± 1 kHz de fréquence et deux à 455± 1,5 kHz de fréquence. Les largeurs de bande pour le modèle à ±1 kHz à −6 dB sont de 5± 1 kHz, 6± 2 kHz et 12± 1 kHz et à 40 dB, elles sont de 16 kHz, 18 kHz et 28 kHz. Pour les types de ±1,5 kHz les largeurs de bandes à −6 dB sont de $16\pm^{3}_{0}$ kHz et $22\pm^{3}_{0}$ kHz et $22\pm^{3}_{0}$ kHz et à −40 db, elles sont de 36 kHz et 42 kHz. Les caractéristiques de bourdonnement sont d'environ 3 dB et le déplacement de fréquence est inférieur à 150 dans 10^{6}/° C. L'entrée de signaux au transformateur est de 5 V au maximum à 455 kHz, la gamme de température de fonctionnement allant de −10 à +70° C et le poids étant d'environ 9 g.

Les éléments de base sont un résonateur en métal à sections et deux transducteurs. Aucun élément soudé n'est utilisé dans le couplage des sections résonantes, ce qui réduit les pertes au minimum. Les transducteurs d'entraînement et de captage sont des disques en céramique polarisés d'une haute efficacité et fixés fermement aux extrémités du résonateur.

Des transformateurs d'équilibrage d'impédance spéciaux pouvant être utilisés à l'entrée et à la sortie du filtre sont également livrables. Ces transformateurs servent à assurer un tranfert d'énergie maxima et un bourdonnement minimum et ils peuvent être obtenus soit pour circuit à lampes soit pour circuit à transistors. Les transformateurs à l'entrée du filtre peuvent recevoir jusqu'à 30 kΩ et à l'extrémité de sortie ils ont une impédance de 800 Ω. Ils sont de petites dimensions afin de s'adapter au filtre.

EE 04 756 *pour plus amples renseignements*

COMMUTATEURS À GALETTE ROTATIFS

A.B. Metal Products Ltd, North West House, 119/127 Marylebone Road, London, N.W.1

(Illustration à la page 191)

La société A.B. Metal Products Ltd présente une nouvelle gamme de commutateurs à galette rotatifs de haute qualité, le Mini-12 (12 directions), le Mini-24 (24 directions) et le PY (22 pinces/12 directions), construits en vue d'une très longue durée d'utilisation fiable et convenant pour les applications militaires et industrielles.

Ces commutateurs comportent un nouveau mécanisme à index utilisant une roue à index en métal fritté et une broche de détente en acier donnant une performance sûre et une longue durée d'utilisation. Le bras d'indexation peut être fixé à l'un ou aux deux côtés du volant à index qui, lorsqu'il est accouplé à une variété de ressorts à tension, fournit une gamme extrêmement étendue de couples de fonctionnement.

Des stators et rotors moulés donnent une résistance élevée à l'isolement. Des pinces de contact doublées d'argent et des lames de rotor en matériau non ferreux à ressort de haute qualité donnent une faible résistance de contact. Les galettes sont à autoespacement avec écrous à entaille assurant un appui total aux pinces de contact.

La résistance à l'isolement est supérieure à 50 000 MΩ; la preuve de tension s'effectue à la température normale et la pression est de 2 kV entre les contacts électriques et les composants normalement mis à la masse; elle est de 1 kV entre les contacts électriques prévus pour l'isolement l'un de l'autre; la résistance de contact est de 4 à 6 mΩ; la capacité de rupture de commutation (charge résistive) est de 50 mA à 300 V c.a./c.c. ou de 500 mA à 30 V c.a./c.c.; la capacité de transmission de courant est de 2 A.

Les types MINI-24 et PY sont livrables avec pivot concentrique.

Des écrans électrostatiques portant des courroies et des commutateurs de secteurs constituent des accessoires standard.

EE 04 757 *pour plus amples renseignements*

POTENTIOMÈTRES BOBINÉS

Ferranti Ltd, Thornybank, Dalkeith, Midlothian, Scotland

(Illustration à la page 191)

De nouveaux perfectionnements du potentiomètre bobiné de précision type 11HL récemment mis au point ont permis à la société Ferranti Ltd de fournir des fonction sinus/cosinus sur groupe individuel. Des éléments allant jusqu'à 6 groupes sont livrables. Le couple n'est que de 3 g/cm par groupe et la conformité est supérieure à ±1 % de crête à crête.

Les valeurs de résistance vont de 50 Ω à 50 kΩ par quadrant et la gamme de température de fonctionnement va de −65° à +150° C. Des alliages de métaux précieux sont utilisés pour tous les frotteurs, bagues collectrices et la plupart des enroulements pour assurer une résolution maxima et un faible bruit et les cosses sont soudées aux tours simples.

Utilisant la méthode d'enroulement par came, la société Feranti Ltd peut également produire facilement d'autres règles trigonométriques ou empiriques telles que la correction de charge.

EE 04 758 *pour plus amples renseignements*

TRANSDUCTEUR DE PRESSION ABSOLUE

Electro Mechanisms Ltd, 218-221 Bedford Avenue, Slough, Buckinghamshire

(Illustration à la page 192)

La société Electro Mechanisms Ltd présente une nouvelle gamme de transducteurs de pression absolue dont le diamètre n'est que de 3,17 cm. Ces dispositifs, entièrement fabriqués en acier inoxydable, sont prévus pour la plupart des gammes de pression de 50 livres/pouce² à 5000 livres/pouce². Des arrêts de surcharge mécanique interne protègent le transducteur et permettent des surcharges de 5 fois la pression nominale avec moins de 0,2 pour cent de déphasage de zéro. Des éléments peuvent être fournis pour fonctionner dans la gamme étendue de température de −320° F à +250° F gardant une très grande précision, inférieure à 0,01 pour cent d'erreur dans la gamme de température compensée. Les éléments sensibles utilisés sont des extensomètres à semi-conducteur qui fournissent une sortie électrique élevée de 150 mV avec faible résistance. Les instruments de mesure et enregistreurs peuvent être entraînés directement par le transducteur, ce qui permet de se passer d'amplificateur qui pourrait produire des erreurs dans le système. La non-linéarité combinée avec l'hystérésis de 0,15 pour cent permet l'emploi dans des systèmes de haute précision.

Ces transducteurs de précision sont fabriqués par la société Schaevitz Bytrex, Inc. de Massachusetts.

EE 04 759 *pour plus amples renseignements*

SYSTÈMES DE CONTRÔLE DE VIBRATIONS

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustration à la page 192)

Ce système (type 1435B) a été étudié pour le contrôle continu et automatique des vibrations opérationnelles de la gamme la plus étendue de matériel et de machines ainsi que pour donner l'alarme en cas de vibrations atteignant une valeur préréglée. Pour faire face aux nombreuses et différentes formes d'exigences opérationnelles, trois modules d'instruments de base sont livrables. Ces modules comprennent le contrôleur de vibrations (type 1435/1B), l'instrument de mesure (type 1435/2) et une gamme de sept filtres (type 1435/3B à 1435/9B). Ces modules peuvent être combinés par branchage dans un chassis prébobiné (type 1435/11B) dans n'importe quel montage voulu pour former des systèmes de n'importe quel mode de fonctionnement.

Le module de contrôle est normalement branché par bobinage aux transducteurs sensibles à la vitesse; ces derniers sont montés à un point choisi de la machine (ou de la structure) soumise au contrôle. Il est cependant possible d'obtenir un signal d'accéléromètre pouvant être injecté dans le contrôleur, un courant continu de 27 V étant prévu pour l'excitation d'émetteur ou de source cathodique. Le contrôleur fournit une sortie électrique continue proportionnelle au déplacement de pointe de la machine dans la gamme de fréquence de 10 Hz à 800 Hz (équivalent à 600 à 48 000 tours/min). Deux relais dans chaque contrôleur peuvent être préréglés pour fonctionner à n'importe quel niveau de sortie voulu; ils peuvent normalement s'ouvrir pour entraîner un signal d'alarme ou se fermer normalement pour déclencher le dispositif de protection contre les pannes. Etant donné que le circuit de relais et celui d'amplification sont séparés, le contrôle continue après le déclenchement du premier relais. Un changement de gain de 10:1 après le premier avertissement peut mettre en oeuvre le second avertissement (ou coupure) fonctionnant à un niveau allant jusqu'à 10 fois le niveau de déplacement du premier. Pour pouvoir contrôler des composants spécifiques, tels que les paliers ou les roues dentées, on peut faire passer l'entrée par un filtre donnant une réponse de fréquence déterminée au contrôleur. Lorsqu'il s'agit d'effectuer l'enregistrement d'entretien de limite de vibrations (par exemple pour indiquer différentes charges sur le matériel de conversion ou les machines-outils) la sortie du contrôleur peut être injectée à un instrument de mesure. Il existe un châssis standard pour montage sur bâti (type 1435/11B) mesurant 48,2 cm × 22,2 cm × 26,6 cm de profondeur dans lequel on peut monter jusqu'à 4 modules, chaque module mesurant 21,59 cm × 10,79 cm × 20,32 cm de profondeur. Il y a, en outre, des connexions extérieurs pour transducteurs, alimentations, contacts de relais, instruments de mesure externes ou enregistreurs à l'arrière du châssis.

Le système peut fonctionner avec un maximum de 4 moniteurs alimentés à partir de 4 transducteurs ou avec n'importe quelle combinaison voulue de contrôleurs, d'instruments de mesure et de filtres jusqu'à un total de 4. L'instrument de mesure comporte 3 voies (7, si nécessaire), le filtre ayant deux voies. Pour les montages n'utilisant que 3 modules ou moins de trois modules, il existe des panneaux de suppression de format modulaire.

L'amplificateur du contrôleur permet les réglages de gains nominaux de 0,001 pouce de crête à 0,020 pouce de crête à crête. La réponse de fréquence normale de 10 Hz à 800 Hz peut être étendue si nécessaire. Elle convient pour l'emploi avec les transducteurs de vitesse ayant des sensibilités de 10 mV pouce/sec à 150 mV pouce/sec. Le signal est intégré intérieurement pour donner une tension proportionnelle au déplacement. La société Dawe fournit également un transducteur à cadre mobile (type 1403/1) pour l'emploi à des températures allant jusqu'à 70° C.

EE 04 760 *pour plus amples renseignements*

RELAIS DE PUISSANCE

Magnetic Devices Ltd, Exning Road, Newmarket, Suffolk

(Illustration à la page 192)

Ce relai de puissance comprenant la bobine et l'assemblage de bride du relais standard type P.O. 3000 mais muni d'une lame de puissance et d'un assemblage de contacts pour constituer un relais à multicontacts d'une plus grande sensibilité que la normale pour la capacité de commutation prévue a été homologué par le Central Electricity Generating Board (certificat No. 173). Ce relais série C.350 (c.c.) est conforme à la catégorie 1 et à la catégorie légère courante couvrant la production, la transmission et les utilisations nucléaires. Le relais série C.350 comprend plusieurs caractéristiques intéressantes dont des lames de contact de conception spéciale réduisant les rebondissement au minimum et prolongeant la durée de vie du relais tandis que des espaces généreux entre les contacts augmentent le potentiel intercontact pouvant être appliqué en toute sécurité.

