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Commentary

AN International Medical Electronics Exhibition was held at Olympia, London, during the last week of July. This was the first exhibition of its kind ever to be held, and it is unlikely that another such exhibition will be held in London for at least a number of years. The reason for this is that the exhibition was held in conjunction with the Third International Conference on Medical Electronics, and it is intended to hold the conferences in a different country each year: thus, last year it was held in Paris, next year it will be held in the U.S.A., and in the following year it will, most likely, be held in the U.S.S.R.

This year the Conference was organized jointly by the Electronics and Communications Section of the Institution of Electrical Engineers, and the International Federation for Medical Electronics. The Institution of Electrical Engineers, of course, needs no introduction but, since the International Federation for Medical Electronics is a very much newer and less well known organization, a few words covering its formation and history may not be out of place.

The Federation owes its being very largely to the pioneering work of Dr. V. K. Zworykin of the U.S.A., who has devoted his retirement to this very worthy task. On one of his trips abroad, on the occasion of the International Conference on Colour Television in Paris, he spoke to a number of his friends on his work on medical electronics in the U.S.A., and was astonished at the interest shown in having such an organization on an international level. At about the same time Dr. Zworykin was invited by Dr. Bronk of the Rockefeller Institute to form a medical electronics laboratory and research centre at the Rockefeller Institute in New York, and one of the functions of the centre was to hold conferences aimed at bringing medical practitioners, electronic engineers, and other medical scientists together. As a result of all this apparent interest it was decided to hold a preliminary conference to explore the need for international co-operation in this field. This conference was held in Paris in 1958, and was attended by fifty scientists from twelve different countries. The enthusiasm of this small but highly representative group was such that it was decided to form an international federation of interested societies and to prepare a second conference. This second conference was held at the new UNESCO headquarters in Paris last summer and was attended by over 400 delegates from 17 countries: it was at this meeting that the federation was legally established. At the recently held third conference in London there were over 700 participants from 20 or more countries. At the exhibition held in conjunction with the conference there were 81 exhibitors representing the manufacturers of some twelve different countries: in addition, there was a number of research exhibits by various medical research centres.

It is apparent, therefore, that there is a great deal of interest in this branch of electronic engineering, and it is gratifying and comforting to observe that it is possible to obtain such widespread and enthusiastic international co-operation on a subject which is wholly for the good and the well being of mankind.

The scope of the applications of electronic engineering is immense and covers both diagnosic and treatment and fundamental research into the exact methods of function of the various organs of the human body. It also covers the health hazards of nuclear radiation and, largely from the point of view of the Armed Services, the effects on the body of unnatural environments, particularly as they pertain to high speed flight and space travel. The bulk of the effort is, however, on research into and diagnosis and treatment of illnesses and diseases.

Among the foremost assets which electronics brings to the medical world are its ability to provide an accurate control of physical quantities and processes and the ease with which instantaneous recordings of transient phenomena can be made. In many cases the electronic circuits employed are of comparatively simple design; indeed they are often no more than appendages, albeit necessary ones, to ingenious mechanical contrivances. The equipment employed requires to have, however, two virtues which are not always the easiest to provide; absolute reliability, and ease of operation, and in this respect the many new electronic components and manufacturing methods are playing a vital part.

From an observation of the Conference and Exhibition it is clear that electronics has already made a very considerable impact on medicine, and equally that the field for future exploration and progress is almost unlimited. From the comments of various persons who attended it would seem that one of the great hindrances to progress is a lack of funds, particularly in the case of hospital appropriations for new equipment. In so far as this country is concerned this may seem a little odd, in view of the large sums of money which are being spent on the Health Services. It is, however, a matter which is somewhat outside the scope of comment for a technical journal and suffice it to observe that, although medical electronics may not be one of the most lucrative fields of endeavour, it is one which can provide its participants with the utmost personal satisfaction.

The Elliott Sheffer Stroke Static Switching System

By P. Kellett*

A system is described in which interconnexions between a number of logic elements, all of the same type (the Sheffer Stroke) permit any logical configuration or switching sequence. The simple rules which restrain the interconnexions are stated. A number of basic interconnexions of general use are described and a comparison made between the 'Sheffer Stroke', the 'nor' and the better known 'and', 'or', 'not' logic elements.

Available power switches are described. These may be controlled by the output of a logic element. Provision is made for connecting the loads in a matrix with considerable economy.

(Voir page 592 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 600)

THE Sheffer Stroke Static Switching System[1] has been developed to fulfil the requirements of many different fields of activity which come under the heading of digital computing technique. The main areas where positive requirements exist are:—

(1) 'On-line' computer peripheral equipment

(2) 'Off-line' ,, ,, ,,

(3) Sequence control and interlock systems in automation and process control in practically all areas of industry.

In the case of (1) and (2) the circuits used in the computer are frequently of such a nature that it is unsuitable or uneconomic when used to control comparatively slow peripheral equipment. In addition, the asynchronous nature of the Sheffer system enables the system designer to extract the maximum speed advantage when coupling together two independent synchronous systems such as a magnetic film or tape store and a synchronous computer.

Off-line control equipment using the Sheffer system requires no 'clock' waveforms and can be completely independent of other equipment. It is possible to design and build very small units which are economically attractive, a difficult proposition with most synchronous systems.

The process control and automation applications are, perhaps, potentially the greatest of all. In the past many problems in this field have been tackled with electromagnetic relays, but the limitations of reliability and speed have only been overcome by the advent of solid state systems of which the Sheffer Stroke system is a very flexible and powerful example. It represents an integrated design philosophy including contactless input devices (photo transistors and photo-voltaic cells) and contactless power switches (100VA with easily obtainable transistors) as well as the inherent simplicity of the basic element which enables logical nets of unlimited complexity to be designed as required.

Realization of the Logic System

Sheffer Stroke Logic and its dual 'Nor' Logic are both systems of logic in which one compound logical element is used to synthesize by interconnexion, all possible logical functions[2]. Before describing the circuit used it is essential to emphasize the fact that a logic element is defined by the circuit function and the digit sign convention chosen. A circuit which is capable of performing a useful logic function must perform one unambiguous circuit function. Such a circuit may perform two logic

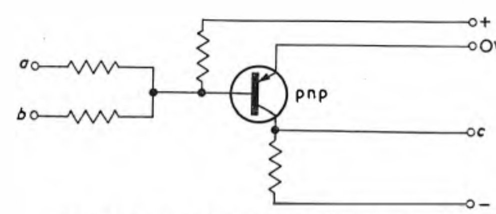

Fig. 1. Basic 'Sheffer Stroke" logic circuit

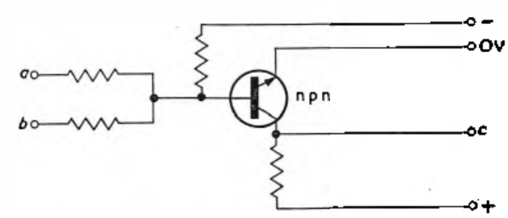

Fig. 2. Basic 'Nor' logic circuit

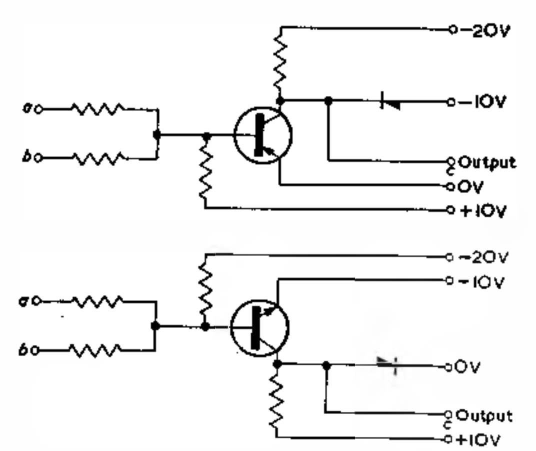

Fig. 3. Compatible 'Sheffer Stroke' and 'Nor' logic circuits

functions which are dual to each other. Contrary to recent practice in dealing with transistor 'nor' and 'Sheffer' logic, but in full accordance with well established computer practice, positive voltage excursions will

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be defined as '1' ('true' or 'up') and negative voltage excursions as '0' ('false" or 'down').

The basic circuit of the Sheffer logic is shown in Fig. 1. If both input terminals a and b are at zero volts potential, then the positive bias applied to the base of the transistor will cause the transistor to 'cut off', and c will assume a potential close to that of the negative supply line. If

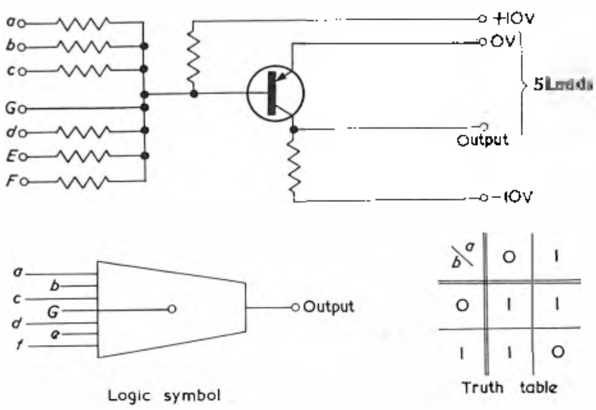


Fig. 4. Circuit and symbol of 'Sheffer Stroke' logic element

a or b or both a and b are connected to some negative potential, the transistor will conduct and 'bottom'. c will assume a potential of zero volts (approx.). Since the negative excursion is defined as '0' and the positive excursion (zero volts) as '1' the behaviour of the circuit may be expressed as a logical function. The output c is '1' when either a or b are not '1'. This statement may be expressed in symbolic form:—

$$c = \bar{a} + \bar{b} = \overline{ab} \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

This expression defines the Sheffer Stroke logic elements. The output voltage excursion of such a circuit, when connected to loads which are identical with the input resistors, can be made compatible with the input requirements. The design methods required are given in detail elsewhere[2,3].

The dual of this circuit is shown in Fig. 2, when an npn transistor is used. This circuit represents a NOR element defined by the Boolean expression

$$c = \bar{a}\,\bar{b} = \overline{a + b} \quad \ldots\ldots\ldots \quad (2)$$

In this form it is difficult to design the dual circuits in such a manner that they are compatible with each other. A possible modification to the simple circuit which makes a compatible design possible is shown in Fig. 3. Experience of designing systems with only one type of element suggests that the additional complexity and expense of working with two different elements greatly reduces the attractiveness of what appears, at first sight, to offer an

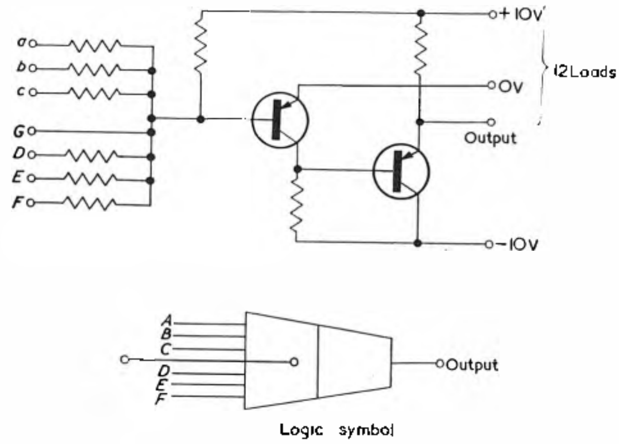


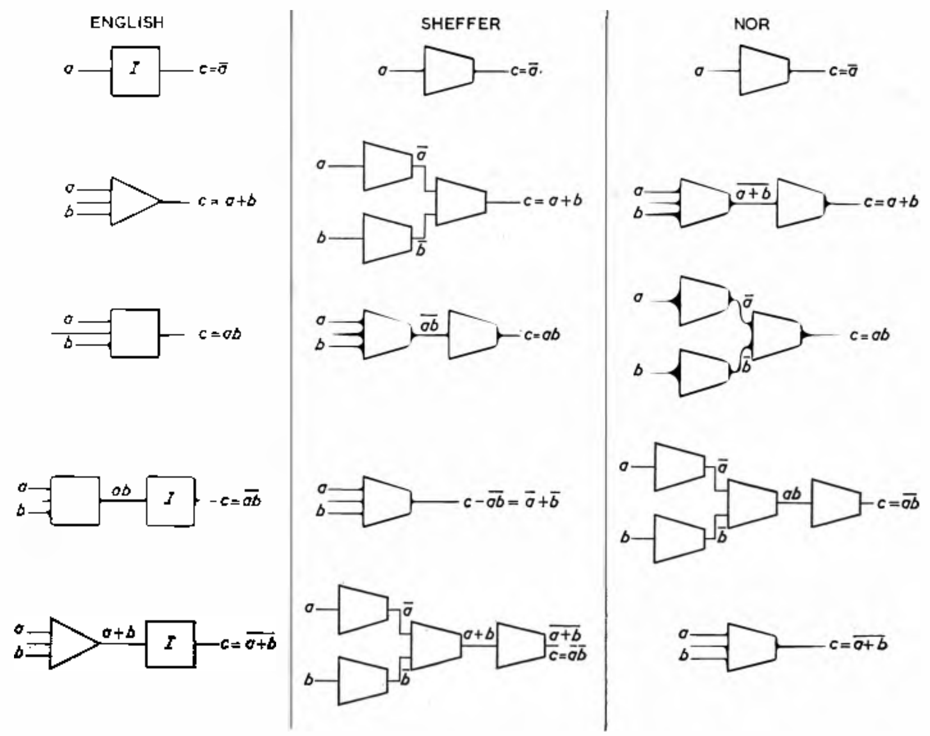
Fig. 5. Circuit and symbol of emitter-follower version of 'Sheffer Stroke' element

extremely powerful logic system. Considerable experience of operating with two particular circuit configurations have established that the reliability of the system is of an extremely high order. These two circuits, together with the symbols which will be used to represent them, are shown in Figs 4 and 5. The circuit shown in Fig. 4 is the most widely used. The provision of six input terminals together with a permissible loading of up to five further elements is adequate for the majority of applications. Cases do, however, arise when a heavier loading is desirable. The circuit shown in Fig. 5 is capable of driving up to 12 loads or in certain cases up to 50. Logically the two circuits are identical, unless output terminals are directly connected together. The difference in this instance will be described later.

Logical Interconnexions

The table shown in Fig. 6 illustrates the equivalence

Fig. 6. Table of logical equivalents



between interconnexions of AND, OR, NOT (English) elements and NOR or Sheffer elements. This table may be used to derive the Sheffer equivalents of known configurations of English logic.