Le montage des contacts est de 2 à 6 pôles avec contact normalement ouvert, et de 2 à 6 pôles avec contact normalement ouvert, normalement fermé ou de permutation et la vitesse de fonctionnement typique est de 70 msec, déclenchement 15 msec. La gamme de températures ambiantes va jusqu'à 35° C, adaptable aux températures plus élevées le cas échéant.

La valeur nominale de contact avec argent fin, convenant pour les charges résistives jusqu'à 10 A à 30 Vc.c. ou 250 Vc.a., et avec Elkonite D54X, convenant pour les charges résistives allant jusqu'à 15 A à 30 Vc.c. ou 250 Vc.a. L'isolement est de 1500 V efficace de la bobine à la terre; 2000 V efficaces contact à contact, et 2500 V efficaces à contact à la terre.

Les cosses de lame de contact conviennent pour les connexions soudées ou pour les réceptacles de la série 250 à connexion rapide. Les cosses de bobine ne conviennent que pour les connexions soudées mais elles peuvent être reliées à deux cosses supplémentaires prévues dans la pile de contact si l'emploi de réceptacles à connexion rapide est voulu.

EE 04 761 *pour plus amples renseignements*

ECHELLES NUMÉRIQUES DÉCIMALES

Moore Reed (Industries) Ltd, Walworth Industrial Estate, Andover, Hampshire

(Illustration à la page 193)

Des échelles électriques de differentes longueurs et résolution peuvent être construites à partir des modules "Digikit" fabriquées par la société Moore Reed & Co. Ltd. Les lectures d'échelles peuvent être affichées directement sur écran numérique ou enregistrées sur bande. Les positions choisies peuvent être déterminées pour les besoins de commande à programme.

En intercalant les commutateurs "Digishift" entre l'échelle et l'affichage, on peut faire glisser électriquement l'échelle ou supprimer le zéro à volonté à n'importe quelle position sans devoir débrancher mécaniquement l'échelle de la partie mobile sous contrôle. L'échelle peut être inversée électriquement à n'importe quel moment ou utilisée pour les lectures de zéro de compensation. Notre gravure montre un assemblage Digikit typique, présentant une longueur d'échelle de 1000 divisions en dix tours de l'arbre d'entrée. Des résolutions de 1000 dans un tour et les longueurs d'échelle dépassant 1 000 000 de divisions peuvent être produites en choisissant les combinaisons appropriées dans les gammes de composants prévues.

EE 04 762 *pour plus amples renseignements*

RÉSISTANCES À FILM DE CARBONE

Distributeurs: Standard Telephones & Cables Ltd, Rectifier Division, Edinburgh Way, Harlow, Essex

(Illustration à la page 193)

Pour répondre à la demande de résistances de précision ayant une haute stabilité et une grande fiabilité à un prix économique, la division des redresseurs de la société Standard Telephones and Cables Ltd a enrichi sa gamme standard de résistances à film de carbone Beyschlag (Allemagne Occidentale) d'une gamme de types de toute première qualité offrant le meilleur rendement possible pour leur prix.

La nouvelle gamme permet à la Division de fournir des résistances à film de carbone d'une haute stabilité et d'une tolérance de ±1 %, d'une valeure nominale de ¼ W et ½ W dans la série IEC E24 de valeurs préférées de 51 Ω à 1 MΩ. Des versions de valeurs identiques ayant une tolérance de ±2 % sont prévues de 10 Ω jusqu'à 1 MΩ et jusqu'à 2,2 MΩ dans le format de ½ W. En outre, une série à ¼ W ayant une tolérance de ±1 % peut être fournie dans des valeurs de E24 de 51 Ω à 68 kΩ. Des versions à tolérance de ±2 % existent de 10 Ω jusqu'à 240 kΩ. Tous ces composants sont nominalement étudiés pour le fonctionnement à 70° C.

En sa qualité de distributeur exclusif pour le Royaume-Uni, la Division des redresseurs de la Standard Telephone and Cables a de grand stocks de résistances à film de carbone Beyschlag et peut fournir tous les types dans les quantités pilote allant jusqu'à 200 dans la plupart des valeurs courantes.

Pour faciliter la construction de son matériel à l'utilisateur, les résistances Beyschlag peuvent être fournies sur bande continue appellée "Autoband", de manière à pouvoir les injecter directement dans des machines automatiques de coupage et de formage utilisées pour préparer les composants de plaquettes de circuit imprimé.

EE 04 763 *pour plus amples renseignements*

HORLOGE NUMÉRIQUES

Darang Electronics Ltd. Restmor Way, Hackbridge Road, Hackbridge, Surrey

(Illustration à la page 193)

L'horloge électronique numérique, modèle 670, constitue la plus récente réalisation dans la gamme des instruments "Digicron" de la société Darang Electronics Ltd.

Conçue pour le minutage général sur les systèmes à 12 ou à 24 heures, la 670 comporte un affichage numérique au néon de grandes dimensions pour la lecture à des distances supérieures à 12 mètres, même dans des conditions d'éclairage ambiant très clair. Elle est enfermée dans un coffret de style moderne avec un choix de quatre couleurs.

La source de minutage provient du courant secteur de 50 Hz ou d'une source de cristal pouvant être prévue moyennant un supplément de prix. En cas de panne de courant, l'horloge ne se remettra en marche que lorsqu'elle est réenclenchée. Il suffit d'une seule commande pour régler le temps et un seul commutateur pour effectuer le programme de l'horloge pour le système à 12 heures ou celui à 24 heures.

EE 04 764 *pour plus amples renseignements*

RÉSISTANCES À COEFFICIENT DE TEMPÉRATURE POSITIF

Thorn Parsons Co. Ltd. 40 Broadway, London, S.W.1

(Illustration à la page 193)

Une nouvelle gamme de "Posistors", c'est à dire de résistances à coefficient de température positif largement accru, a été mise au point par la société Thorn Parsons Co. Ltd.

Le Posistor se caractérise par le fait que sa résistance demeure quasi constante jusqu'à une température de seuil mais augmente très rapidement aux températures plus élevées. Cette caractéristique ainsi que la faculté de pouvoir régler la température de seuil en faisant varier la composition chimique du dispositif rend le Posistor idéel pour de nombreux systèmes de contrôle et de sécurité, particulièrement pour la protection des moteurs électriques.

Les moteurs électriques modernes ont été conçus pour fonctionner à des températures proches du point auquel l'endommagement peut se produire, ce point étant beaucoup plus rapproché que cela n'a été le cas jusqu'ici. Pour la protection thermique de ces moteurs, le Posistor offre un degré élevé de fiabilité. Ses dimensions réduites (en coffrets spéciaux ayant jusqu'à 1 mm de diamètre) permettent de le loger dans le bobinage du moteur, directement "au point chaud", donnant ainsi des réponses beaucoup plus rapides aux conditions dangereuses que les commutateurs bimétalliques employés jusqu'ici. Sa constante de temps thermique d'environ 3 sec et l'abscence de parties mobiles comportes des avantages supplémentaires.

Pour les applications à moteur, les dispositifs sont habituellement isolés à la résine d'époxyde. La plupart des demandes sont pour des Posistors et des contrôleurs fonctionnant à des températures de commutation de 120° C (norme britannique 2757, isolement Classe E) et 130° C (Classe B). D'autres types d'isolement peuvent être utilisés et il existe des Posistors pour des températures de fonctionnement dans la gamme de 95° C à au-dessus de 165° C.

En plus du contrôle de température, les Posistors peuvent être utilisés pour limiter les courants ou pour percevoir les changements dans l'environnement du dispositif, par exemple des changements de liquide/vapeur, les variations de débit, etc.

EE 04 765 *pour plus amples renseignements*

ALIMENTATIONS

Irwin & Partners Ltd. Kibworth Street, Dorset Road, London, S.W.8

(Illustration à la page 193)

Cette nouvelle gamme de blocs d'alimentation a été étudiée en vue de fournir des blocs de banc d'essai autonomes, universels et d'un prix raisonable à l'intention des laboratoires, écoles etc. La tension de sortie est à variation infinie dans la gamme choisie par un tranformateur rotatif et sur toutes les basses gammes la sortie est isolée de l'alimentation secteur par un tranformateur à part. Sept modèles sont prévus, variant dans la combinaison de gamme de tension et de perfectionnement.

Sur tous les modèles, à l'exception des modèles A.1 et B.1, le choix d'une gamme de tension particulière s'effectue au moyen d'un commutateur rotatif, ce commutateur étant verrouillé mécaniquement au commutateur de sortie pour éviter la possibilité d'un changement à une gamme de tension plus élevée, la charge étant toujours en circuit.

Le redressement des gammes de tension continue s'effectue au moyen de diodes au silicium. Afin de maintenir une régulation raisonnable un choke à faible résistance est utilisé.

Tous les instruments de mesure sont à cadre mobile et à plusieurs gammes et ils sont automatiquement branchés sur la gamme appropriée par le commutateur de gammes de sortie. Lorsque des instruments de mesure ne sont pas prévus, des douilles permettent le branchement sur les instruments de mesure extérieurs. Dans le cas de l'ampèremètre un commutateur de pont est également fixé à l'appareil.

Ces blocs sont actuellement prévus dans les modèles suivants:

Modèle A1—Sorties 0-250 Vc.a. à 1 A
0-250 Vc.c. à 1 A
—non filtré
Sans instrument de mesure

Modèle B1—Sorties 0-250 Vc.a. à 1 A
0-250 Vc.c. à ½ A
—filtré
Sans instrument de mesure

Modèle C1—Sorties 0-250 Vc.a. à 1 A
0-250 Vc.c. à ½ A
—filtré
Muni d'un voltmètre et d'un ampèremètre

Modèle D1—Sorties 0- 25 Vc.c. à ½ A
—filtré
0-250 Vc.c. à ½ A
—filtré
0- 25 Vc.a. à 1 A
0-250 Vc.a. à 1 A
Avec voltmètre et ampèremètre

Modèle E1—Sorties 0-250 Vc.a. à 1 A
5- 25 Vc.c. à ½ A
—stabilisé
0-250 Vc.c. à ½ A
—filtré
Sans instrument de mesure

Modèle F1—Sorties 0- 25 Vc.a. à 1 A
0-250 Vc.a. à 1 A
5- 25 Vc.c. à ½ A
—stabilisé
0-250 Vc.c. à 1 A
—filtré
Avec voltmètre et ampèremètre

Modèle H1—Sorties 0- 12 Vc.c. à 1A
—filtré
0-500 Vc.c. à 100 mA
—filtré
5 Vc.a. à 3 A
6,3 Vc.a. à 3 A
Avec voltmètre et ampèremètre

EE 04 766 *pour plus amples renseignements*

ALIMENTATION E.H.T.

Linstead Electronics Ltd. 35C Newington Green, London, N.16

(Illustration à la page 194)

Cet instrument fournit de 0 à 6 kV à variation continue, indiqués sur un instrument de mesure. Il fournit en outre des courants maximum d'environ 1 mA à 5 kV, 1,3 mA à 4 kV, 2 à 3 mA aux tensions inférieures. A 3 A, 6,3 V l'alimentation de chauffage fournit la puissance pour une gamme étendue de tubes cathodiques ou autres tubes à cathode chaude. Les deux côtées du bloc peuvent être mis à la masse et l'alimentation de chauffage peut être isolée pour l'emploi à très haute fréquence si nécessaire. La très haute tension s'obtient d'un oscillateur haute fréquence, n'exigeant qu'un filtrage réduit de capacitance et donnant un faible courant de court-circuit. Il n'est donc pas susceptible de produire des chocs dangereux et de longue durée.

EE 04 767 *pour plus amples renseignements*

POTENTIOMÈTRES

A.B. Metal Products Ltd, North West House, 119-127 Marylebone Road, London, N.W.1

(Illustration à la page 194)

La société A.B. Metal Products Ltd, vient d'annoncer la création d'une nouvelle gamme de potentiomètres à composition de carbone, d'une diamètre de 23,8 mm, d'une puissance nominale de ½ W. Il s'agit en effet des potentiomètres de la série 45.

Un nouveau dessin mécanique et l'introduction des méthodes de production les plus modernes ainsi que l'assemblage mécanisé permettent de fabriquer des commandes de très haute qualité qui peuvent être commercialisées à un prix très économique.

La gamme de résistances standard va de 500 Ω à 10 MΩ (linéaire, concentration, semi-concentration, concentration inversée et autres lois). Il sont fournis avec des prises simples ou doubles, montage à douille standard ou à fiche, circuit imprimé ou borne bobinée ainsi qu'avec commande concentrique.

La souplesse mécanique des potentiomètres de la série 45 permet un nombre presque illimité de combinaisons, offrant ainsi d'énormes possibilités aux réalisateurs. Les broches moulées constituent des accessoires normaux. Les commandes peuvent être fournies avec plaque frontale plastique pour montage simple à fiche ainsi qu'avec des bornes pour connexion bobinée qui facilite l'installation et le remplacement de composants et résulte en une économie appréciable de temps et de coût.

Les commandes peuvent être munies d'un commutateur secteur convenant pour le fonctionnement dans les conditions de courant très élevé qui sont habituellement le cas avec les circuits de puissance à redresseur au silicium.

EE 04 768 *pour plus amples renseignements*

AMPLIFICATEUR DIFFÉRENTIEL À OSCILLOSCOPE

Hewlett-Packard Ltd, 224 Bath Road, Slough, Buckinghamshire

(Illustration à la page 194)

Le nouvel amplificateur oscilloscopique à fiche vertical à couplage de tension continue de la société Hewlett-Packard mesure les signaux de courant alternatif au niveau des microvolts par rapport au niveau élevé de tension continue, sans dérive aucune. Tant les mesures de tension continue que celles de tension alternative peuvent être effectuées avec une précision de 0,4 %. La fiche a une sensibilité maxima de 50 μV/cm. Grâce au nouveau dispositif de dérive nulle Hewlett-Packard et à la compensation de courant continu étalonné à haute résolution, l'appareil effectue facilement certaines mesures oscilloscopiques qui étaient impossibles auparavant et facilite grandement d'autres mesures. Le nouvel amplificateur différentiel HP, modèle 1406 A, à fiche s'accorde avec le bâti oscilloscopique modèle HP 140 A et il est même plus utile avec le nouvel oscilloscope d'emmagasinement à persistance variable modèle HP 141 A qui vient d'être introduit.

Le dispositif de compensation étalonné du 1406 A en fait un voltmètre précis. La lecture va jusqu'à quatre chiffres significatifs, et les voyants lumineux du panneau frontal indiquent la décimale. L'amplificateur fournit jusqu'à 10 V de compensation sur la gamme de sensibilité maxima (50 μV/cm). Cet indice monte à 1000 V sur les gammes moins sensibles.

L'entrée de l'amplificateur 1406 A peut être utilisée différentiellement sans compensation. Dans le fonctionnement différentiel, les signaux de mode commun parasites, tels que le bourdonnement de 50 Hz sont rejetés par 85 dB ou davantage.

La largeur de bande maxima de amplificateur différentiel est de 400 kHz. Lorsque les composantes de signaux haute fréquence sont indésirables, des filtres incorporés réduisent la largeur de bande à 100, 25 ou 5 kHz au moyen d'une commande de panneau frontal. L'appareil comporte également un amplificateur isolant incorporé pour entraîner le matériel externe tel q'un enregistreur X-Y ou un magnétophone.

La connexion de masse de l'amplificateur peut être soit flottante vers le circuit mesuré soit reliée au châssis oscilloscopique selon les besoins à l'aide d'un commutateur de panneau frontal. On élimine ainsi les circuits de terre et on obtient des images claires et d'une haute précision.

EE 04 769 *pour plus amples renseignements*

APPAREIL REFROIDISSEUR

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

(Illustration à la page 195)

La division Lektrokit de la société A.P.T. Electronic Industries Ltd, vient d'annoncer un nouvel élément à sa gamme de composants dans les systèmes de construction Lektrokit.

Portant la référence LKU.511 cet appareil a été principalement conçu comme refroidisseur pour matériel électronique, particulièrement pour améliorer l'efficacité des sources froides de transistors ou pour abaisser le niveau de température à l'intérieur d'assemblages, afin d'assurer un fonctionnement sûr et une stabilité générale. Bien que les dimensions de l'appareil permettent de l'incorporer au système de construction à châssis et à bâti, il peut être logé à l'intérieur de n'importe quel matériel électronique.

Le logement mesure 12,7 × 12,7 cm de large à travers l'ouverture et 11,4 cm de profondeur. Les trous de fixation permettent d'utiliser le refroidisseur comme élément de remplacement pour la plaque latérale normale de n'importe quel assemblage Lektrokit avec les rails boulonnés sur ce dernier.

En cours de fonctionnement, l'appareil fournit un déplacement d'air d'environ 56 cm³/h dans des conditions de flux d'air raisonnablement frais. Les lames du ventilateur sont actionnées par un moteur fractionnel de 200 h.p/250 V c.a. Il est muni de douilles à fiches (2 × 2BA) à la face d'entrée d'air pour pouvoir monter l'appareil sur les panneaux de ventilation Lektrokit numéro 2 et 3.

L'appareil s'insère dans le même espace qu'un assemblage de châssis Lektrokit normal en coupant les rails LK.201 au trois quarts de leur longueur et en fixant le refroidisseur à l'un des côtés. De cette manière, il peut être fixé dans un bâti standard de 48,26 cm à l'aide de supports de montage normaux Lektrokit (LK.601) afin d'assurer un montage affleurant.

EE 04 770 *pour plus amples renseignements*

INDICATEUR DE VIDE

Edwards High Vacuum Ltd, Manor Royal, Crawley, Sussex

(Illustration à la page 195)

La société Edwards High Vacuum Ltd, a réalisé un nouveau vacumètre Pirani, modèle 9, couvrant la gamme de pressions de 500 à 10^{-3} torr. L'échelle comporte trois gammes ainsi que la commutation automatique d'une gamme à la gamme suivante, des lampes de couleur sur le panneau frontal indiquant la gamme en cours de fonctionnement. Ce vacumètre a été spécifiquement conçu pour la mesure à faible pression dans une variété de processus chimiques métallurgiques et autres qui, bien que n'étant pas exécutés au vide le plus élevé, exigent néanmoins un indicateur électrique qui permette l'enregistrement et la commutation externe.

Le nouvel instrument s'averera également utile dans une gamme étendue d'applications de laboratoire exigeant un vacumètre à plusieurs têtes.

Un régulateur de tension à pont de Wheatstone et un amplificateur différentiel forment le circuit de commande de base qui se caractérise par une résistance de compensation de température bobinée sur les têtes indicatrices. Ce circuit réduit considérablement les effets des variations de température ambiante sur l'étalonnage d'indication. L'inclusion du régulateur de tension élimine en outre la nécessité de tout réglage de tension manuelle. Les boutons-poussoirs sur le panneau frontal permettent de choisir n'importe laquelle des quatre têtes de mesure de la série 9 qui peuvent être reliées au bloc de commande. Les têtes de mesure de la série 9 sont livrables en version pour connexion au système de vide sous verre ou sous métal et elles sont électriquement interchangeables. Les bornes sur le panneau arrière permettent d'entraîner un enregistreur potentiométrique ou un cadre mobile; en outre une alimentation à faible impédance 19 Vc.c. permet d'entraîner une plume d'enregistrement auxiliaire pour indiquer les points de changement de gammes durant l'enregistrement et les contacts de puissance pour la commande de processus. La construction modulaire et les circuits à transistors assurent à l'instrument une compacité qui facilite grandement l'entretien.

EE 04 771 *pour plus amples renseignements*

ASSEMBLAGES À PLUSIEURS DIODES

SGS-Fairchild Ltd, Planar House, Walton Street, Aylesbury, Buckinghamshire

Ces nouveaux assemblages à plusieurs diodes contiennent jusqu'à 16 copeaux dans un transistor TO-5 et comprennent plusieurs combinaisons de ponts à diode, de matrices, de réseaux et de modulateurs annulaires. La société SGS-Fairchild Ltd, offre également un service spécial pour répondre aux demandes de combinaisons supplémentaires pour circuits individuels. L'existence d'une gamme aussi étendue d'assemblages standard a été rendue possible grâce aux caractéristiques du processus planaire au silicium employé par la société SGS-Fairchild. Il permet de produire des diodes à haute conductance et à action rapide d'une uniformité et d'une fiabilité exceptionnelle et insensible à l'usage prolongé dans des conditions d'environnement et d'utilisation les plus astreignantes.

Ces diodes multiples SGS-Fairchild sont encapsulées dans des boîtes TO-18 et TO-5 donnant des dissipations de puissance maxima de 400 mW pour chaque élément. Des températures de fonctionnement et d'emmagasinement allant jusqu'à 200° C peuvent être obtenues et, étant donné que tous les éléments se trouvent dans la même enveloppe, la conductibilité thermique est supérieure à celle obtenue avec d'autres diode et moins affectée par les facteurs externes.

Ces dispositifs présentent en outre l'avantage d'un format réduit car on peut loger jusqu'à 16 diodes dans une seule boîte TO-5. Cela permet des frais d'assemblage considérablement plus réduits étant donné que plusieurs interconnexions sont déjà faites: le dispositif à 16 copeaux ne nécessite que dix connexions au lieu des 32 exigées par les diodes distinctes.

Les types BAX45 à BAX51 sont des matrices à diodes épitaxiales planaires au silicium pour les applications d'entraînement à noyau à action rapide dans les systèmes de calculatrices à mémoire. Il vont des éléments à deux diodes avec trois conducteurs à ceux à 16 diodes et dix contucteurs BAX51.

Les modèles BAX52 et BAX53 sont des ponts à diodes pour les applications à action rapide, contenus dans des boîtes TO-5 et TO-18 respectivement. Les copeaux à diodes ne sont pas équilibrés mais ils sont hermétiquement scellés dans leur encapsulation.

Les modèles BAX54 à BAX55 sont des modulateurs annulaires à diodes épitaxiales planaires au silicium conçues pour les applications dans le matériel multivoies à action rapide. Ils conviennent particulièrement pour l'emploi dans les centraux téléphoniques électroniques.