The derivation of an exclusive OR, or anti-coincidence unit is shown in Fig. 7(a), (b) and (c). The derivation of (c) from (b) in which the two elements are eliminated is obvious from an inspection of the Boolean indentities shown on the diagram. The final diagram still appears uneconomic in terms of the number of elements when compared with the original configuration of 7(a). When

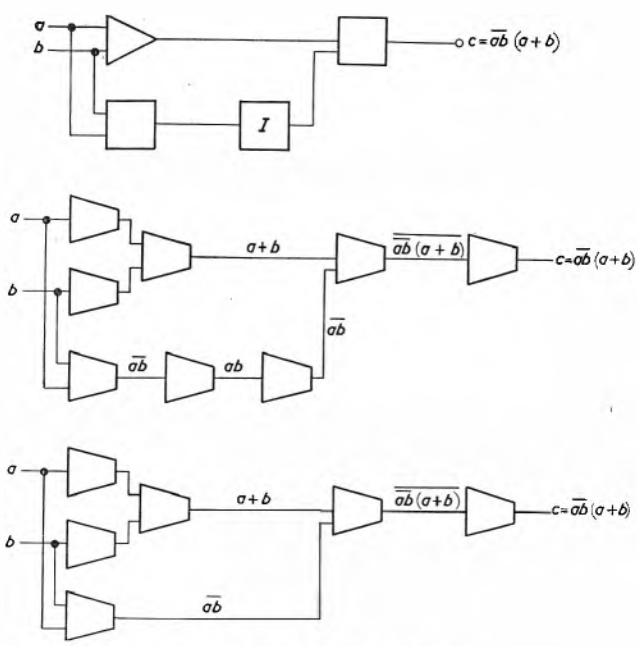


Fig. 7. Exclusive 'or' derivation

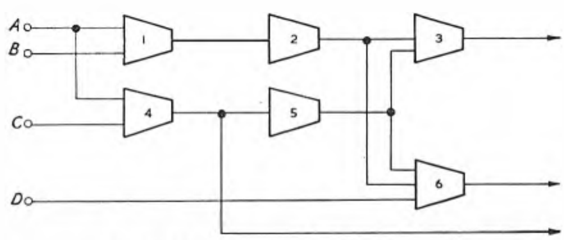


Fig. 8. Application of redundancy theorems

small configurations are equated in this manner, an apparent extravagance is often seen. This effect is quite untypical of an actual system. In general, an extensive logical net can be equated using NOR and Sheffer logic using no more elements than would be used in an 'English logic' system. An informal proof of this concept is given by W. D. Rowe[2].

The idea that any logical net composed of Sheffer elements will contain a large number of elements used solely as invertors is manifestly untrue. Two theorems of use in eliminating redundant elements illustrate this latter point.

Theorem 1

Any element within a net with only one input is redundant.

Theorem 2

Any element in which the output terminal is unloaded is redundant.

In the logic diagram of Fig. 8, which represents a section of logic net, the redundancy is not obvious. The mechanical application of the two rules, in order, leads to Fig. 9 and finally Fig. 10. The two theorems are of general application to both NOR and Sheffer logic. The

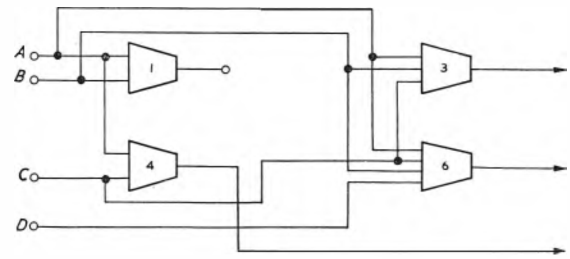


Fig. 9. Application of redundancy theorems

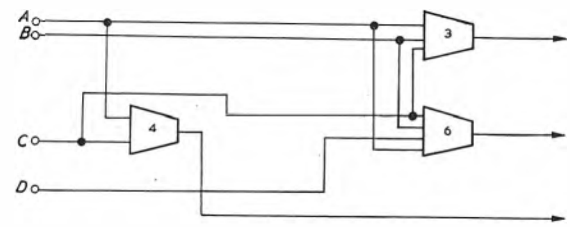


Fig. 10. Application of redundancy theorems

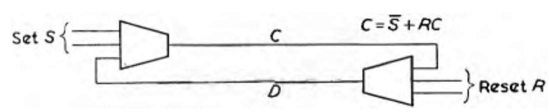
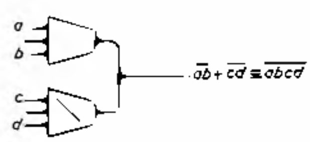


Fig. 11. One bit storage configuration



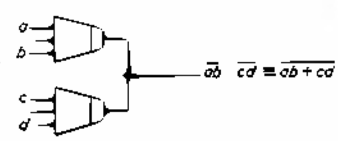


Fig. 12. Direct interconnexion of output terminals (*single elements*)
Fig. 13. Direct interconnexion of output terminals (*double elements*)

derivation of Fig. 7(c) from 7(b) follows also from the purely mechanical application of these two theorems.

Since the element exhibits both power and voltage gain, the interconnexion shown in Fig. 11 is permissible. This configuration performs the function of one bit storage. Two outputs inverted with respect to each other are available and setting may be accomplished by bringing 'down' one of the five input terminals so labelled. Resetting is similarly accomplished. In Boolean terminology:—

$$C = \overline{S\,D}$$

$$D = \overline{R\,C}$$

$$C = \overline{S\,\overline{RC}}$$

$$= \overline{S} + R\,C$$

This expression, characteristic of the storage function, is worthy of note.

The direct interconnexion of output terminals is the only important connexion remaining. This interconnexion is not, strictly speaking either NOR or Sheffer logic, and cannot be predicted from the logic function. Study of the two circuits of Figs. 4 and 5 shows that direct connexion of the 'single' transistor circuit (a 'single'

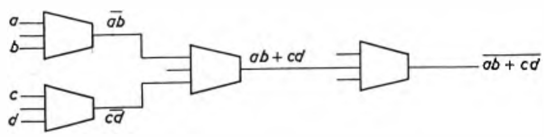


Fig. 14. Orthodox interconnexion

element) results in an OR or 'mix' function, whereas direct connexion of the two transistor circuit (the 'double' element) produces an AND or 'gate' function. This type of interconnexion together with the pertinent Boolean expression, is illustrated in Figs. 12 and 13. The logical function of Fig. 13 may of course be produced by orthodox interconnexion of 'single" elements, as illustrated in Fig. 14. In both cases the restrictive rules relating to the number of loads permitted is unchanged, each common output being considered as a single load. In addition, in the case of 'single' elements, the negative supply is connected to only one of the common elements, the diagonal line through the element signifying—'no negative supply'.

The derivation of Sheffer nets by use of the table shown in Fig. 6 represents a useful approach to logic design. Various algebraic notations have been introduced as an aid to logic design using the Sheffer element[4]. Special notations are not essential, however, and the following method of designing with Boolean algebra covers many requirements.

Consider the basic expression for a Sheffer element, with output = $\overline{A.B.C.}\ldots$ where $ABC\ldots$.represent the inputs to that element. Taking now any Boolean expression, it must be reduced to this basic form. $A.B.C.\ldots$ may themselves be complex and must each be reduced to the same form. The final algebraic expression will define a unique interconnexion of Sheffer elements. This reduction of Boolean expressions to 'bar product' form is an easy mechanical process.

As an example consider the 'exclusive OR' configuration of Fig. 7. The Boolean expression given in Fig. 7 is:—

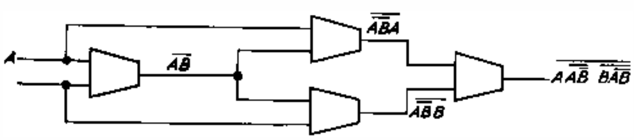


Fig. 15. Economical exclusive 'or'

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

The reduction to 'bar products' is shown below:—

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

$$= \overline{\overline{\overline{AB}\;\overline{\overline{(A + B)}}}}$$

$$= \overline{\overline{\overline{AB}\;\overline{\overline{A}\;\overline{B}}}}$$

This expression defines uniquely the interconnexion shown in Fig. 7(c). The total number of elements is obviously equal to the number of 'bars', i.e. six. It is interesting to consider rearrangements of the algebraic expressions in order to reduce the number of 'bars'.

For example:

$$\overline{AB}\,(A + B)$$

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

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

$$= \overline{\overline{A\overline{B}}\qquad\overline{\overline{A}B}}$$

This expression can be realized with only five elements, i.e. the number of bars in the final expression.

Further rearrangement gives:—

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

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

$$= \overline{\overline{A\,\overline{A\,B}}\;.\;\overline{B\,\overline{A\,B}}}$$

Again there are five bars, but since one bar term, AB, appears twice, this expression can be realized with only four elements as shown in Fig. 15.

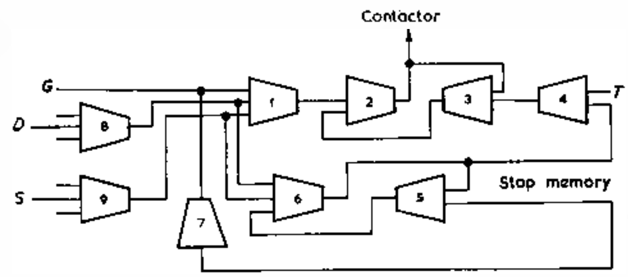


Fig. 16. Motor control logic

A combination of the block diagram logical approach and the algebraic approach, together with the use of redundancy theorems provides a versatile technique of logic design. Of particular interest is the combination of algebra and redundancy theorems, since both are completely mechanical the possibility of automated logic design is apparent.

An Application of the Logic System

A simple example of the use of Sheffer logic is illustrated in the example given below. A reliable start-stop control for part of an automatic assembly line was required. The specification for this control system was as follows:

The motor must start when the start button is operated providing the safety cage is closed, and the stop signal is not also present. The motor must stop at time T after being instructed to stop or after the safety cage has been opened. In the first applications signals were derived from cam operated micro-switches, but in later models, photo sensitive cells were used.

The logic diagram of the system which fulfils this specification is illustrated in Fig. 16. The control action required implies that provision for 'remembering' two states is required. The motor must 'remember' that it has been instructed to 'run,' while a 'stop' instruction must not be obeyed until time T. These two storage functions are performed by elements S_2, S_3 and S_5, S_6. The functions of the remaining elements may be seen on inspection of the diagram.

The two elements (S_8) and (S_9) are redundant providing input signals corresponding to $\bar{S}$ and $\bar{D}$ are made available instead of S and D. This is feasible with either contact or photocell inputs.

It is possible to formulate the Boolean algebraic expressions for the control function, and by algebraic manipulation, express the equation in a form applicable to a Sheffer logic system. The motor must run (R) when it is running and is not remembering to stop, S_M at time T. Or it must run when told to do so (G) and when the safety cage, D, is not open and it is not being told to stop (S).

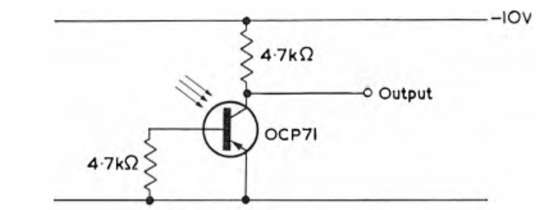


Fig. 17. Photo-transistor input circuit

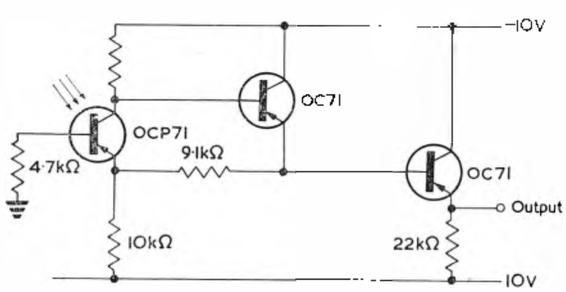


Fig 18. Photo-transistor Schmitt circuit

$$R = RS_MT + G\,D\,S \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad \text{(a)}$$

$$S_M = S_MG + D + S \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad \text{(b)}$$

These two expressions may be written as:

$$R = \overline{\overline{RS_M\bar{T}}\;\overline{G\,\bar{D}\,\bar{S}}} \quad \ldots\ldots\ldots\ldots \quad \text{(c)}$$

$$S_M = \overline{S_M\,\bar{G}\,\bar{D}\,\bar{S}} \quad \ldots\ldots\ldots\ldots\ldots \quad \text{(d)}$$

The Sheffer logic diagram may be constructed directly from these equations by remembering only the output expression for a Sheffer element $C = \overline{AB}$ (equation (1)). The reduction of algebraic equations to this form is purely mechanical and frequently forms a useful method of design, or a check on an existing design. Such algebraic manipulation does not give the most economic system possible in many cases, but it will give, as a result of the application of mechanical rules, a solution, which fulfils the control specification. Application of the 'theorems' of redundancy stated previously may then be applied to the logic diagram. This combination of algebraic manipulation and application of the theorems to the diagram so produced, forms a completely mechanical method of design which designers, unfamiliar with the more sophisticated methods of logical reduction, should find extremely useful.

Realization of Input and Output Circuits

The Static logic system described, must, of necessity, have means of communication with the outside world. Signals from external devices, frequently mechanical, must be capable of operating the logic system, while output signals from the logic are required to operate mechanical devices through solenoids or motors, or to operate lamps, etc.

Input signals from contact switches may be connected to switch 10V lines which may operate the logic direct.

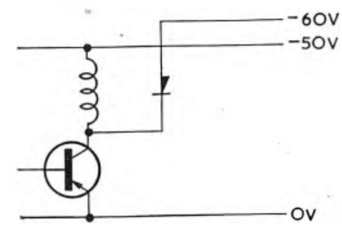


Fig. 19. Simple power switch

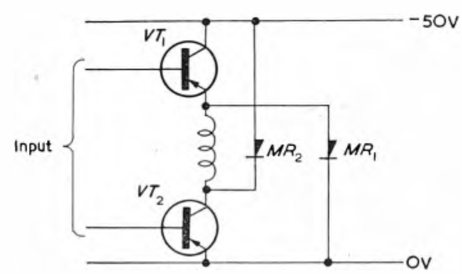


Fig. 20. Basic circuit of actual power switch

Fig. 21. Equivalent circuit of power switch

Greater reliability can usually be obtained by using photo transistors as input devices. A typical circuit is shown in Fig. 17. This circuit supplies a signal identical to that obtained from a 'single' element, operating quite satisfactorily from an under-run 3W lamp. Occasions do arise when it is necessary to obtain a more definite 'threshold' than the latter circuit can provide. A photo transistor 'Schmitt' circuit of the type shown in Fig. 18 has proved useful in such applications. It is probable that in the future greater reliability will be achieved with photo-voltaic cells, and circuits using these devices are under investigation.

The most common type of output devices are solenoid operated, and the drive requirements frequently give rise to a number of problems. One particular application which has been studied involves the driving of any one solenoid out of 32, the drive requirements being 50V at 1·1A. Rapid recovery of the solenoid is essential. This implies that an overshoot of voltage must be permitted to build up across the coil. At the time of this study, the only transistor capable of supplying the current, had a maximum collector rating of 60V. The obvious safe 'circuit' shown in Fig. 19 is thus limited to a 10V overshoot which was insufficient for reliable operation of the device. In order to achieve a larger overswing, without exceeding the permitted voltage rating of the transistor, the circuit shown in simplified form in Fig. 20 was developed.

Transistors VT_1 and VT_2 are switched on and off together and the voltage across each transistor is prevented by the diodes from exceeding the supply voltage of 50V. That this circuit allows 100 per cent overshoot to develop across the coil may be understood from a study of the equivalent circuit of Fig. 21.