Enfin, il y a un choix de 18 réseaux de diodes épitaxiales planaires au silicium SFS-Fairchild, type BAX56 à BAX73. Ces derniers ont été conçus pour les applications d'entraînement à noyau à action rapide et ils sont offerts sous forme de réseaux cathodique communs ou anodiques communs dans des combinaisons de 2 à 5 diodes.

EE 04 772 *pour plus amples renseignements*

Résumés des Principaux Articles

"Marge de bruit" dans les systèmes électrologiques binaires par E. L. Jones

Résumé de l'article aux pages 142 à 147

Le terme de "marge de bruit" (ou "immunité de bruit") tel qu'utilisé dans les systèmes électrologiques binaires, a été défini de différentes façons. Chaque définition comporte un certain nombre d'hypothèses. Certaines de ces hypothèses ont été explicitement reconnues mais d'autres ne l'on pas été. Dans de nombreux cas des variantes d'hypothèses conduiraient à des définitions fondamentalement différentes de la marge de bruit; mais dans d'autres cas les hypothèses sont formulées uniquement pour simplifier le problème, ainsi que son exposé, ou encore pour faciliter la mesure, sans affecter les considérations de base en question.

Cet article s'efforce d'expliciter les diverses hypothèses en la matière et d'indiquer celles qui sont fondamentales et introduites uniquement pour des raisons de simplicité. Lorsque des hypothèses fondamentales différentes méritent d'être considérées, les raisons portant à préférer l'une d'elles par rapport à l'autre sont indiquées.

Les considérations sont pour le moment limitées aux marges de bruit (très faible fréquence) "statique" étant donné que jusqu'à ce que ces marges soient parfaitement comprises et qu'on se soit mis d'accord sur leur définition et leur interprétation, il est inutile d'essayer d'étudier les marges de bruit (transitoires) "dynamiques".

Des moyens graphiques sont utilisés pour déterminer et exprimer les marges de bruit statique d'éléments logiques intégrés.

Un détecteur sensible aux phases de commutation de courant et son application à un voltmètre de composantes résolues
par J. G. Lacey

Résumé de l'article aux pages 148 à 151

L'auteur décrit un détecteur sensible aux phases qui utilise des transistors comme commutateur de courant pour obtenir une meilleure linéarité d'amplitude et fournir une gamme de fréquences de fonctionnement étendue. Le signal de référence sinusoïdale est converti en une onde carrée par un circuit utilisant une diode-tunnel comme commutateur hystérétique et employant un circuit asservi pour obtenir un rapport unitaire marque/espace. L'auteur décrit également un voltmètre à composante résolue utilisant ces détecteurs qui affichent simultanément les composantes de quadrature et en phase d'un signal sinusoïdal de courant alternatif par rapport à un signal de référence connexe. L'instrument fonctionne dans la gamme de fréquence de 20Hz à 150kHz.

Boîtes de décades à sortie binaire par D. P. Franklin

Résumé de l'article aux pages 152 à 160

Des circuits ont été réalisés pour la conversion décimale/binaire dans lesquels la logique de combinaison voulue est effectuée entièrement de manière passive. La fonction ET s'effectue à travers la connexion en série de contact de commutateurs à dix directions multipolaires cependant que la fonction OU se fait par l'emploi de routes parallèles, souvent au moyen de diodes.

Deux modes de fonctionnement sont décrits dans lesquels les entrées aux éléments logiques ont, ou n'ont pas, préalablement été mis sous forme binaire.

Un bloc d'alimentation à haute tension stabilisé de secteur utilisant des redresseurs d'avalanche commandée par R. W. J. Barker

Résumé de l'article aux pages 161 à 165

Une nouvelle technique utilisant des redresseurs d'avalanche commandée pour produire une alimentation à haute tension de faible puissance stabilisée de secteur est décrite.

On peut ainsi calculer la performance de l'alimentation par l'emploi de la configuration double à tension conventionnelle ainsi qu'une méthode pour spécifier ces composants. Les graphiques produits permettent l'évaluation des équations qui en résultent.

Le rendement pratique d'un bloc d'alimentation de laboratoire utilisant le processus qui en résulte est comparé avec la performance théorique.

Un dispositif électromécanique pour mesurer la valeur moyenne de marche par K. G. Ng et K. K. Murthy

Résumé de l'article aux pages 169 à 171

Les auteurs décrivent un nouvel appareil électromécanique pour mesurer les valeurs moyennes de marche. Il s'agit d'un dispositif peu coûteux de construction simple et d'une grande fiabilité de fonctionnement. Il est particulièrement utile lorsqu'on veut simuler sur une calculatrice analogique la moyenne de marche de signaux.

Détermination de la non linéarité par la simulation analogique par G. W. Holbrook et J. S. Collins

Résumé de l'article aux pages 172 à 173

La plupart des dispositifs non-linéaires peuvent être décrits par une série de puissances reliant les courants d'entrée et de sortie ou les tensions. Les coefficients des termes précédents et significatifs peuvent être obtenus en synthétisant la forme d'onde de sortie du dispositif au moyen d'une séquence de circuit intégrateur utilisant un signal d'entrée de rampe. La sortie de chaque circuit intégrateur peut être utilisée pour fournir un report direct du coefficient correspondant dans une gamme de tensions de polarisation.

La modulation d'impulsions de pente variable par O. E. Kruse, D. L. Lewis et J. K. Jackson

Résumé de l'article aux pages 174 à 176

Cet article décrit un système de modulation comportant deux canaux d'information portés par le même train d'impulsions triangulaires; l'un des canaux est porté par les pentes d'établissement et l'autre par les pentes d'affaiblissement.

Un intégrateur d'appareil de concentration de données et un comparateur de limite Alcom par R. J. S. Ruckman

Résumé de l'article aux pages 177 à 183

Cet article décrit un dispositif d'intégration par la méthode de comptage, l'intégration tenant compte des décalages de zéro ainsi que des valeurs positives et négatives. Les limites d'alarme sont comparées avec les valeurs intégrées pendant que s'effectue le comptage et on peut obtenir des renseignements permettant de déterminer si une variable donnée se trouve en état d'alarme ou non.

NEUE AUSRÜSTUNGEN

Übersetzung der Seiten 190 bis 195

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Vorverstärker mit hohem Auflösungsvermögen

High Volt Linear Ltd, 67 Dudley Street, Luton, Bedfordshire

(Abbildung Seite 190)

Der FET-Vorverstärker 118 ist eine Ergänzung des ORTEC-Programms ladungsempfindlicher Verstärker. Der in erster Linie für Gammaspektrometrie mit hoher Auflösung bestimmte Vorverstärker ist für Verwendung mit gekühlten Halbleitermessköpfen niedriger Kapazität und Streuverluste ausgelegt, wie z.B. die Serie ORTEC 8000 oder andere ähnliche lithiumdotierte Germanium-Messköpfe. Die Ladungsverstärkung, die auf $\times 1$ oder $\times 10$ umstellbar ist, ist in der $\times 10$-Position 365 mV/MeV(Ge), und das Auflösungsvermögen ist besser als 1,25 keV. Der Vorverstärker 118 ist mit dem modularen Instrumentierungssystem ORTEC 400 kompatibel und kann entweder direkt vom linearen Verstärker 410 oder von einem getrennten Netzanschluss 115 betrieben werden.

Die Aussenabmessungen des Vorverstärkers 118 sind 46 $\times$ 102 $\times$ 152 mm; bei Verwendung mit ORTEC-Messköpfen kann er so an den Kryostaten angebaut werden, dass die Kapazität des Vorverstärkereingangs auf ein absolutes Minimum reduziert wird.

EE 04 751 *für weitere Einzelheiten*

Miniaturpotentiometer

Davall Electronics Ltd., Rothersthorpe Avenue, Northampton

(Abbildung Seite 190)

Das Davall-Miniaturpotentiometer 72 mit gepresster Schicht soll eine wesentlich längere Lebensdauer haben als die herkömmlichen mit aufgespritzter Schicht, die in der kommerziellen Elektronik zur Zeit üblich sind.

Isolierkörper, Vollkohleschicht und Klemmen sind beim Typ 71 in ein integriertes Bauelement gepresst. Diese Bauart gewährleistet hohe Stabilität in allen Feuchtigkeits-, Temperatur- und Vibrationszuständen. Während der ganzen Lebensdauer ist der Rauschpegel viel niedriger als der von Potentiometern mit aufgespritzter Schicht und reduziert mit Benutzung. Die Nennbelastbarkeit des Typs 72 ist bei 55° C 0,5 W (linear) und 0,25 W (nichtlinear). Widerstandswerte sind zwischen 50 Ω und 2 MΩ (linear) lieferbar. Die Höchstbetriebsspannung ist 500 V—.

Das Potentiometer kann mit drei Fahnen direkt in Druckschaltungen mit 2,54 mm Rastermass eingelötet werden und ist kompakt genug für Aussenabmessungen von 19 mm Durchmesser und 25,4 mm Tiefe (ohne Lötfahnen).

EE 04 752 *für weitere Einzelheiten*

Trennverstärker

Ether Ltd, Caxton Way, Stevenage, Hertfordshire

(Abbildung Seite 190)

Der Trennverstärker BA1 wurde in erster Linie für die Verbindung zwischen piezoelektrischen Beschleunigungsgebern und ihren Mess- und Registriereinrichtungen entwickelt, bei denen es zur Minderung des Belastungseffektes auf Eingangsimpedanzen in der Grössenordnung von 1 GΩ ankommt. Man kann ihn für Signalnachweis auch dort einsetzen, wo es auf vernachlässigbar niedrige Belastung ankommt, vorausgesetzt dass die Eingangssignalamplitude innerhalb der gegebenen Grenzen liegt.

Das Gerät ist ein zweistufiger Verstärker in Festkörpertechnik mit Nennverstärkung von 1:1 und einem ungewöhnlich hohen Impedanzwandlerverhältnis, für das 40 $\times$ 10^6 typisch ist. 100 Prozent Spannungsgegenkopplung über beide Stufen erhöht die Impedanz des Eingangs und reduziert die des Ausgangs. Die erste Stufe ist ein FET mit hochohmigem Eingang, die zweite ein Silizium-Legierungsflächentransistor, der den erforderlichen Ausgangsstrom abgibt.

EE 04 753 *für weitere Einzelheiten*

Magnetband-Pendeleinrichtung

James Scott (Electronic Engineering) Ltd, Carntyne Industrial Estate, Glasgow, E.2

(Abbildung Seite 190)

Bei der Aufzeichnung von Versuchsdaten oder anderer Information mit Hilfe des Bandgerätes der Messausrüstung kann die so aufgezeichnete Information über grosse Bandlängen verteilt sein, während die für den Anwender wichtige Information nur einen kleinen Teil dieser Bandlänge in Anspruch nimmt. Ein weiteres Problem tritt auf, wenn der Benutzer diesen Teil des Bandes viele Male hintereinander lesen will.