In (a) both transistors are 'on' and the solenoid is effectively directly across the supply, and X is 50V negative with respect to Y. The diodes are both reverse biased and may be ignored. An instant after switching off, the reverse voltage across the solenoid will drive X positive and Y negative. The two diodes will constrain the solenoid with X at 0 volts and Y at -50V. Thus 100 per cent overshoot has been permitted. The voltage across the transistors has, however, been constrained rigidly within their ratings.

The use of two power transistors per coil does not lead to an uneconomic configuration since it is possible to arrange the 32 solenoids in a 4×8 matrix driven by 12 power transistors. It is necessary to place a diode in series with each solenoid in order to block unwanted paths through the matrix, as illustrated diagrammatically in Fig. 22.

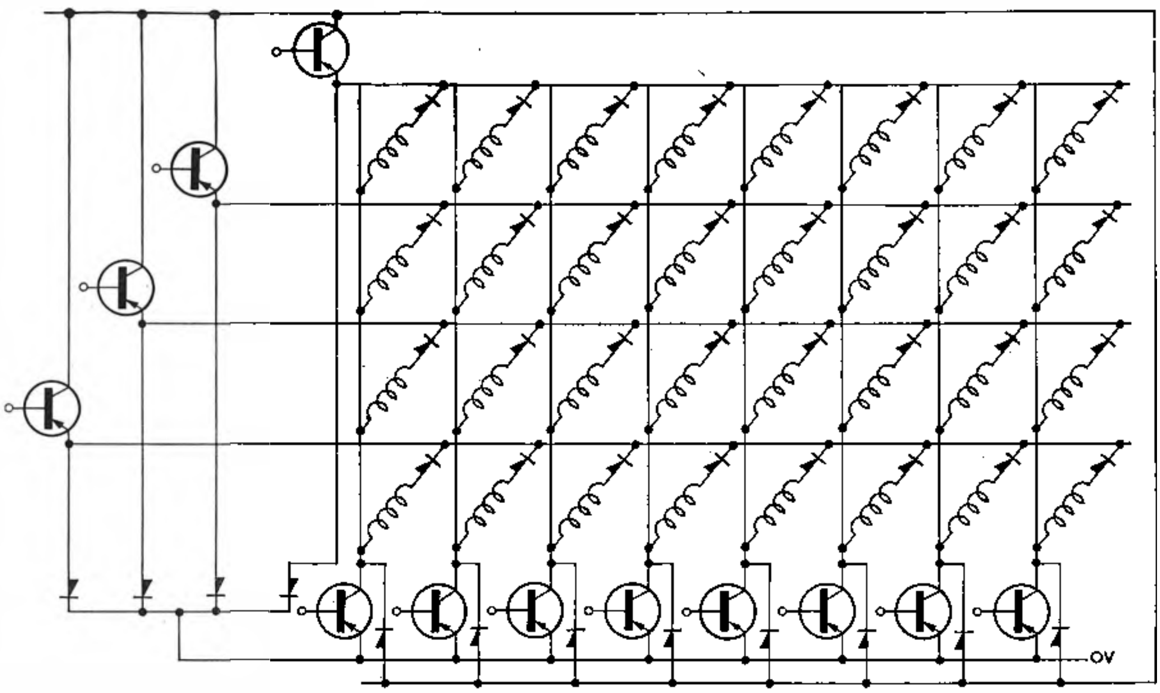


Fig. 22. 4 × 8 power matrix

The amplifiers fall naturally into two groups, 'top amplifiers' and 'bottom amplifiers.' They have the same dimensions as the logic elements by which they are controlled. Where fast recovery is not required, it is permissible to use only one amplifier of either type.

A matrix similar to the one described, was used in 'Ella' (the machine used by the BBC Television to analyse the 1959 General Election Results), to control the 40 'in-line' decimal indicators on which the latest analysis was displayed.

One further output is often required in the form of small indicator lamps. No special circuits are required since the 'double' logic element is capable of driving small lamps directly.

Conclusions

It is suggested that the system described offers an extremely sophisticated approach to the design of industrial control and computer peripheral equipment systems. Modifications and extensions can very easily be carried out, and the reliability and speed compare very favourably with relays or thermionic valves. Since the system is completely asynchronous very small installations are economically feasible. The flexibility is such that Colleges and Universities should find the system useful, not only in teaching of electronics, logic and computing techniques, but also in research activities in many fields.

The system design methods described offer alternative approaches to the problem, including an extremely useful technique employing Boolean algebra together with a means of eliminating redundant elements.

An extension to the system which enables the formation of differentiated waveforms greatly extends the versatility of the system. Counting, 'shifting' and timing operations may be carried out in a very economical manner. Since the whole system is dependent only on the performance of the simple basic circuit, the overall reliability is of a very high order. Circuits of the type described in this article are now being marketed under the trade name 'Mini-log'.

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Systems Using Transistors and Transductors

By D. A. Ramsay*, B.Sc. (Hons.)

(Part 2)

(Voir page 526 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 531)

Transistor Control Element

IF a resistance replaces the voltage source control element across x-y in the circuit of Fig. 4(a), then over a negative half-cycle of v_a a distorted half-cycle of voltage in opposition to v_a is generated across x-y by the resetting current flowing through this resistance. Notice that for this voltage to be significant the value of this resistance must be of the same order as the unsaturated impedance of the winding and therefore of an order higher than R_L. To a good approximation the flux reset depends only on the integral over the reset half-cycle of the voltage across x-y, so that a resistance control element will cause nearly the same output voltage as a voltage control element provided this voltage integral is the same in both cases.

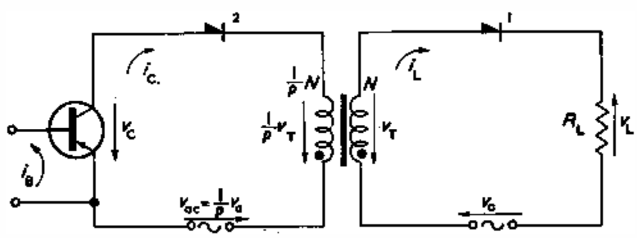

Fig. 7. Half-wave flux reset magnetic amplifier

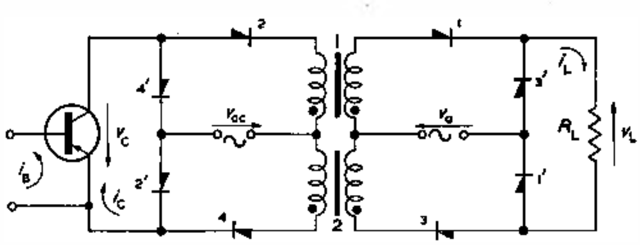

Fig. 8. Full-wave flux reset magnetic amplifier

The effective resistance of a triode or pentode valve or of a transistor can be varied by an independent electrical signal and these devices may be used as the control element of a flux-reset magnetic amplifier[7]. Fig. 4(b) shows how a transistor in a common emitter connexion may replace the voltage source control element in Fig. 4(a). Base and emitter may be interchanged to give a common base configuration but the common emitter connexion is likely to be the more useful. If i_B is increased so that the transistor is taken from cut-off to saturation then v_c will diminish from being almost identically equal to v_a over a reset half-cycle to being almost zero, and V_L will reduce from $\frac{1}{2}V_a$ to almost zero. In this simple circuit the signal and power gains of the transistor are effectively cascaded with those of the magnetic amplifier.

In the circuit of Fig. 4 the output and reset branches are closed through a common a.c. supply and a common winding on the transductor element. Fig. 7 shows a circuit equivalent to that of Fig. 4 but with the output and reset branches closed through separate windings and separate a.c. supplies. The terminal of a winding which is positive for increasing flux is indicated by a dot. If ρ is the ratio of the output to reset turns, voltages in the reset circuit have values $1/\rho$ times the values of corresponding voltages in the single winding circuit so that for the two-winding circuit (compare equation (12)),

$$V_L = \rho\ V_o \qquad \ldots\ldots\ldots\ldots\ldots \quad (14)$$

In a two-winding amplifier the output is electrically isolated from the input, the voltage gain may be greater than unity, and with a transistor control element the reset circuit may be designed to operate at the voltage level best suited to the transistor, whatever the output voltage level.

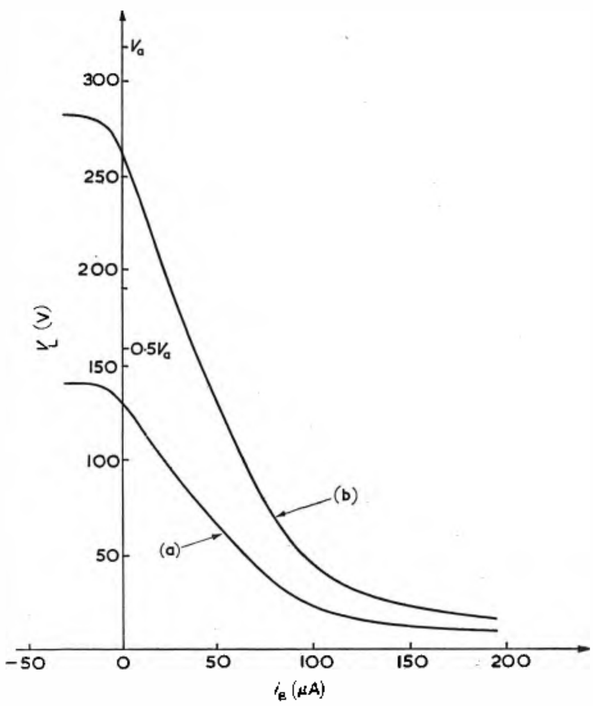

Fig. 9. Mean load voltage (V_L)/transistor base current (i_B) curves
(a) Half-wave circuit
(b) Full-wave circuit

The two circuits so far described are classed as 'half-wave' magnetic amplifiers since their outputs are essentially the half-wave rectified supply voltage with each output half-cycle suppressed from zero electrical angle to the firing angle θ.

Fig. 8 shows how two half-wave circuits may be combined to give a full-wave rectified output voltage, every half-cycle being suppressed from zero to the firing angle. A positive half-cycle is an output half-cycle for core 1 and at the same time a reset half-cycle for core 2. For the same supply voltage the mean output voltage from the full-wave circuit is twice that from the half-wave circuit. Fig. 9 shows the type of output/input curves which are obtained with transistor controlled half-wave and full-wave flux-reset magnetic amplifiers.

The full-wave flux-reset amplifier responds to signal

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frequencies in the range zero to one half the supply frequency; at the latter signal frequency the amplifier lag is 90° and its voltage gain is 3dB below its gain at low frequency.

Design

If a measured V_L/V_c curve of any of the flux-reset magnetic amplifiers described above is plotted against axes scaled in terms of V_L/AN (volt/cm² turn) and $V_c/A(N/\rho)$ and this normalized curve is corrected to cater for the effects of distributed resistance and rectifier imperfections in the actual circuit used (to the accuracy required for design purposes this can usually be easily done) then the resultant curve is characteristic of the core material at the supply frequency used. This normalized curve can then be used to determine the values of A, N, ρ required in an amplifier to provide a specified V_L/V_c relationship.

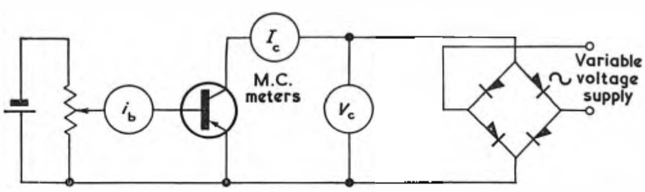


Fig. 10. Circuit for measuring mean value transistor characteristics

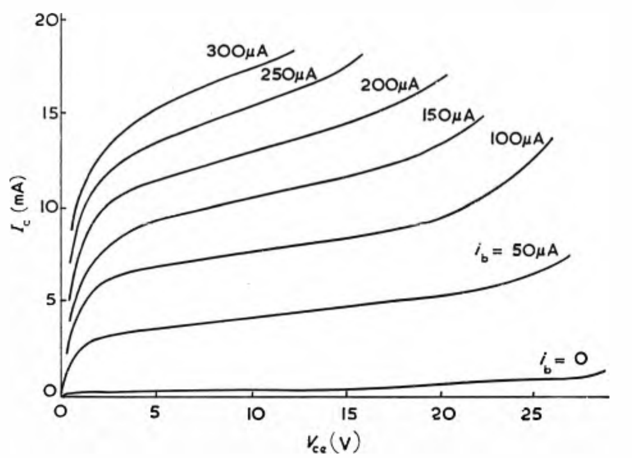


Fig. 11. Transistor mean value characteristics (Brimar TJ3; 2 400c/s; 18°C)

If the control element of the Ramey magnetic amplifier is a transistor, the relationship between V_c and i_b is required in addition to the V_L/V_c curve, before the V_L/i_b curve can be derived. With a transistor control element the control circuit consists essentially of the rectified supply voltage applied across a non-linear inductance —the control winding of the transductor element, in series with a non-linear resistance—the transistor emitter-collector. The detailed behaviour of this circuit is highly complex. However, a simple graphical construction will now be described which yields V_c, the mean value of the voltage across emitter-collector of the transistor control element, as a function of the base current i_b, for a given value of reset supply voltage V_{ac}. This construction uses independently measured mean value characteristics of the transductor element and of the transistor control element.

The required mean value transistor characteristics are measured with the circuit shown in Fig. 10. The alternating supply voltage is full-wave rectified and then impressed across emitter-collector causing a distorted full-wave rectified collector current. Moving-coil meters are used to measure the mean values V_{ce}, I_c, of this voltage and this current. Using this circuit a family of curves of the form shown in Fig. 11 may be plotted. (Subscripts in capitals denote mean values of the variables.) This family of curves is characteristic of the transistor at the supply frequency used.

The transductor element characteristics required are given by the I_T/V_T mean value magnetization curve (Fig. 2), measured—or derived from a normalized magnetization curve for the material—for the reset winding (N/ρ turns).

If two transductor elements with this known I_T/V_T magnetization curve and a transistor whose I_c/V_{CE} curves are those given in Fig. 11 are used to form a full-wave flux-reset magnetic amplifier with the circuit shown in Fig. 8, then the mean value V_c of the voltage across collector-emitter of the transistor in this circuit can be derived as a function of the input variable i_B by the 'load-line' construction shown in Fig. 12.