Die Bandpendeleinrichtung macht vom Fernsteuerungsanschluss des Bandgerätes Gebrauch und wird—nach elektronischem Markieren des gewählten Bandabschnittes durch die Bedienung—diesen vorgewählten Abschnitt lesen, das Bandgerät am Ende dieses Abschnittes stoppen, den Rücklauf zum Anfang des Abschnittes einleiten, diesen Abschnitt automatisch noch einmal wiedergeben, usw. Diese Vorgänge werden automatisch wiederholt, ohne dass die Bedienung die Bandgerätsteuerung berührt. Nach Beendigung der Auswertung braucht man den Schalter der Pendeleinrichtung lediglich von "automatisch" auf "manuell" umzustellen. Die manuellen Steuerungselemente des Bandgerätes werden dann wieder wirksam, und das Gerät kann abgestellt werden.

Das Markieren des gewünschten Bandabschnittes erfolgt mit der Pendeleinrichtung und unter Benutzung eines Kanals des Bandgerätes; wenn die Bedienung eine Taste drückt, wird in diesem Kanal ein Impuls feststehender Länge aufgedrückt. Dieser Impuls wird vorzugsweise bei der Bandgeschwindigkeit aufgezeichnet, bei der die Wiedergabe erfolgen soll.

Die Pendeleinrichtung ist ein in sich geschlossenes Gerät mit eigenem Netzteil, lässt sich aber auch von einer externen 28-V-Gleichspannungsquelle treiben.

Es kann mit Geschwindigkeiten von 4,8 cm/s bis zu 152 mm/s gependelt werden. Das Gerät ist 292 mm breit, 280 mm hoch und 343 mm tief.

EE 04 754 *für weitere Einzelheiten*

Hochintensität-Mikrofon

D. J. Birchall, 5 Leys Avenue, Letchworth, Hertfordshire

(Abbildung Seite 191)

Diese Kristallmikrofonkapseln gibt es mit Durchmessern von 9,5 bis zu 35 mm und mit Ladungsempfindlichkeiten von

40 pC bis zu 2000 pC je 7 × 10^4 Dyn/cm². Diese Elemente sind für Betrieb in Umgebungen mit hohem Raumdruck und Temperaturen bis zu 250° C geeignet und gegen Vibration unempfindlich. Das Gehäuse ist aus rostfreiem Stahl gefertigt, die Membrane angeschweisst. Die Abbildung zeigt ein angeflanschtes Modell von 35 mm Durchmesser.

EE 04 755 *für weitere Einzelheiten*

Mechanische Filter

Vertrieb: Impectron Ltd, Impectron House, 125 Gunnersbury Lane, London, W.3

(Abbildung Seite 191)

Mechanische Miniaturfilter und Anpassungsübertrager der Toko Inc, Tokio, Japan, können nunmehr von Impectron Ltd bezogen werden. Sie sind hauptsächlich für hochwertige Röhren- und Transistorempfänger, selektive Radiorufsysteme, Funkfernsprecher und ähnliche Ausrüstungen bestimmt. Die Selektivität ist ausserordentlich gut und soll der herkömmlichen LC-Schaltung wesentlich überlegen sein. Die Abmessungen, über die Anschlussstifte um 4 mm herausragen, werden als 42 mm lang, 10,7 mm breit und 10,2 mm tief gegeben.

Insgesamt sind fünf Modelle lieferbar, und zwar drei mit 455 ±1 kHz und zwei mit 455 ±1,5 kHz. Die Bandbreiten der ±1-kHz-Ausführungen sind bei −6 dB 5 ±1 kHz, 6 ±2 kHz und 12 ±1 kHz, bei −40 dB 16 kHz, 18 kHz bzw. 28 kHz. Für die 1,5-kHz-Ausführung sind die Bandbreiten bei −6 dB 16^{+3}_{-0} kHz und 22^{+3}_{-0} kHz und bei −40 dB 36 kHz bzw. 42 kHz. Die Welligkeit ist innerhalb 3 dB und Frequenzverschiebung unter 150 × 10^{-6}/°C. Die höchste Signaleingabe in den Wandler ist 5 V bei 455 kHz, der Betriebstemperaturbereich −10°...+70°C und das Gewicht etwa 9 g.

Grundelemente sind ein unterteilter Metallstabresonator und zwei Wandler. Die Kupplung der Resonanzabschnitte erfolgt ohne geschweisste Teile, so dass Verluste auf ein Mindestmass reduziert werden. Die Treiber- und Abtastwandler bestehen aus leistungsfähigen, polarisierten Keramikscheiben, die zuverlässig an den Resonatorenden befestigt sind.

Spezial-Anpassungsübertrager sind für Anwendung als Filtereingang und -ausgang lieferbar. Sie gewährleisten maximale Energieübertragung, niedrigste Welligkeit und können für Röhren- oder Transistorschaltungen bestellt werden. Übertrager für den Filtereingang sind für bis zu 30 kΩ ausgelegt, die für den Ausgang haben 800 Ω Impedanz. Ihre kleinen Abmessungen entsprechen denen der Filter.

EE 04 756 *für weitere Einzelheiten*

Drehschalter

A.B. Metal Products Ltd, North West House, 119/127 Marylebone Road, London, N.W.1

(Abbildung Seite 191)

A.B. Metal Products Ltd hat eine neue Baureihe hochwertiger Drehschalter eingeführt, die auf Langzeit-Zuverlässigkeit gebaut und für professionelle sowie militärische Anwendungszwecke geeignet sind: MINI-12 (12 Schaltstellungen), MINI-24 (24 Stellungen) und PY (22 Klammern/12 Stellungen).

Das neu durchkonstruierte Rastwerk hat einen Schaltkranz aus gesintertem Eisen und gehärtete Rastbolzen, die zuverlässiges Arbeiten und lange Lebensdauer gewährleisten. Der Rasthebel kann entweder auf einer Seite oder beiden Seiten des Schaltkranzes angebracht werden, was mit den verschiedensten Spannfedern eine äusserst breite Auswahl von Achsendrehmomenten ergibt.

Gepresste Statoren und Rotoren ergeben hohen Isolationswiderstand. Schwer versilberte Kontaktklammern und Rotormesser aus hochwertigem NE-Federwerkstoff ergeben niedrigen Kontaktwiderstand. Abstand haltende Schaltebenen mit Kronierung über die gesamte Tiefe versteifen die Kontaktklammern.

Der Isolationswiderstand liegt über 50 GΩ; bei normaler Temperatur und normalem Druck ist die Überschlagfestigkeit zwischen elektrischen Kontakten und üblicherweise geerdeten Teilen 2 kV und zwischen Kontakten, die voneinander isoliert sein sollen, 1 kV; der Kontaktwiderstand ist 4...6 mΩ, der höchste Schaltstrom bei ohmscher Last 50 mA bei 300 V≃ oder 500 mA bei 30≃, die Strombelastbarkeit 2 A.

Die Typen MINI-24 und PY sind mit Doppelachse lieferbar.

Elektrostatische Abschirmung, Lagerschellen und Netzschalter sind Standardzubehör.

EE 04 757 *für weitere Einzelheiten*

Drehpotentiometer

Ferranti Ltd, Thornybank, Dalkeith, Midlothian, Schottland

(Abbildung Seite 191)

Weiterentwicklung der vor kurzem eingeführten Präzisionsdrahtpotentiometer 11HL gestattet Ferranti Ltd, Einzelheiten mit Sinus-Kosinus-Funktion herzustellen. Sie sind in Einzel- oder Mehrfachausführung mit bis zu sechs Einheiten lieferbar. Das Drehmoment ist nur 3 g.cm per Einheit, und die Übereinstimmung mit der Nennkurve ist besser als ±1 Prozent Spitze-zu-Spitze.

Widerstandswerte gibt es von 50 Ω bis zu 50 kΩ per Quadrant, und die Betriebstemperatur ist −65°...+150° C. Alle Schleifer, Schleifringe und die meisten Wicklungen sind zwecks besserer Auflösung und niedrigen Rauschens aus Edelmetallen hergestellt; Abgriffe sind jeweils an eine Windung geschweisst.

Auf Grund ihrer Spezialwickelmethoden kann Ferranti Ltd auch andere trigonometrische und empirische Kurvenverläufe für beispielsweise Belastungsausgleich herstellen.

EE 04 758 *für weitere Einzelheiten*

Geber für absoluten Druck

Electro Mechanisms Ltd, 218-221 Bedford Avenue, Slough, Buckinghamshire

(Abbildung Seite 192)

Eine neue Baureihe von Gebern für absoluten Druck mit nur 31,8 mm Durchmesser wird von Electro Mechanisms bekanntgegeben. Die völlig aus rostfreiem Stahl hergestellten Messwandler sind für die meisten Druckbereiche von 3,5...350 kg/cm² lieferbar. Interne mechanische Überlastanschläge schützen die Geber und gestatten Belastung mit bis zu 5mal Nenndruck mit weniger als 0,2 Prozent Nullpunktänderung. Die Geber sind für Betrieb im Temperaturbereich −196° C...+121° C lieferbar, in dem sie ihre hohe Genauigkeit beibehalten. Über den temperaturkompensierten Bereich beträgt der Fehler weniger als 0,01 Prozent. Die angewendeten Messelemente sind Halbleiter-Dehnungsstreifen, die bei niedrigem Widerstand die hohe elektrische Ausgangsspannung von 150 mV abgeben. Der Geber kann direkt Mess- und Registriergeräte treiben, so dass Verstärker, die Fehlerquellen darstellen können, überflüssig werden. Die kombinierte Nichtlinearität und Hysterese von 0,15 Prozent gestattet Anwendung in hochgenauen Systemen.

Diese Druckgeber werden von Schaevitz Bytrex Inc in Massachusetts hergestellt.

EE 04 759 *für weitere Einzelheiten*

Schwingungsmonitorsystem

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Abbildung Seite 192)

Das System 1435B ist für laufende und automatische Überwachung der im Betrieb auftretenden Schwingungen einer breiten Auswahl von Anlagen und Maschinen bestimmt und löst entweder einen Alarm oder einen Selbstschutz aus, wenn Schwingungen einen vorgewählten Wert überschreiten. Drei Grundinstrumentmoduln gestatten Anpassung an die verschiedensten Betriebsbedingungen. Es gibt als Moduln den Schwingungsmonitor 1435/1B, Messgerät 1435/2 und sieben Filter 1435/3B... 1435/9B. Sie können durch Einstecken in ein vorverdrahtetes Chassis 1435/11B in jeder gewünschten Anordnung als den vorgezogenen Betriebsbedingungen entsprechendes System zusammengeschaltet werden.

Der Monitor-Modul wird normalerweise durch Leitungen mit dem in einem gewählten Punkt der Maschine oder Struktur montierten geschwindigkeitsempfindlichen Messwandler verbunden. Jedoch kann man auch das Signal eines Beschleunigungsgebers in den Monitor speisen, dem 27 V – für die Erregung von Gebern oder Quellengebern entnommen werden können. Das laufend abgegebene elektrische Ausgangssignal des Monitors ist der Spitzenamplitude der Maschine im Frequenzbereich 10 Hz...800 Hz (600...48 000 UPM entsprechend) proportional. In jedem Monitor gibt es zwei Relais, die sich für Ansprechen bei jedem gewünschten Ausgangspegel einstellen lassen. Sie können entweder mit Arbeitskontakten zur Auslösung eines Alarms oder mit Ruhekontakten für Eigenschutzeinrichtungen bestückt werden. Da Relais- und Verstärkerschaltungen getrennt sind, wird auch nach Auslösung des ersten Relais weiter überwacht. Eine Verstärkungsänderung von 10:1 nach der ersten Warnung gestattet Veranlassung der zweiten Warnung oder des Abschaltens bei bis zu zehnfachem Amplitudenhub.

Für die Überwachung eines bestimmten Teiles, beispielsweise eines Lagers oder eines Zahnrades, kann der Eingang durch ein Filter gespeist werden, das dem Monitor einen bestimmten Frequenzgang gibt. Für die überwachungsmässige Erfassung der Schwingungsgrenzen, z.B. Anzeige der veränderlichen Lasten einer Fertigungsanlage oder Werkzeugmaschine, kann der Ausgang des Monitors mittels eines Messgerätes angezeigt werden. Ein lieferbares Chassis 1435/11B für Einbau in 19″-Gestelle ist 222 mm hoch und 267 mm tief. Es kann vier Moduln von 216 × 108 × 203 mm tief aufnehmen. Messwandler, Stromversorgungen, Relaiskontakte, externe Mess- und Registriergeräte werden hinten angeschlossen.

Das System kann mit bis zu vier, von vier Messwandlern gespeisten Monitoren betrieben werden, oder mit jeder beliebigen Kombination von Monitoren, Messgeräten und Filtern bis zu insgesamt vier. Das Messgerät hat drei, auf Wunsch sieben Kanäle, das Filter zwei. Für Anordnungen, in denen drei oder weniger Monitoren erforderlich sind, gibt es Leerfelder in Modulabmessungen.

Der Monitorverstärker kann bei Nennverstärkung für 0,025 mm Spitze zu Spitze bis zu 0,5 mm Spitze zu Spitze eingestellt werden. Der übliche Frequenzgang zwischen 10 und 800 Hz lässt sich erweitern. Das Gerät ist für Geschwindigkeitsgeber mit Empfindlichkeiten von etwa 4 mV cm/s bis zu 60 mV cm/s geeignet. Die interne Integrierung ergibt eine der Amplitude proportionale Spannung. Dawe liefert auch einen Drehspulgeber 1403/1 für Anwendung in Temperaturen bis zu 70° C.

EE 04 760 *für weitere Einzelheiten*

Starkstromrelais

Magnetic Devices Ltd, Exning Road, Newmarket, Suffolk

(Abbildung Seite 192)

Dieses aus Spule und Joch des Standard-Relaistypes P.O. 3000 bestehende Starkstromrelais ist mit Starkstromkontaktmessern und -kontakten bestückt, wodurch ein Mehrkontaktrelais mit höherer Empfindlichkeit als für diese Schaltkapazität üblich ist, entsteht, das von den britischen Elektrizitätswerken mit Zertifikat Nr. 173 zugelassen wurde.

Das Relais C.350 für Gleichstrom genügt den Anforderungen der Klasse I und der Schwachstromklasse für Anwendung in Erzeugung, Übertragung und Kerntechnik.

Die Serie C.350 hat verschiedene interessante Konstruktionsmerkmale, darunter speziell entwickelte Messerkontakte, die Prellen auf ein Minimum reduzieren und die Lebensdauer des Relais verlängern, sowie reichlich bemessene Abstände der Kontakte, die Anlegen eines maximalen zulässigen Potentials zwischen Kontakten gestatten.

Die Relais können 2...6polig, mit Arbeits-, Ruhe- oder Umschaltkontakten bestückt und mit einer typischen Ansprechzeit von 70 ms und Abfallzeit von 15 ms geliefert werden. Umgebungstemperaturen bis zu 35° C sind zulässig, Anpassung an höhere Temperaturen ist auf Kundenwunsch möglich.

Die Feinsilberkontakte sind für ohmsche Belastung bis zu 10 A bei 30 V – oder 250 V∼, mit Elkonite D54X für ohmsche Belastung bis zu 15 A bei 30 V– oder 250 V∼ bemessen. Die Isolation zwischen Spule und Erde ist 1500 V_{eff}, von Kontakt zu Kontakt 2000 V_{eff}, zwischen Kontakt und Erde 2500 V_{eff}.

Die Kontaktmessverfahren sind für Lötverbindungen oder "Schnellverbindungsbuchsen" Baumuster 250 geeignet. Spulenfahnen eignen sich nur für Lötverbindungen, können aber an zwei Leerfahnen des Kontaktpaketes angeschlossen werden, wenn schneller Buchsenanschluss erwünscht ist.

EE 04 761 *für weitere Einzelheiten*

Dezimal-Ziffernskalen

Moore Reed (Industries) Ltd, Walworth Industrial Estate, Andover, Hampshire

(Abbildung Seite 193)

Die von Moore Reed & Co Ltd hergestellten "Digikit"-Moduln gestatten den Aufbau elektrischer Skalen von unterschiedlicher Länge und Auflösungsvermögen. Die Skalenanzeige kann entweder digital erfolgen oder auf Magnetband usw. registriert werden. Positionen für programmierte Steuerung sind vorwählbar.

Durch Anordnung eines "Digishift"-Schalters zwischen der Skala und der Anzeige wird elektrisches Verschieben zu jeder gewünschten Position oder Nullunterdrückung möglich, ohne dass mechanisches Abkuppeln der Skala von den kontrollierten beweglichen Teilen notwendig wird. Die Skala kann auch jederzeit elektrisch umgekehrt oder für Anzeige von Werten mit Nullpunktunterdrückung eingesetzt werden.

Die Abbildung zeigt einen typischen "Digikit"-Zusammenbau für 1000 Teilungen für 10 Umdrehungen der Eingangswelle. Auflösungen bis zu 1000 je Umdrehung und Skalenlängen von mehr als 1 000 000 Teilungen lassen sich durch geeignete Kombination von vorhandenen Komponenten erstellen.

EE 04 762 *für weitere Einzelheiten*

Kohleschichtwiderstände

Vertrieb: Standard Telephones & Cables Ltd, Rectifier Division, Edinburgh Way, Harlow, Essex

(Abbildung Seite 193)

Die Rectifier Division der Standard Telephones and Cables Ltd hat ihr Standardsortiment der westdeutschen Beyschlag - Kohleschichtwiderstände durch hochwertige Qualitätstypen ergänzt, die das bestmögliche Leistung : Preis-Verhältnis bieten sollen, um dem Bedarf an Präzisionswiderständen hoher Konstanz und Zuverlässigkeit bei wirtschaftlichem Preis zu genügen.

Der Geschäftsbereich kann nunmehr hochkonstante Kohleschichtwiderstände mit ± 1 Prozent Toleranz, ¼ W und ½ W Belastbarkeit in den bevorzugten Werten der IEC-E24-Reihe zwischen 51 Ω und 1 MΩ liefern. Ausführungen mit gleichen Kenndaten, aber ±2 Prozent Toleranz sind von 10 Ω bis zu 1 MΩ, in der ½-W-Grösse bis zu 2,2 MΩ zu haben. Ausserdem gibt es einen Typ für ⅛ W mit ± 1 Prozent Toleranz in E-24-Werten von 51 Ω bis zu 68 kΩ und mit ±2 Prozent Toleranz von 10 Ω bis zu 240 kΩ. Diese Bauelemente sind alle für Betrieb bei 70° C bemessen.

Als Alleinvertreter in Grossbritannien hält die Rectifier Division der STC Beyschlag-Kohlewiderstände in grösseren Mengen auf Lager und kann in den meisten Werten Erprobungsquantitäten bis zu 200 vom Lager abgeben.

Beyschlag-Widerstände werden zur Unterstützung gerätebauender Kunden unter dem Namen "Autoband" in Bandverpackungen geliefert, so dass sie direkt von der Packung in die automatische Schneid- und Biegemaschine für die Anschlussdrähte, die die Bauelemente für die Bestückung gedruckter Schaltungen vorbereiten, gehen können.

EE 04 763 *für weitere Einzelheiten*

Digitaluhr

Darang Electronics Ltd, Restmor Way, Hackbridge Road, Hackbridge, Surrey

(*Abbildung Seite 193*)

Die elektronische Digitaluhr Modell "Six-Seventy" ist die neuste im "Digicron"-Programm der Darang Electronics Ltd.

Das für allgemeines Zeitmessen in 12- oder 24-Stunden-Systemen ausgelegte Modell "Six-Seventy" hat eine digitale Grossanzeige, die selbst bei heller Umgebungsbeleuchtung aus über 12 m Entfernung lesbar ist. Das modern gestaltete Gehäuse gibt es in vier Farben.

Die Zeitzählung wird durch die 50-Hz-Netzfrequenz oder eine gegen Aufschlag eingebaute externe Kristallquelle gesteuert. Bei Ausfall der Netzspannung läuft die Uhr nicht wieder an, bis sie zurückgestellt ist. Die Zeit wird mittels Einknopfbedienung eingestellt, und Betätigung von nur einem Schalter programmiert die Uhr entweder für das 12- oder das 24-Stunden-System.

EE 04 764 *für weitere Einzelheiten*

Widerstände mit positivem Temperaturkoeffizienten

Thorn Parsons Co. Ltd, 40 Broadway, London, S.W.1

(*Abbildung Seite 193*)

Thorn Parsons Co Ltd kündigt neue Posistors—Widerstände mit wesentlich erhöhtem positiven Temperaturkoeffizienten—an.

Eigenschaften des Posistors sind ein fast konstanter Widerstand bis zur Schwellentemperatur und sehr starker Anstieg bei höheren Temperaturen. Durch diese Eigenschaft sowie die Möglichkeit, die Schwellentemperatur durch Änderung der chemischen Komposition zu beeinflussen, ist der Posistor für viele Schutz- und Überwachungssysteme, vor allem Schutz von Elektromotoren, geeignet.

Moderne Elektromotoren laufen bei Temperaturen, die dem Wert, bei dem Schaden entsteht, viel näher liegen als früher üblich. Zum Schutz dieser Motoren gegen Überhitzen bietet der Posistor einen hohen Zuverlässigkeitsgrad. Kleine Abmessungen—in Spezialgehäusen bis zu 1,6 mm Durchmesser herunter—gestatten Einbau in die Motorwicklungen, d.h. direkt an der heissen Stelle, so dass sie schneller auf Gefahrenzustände ansprechen als die bisher angewendeten Bimetallschalter. Die thermische Zeitkonstante ist ungefähr 3 s, und weitere Vorteile werden durch die Abwesenheit beweglicher Teile erzielt.

Für Einbau in Motoren werden die Elemente üblicherweise mit Epoxydharz isoliert. Der grösste Bedarf besteht für Posistors und Regelgeräte, die bei 120° C (britische Norm BS. 2757 Isolierung Klasse E) und 130° C (Klasse B) schalten. Posistors sind mit anderen Isolierungsarten und für Betriebstemperaturen von 95°...165° C lieferbar.