In this construction the I_T/V_T curve of Fig. 2 is superimposed on the I_c/V_{ce} curves of Fig. 11, the origin of

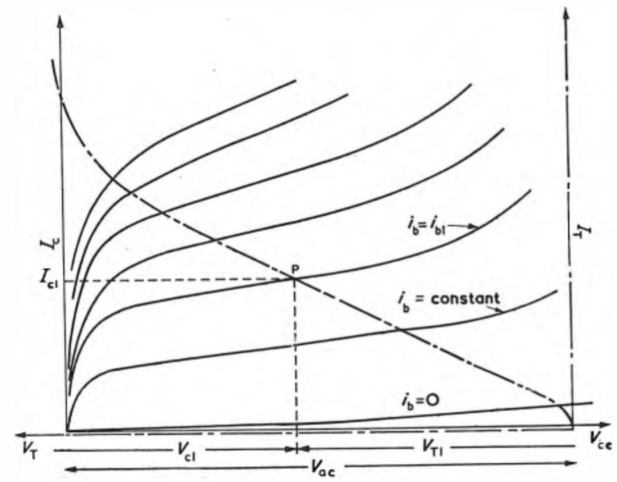


Fig. 12 Load-line construction for deriving the mean control voltage V_c as a function of base current i_B

the I_T/T_T curve being set at $V_{ce} = V_{ac}\,(= 1/\rho\, V_a = 1/\rho\, V_s$, normally). The operating point of the control circuit for $i_b = i_{b1}$ is then given by the point P, the intersection of the I_T/V_T curve with the I_c/V_{ce} curve for $i_b = i_{b1}$. At this point, V_c has the value V_{c1}, the mean emitter-collector current is I_{c1} and the mean value over a reset half-cycle of the voltage across a reset winding is V_{T1}.

The above construction thus gives V_c as a function of i_b. If the V_L/V_c curve (Fig. 6(b)) corresponding to the values of ρ, N, A for the transductor elements is known (and it may be derived from the normalized V_L/V_c curve described at the beginning of this section) then the V_L/i_b curve for the transistor controlled circuit may be predicted.

The above graphical method of predicting the output/input curve of a flux-reset magnetic amplifier with a transistor control element may be adapted for use with any dissipative control element; in particular for a triode or a pentode thermionic valve control element, where the input variable is the grid-cathode voltage v_g.

The theoretical justification for the procedure described above is involved and will not be pursued here. To check its validity, the method was carefully applied to predicting the V_L/v_g curve of an experimental pentode-controlled flux-reset magnetic amplifier with H.C.R. core material. Fig. 13 shows that the measured V_L/v_g curve quite closely agreed with the predicted curve. The method has in fact proved satisfactory in designing a wide range of valve and transistor controlled magnetic amplifiers.

Design Formulae

The determination of the optimum parameters of an amplifier to satisfy a given specification requires design formulae in addition to the graphical method described above. Some of these formulae will now be given, specifically for the full-wave amplifier with common-emitter transistor control element (Fig. 8). A number of factors which are important in practice (e.g. winding resistance, rectifier characteristics, temperature effects on the transistor) are entirely neglected here; they can all in fact be adequately accounted for by fairly simple developments of the methods outlined in this article.

The saturation voltage V_s of the transductor elements should be equal to the mean value V_a of the output circuit supply voltage which, in turn, is equal to the

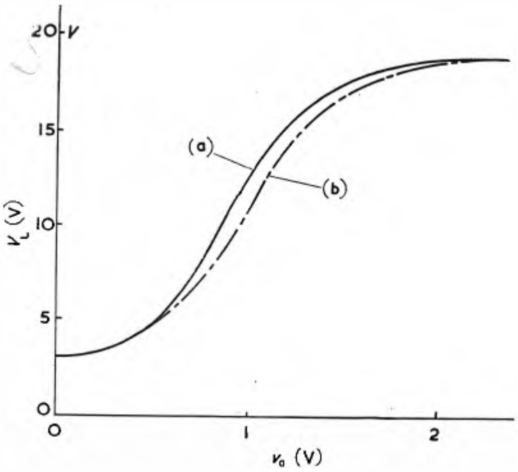


Fig. 13. Mean load voltage (V_L)/grid voltage (v_g) curve of a flux-reset magnetic amplifier element with pentode control element
(a) Observed
(b) Predicted

maximum load voltage $V_{L(pk)}$, say. Using equation (6), V_s can be expressed in terms of ω and the parameters of the transductor elements to give:

$$V_a = V_{L(pk)} = V_s = 2/\pi \,.\, 10^{-8}\ \omega\ AN\ B_s \quad \ldots\ldots \quad (15)$$

For maximum output linearity and gain, the reset circuit should operate the transistor up to its maximum permissible voltage, $V_{ce(pk)}$ say; this determines V_{ac}, the mean value of the reset circuit supply voltage:

$$V_{ac} = 2/\pi\ V_{ce(pk)} \quad \ldots\ldots\ldots\ldots \quad (16)$$

The ratio of output turns to reset turns is now given by:

$$\rho = V_a/V_{ac} = 10^{-8} \,.\, \frac{\omega\, AN\, B_s}{V_{ce(pk)}} \quad \ldots\ldots\ldots\ldots \quad (17)$$

To obtain an expression for the gain of the amplifier in terms that are characteristic of the transistor and of the core material, two further parameters are used. First, the mean value current gain of the transistor at the given supply frequency is defined as (Fig. 11):

$$\beta_{(mn)} = \left(\frac{\partial I_c}{\partial i_B}\right)_{V_{ce}} = \text{const.} \quad \ldots\ldots\ldots\ldots\ldots \quad (18)$$

$\beta_{(mn)}$ varies with V_{ce} and I_c but is usually approximately constant over most of the operating range, having a value of say 10 to 60 depending on the type of transistor. Second, the slope of the normalized mean value magnetization curve of the core material at the given supply frequency is denoted by $x_{(mn)}$. The slope of the I_T/V_T curve for a particular transductor element with parameters $A, N/\rho, l$ can be derived from $x_{(mn)}$ by the equation:

$$dV_T/dI_T = x_{(mn)} \,.\, AN^2/\rho^2 l \quad \ldots\ldots\ldots\ldots \quad (19)$$

$x_{(mn)}$ varies with peak core flux density but again is approximately constant over most of the operating range.

Referring to Fig. 12 and using equations (14), (17), (18) and (19) the following expression can be derived:

$$dV_L/di_b = -\,10^8 \,.\, \frac{N\, V_{ce(pk)}}{\omega\, l\, B_s} \,.\, x_{(mn)}\, \beta_{(mn)} \quad \ldots\ldots \quad (20)$$

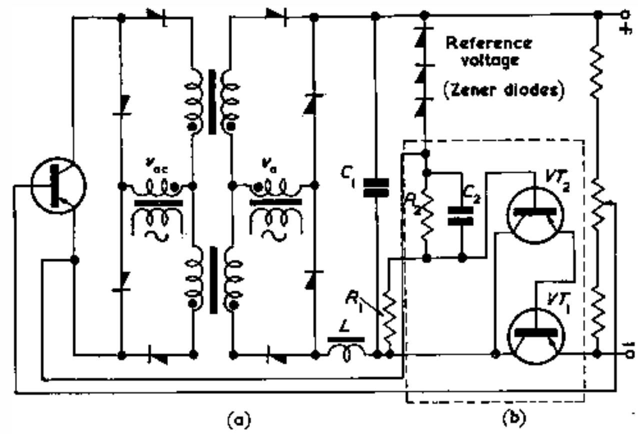


Fig. 14. (a) Full-wave transductor stabilized d.c. power supply
(b) Transistor output-ripple reducer

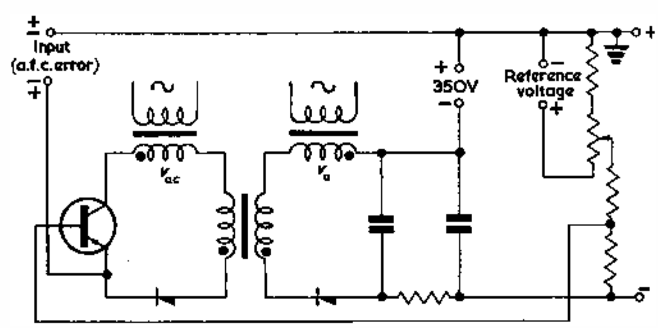


Fig. 15. Half-wave transductor stabilized variable d.c. voltage supply

If the input signal to the transistor base-emitter circuit is a voltage v_i and the total resistance in this circuit is r_i, then the voltage gain of the amplifier is given by:

$$dV_L/dv_i = -\,10^8 \,.\, \frac{N\, V_{ce(pk)}}{r_i\omega\, l\, B_s} \,.\, x_{(mn)}\, \beta_{(mn)} \quad \ldots\ldots \quad (21)$$

If $V_{L(pk)}$ and dV_L/dv_i are specified then A and N can be found from equations (15) and (21). (l is given by the size of lamination or toroid used in the tranductor element.)

Overvolting and Unbalancing

The flux-reset magnetic amplifier with transistor control element is particularly suitable for the stabilization of d.c. supplies. In Fig. 14 the output of a full-wave flux-reset magnetic amplifier, which is the full-wave rectified supply voltage suppressed over an initial part of each half-cycle, is smoothed by an LC filter to provide a d.c. voltage. This voltage is compared with a voltage reference (shown as a stack of Zener silicon diodes, but it could be a gas discharge tube) and the error is fed back to the input of the transistor control element. In the circuit of Fig. 15 a half-wave flux-reset magnetic amplifier is used with an

RC smoothing network; this circuit has advantages in a high voltage, low current supply. Fig. 15 shows also how the stabilized output voltage may be varied by an input voltage signal, about a value preset by the potentiometer. This circuit has been used as a voltage supply unit for the reflector electrode of a klystron oscillator (cathode at -350V), the stabilized output voltage being varied by an automatic frequency control error voltage.

The amplifier in a voltage stabilizer is required only to have a control range equal to the maximum output variation (due to supply and load changes, etc.) which would have been present with the feedback loop open. This output variation is usually only a fraction of the nominal output voltage.

In fact, the transductor elements in Figs. 14 and 15 need only be designed to have a saturation voltage V_s equal to this range of open-loop variation. V_a may then be increased or 'overvolted'[8] beyond V_s until the centre of the control range has been raised to the nominal output voltage; the amplifier output voltage can then be varied $\pm V_s/2$ about this nominal voltage.

Since core volume is approximately proportional to V_s, overvolting a stabilizing transductor in this way can reduce its bulk by a factor of say about a half to a quarter.

So far it has been postulated that the voltage in the reset circuit must balance the voltage in the output circuit, i.e. that $V_{ac} = V_a/\rho$. When V_a is increased to be greater than V_s, V_{ac} need not then be increased correspondingly, so that the magnetic amplifier operates in an unbalanced condition. By unbalancing an overvolted flux-reset amplifier it is possible to more than double the voltage gain. To ensure that the peak voltage across the transistor does not exceed the permissible value of $V_{ce(pk)}$, V_{ac} must however satisfy the following conditions:

$$K_{ac} = 1 \text{ for } 1 \leqslant K_a \leqslant 2 \text{ and } K_{ac} = \frac{K_a}{1 + \dfrac{K_a}{2\sqrt{(K_a - 1)}}} \text{ for } K_a > 2,$$

where $K_a = V_a/V_s$ and $K_{ac} = \dfrac{\rho V_{ac}}{V_s}$.

Transistor Ripple Reducer

If in the circuit of Fig. 14 the boxed part (b) is omitted, then the output voltage is stabilized only by the feedback loop formed by the error circuit and the transistor controlled magnetic amplifier. The magnetic amplifier bandwidth is half the frequency of the a.c. supply and the loop bandwidth is reduced from this by the filter in series with the magnetic amplifier output. This stabilization loop cannot therefore reduce output variations at a frequency above, say, about one quarter of the supply frequency.

In particular, the level of the output rectification ripple in the above circuit is entirely determined by the attenuation of the *LC* filter. This level has in fact, in a number of applications, proved adequate in stabilized supplies for transistor and valve circuits.

However the output ripple from the filter may be reduced by an order or more, or alternatively for the same output ripple level, the bulk of the *LC* filter may be diminished, by incorporating the simple transistor stabilizing circuit shown in Fig. 14(b). This comprises an emitter-follower, or as shown in the diagram, a compound emitter-follower, fed through the high-pass filter network formed by R_1, R_2, C_2. This filter is designed to have a cut-off frequency which is somewhat below the cut-off frequency of the low-pass filter formed by the magnetic amplifier in series with the *LC* filter. Thus fluctuations in the output voltage at a frequency within the bandwidth of the magnetic amplifier stabilization loop do not cause an input signal to the transistor stabilizer, whereas at frequencies higher than this they do. In short, the addition of the transistor stabilizer extends the total stabilization cut-off frequency from about one quarter the supply frequency to a very much higher frequency.

The transistor ripple reducer adds only three or four components (VT_2 may be omitted at the expense of ripple reduction) to the power supply circuit, viz. VT_1, VT_2, R_2, C_2. The collector current rating of transistor VT_1 must exceed the maximum output current of the power supply but only a low voltage need appear across collector-emitter (of the order of the ripple voltage across C_1). Transistor VT_2 and resistor R_2 require only very low power and current ratings. For supply frequencies up to about 2kc/s, C_2 requires to be a fairly large value capacitor (of the order of 10μF say). However provided the reference voltage is made large enough, to within say 20V

Fig. 16. Transductor stabilized power packs

(a) −10V, 200mA, (b) +25V, 200mA, (c) +125V, 100mA, (d) −350V, 65mA (before potting), (e) −780V, 2·5mA (before potting)

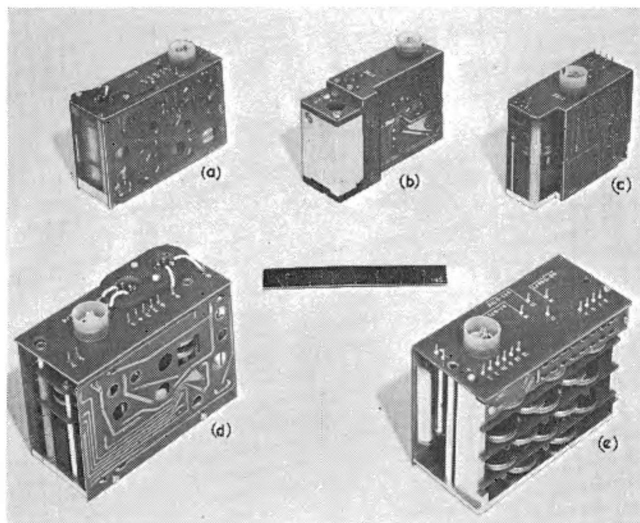
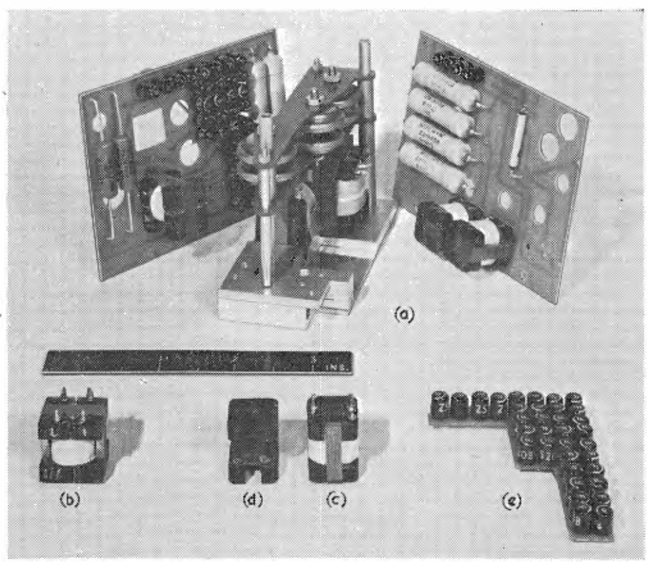


Fig. 17. Details of the −350V supply

(a) −350V block, (b) reset transformer, (c) a transductor element, (d) transductor element bobbin, (e) bridge rectifier stack



of the output voltage, a small electrolytic capacitor can be used.

The loss in the transistor ripple reducer need only be quite small and the reduction in power pack efficiency which it occasions should generally be of the order of a few per cent. For example in a 350V, 100mA supply the r.m.s. voltage across VT_1 collector-emitter was about 5V. Therefore VT_1 dissipated about 0·5W compared with the 35W output power.

If R_2 and C_2 are omitted, the transistor stabilizing circuit reduces to the transistor equivalent of the well-known series-valve stabilizer. In principle the transistor circuit then by itself completely stabilizes the supply and the magnetic amplifier can be removed. In this case, however, the voltage across the transistor collector-emitter is greatly increased, up to say one third of the output voltage, so that voltage supplies above a certain level (say about 200V with the best transistors available at present) cannot be stabilized solely by the transistor circuit. Further, in the d.c. coupled series-transistor stabilizer there is an average dissipation in the series transistor which is an appreciable fraction of the output power whereas the magnetic amplifier controls by a switching action which inherently results in low loss.

A Transductor Stabilized Power Pack

The circuits and design methods described above have been applied to the development of a power pack supplying five stabilized d.c. voltages from a 2400c/s a.c. power supply. Four of these supplies (−10V, 200mA; +25V 200mA; +125V, 100mA; −350V, 65mA) have circuits of the form shown in Fig. 14(a) but use miniature neon tube voltage references; their output ripple levels were adequate without transistor output ripple reducers. Fig. 15 shows the circuit of the fifth supply (maximum output −780V, 2·5mA).

The output power control range in each of the five amplifiers is about 50 times the maximum dissipation in the transistor control element. The overall power gains of the amplifiers are about 15 000 and their voltage gains range from ×30 to ×500. The overall efficiency of this power pack is about 50 per cent. An earlier power pack had an efficiency of 75 per cent, but for this application the efficiency was lowered to reduce size.

Fig. 16 shows the engineered power pack. Each of the five supplies is largely self-contained in an encapsulated block. A hole in the base plate of each block contains the single germanium junction 100mW transistor in each line. Printed circuit boards form vertical sides to a block.

Fig. 17 shows constructional details of a block. Transductor element cores are formed of 0·004in thick, $\frac{3}{4}$in by $\frac{3}{8}$in, U-shaped laminations in H.C.R.; these laminations are contained in a high-cheeked bobbin which obviates core clamps and makes possible resin-casting without magnetic deterioration[9]. Wire diameters down to 0·0016in (48 s.w.g.) are used. All rectifiers are composed of silicon diodes mounted on printed circuit boards. Capacitors of the tantalum electrolytic type are used exclusively.

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(*To be continued*)

The West German Public Radio Telephone Service

AN extensive frequency modulated mobile radio telephone service in the v.h.f. band is now being installed in West Germany and when complete it is expected to offer facilities between fixed subscribers on the German telephone network and mobile users located anywhere within the Republic. The system is intended for use on ships on the island waterways as well as road vehicles.

Under German Post Office regulations, all fixed v.h.f. radio installations for whatever purpose, are part of the public telephone network, but mobile installations are regarded as the responsibility of the user who must purchase apparatus of an approved type and undertake regular maintenance before a licence can be granted.

The present number of fixed installations forming part of the new public mobile telephone network and their approximate areas of coverage are shown in Fig. 1 from which it is seen that the major portions of the industrial areas and the Rhine are already covered. It is planned to extend the present mobile telephone network to cover the whole of the Federal Republic with the two-way system and, at a later stage, to introduce a simple one-way call system.

At present, eight channels are provided between the fixed station and the mobile user who can subscribe to one or all eight channels as desired.

A selective calling system is used to ensure privacy of communication and to prevent the use of any channel by more than one mobile user at a time in any given area.

Mobile Equipment

The mobile equipment consists essentially of a frequency-modulated transmitter and receiver for duplex operation in the 156-174Mc/s band, together with the associated selective calling unit, control unit and power supplies. The various units are interconnected by multicore cables and are mounted on anti-vibration mountings.

The present equipment has been designed and manufactured by Tekade of Nurnberg and has received type approval by the Central Office for Communications (Fernmeldetechnische Zentralamt).

TRANSMITTER

The mobile transmitter is a crystal controlled frequency-modulated ten valve unit with an output of 8W in the 156-174Mc/s band. A maximum of eight switched frequency channels are provided with a minimum spacing of 50kc/s within a total range of 800kc/s.

RECEIVER

This is an eleven valve unit of conventional design, with a sensitivity of better than 1μV for a 20dB signal-to-noise ratio.

The audio output is 1W.

SELECTIVE CALL UNIT

The selective calling system is operated on a 20/4 arrangement in which four specified audio frequencies out of the total of twenty used in the complete service are used to frequency modulate the v.h.f. carrier, which allows a total of 4 845 call 'numbers' to be composed.

These frequencies are demodulated by the selective call unit and the resulting output applied to a limiting low noise amplifier which is connected to a filter circuit comprising four relays, each of which is tuned to an assigned call frequency.

The selective call unit is transistorized and comprises three plug-in sub-units as follows:

(1) A 2 280c/s filter for operation of a 'disengaged' signal.
(2) A decoder consisting of four tuned relays.
(3) A control relay unit.

CONTROL UNIT

A two valve channel oscillator with its associated crystals is mounted in the control unit, which carries the handset.

The following controls are mounted on the unit:

(1) On-off switch for power supplies with pilot lamp.
(2) Selector switch for a maximum of eight channels.
(3) 'Call' signal lamp and buzzer.
(4) 'Line available' signal lamp.

POWER SUPPLIES

With the exception of the output stage in the transmitter, sub-miniature valves are used throughout and normally power supplies for the operation of the equipment can be obtained from a vibrator, or alternatively, from a transistorized power pack running from the vehicle's 12V battery. Equipment is also available for operation on 6 and 24V d.c. and 110-220V a.c.

Maintenance

A special service unit can be connected to the mobile equipment to enable routine checks to be carried out. In addition to the measurement of voltages and currents throughout the equipment, the transmitter output power can also be obtained with the use of the service unit.

Calling

Calling to and from the mobile user is carried out in the conventional manner.

An outgoing call from the mobile user is originated merely by raising the handset. When the call has been established, other stations operating on the assigned channel in that area are 'blocked' for the duration of the call to ensure privacy of communication. Calling from a mobile station to the fixed station is, of course, only possible when a channel is available and this is indicated by a green lamp in the operating unit.

An incoming call to a mobile user is also carried out

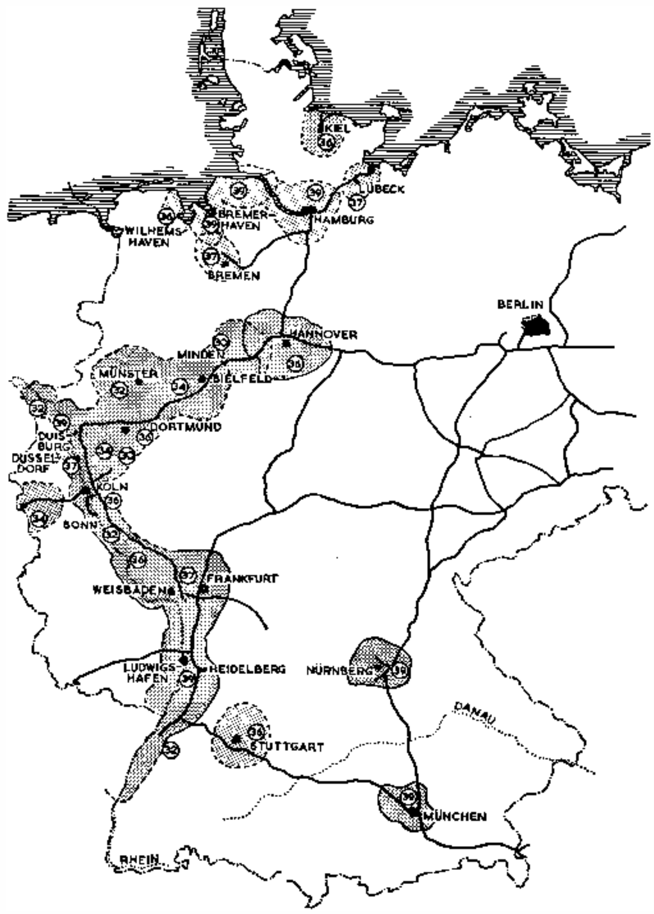


Fig. 1. The present coverage of the system

Areas enclosed in full lines indicate measured ranges while dotted lines indicate estimated ranges

in the conventional manner, but in addition to giving the number of the wanted subscriber, the area in which he is believed to be is required so that the appropriate transmitter may be used. Indication of an incoming call is given to the mobile user by the buzzer and call signal lamp on the control unit.

Public One-Way Mobile Call System

Considerable use has been made of the mobile radio telephone system, but there is a need for a simple one-way calling system and the German Post Office, in conjunction with Tekade, have developed a prototype system which is now undergoing field tests.

The one-way calling system is designed to operate side by side with the existing v.h.f. mobile telephone service and the mobile equipment consists of a simple receiver with a selective calling unit. An incoming call to a mobile user is received in the same manner as before and is indicated by a buzzer and a 'call' signal lamp.

The mobile user is then required to use the nearest public telephone where, after identification, the call is put through to him.

The 'call' signal lamp remains operated until reset manually and is intended to indicate that a call has been received should the driver have been temporarily absent from the vehicle.

A second signal lamp is provided which is effectively an indicator of field strength of the local transmitter in the area of the mobile user. The lamp is extinguished when the field strength is below the minimum and so serves as a warning to the driver not to park his vehicle in unsuitable areas.

Frequency Division Circuits for Musical Instruments

By Alan Douglas, M.I.R.E., M.Amer.I.E.E.

The distinguishing features of this class of circuit are, the division factor is always two; the maximum simplicity, for economic reasons; and the requirement that the waveform should be suitable for steady tone synthesis. It is difficult to meet all these conditions in a simple circuit.

(Voir page 592 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 600)

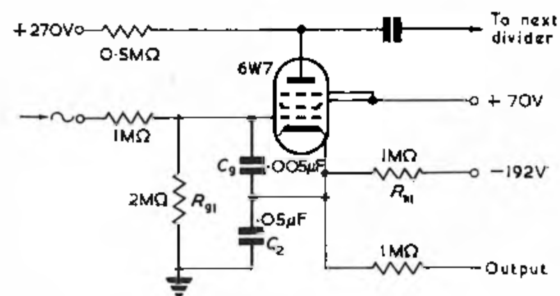
ALTHOUGH frequency division is still widely used in organesque keyboard instruments, this is only justified by the publicity departments of the respective makers. The usual plea that 'only the top octave has to be tuned' does not result in much saving either in time or money, as compared with a free-phase or independent oscillator system. There is little else to commend the method except the fact that from one chain of dividers, many pitches can be simultaneously derived. This makes the instrument appear to be larger than it really is.

The object of frequency division, of course, is to provide as many octaves of notes below the frequency of the oscillators feeding the dividers as is economically or tonally desirable. Because only this oscillator need be really stable, good quality components can be used and as nearly all dividers are of an *RC* type, the cost per octave is low compared with that of separate *LC* oscillators per note.

There are three main types of divider. Those employing gas tubes; those using an *RC* timing system; and blocking oscillator circuits. On the face of it, gas tubes would appear useful. It is comparatively easy to obtain the very desirable sawtooth waveform and there may be no heater current. Yet all attempts to use neon tubes have ended in failure. Some of the reasons for this were recently outlined in this journal[1]. One could quite well use thyratrons, and this is done in one instrument[2]; but they are expensive and require a hard valve as a buffer between each stage. So the cost is too high for full-scale polyphonic instruments.

By far the largest class of divider makes use of a timing system which is sometimes based on the principle that one incoming pulse unblocks a valve grid and allows conduction; the grid potential then gradually floats towards the blocking potential in such a way that the second incoming pulse will not be able to trigger it, but by the time the third pulse arrives, it has regained the correct conditions and so conducts once more. There are many variations of this kind of circuit, mostly developed with two objects in view; one, to be able to respond to some semitones of pitch, so that the divider constants will not have to be of too close a tolerance to make manufacture or servicing difficult; two, to be capable of accepting some variation of drive power level, so that small output variations from the oscillator will not either fail to trigger or overload the divider circuit.

With a well-designed system it is quite true that all octaves will be accurately in tune with the drive oscillator, but this fact does much to rob such instruments of any richness of tone quality due to the small inaccuracies in tuning which give this quality to conventional organs. Against this it must be noted that if the drive oscillator fails, some divider systems will free-run at their own natural frequency, which is usually different from the locked frequency. The top divider then acts as a driver for the rest so that some octaves of spurious notes are produced—a most disconcerting effect from the performer's point of view.



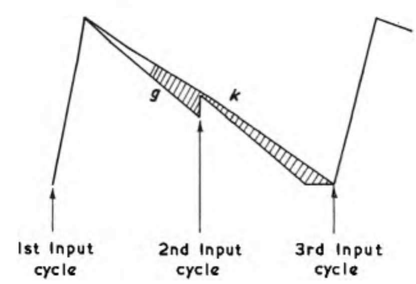


Fig. 1. (a) Single valve divider
(b) Waveforms from above circuit

One form of simple divider is shown in Fig. 1(a). The small standing anode current of the valve causes a drop across R_{k1} of 198V, thus opposing the −192V fixed bias supply and leaving a bias of 6V on the valve. The incoming signal is sharpened to form a triangular pulse, which causes the grid to become suddenly positive. The resulting flow of anode current increases the drop across R_{k1}, and makes the cathode become suddenly positive as well. This change in potential charges C_g which makes the grid even more positive. So when the valve is triggered, both cathode and anode potentials become so similar that no anode current can flow. C_2 is then charged to a similar voltage, represented by the peak of the wave in Fig. 1(b).

The grid and cathode now begin to 'float' towards earth, but not at the same rate. The grid potential (*g* in Fig. 1(b)) drops at a rate fixed by leakage from earth through R_{g1} into C_g and C_2 in series. The cathode potential (*k* in Fig. 1(b)) drops at a rate fixed by the leakage from the −192V line through R_{k1} into C_2. The time-constants are calculated so that *g* 'floats' down more rapidly than *k*. Thus when the second input cycle starts to raise the grid potential again, it will not exceed the cathode voltage and the valve remains cut off. The shaded area in Fig. 1(b) indicates the region where the grid-to-cathode voltage is sufficient to hold the valve cut off. The third input cycle, however, finds the grid voltage low enough to trigger the valve and the process starts again.