Posistors können ausser für Temperaturüberwachung auch für Strombegrenzung und das Fühlen von Änderungen in der Umgebung von Elementen wie z.B. Feuchtigkeits-Dampfübergang, die Rate von Durchströmungsänderungen oder Flammenrückschlag, eingesetzt werden.

EE 04 765 *für weitere Einzelheiten*

Stromversorgungen

Irwin & Partners Ltd, Kibworth Street, Dorset Road, London, S.W.8

(*Abbildung Seite 193*)

Zielsetzung für die neuen Geräte war die Entwicklung universeller, in sich geschlossener Arbeitsplatz-Stromversorgungen für Labors, Schulen usw. zu mässigen Preisen. Die Ausgangsspannung ist in dem mittels Regelumspanner gewählten Bereich stufenlos einstellbar, und in allen niedrigen Bereichen ist der Ausgang durch einen getrennten Umspanner vom Netz isoliert. Es werden sieben Modelle mit verschiedenen Spannungsbereichkombinationen und Verfeinerungen hergestellt.

An allen Geräten, mit Ausnahme der Modelle A.1. und B.1., wird der gewünschte Spannungsbereich mittels eines Drehschalters gewählt, der mechanisch mit dem Ausgangsschalter verriegelt ist, so dass man nicht mit angeschlossener Last auf einen höheren Spannungsbereich umschalten kann.

Die Gleichrichtung der Gleichstrombereiche erfolgt mit Hilfe von Siliziumdioden. In der Glättung wird zur Erzielung angemessener Regelung eine niederohmige Drossel verwendet.

Alle Messgeräte sind Vielbereich-Drehspulinstrumente, die durch den Ausgangsbereichschalter automatisch auf den richtigen Bereich umgeschaltet werden. Wo keine Messgeräte eingebaut sind, gestatten Buchsen Anschluss externer Instrumente. Für Strommesser ist ein Überbrückungsschalter eingebaut.

Zur Zeit sind folgende Geräte lieferbar:

Modell A1: Ausgänge 0...250 V~ bei 1 A
0...250 V— bei 1 A
ungeglättet
Keine eingebauten Messgeräte

Modell B1: Ausgänge 0...250 V~ bei 1 A
0...250 V— bei ½ A
geglättet
Keine eingebauten Messgeräte

Modell C1: Ausgänge 0...250 V~ bei 1 A
0...250 V— bei ½ A
geglättet
Eingebaute Strom- und Spannungsmesser

Modell D1: Ausgänge 0... 25 V— bei ½ A
geglättet
0...250 V— bei ½ A
geglättet
0... 25 V~ bei 1 A
0...250 V~ bei 1 A
Eingebaute Strom- und Spannungsmesser

Modell E1: Ausgänge 0...250 V~ bei 1 A
5... 25 V— bei ½ A
stabilisiert
0...250 V— bei ½ A
geglättet
Keine eingebauten Messgeräte

Modell F1: Ausgänge 0... 25 V~ bei 1 A
0...250 V~ bei 1 A
5... 25 V— bei ½ A
stabilisiert
0...250 V— bei 1 A
geglättet
Eingebaute Strom- und Spannungsmesser

Modell H1: Ausgänge 0... 12 V— bei 1 A
geglättet
0...500 V— bei 100 mA
geglättet
5 V~ bei 3 A
6,3 V~ bei 3 A
Eingebaute Strom- und Spannungsmesser

EE 04 766 *für weitere Einzelheiten*

Höchstspannungs-Stromversorgung

Linstead Electronics Ltd, 35C Newington Green, London, N.16

(*Abbildung Seite 194*)

Die Ausgangsspannung ist von 0 ... 6 kV stufenlos regelbar und wird auf einem Instrument angezeigt. Die Höchstströme sind etwa 1 mA bei 5 kV, 1,3 mA bei 4 kV, 2...3 mA bei niedrigeren Spannungen. Für die Heizung von Elektronenröhren und anderen Glühkathodenröhren werden 3 A bei 6,3 V abgegeben. Jedes Ende der Höchstspannung kann geerdet und die Heizspannung auf Wunsch bei Anwendung der Höchstspannung isoliert werden. Die Höchstspannungsquelle ist ein HF-Oszillator, was niedrige Glättungskapazitäten erfordert und niedrige Kurzschlussströme ergibt. Das Gerät kann daher keinen anhaltenden und gefährlichen Schlag verursachen.

EE 04 767 *für weitere Einzelheiten*

Potentiometer

A.B. Metal Products Ltd, North West House, 119-127 Marylebone Road, London, N.W.1

(*Abbildung Seite 194*)

A.B. Metal Products Ltd kündigt neue Kohleschicht-(Composition-) Potentiometer der Serie 45 an, die mit 1/2 W belastbar sind und einen Durchmesser von 23,8 mm haben.

Die neue Konstruktion und Einführung modernster Fertigungsmethoden mit automatischer Montage ermöglichen Fertigung mit gleichbleibender hoher Qualität und Abgabe zu äusserst wirtschaftlichen Preisen.

Die Potentiometer werden in Standardwerten von 500 Ω bis zu 10 MΩ mit linearen, logarithmischen, halblogarithmischen, negativlogarithmischen und anderen Regelkurven mit ein oder zwei Anzapfungen, mit Zentralbefestigung, Einsteck- oder Würgemontage, Standardlötfahne, Anschlüssen für Druckschaltungen oder Wickelverbindungen, in Tandem- oder konzentrischer Ausführung geliefert.

Die mechanische Vielseitigkeit der Serie 45 gestattet fast unbegrenzte Kombinationen, was dem Entwurfsingenieur enormen Spielraum gibt. Die durchweg mit gepressten Achsen ausgerüsteten Regler sind mit integraler Frontplatte für Einsteckmontage und Anschlüssen für Wickelverbindung lieferbar. Das erleichtert Einbau und Auswechseln der Bauelemente, was zu wesentlichen Einsparungen an Zeit und Geld führen kann.

Die Potentiometer können mit Netzschaltern ausgerüstet werden, die für Betrieb mit den in siliziumgleichrichterbestückten Stromkreisen auftretenden hohen Stromstössen geeignet sind.

EE 04 768 *für weitere Einzelheiten*

Differentialverstärker für Oszillografen

Hewlett-Packard Ltd, 224 Bath Road, Slough, Buckinghamshire

(Abbildung Seite 194)

Ein neuer galvanisch gekoppelter, vertikaler Oszillograf-Verstärkereinschub von Hewlett-Packard misst ohne Drift den Mikrovoltpegel von Wechselstromsignalen in Gegenwart hoher Gleichstrompegel. Sowohl Wechsel- wie Gleichstrommessungen werden mit 0,4 Prozent Unsicherheit durchgeführt. Der Einschub hat eine Höchstempfindlichkeit von 50 μV/cm. Mit Hewlett-Packards neuer Nulldrift und der mit hoher Auflösung unterteilten Gleichstromverschiebung kann man leicht Oszillografenmessungen durchführen, die früher unmöglich waren, und andere lassen sich bequemer ausführen.

Der neue HP-Differentialverstärkereinschub 1406A passt in das HP-Oszillografen-Hauptgestell 140A und ist mit dem vor kurzem eingeführten HP-Speicheroszillografen 141A mit veränderlicher Nachleuchtdauer noch nützlicher.

Durch die geeichte Verschiebeeinrichtung wird das Modell 1406A zu einem genauen Voltmeter. Die vier bedeutsamsten Ziffern werden angezeigt, und Lampen auf der Frontplatte geben die Kommaposition. Im Höchstempfindlichkeitsbereich (50 μV/cm) gibt der Verstärker Verschiebespannungen bis zu 10 V ab, die in den weniger empfindlichen Bereichen bis auf 1000 V heraufgehen.

Der Eingang des Verstärkers lässt sich differential ohne Verschiebung benutzen. Bei Differentialbetrieb werden Störspannungen wie z.B. 50-Hz-Brummen mit 85 dB oder mehr unterdrückt.

Die maximale Bandbreite des Differentialverstärkers ist 400 kHz. Wenn hochfrequente Signalkomponenten unerwünscht sind, kann die Bandbreite von der Frontplatte aus auf 100, 25 oder 5 kHz reduziert werden.

Ausserdem ist ein Trennverstärker zum Treiben externer Geräte wie beispielsweise X-Y-Schreiber oder Magnetbandgerät eingebaut.

Die Erdung des Verstärkers kann mittels Frontplattenschalter entweder nur an die Umschaltung oder an das Oszillografenchassis gelegt werden. Das hilft bei der Beseitigung von Erdschleifen zur Erzielung klarerer Bilder und höherer Messgenauigkeit.

EE 04 769 *für weitere Einzelheiten*

Kühlaggregat

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

(Abbildung Seite 195)

Der Bereich Lektrokit der A.P.T. Electronic Industries Ltd kündigt eine Ergänzung des Bauelementesortiments für ihr Gestell- und Chassis-Bausystem Lektrokit an.

Der Bauteil LKU.511 wurde in erster Linie als Kühlaggregat für elektronische Ausrüstungen entwickelt, in denen er zur Erhöhung des Wirkungsgrades von Transistor-Kühlkörpern und durch Herabsetzung der Temperatur innerhalb der Ausrüstung zu höherer Betriebssicherheit und allgemeiner Stabilität beiträgt. Die Abmessungen sind so gewählt, dass das Kühlaggregat in das Chassis- wie auch das Gestellsystem passt, ausserdem jedoch in beliebige elektronische Geräte eingebaut werden kann.

Die Gehäuseabmessungen sind 127 × 127 mm über der Öffnung und 114 mm tief. Befestigungslöcher sind vorgesehen, so dass das Aggregat mit angeschraubten Chassisschienen als Ersatz der normalen Seitenplatte jeder Lektrokit-Konstruktion eingebaut werden kann.

In Betrieb werden bei einigermassen freiem Luftstrom etwa 56 m^3/h verdrängt. Die Ventilatorflügel werden durch einen Kleinstmotor getrieben, der an 200...250 V~ angeschlossen wird. Der Luftsaugstutzen ist mit zwei Gewindebuchsen für Montage an Lektrokit-Belüftungsfelder Nr. 2 und 3 ausgerüstet.

Durch Kürzen der Schienen LK.201 auf drei Viertel ihrer Länge und Anbau des Kühlaggregates an eine Seite passt die Ausrüstung in denselben Raum wie das normale Lektrokit-Chassis. Auf diese Weise kann man es mit den Lektrokit-Montagewinkeln LK.601 entweder bündig oder hervorstehend in ein 19"-Standardgestell einbauen.

EE 04 770 *für weitere Einzelheiten*

Vakuummesser

Edwards High Vacuum Ltd, Manor Royal, Crawley, Sussex

(Abbildung Seite 195)

Edwards High Vacuum Ltd hat einen neuen Pirani-Vakuummesser Modell 9 für den Druckbereich 500...10^{-3} Torr angekündigt. Die Skala hat drei Bereiche mit automatischem Umschalten von einem Bereich zum nächsten, wobei farbige Lampen auf der Frontplatte den eingeschalteten Bereich anzeigen. Das Instrument wurde speziell für Niederdruckmessungen in Verfahren der chemischen, metallurgischen und verwandten Industrien entwickelt, die zwar nicht im höchsten Vakuum durchgeführt werden, aber doch elektrische Messgeräte erfordern, die Registriergeräte und externe Schalteinrichtungen treiben können. Das neue Instrument dürfte auch bei vielen Laboranwendungen nützlich sein, die einen Vakuummesser mit mehreren Messköpfen verlangen.