This divider has the advantage that it requires only one valve and cannot oscillate on its own if the drive fails;

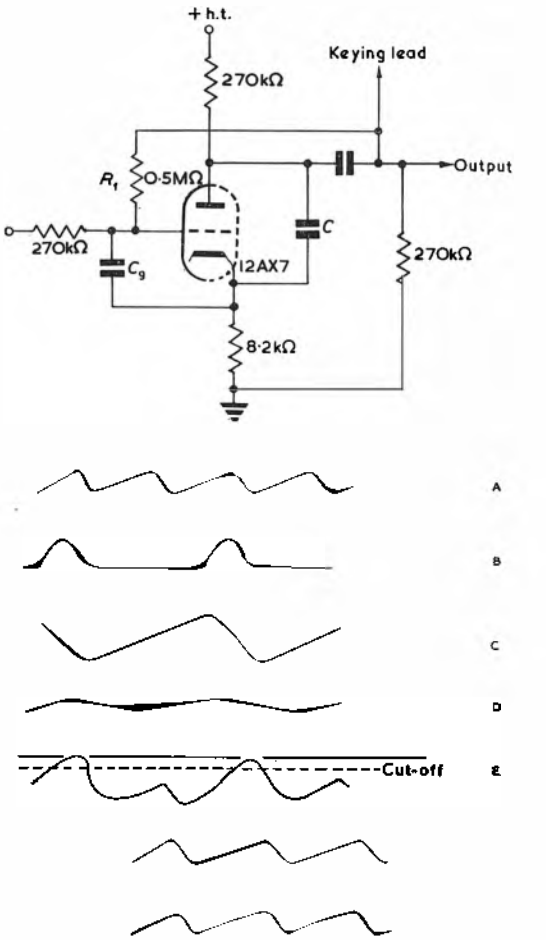


Fig. 2. **(a) Triode divider**
(b) Waveforms from above
(c) Permissible waveform degradation

it has the drawback that timing is very critical, and any change in bias or h.t. line voltages, input drive or component tolerances will affect the performance. Were the cathode to 'float' down too quickly, or the grid too slowly, the difference between their potentials may be so small that a second input signal could trip the valve. No division would then take place. Equally, if the position were reversed, the valve might remain cut off over more input cycles, so that division might, for example, take place by a factor of four. The waveform of Fig. 1(b) is very useful for tonal synthesis.

Another simple divider is shown in Fig. 2(a). This is rather ingenious in that a sawtooth waveform is delivered even if the incoming signal takes some other form. When the drive signal causes the valve to conduct, the fall in anode voltage occasioned thereby is fed back to the control grid, delayed in time by the constants of R_f and C_g. Thus when the second input wave arrives the grid potential is more negative, due to the fed-back signal, than when the first peak arrived. The amplitude of the incoming wave is not great enough to overcome this additional bias, so the second wave causes little or no conduction. The anode capacitor is therefore not discharged, on the contrary it continues to charge. By the time a third input peak arrives, the bias has returned to its original value, the valve conducts, and C_2 is quickly discharged. The resulting anode voltage is fed back to the grid, blocking it against the fourth input cycle, and so on. The waveforms in the various parts of the circuit during division are shown in Fig. 2(b). A is the input waveform; B the anode current waveform; C, the voltage across C_2; D, the signal fed back to the grid; and E, the composite waveforms (sum of A + D) on the grid. The ability to conduct only on alternate input signals is clearly shown in this latter illustration. The component tolerances are moderately critical, but they can vary slightly as indicated in Fig. 2(c). It is very difficult to detect the small second harmonic content at this amplitude.

Another kind of phase shift divider is shown in Fig. 3. Here, an additional valve is used to isolate the keying load and to drive the next stage. The time-constants of C_1, C_2 and C_3 have to be carefully adjusted to maintain locking and the drive power can be controlled by R_1 and

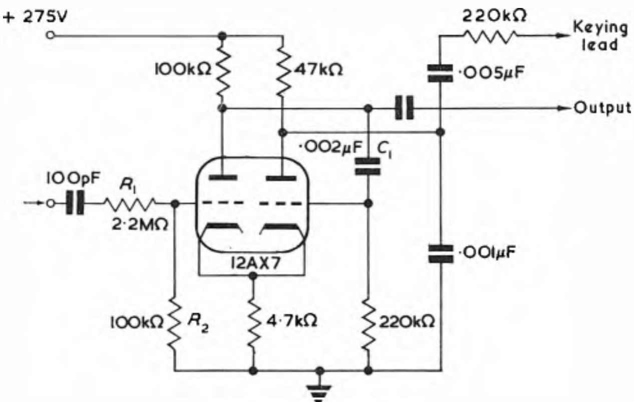


Fig. 3. Phase shift divider

R_2. The waveform from this divider is not particularly good.

The old-established Eccles-Jordan circuit is favoured by some designers, more especially where an approximately square output waveform is desired. The necessity for a double triode increases the cost and it is not easy to operate on the characteristic waveform to produce acceptable tone qualities. The main virtues of this circuit are that it will accept wide variations in drive voltage and frequency, and considerable latitude in component tolerances. A satisfactory version is shown in Fig. 4. To effect good triggering, the input waveform should be a sharp pulse; V_1 performs this function. The action of the circuit is well known to most readers, but as a number of musically inclined readers have a more limited knowledge, this may be summarized as follows. Suppose for example that the first triode is conducting and the second is cut off. A negative input pulse impressed on both grids cannot further cut off the second one, but will cut off the first. This action produces a positive pulse at the anode of the first triode, which is then applied to the grid of the second triode through its feedback connexion. This triode conducts suddenly, producing a negative pulse at the anode, which is applied to the first triode grid through its feedback circuit, so ensuring that the first triode remains cut off. Thus the initial state is reversed, so that the next input pulse will act on the second triode, cutting it off again while making the first conductive. Thus two input

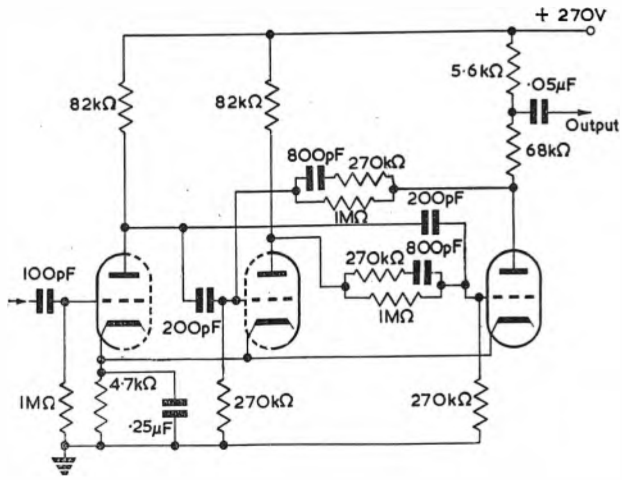


Fig. 4. Modified Eccles Jordan divider

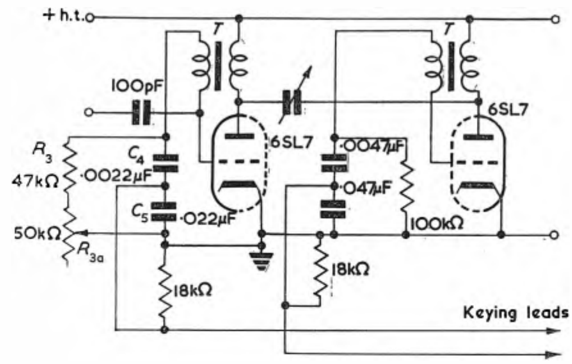


Fig. 5. Magnetic divider

cycles are required to provide one output cycle. Such dividers may be cascaded to quite an extent.

With the exception of this circuit, all the other forms of divider not so far mentioned and which employ *RC* networks require very close tolerance components, which sometimes drift out of their working range. Consequently the attractions of blocking oscillators are considerable. They are stable, generate large pulses, and in most cases are fully cut off in the absence of a signal. Chains of such dividers can be magnetically or capacitively coupled. An example of the latter is shown in Fig. 5. The pulse transformer *T* is so connected that a positive pulse at the anode will place a negative pulse on the grid, and vice versa. Supposing that a small random positive signal appears on the grid when switching on. Because of the high gain of the valve this results in a much larger negative signal at the anode. This is fed back to the grid by the transformer, making the grid more positive so rapidly that a large amplitude pulse of grid current flows. This current flows through R_3 and R_{3a} in series, causing a considerable voltage drop across these, the negative component of the voltage being at the grid. This negative pulse charges C_4 and C_5 in series, the voltage resulting therefrom cutting off the valve. This fast negative charge forms the vertical edge of the sawtooth wave, it is fast because of the high amplification of the valve. The valve is now cut off, whereupon the capacitors discharge through the resistors; the time-constant formed thereby giving the slope to the sawtooth. As soon as the capacitors have discharged enough to raise the grid voltage above cut-off, the action of the circuit starts again.

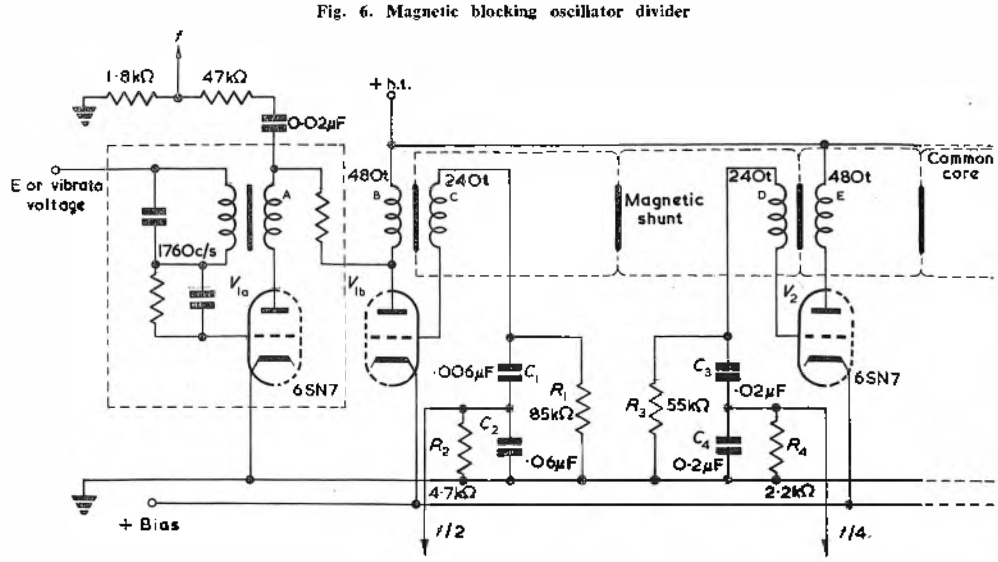
Since the sawtooth wave exists across the two capacitors in series, the reactance of the mid-point may be made low by suitable choice of the capacitance ratios. If, for example, C_5 is ten times larger than C_4, it has one-tenth the reactance, so the output impedance is low and there is little chance of the signal being spuriously fed back to the grid. These dividers can run free, but the time-constants are chosen to be very close to the locked frequency, and synchronizing pulses are fed to the valves through coupling capacitors C_6, etc. These latter are adjustable to compensate for component tolerance drift, etc.

There are variations of the foregoing circuit in use, but without doubt the most satisfactory magnetic divider is that shown in Fig. 6. One of the master oscillators for A = 1 760c/s is shown at the extreme left, and is contained in a screening can shown dotted. 1 760c/s is directly obtained through the impedance-adjusting network at the top of the circuit. Oscillations in winding A are transferred to B, thus inducing similar voltages in C by virtue of the tight coupling.

Now the grid of V_2 is biased so that this circuit cannot free-run. Capacitor C_1 and resistor R_1 are chosen to have a discharging time greater than *f*, so that when a sufficient signal level is obtained from the transformer conduction takes place only on alternate pulses therefrom. The frequency of $f/2$ is then 880c/s. C_2 is large in value and provides a low impedance path for the signal circuit, while R_2 forms an escape in the event of any d.c. component which might appear as a result of leakage in C_1. C_2 R_2 also assist in decoupling the signal circuit.

Because transformer D/E and any subsequent ones are wound on a common laminated core, enough energy is transferred from one coil to the next to supply several dividers. Moreover, through the introduction of mag-

Fig. 6. Magnetic blocking oscillator divider



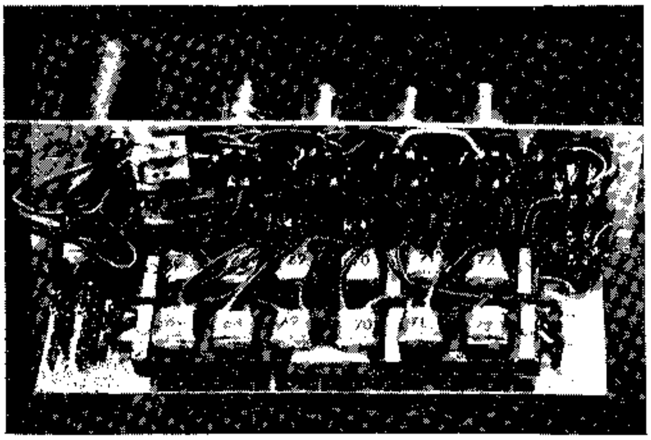


Fig. 7. Six cascaded dividers

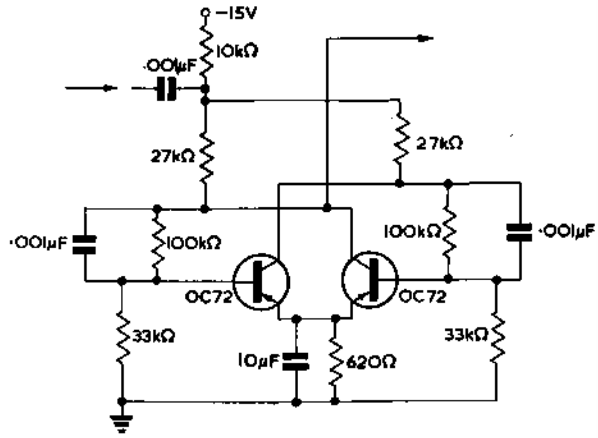


Fig. 8. Transistor frequency divider

netic shunts between the coils as shown, the amount of coupling can be suitably regulated. By using a magnetic system, the vagaries of capacitors when supplied with high energy pulses are avoided and this device is outstandingly stable. A photograph of six cascaded dividers is shown in Fig. 7, illustrating their small size.

Quite an amount of work is going on in the matter of transistor divider circuits, but so far no comparatively or competitively simple circuit giving the preferred sawtooth waveform has proved suitable for manufacture. Beyond length of life, there are not great advantages in an instrument which must necessarily be large in size.

Fig. 8 shows a successful divider yielding a square wave without overshoot, and which is suitable for a simple melodic instrument. It will divide over a number of octaves with the component values shown.

Divider organs were more valuable some years ago when some of the elements required for free-phase oscillators had not attained the stability they have today. In the better examples, tonal synthesis is fairly good for individual stops, but as must follow from the locked phase relationship, massive combinations of organ voices lack 'body' and are not to be compared with the effect obtainable from instruments using independent oscillators for each note.

But, even assuming good circuit design, the greatest difficulty with divider systems lies in keying the signal. There is no alternative, since all dividers must operate constantly. It is safe to say that no frequency divider organs simulate the speech of organ pipes in spite of much ingenuity; the best that can be expected is complete transient suppression.

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A Pulse Measuring Unit

A requirement existed for the accurate measurement of a pulse over a wide range of temperatures. Present known methods of achieving this were not suitable due to size, complexity or stability so the following system was designed.

Fig. 1 gives the logical circuit of the unit, where the pulse to be measured is fed into an amplifier and shaper in order to give the pulse a suitable rise-time. The pulse is then passed into the delay line whose length is the required delay time plus half the overall tolerance. ($d + \frac{1}{2}t$).

When the leading edge of the pulse reaches (a) a 'half set' is put on G_1 and a 'full set' on G_2.

If the duration of the pulse is greater than $d - \frac{1}{2}t$ a further half set is available from (b). These will open gate G_1, causing G_2 to be inhibited and an 'accept' answer to be read.

If the pulse is less than $d - \frac{1}{2}t$ G_1 will not open, therefore a 'reject' will be read from G_2.

If the pulse is greater than $d + \frac{1}{2}t$ then although G_1 will get two half sets, it will be inhibited by c, causing a 'reject' to be read from G_2.

Should it be required to read the actual pulse duration, the outputs from b, c, e, f etc., can be taken to a read-out unit where the number of sections having voltage on them, when the leading edge arrives at (a), can be simply indicated.

For a unit to measure 15msec ± 1msec to an accuracy of ±50μsec the unit would contain 160 delay sections, whose size would be approximately 20in³. The associated gates would take up approximately 4in³, making a unit of approximately 6in × 2in × 2in.

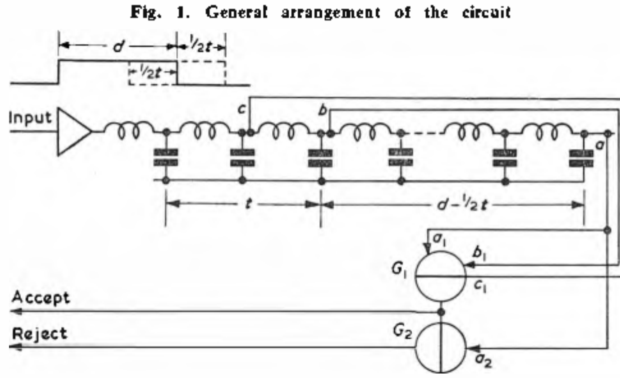
This compares very favourably with present known methods of obtaining this accuracy and can be made stable over a wide temperature range.

The units lower limitation is the $f\alpha$ of the gate transistors, therefore pulses of the order of ones of microseconds may be measured.

The upper limit is given by the size of the delay unit so that pulses up to tens of milliseconds may be measured.

The unit has been primarily designed for use on 'Go/No Go' test equipment, but it is considered the use may be extended to the computer field or any pulse measuring device.

Fig. 1. General arrangement of the circuit



* A communication from E M I. Ltd.

The Minimization of the Effect of Drift in D.C. Analogue Computers

By E. T. Emms*, B.Sc., A.R.C.S., A.M.I.E.E. and K. H. Brinkmann*

The effects of d.c. amplifier drift on d.c. analogue computers are explained, and methods of choosing scaling factors and apportioning gains to minimize these effects are given.

A number of guidance rules are formulated which will help the systems designer to make the greatest use of his equipment.

(Voir page 592 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 600)

THE purpose of this article is to define a number of rules which give guidance in the choice of scaling factors in d.c. analogue computing chains which if followed enable the systems designer to minimize the effects of d.c. amplifier drift. The reasons why drift occurs in d.c. amplifiers, and the methods of reducing such drift, are not considered to be within the scope of this article.

It is normal to arrange the output of a d.c. amplifier to be zero for zero input. This condition is set up by a 'zero' control. However, it is found that the zero output-zero input condition is not maintained with the passage of time, the output drifting away from zero. This drift is characterized by a short-term and a long-term effect. The short-term effect (minute-to-minute) is usually quite small and is normally not troublesome. The long-term effect (hour to hour and day to day) may be troublesome. It is usual to specify the expected drift over a given period, e.g. 8 hours. The most convenient way of specifying the drift is to specify it as that voltage which when applied to the input of the amplifier would restore its output to zero. This voltage is called the drift voltage δ of the amplifier and may be either positive or negative.

Definition of Drift Gain

The d.c. amplifier is invariably used with feedback. The common arrangement when a number of signals are to be fed to the amplifier is shown in Fig. 1. The feedback impedance may be:

(a) A simple resistance $Z(p)=R$ in the case of a summing amplifier,

(b) A capacitor $Z(p) = 1/pC$ in the case of an integrating amplifier.

Equating the current flow into the point A (Fig. 1):

$$(\alpha/R)(V_\alpha - V') + (\beta/R)(V_\beta - V') + \dots \dots \dots = \frac{(V'-V_0)}{Z(p)}$$

and the gain equation of the amplifier is:

$$-V_0 = G(V' + \delta)$$

hence, assuming G to be very large:

$$-V_0 = (Z(p)/R)\,\Sigma r V_r + \delta\,\{1 + (Z(p)/R)\ \Sigma r\} \quad \dots (1)$$

The first term on the r.h.s. represents the desired summed output due to the input signals. The second term represents the output due to drift. If the amplifier is a summing amplifier $Z(p) = R$ then the drift output is $(1 + \Sigma r)$ times the drift voltage δ. If the amplifier is an integrator $Z(p) = 1/pC$ then the drift output is:

$$(1 + (\Sigma r/pCR))\,\delta$$

and if $Z(p)/(R/r)$ or r/pCR is called the 'gain' to an input V_r, the following general rule can be formulated:

The drift output is given by the drift voltage δ multiplied by the sum of all the individual external gains plus one.

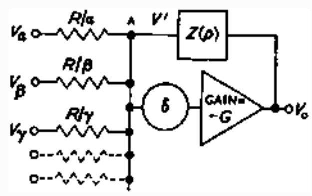


Fig. 1. Normal arrangement of a computer amplifier

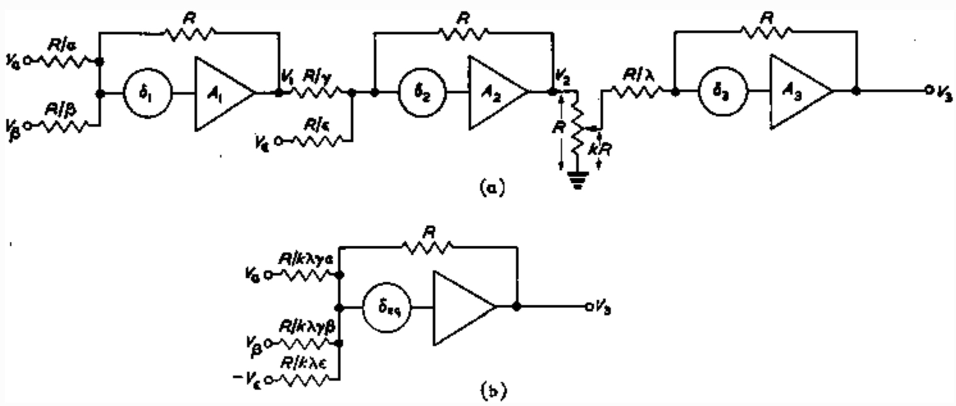


Fig. 2. Derivation of the equivalent single amplifier stage

This multiplier (the sum of the gains plus one) is called the *Drift Gain*.

Reduction of a Chain of Summing Amplifiers to an Equivalent Single Amplifier

If an odd number of summing amplifiers is cascaded, then for the purpose of drift analysis they can always be reduced to an equivalent single summing amplifier. For example consider the arrangement of Fig. 2(a). Applying equation (1) to each stage in turn and solving for V_3:

$$-V_3 = k\lambda\gamma\alpha V_\alpha + k\lambda\gamma\beta V_\beta - k\lambda\varepsilon V_\varepsilon + k\lambda\gamma(1+\alpha+\beta)\delta_1 - k\lambda(1+\gamma+\varepsilon)\delta_2 + (1+\lambda)\delta_3$$

The first three terms represent the outputs due to the

* *General Precision Systems Ltd.*

input signals, and the last three the outputs due to drifts. The system may clearly be replaced by the equivalent single stage shown in Fig. 2(b) where:

$$\delta_{eq} = \frac{k\lambda\gamma(1+\alpha+\beta)\delta_1 - k\lambda(1+\gamma+\epsilon)\delta_2 + (1+\lambda)\delta_3}{1 + k\lambda\gamma\alpha + k\lambda\gamma\beta + k\lambda\epsilon}$$

A word of warning is necessary. According to this equation, the drift due to amplifier (2) apparently tends to cancel that due to amplifiers (1) and (3). This is so if the

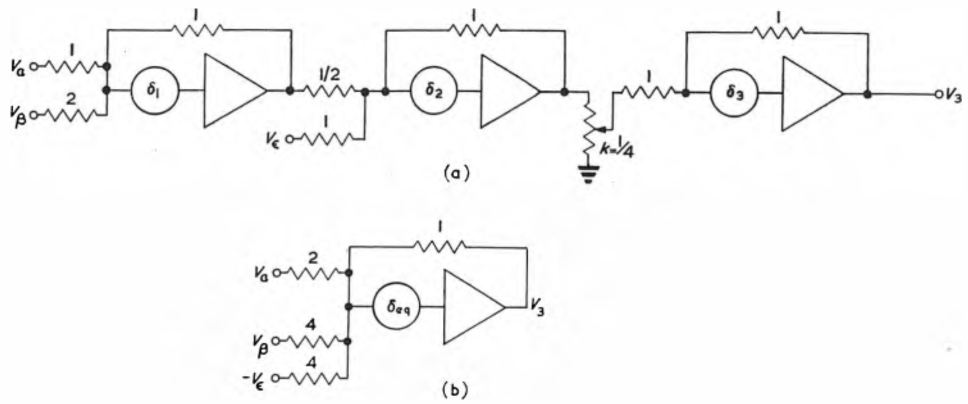


Fig. 3. Example of the equivalent single amplifier stage

amplifiers all drift the same way, i.e. δ_1, δ_2 and δ_3 are all of the same signs, but the amplifiers do not necessarily drift all the same way.

Fig. 3(a) shows a specific numerical example. If the drift voltages are equal and of the same sign, then the equivalent drift voltage is equal to 9/8 of the drift voltage of each amplifier, while if the pessimistic view is taken that $\delta_1 = \delta_3 = -\delta_2$ this figure becomes 17/8.

Proportioning Gain Throughout a Cascade of Summing Amplifiers

The following problem often arises: a number of signals are to be fed into a cascade of summing amplifiers at various points. The gains of these signals to the output must be constant. This leaves the problem of apportioning gains in the hands of the designer—how shall he best apportion these gains in order that the drift output should be minimized?

Consider Fig. 4. Applying equation (1):

$$-V_2 = -\alpha\beta V_\alpha - \beta\gamma V_\gamma + \lambda V_\lambda - \delta_1\beta(1+\alpha+\gamma) + \delta_2(1+\beta+\gamma)$$

The drift output is:—

$$-V_2\,(\text{drift}) = -\delta_1\beta(1+\alpha+\gamma) + \delta_2(1+\beta+\lambda)$$

If the amplifiers drift the same way by the same amount $(+\delta_1 = +\delta_2)$, then:

$$-V_2\,(\text{drift}) = -\delta_1(\alpha\beta + \beta\gamma - \lambda - 1)$$

The bracketed term is constant since $\alpha\beta$, $\beta\gamma$ and λ are all fixed. If, however, the amplifiers drift in opposite directions and one takes the pessimistic view that $-\delta_2 = +\delta_1$, then:

$$-V_2\,(\text{drift}) = -\delta_1(\alpha\beta + \beta\gamma + \lambda + 1 + 2\beta)$$

All terms within the brackets are fixed except β. To minimize the drift β is required to be as small as possible and therefore α and γ should be as large as possible maintaining $\alpha\beta$ and $\beta\gamma$ constant. Thus it may be concluded:

In a cascaded chain of summing amplifiers apportion the majority of the gain as far forward as possible to minimize drift. The limitation is set by the possibilty of overloading the forward stages.

As an example if $\alpha\beta = \beta\gamma = \lambda = 1$ then with $\delta_1 = -\delta_2$ the output drift would be $-\delta_1(4 + 2\beta)$ so that if $\beta = 1/2$ this becomes $-5\delta_1$ whereas if $\beta = 3$ it is $-10\delta_1$.

Pitfalls to be Avoided

(a) It is sometimes tempting for various reasons to attenuate an input signal using a potentiometer before application to an amplifier. Consider Fig. 5(a). Solving this network it is found $-V_o = V_1 + 3\delta$. The signal gain is therefore unity and thus corresponds to the simple circuit shown in Fig. 5(b) which has a drift output of only 2δ. Thus—

If possible avoid attenuating signals before summing into amplifiers.

(b) If the physical layout of the summing components is poor it is possible to introduce appreciable leakage at the input of the amplifier (Fig. 6). Here the resistor R/β represents leakage. The output is then:

$$-V_o = \alpha V_1 + \delta(1+\alpha+\beta)$$

If the feedback resistor is high (say 3MΩ) and the leakage resistance is low (say 1MΩ) then $\beta = 3$ and an appreciable increase in drift may result.

When a large value of feedback resistance is used, avoid introducing leakage at the input grid of the amplifier. If a choice of impedance level is possible, keep the feedback resistor down to 1MΩ or less.

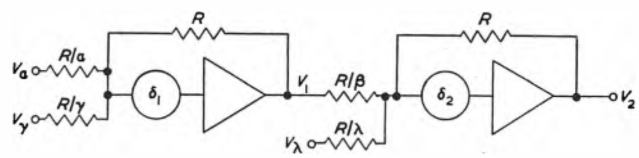


Fig. 4. A typical two-stage computer network

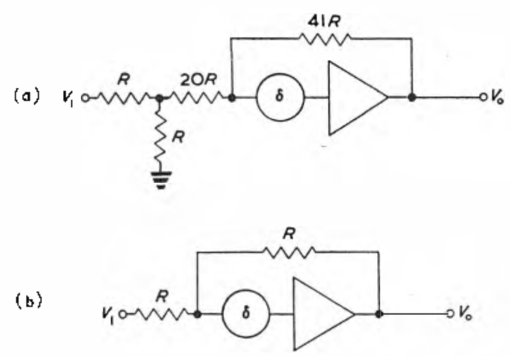


Fig. 5. Example of a common pitfall in circuit design

Drift in Integrator Loops

The most common loop employed in computers is that which contains one pure integrator. Thus a computer for solving problems on car suspensions would involve a minimum of five such loops. These loops would be interconnected but the calculation of drift in such cases is only increased in magnitude but not complexity. Thus it is justifiable to consider the case of a single loop, i.e. a loop containing one integrator and an even number of summing amplifiers. Any odd number of summing amplifiers may be reduced to an equivalent single stage, so the basic circuit shown in Fig. 7 will be considered.