Ein Spannungsregler in Wheatstone-Brückenschaltung und ein Differentialverstärker bilden die Grundprüfanordnung, in die ein um die Messköpfe gewickelter, temperaturabhängiger Kompensationswiderstand einbegriffen ist. Diese Schaltung reduziert den Einfluss der Schwankungen in der Umgebungstemperatur auf die Eichung des Instrumentes wesentlich. Durch den Spannungsregler wird manuelles Justieren der Spannung überflüssig. Drucktasten auf der Frontplatte gestatten die Wahl eines beliebigen der vier Messköpfe der Baureihe 9, die man an das Gerät anschliessen kann. Die Messköpfe der Baureihe 9 kommen in Ausführungen für Metall- oder Glasvakuumsysteme und sind elektrisch austauschbar.

Auf der Rückseite gibt es Buchsen für Anschluss eines Drehspul- oder Kompensationsschreibers, einer niederohmigen, schaltbaren 19-V-Gleichstromspeisung für einen Hilfsschreibstift zur Kennzeichnung der Bereichschaltung während des Registrierens, sowie Starkstromkontakte für Verfahrensreglung. Modulare Konstruktion und Transistorschaltungen durchweg ergeben ein kompaktes Gerät, das auch den Service erleichtert.

EE 04 771 *für weitere Einzelheiten*

Mehrdiodenbausteine

SGS-Fairchild Ltd, Planar House, Walton Street, Aylesbury, Buckinghamshire

Diese neuen Mehrdiodenbausteine enthalten bis zu 16 Halbleiterscheiben in einem TO-5-Gehäuse, unter anderem viele Kombinationen in Diodenbrücken, Matrizen, Gruppen und Ringmodulatoren. Ausserdem ist die Firma bereit, den Bedarf an zusätzlichen Kombinationen für spezielle Schaltungen zu decken.

Die Greifbarkeit der Bausteine in so

breiter Auswahl wird durch die Anwendung des Silizium-Planarverfahrens bei SGS-Fairchild ermöglicht. Nach diesem Verfahren werden schnelle Dioden mit hohem Leitwert mit ungewöhnlicher Gleichförmigkeit und Zuverlässigkeit gefertigt, die durch langzeitigen Einsatz unter schwierigsten Betriebs- oder Umgebungsbedingungen nicht beeinflusst werden.

Diese Mehrdioden der SGS-Fairchild werden in TO-18- und TO-5-Gehäusen mit Belastbarkeiten von bis zu 400 mW je Element geliefert. Betriebs- und Lagertemperaturen bis zu 200° C sind zulässig, und da alle Elemente in einem Gehäuse untergebracht sind, ist die thermische Leitfähigkeit besser als mit diskreten Dioden und weniger durch externe Faktoren beeinflusst.

Da man bis zu 16 Dioden in einem TO-5-Gehäuse unterbringen kann, haben diese Bausteine auch den Vorteil kleiner Abmessungen. Daraus ergeben sich ausserdem wesentliche Einsparungen an Montagekosten, denn viele Verbindungen sind bereits hergestellt; statt der 32 Anschlüsse bei Verwendung von diskreten Dioden sind für den 16-Element-Baustein nur 10 erforderlich.

Typen BAX45 bis zu BAX51 sind Siliziumdiodenmatrizen in Planar-Epitaxialtechnik für schnelle Kerntreiber in Computer-Speichern. Der kleinste ist ein Zwei-Dioden-Baustein, der grösste ist der BAX51 mit 16 Dioden und 10 Anschlüssen.

BAX52 und BAX53 sind Diodenbrücken für Anwendungen, die hohe Geschwindigkeit erfordern, und kommen in TO-5- bzw. TO-18-Gehäusen. Die Diodenscheibchen sind nicht zusammengehörig, sondern einzeln hermetisch dicht verkapselt.

BAX54 und BAX55 sind Siliziumdioden-Ringmodulatoren in Planar-Epitaxialtechnik für Anwendung in schnellen Mehrkanalausrüstungen. Sie sind vor allem für Verwendung in elektronischen Fernsprechämtern geeignet.

Ausserdem gibt es noch ein Sortiment von 18 Siliziumdiodenanordnungen BAX56 bis BAX73 in SGS-Fairchild-Planar-Epitaxialtechnik. Sie wurden als Treiber für schnelle Kernmatrizen entwickelt und als Anordnung mit gemeinsamer Anode oder gemeinsamer Kathode in Kombinationen von zwei bis acht Dioden angeboten.

EE 04 772 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

"Rauschgrenzen" in binärelektrologischen Systemen von E. L. Jones

Zusammenfassung des Beitrages auf Seite 142-147

Der Ausdruck "Rauschgrenze" oder "Rauschfestigkeit" wurde in bezug auf binärelektrologische Systeme vielseitig definiert. Jede Definition geht von einer gewissen Anzahl von Voraussetzungen aus, von denen einige deutlich erkannt wurden, andere nicht. In vielen Fällen würden alternative Voraussetzungen zu grundsätzlich abweichenden Definitionen der Rauschgrenze führen. In anderen Fällen werden Voraussetzungen nur zur Vereinfachung des Problems, zur Erleichterung der Darlegung oder der Messung eingeführt, ohne die in Frage stehenden Grundbetrachtungen zu beeinflussen.

Der Beitrag versucht, diese Voraussetzungen klarer zu bestimmen und aufzuzeigen, welche grundsätzlich sind und welche nur zur Vereinfachung eingeführt wurden. Wo alternative Voraussetzungen in Frage kommen, werden die Gründe für den Vorzug der einen gegenüber der anderen besprochen.

Zunächst werden die Betrachtungen auf "statische" (Tiefstfrequenz-) Rauschgrenzen beschränkt, da angenommen wird, dass es zwecklos ist, die Betrachtung "dynamischer" (schnellvergehender) Rauschfrequenzen in Betracht zu ziehen, bevor man die statischen Rauschgrenzen gründlich versteht und sich über ihre Definition und Deutung einig ist.

Für Bestimmung und Ausdruck statischer Rauschgrenzen in integrierten logischen Bausteinen werden grafische Hilfsmittel gegeben.

Ein stromschaltender, phasenempfindlicher Gleichrichter und seine Anwendung in einem Voltmeter für aufgelöste Komponenten von J. G. Lacey

Zusammenfassung des Beitrages auf Seite 148-151

Ein phasenempfindlicher Gleichrichter wird beschrieben, in dem Transistoren als Stromschalter Anwendung finden, um die Amplitudenlinearität zu verbessern und den Betriebsfrequenzbereich zu erweitern. Das sinusförmige Bezugssignal wird durch eine Schaltung in eine Rechteckwellenform umgewandelt, in der eine Tunneldiode als hysteretischer Schalter wirkt und mittels einer Regelschleife ein Tastverhältnis von 1:1 erreicht wird. Ausserdem wird ein Voltmeter für aufgelöste Komponenten mit diesen Gleichrichtern beschrieben, das gleichzeitig die gleichphasigen und Quadraturkomponenten eines sinusförmigen Wechselstromsignals in bezug auf ein zugehöriges Bezugssignal darstellen. Der Frequenzbereich des Gerätes ist 20 Hz . . . 150 kHz.

Dekadenkästen mit binärem Ausgang von D. P. Franklin

Zusammenfassung des Beitrages auf Seite 152-160

Schaltungen für Dezimal-Binärumwandlung werden entwickelt, in denen die erforderliche kombinatorische Logik völlig passiv arbeitet. Die UND-Funktion wird durch Serienschalten von Kontakten mehrpoliger 10stufiger Schalter erfüllt, die ODER-Funktion durch Verwendung von Parallelwegen, oft über Dioden.

Zwei Arbeitsmethoden werden beschrieben, bei denen die Eingänge zu den logischen Einheiten entweder erst in Binärform gebracht werden oder nicht.

Eine netzstabilisierte Hochspannungs-Stromversorgung mit gesteuerten Lawinengleichrichtern von R. W. J. Barker

Zusammenfassung des Beitrages auf Seite 161-165

In einer neuen Methode zur Konstruktion einer netzstabilisierten Hochspannungs-Stromversorgung niedriger Leistung finden gesteuerte Lawinengleichrichter Verwendung.

Die Berechnung der Eigenschaften einer solchen Stromversorgung mit Hilfe der üblichen Stabilivoltanordnung und die Spezifikation der Bauelemente werden abgeleitet, und gegebene Kurven erleichtern die Beurteilung der abgeleiteten Gleichungen.

Die praktische Leistung eines Labormodells der nach dem abgeleiteten Entwurfsverfahren gebauten Stromversorgung wird mit der theoretischen Leistung verglichen.

Elektromechanische Bildung des laufenden Mittelwertes von K. G. Ng und K. K. Murthy

Zusammenfassung des Beitrages auf Seite 169-171

Eine neue elektromagnetische Einrichtung zur laufenden Bildung des Mittelwertes wird beschrieben. Sie ist preiswert, einfach zu bauen und zuverlässig in Betrieb und ist besonders dort nützlich, wo laufende Signalmittelwerte mit einem Analogrechner simuliert werden sollen.

Bestimmung von Nichtlinearität mittels Analogsimulation von G. W. Holbrook und J. S. Collins

Zusammenfassung des Beitrages auf Seite 172-173

Die meisten nichtlinearen Elemente können durch eine auf die Ausgangs- und Eingangsströme oder -spannungen bezogene Potenzreihe beschrieben werden. Die Koeffizienten der ersten und bedeutsamsten Glieder kann man durch Synthesieren der Ausgangswellenform des Elementes mittels einer Serie integrierender Schaltungen und Anwendung einer Kippschwingung als Eingangssignal erreichen. Der Ausgang jeder Integratorschaltung kann dazu dienen, den korrespondierenden Koeffizienten direkt über einem Vorspannungsbereich aufzutragen.

Kombinations-Pulsmodulation mittels veränderlicher Steilheit der Vorder- und Hinterflanke
von O. E. Kruse, D. L. Lewis und J. K. Jackson

Zusammenfassung des Beitrages auf Seite 174-176

Der Beitrag beschreibt ein Modulationssystem, nach dem zwei Informationskanäle von derselben Dreieck-Impulsreihe getragen werden. Für einen Kanal ist die Steilheit der Vorderflanke, für den anderen die der Hinterflanke der Träger.

Ein Untersetzer und Alarmgrenzenkomparator für Datenerfassungsanlagen von R. J. S. Ruckman

Zusammenfassung des Beitrages auf Seite 177-183

Der Beitrag beschreibt eine nach der Zähltechnik arbeitende Untersetzungsmethode, nach der bei der Untersetzung Nullpunktverlagerung und negative sowie positive Werte berücksichtigt werden. Gleichzeitig und während der Untersetzung werden Alarmgrenzwerte mit den untersetzten Werten verglichen und dabei festgestellt, ob eine bestimmte Variable in "Alarmzustand" oder "Kein Alarmzustand" ist.