Solving for the signal behaviour:

$$-V_3(p) = \frac{\Sigma r V_r/\alpha - \Sigma s V_s/\alpha\beta + \Sigma t V_t/\alpha\beta}{pCR/\alpha\beta + 1}$$

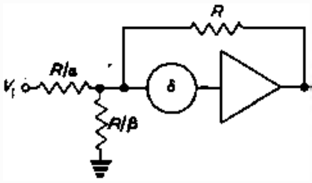


Fig. 6. The avoidance of leakage at the amplifier input

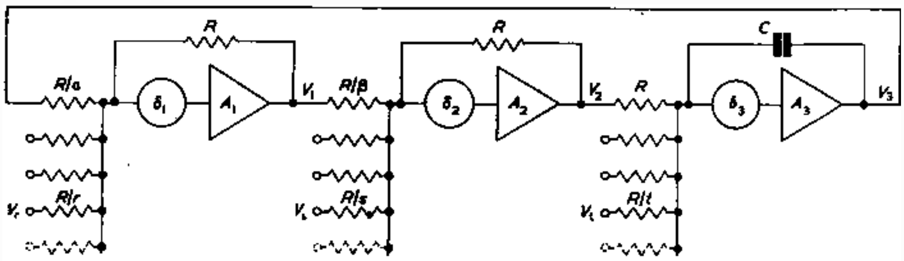


Fig. 7. A typical three-stage closed integrator loop

Thus the loop time-constant is $CR/\alpha\beta$ and the steady state output is:

$$-V_3 = \frac{\beta\Sigma r V_r - \Sigma s V_s + \Sigma t V_t}{\alpha\beta}$$

Solving the circuit for the drift output (in the steady state):

$$-V_3\,(\text{drift}) = \frac{\beta(1+\alpha+\Sigma r)\delta_1 - (1+\beta+\Sigma s)\delta_2 + (1+\Sigma t)\delta_3}{\alpha\beta}$$

The signal gains are fixed by other considerations, i.e. all terms of the type $\beta r/\alpha\beta$, $s/\alpha\beta$ and $t/\alpha\beta$ are fixed. Further the time-constant is normally fixed so that $\alpha\beta$ is fixed (although α and β may be varied with $\alpha\beta$ constant). Thus the only terms not fixed in the expression for the drift output are:

$$\beta\delta_1 - \beta\delta_2 = \beta\ (\delta_1 - \delta_2).$$

If the amplifiers A_1 and A_2 drift equally these terms cancel but if they drift in opposite directions they add and are proportional to β. Thus to minimize drift β should be small and α large with $\alpha\beta$ constant.

To minimize drift at the integrator output of an integrator loop the majority of the loop gain should be concentrated in the leading amplifier of the loop (that to which feedback from the integrator is taken). The limitation is the possibility of overloading.

The case has been considered where RC was fixed and therefore $\alpha\beta$ was fixed for a given time-constant. The effect may now be investigated of varying RC and $\alpha\beta$ so that their ratio is fixed and so that the time-constant is fixed. The expression for drift may be written as:

$$-V_3\,(\text{drift}) = \delta_1 + \frac{\delta_1(1+\Sigma r) - \delta_2}{\alpha} + \frac{(1+\Sigma t)\delta_3 - (1+\Sigma s)\delta_2}{\alpha\beta}$$

It will be seen that if $\alpha\beta$ (and correspondingly RC) is increased the last term may be reduced. The limitation occurs when practical difficulties are encountered. Thus an increase of C may be impractical if one is to employ low leakage, low dielectric hysteresis capacitors. An increase of R is limited by the grid current input characteristics of the amplifier.

As an example consider the circuits shown in Figs. 8(a), 8(b) and 8(c) each of which has a loop time-constant of 1sec.

Fig. 8(a): The majority of the gain is in the second stage. In the steady state:

$$-V_3 = V_\alpha - V_\beta + 4\delta_1 - 4\delta_2 + \delta_3$$

Fig. 8(b): The majority of the gain is in the first stage. In the steady state:

$$-V_3 = V_\alpha - V_\beta + 5\delta_1/2 - 5\delta_2/2 + \delta_3$$

Fig. 8(c): This is identical to Fig. 8(b) except that RC and the loop gain have been correspondingly increased. This necessitates rescaling of the signal inputs to give the same output. In the steady state:

$$-V_3 = V_\alpha - V_\beta + 21\delta_1/10 - 13\delta_2/10 + \delta_3/5$$

The signal behaviour is identical in all cases but the drift outputs differ.

By taking four possible cases Table 1 may be constructed.

Case A: All amplifiers drift the same way the same amount:

$$\delta_1 = \delta_2 = \delta_3 = \delta$$

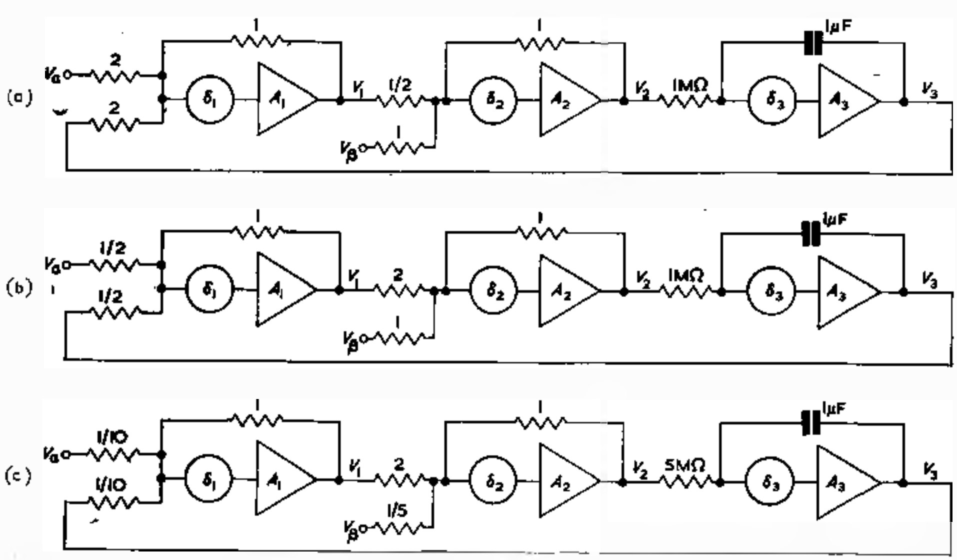
Case B: A_2 drifts the opposite way to A_1 and A_3:

$$\delta_1 = -\delta_2 = \delta_3 = \delta$$

Case C: A_3 drifts the opposite way to A_1 and A_2:

$$\delta_1 = \delta_2 = -\delta_3 = \delta$$

Fig. 8. Three circuits equivalent in signal performance but having different drift behaviour



Case D: A_1 drifts the opposite way to A_2 and A_3:

$$\delta_1 = -\delta_2 = -\delta_3 = \delta$$

TABLE 1

CASE	MAGNITUDE OF DRIFT AT INTEGRATOR OUTPUT		
	Fig. 8(a)	Fig. 8(b)	Fig. 8(c)
A	δ	δ	δ
B	9δ	6δ	$3{\cdot}6\delta$
C	δ	δ	$0{\cdot}6\delta$
D	7δ	4δ	$3{\cdot}2\delta$

These results illustrate the necessity for obeying the rules quoted.

Loop Zeroing

A standard method of zeroing amplifiers is to disconnect the computing components and replace them by a gain network with the input resistor grounded. It is the authors' opinion that a superior method for integrator loops is that referred to as loop zeroing. This procedure has the advantage that no switching at the input grid is necessary, and small differences in earth potential in a computer are automatically compensated for.

The principle of the method may be deduced by examination of the equations for the three amplifier loops. Consider all input signals grounded. Then, for the steady state drift outputs:

$$-V_1 = \frac{(1+\beta+\Sigma s)\delta_2 - (1+\Sigma t)\delta_3}{\beta}$$

$$-V_2 = (1+\Sigma t)\,\delta_3$$

$$-V_3 = \frac{\beta(1+\alpha+\Sigma r)\delta_1 - (1+\beta+\Sigma s)\delta_2 + (1+\Sigma t)\delta_3}{\alpha\beta}$$

From the expression for V_2 it is seen that if the output of A_2 is measured and the zero control of A_3 adjusted until $V_2 = 0$ then $\delta_3 = 0$. Proceeding forward, adjustment of the zero control of A_2 to obtain zero out of A_1 will make $\delta_2 = 0$. Finally the zero control of A_1 may be adjusted to give zero from A_3. This makes $\delta_1 = 0$.

To zero an integrator loop first ground all inputs. Observe the output of the amplifier preceding the integrator and adjust the integrator zero control for zero output. Move the meter forward one amplifier and adjust the zero of the amplifier preceding the integrator for zero indication. Proceed in this fashion, moving the meter forward and adjusting the zero control of the following amplifier.

Note that the sensitivity of the adjustment (i.e. the meter deflexion for a given rotation of the zero control) varies at each setting unless it so happens that the coefficients of $\delta_1\delta_2\delta_3$, etc., are comparable. This is not a disadvantage. If the loop is well designed from the drift point of view the middle stages of the loop will show the greatest sensitivity to adjustment.

Conclusions

The necessary approach to calculate the drift outputs in a computer has been given, and as a result a number of guidance rules have been formulated which will help the systems designer to make the greatest use of his equipment.

Acknowledgments

Acknowledgments are due to the Directors of General Precision Systems Ltd and Mr. G. Hellings, Head of Research, for permission to publish this article.

Automated Television Assembly

The production lines at the General Electric Company's television factory at Coventry are claimed to be the most fully automated in the world, with a high degree of mechanization at all stages of assembly and testing. Capable of producing two models every minute, the new methods have achieved an increase of 100 per cent in output and a 50 per cent saving in space since their introduction.

The introduction of printed wiring panels to the industry during the last few years is the key to many of the automatic and semi-automatic techniques now possible. At Coventry, the process starts with copper-clad plastic (phenolic) board being guillotined to size and pierced with two master holes for accurate location during later operations. A photo-resist is sprayed on the copper surface, and the panels are placed under photographic negatives in printing machines and held in place by vacuum. After exposure to ultra-violet light, the boards are developed by the action of a trichlorethylene spray to remove the unexposed photo-resist.

The next operation is to remove the unwanted copper from the panel, using a unique machine which carries out all the processes of etching, washing and drying. Panels are automatically conveyed through this machine, which first sprays ferric chloride on to the copper surface. After the residual acid has been cleared by high pressure water sprays, the boards pass through a drying oven.

Revolving brush scrubbers remove the remaining photo-resist to prevent interference with the subsequent soldering operations. The 300 holes needed for mounting components are then punched. When degreased, the panel is selectively varnished to expose only the areas of copper required for soldering. For convenience in servicing, the component reference numbers are printed on the reverse side by a silk-screen process.

Assembly of panels is in three operations: automatic component insertion, some manual assembly (such as transformers), and dip soldering. The automatic insertion covers most of the resistors, valveholders and capacitors.

The boards are first fed into rows of machines, which incorporate indexing arrangements to progress them from station to station at intervals of a few seconds. At each station from one to four components are inserted. At the completion of the automatic assembly the panels are released on to a moving belt which feeds the manual operators.

After inspection, completely assembled panels are passed through a mechanical soldering machine. A steadily moving chain conveyor picks up each board presented to it and moves it through the fluxing and dip soldering stages. There is then a careful inspection of the soldering, valves are fitted, and the panels pass to the testing machines.

The equipment in use at Coventry for testing i.f. panels breaks down the work into a series of operations, each of which takes the same time, in this case about 25sec. The machine warms up the valves prior to testing and automatically positions panels at various stations.

The printed panels are inserted into a series of identical trolleys and connexions are made by spring-loaded plungers and plugs and sockets. The trolleys are indexed into the test positions by air cylinders and automatic lifts are arranged at each end of the machine so that the trolleys circulate.

A card is attached to each trolley, and passes through a punching head at each station. If there is a fault, a relay automatically punches a hole in the card.

The main production lines start with the metal chassis framework and finish with the receiver in its carton. Assembly operations are carried out on long indexing tracks which position the chassis in front of each operator in turn. Manufactured items including printed panels, turrets, deflector coils and cable forms are fed to the appropriate positions by colour-coded trays on several conveyors. After the inspection and testing stages the chassis are fitted into cabinets which are supplied complete with protective glass from another overhead conveyor.

An Introduction to the Ternary Code Number System

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This article reviews the information concerning ternary arithmetic, gives a representation of the ternary system and its conversion to and from the binary and decimal numerical systems.

(Voir page 592 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 600)

NUMBERS can be represented in various ways which are designed to accomplish different purposes. The most common number system in daily use is the dozen, due mainly to the fact that twelve can be divided by more numbers than any other small integer. That which has the greatest advantage over the others is the decimal number system, which is said to have originated to assist counting with the aid of the fingers, as the decimal system consists of ten digit symbols.

In switching circuits and especially in the computer field, the binary code, where only two digits are used, has proved to be the most effective and most of the computers so far built use this system. The binary code system has a base of two and thus only two symbols are required, viz., zero and one. These two symbols are ideal for describing two state switching circuits by representing the 'on' condition by one and the 'off' condition by zero.

The number of digit symbols in each system is called the 'radix' of the system. The decimal system has a radix of 10 while the binary has a radix of 2. Any number N can be described by any numerical code system using the following function:

$$N = F_0R^0 + F_1R^1 + F_2R^2 + \ldots F_nR^n = \sum_{m=0}^{m=n} F_mR^m$$

R represents the radix which the number system uses and the integer F_m represents the symbols of the system. The value of the integer in each system must be smaller than the radix itself.

$$0 \leqslant F_m \leqslant (R-1)$$

In practice it is usual to write only the integers, one after the other, leaving out the radix. The order in which the integers appear is from right to left, which is in the opposite direction to that shown above.

$$N = F_n \ldots F_3F_2F_1F_0$$

The maximum amount of information which can be represented by any numerical system is $N = R^n$ and the corresponding amount of equipment which is required to represent that number will be $E = nR$. The minimum amount of equipment required to represent this number will occur when $R = e \simeq 2{\cdot}718$. This is derived by taking the minimum for $dE/dR = 0$ when $E = nR = (\ln N/\ln R)R$.

As stated before, the binary code system is most commonly used in computers and other switching circuits. The main disadvantage of this system is the amount of equipment required to represent each numerical number. A better code system could be that of the ternary code system, which has a radix of three, since this is the nearest whole number to $R = e$, that which gives the minimum equipment required. The three symbols of this system, zero, one and two, could represent the three states of an electronic switching circuit, plus, zero and minus. This article gives an introduction to the arithmetic of the ternary numerical code system.

Conversion into Ternary Codes

A number written in the ternary code system is expressed in the following form:

$$N = F_0 3^0 + F_1 3^1 + F_2 3^2 + \ldots + F_n 3^n = F_0 + F_1 3 + F_2 9 + F_3 27 + \ldots + F_n 3^n$$

For example the number 41 takes the form:

$$2.3^0 + 1.3^1 + 1.3^2 + 1.3^3 = 2 + 3 + 9 + 27 = 41$$

which is written in the ternary form as 1112.

The conversion of any decimal number N_{10} into the corresponding ternary number is effected by finding each integer separately.

The decimal number is divided by the radix 3 and the remainder noted. The quotient (N^1) of the first division becomes the new dividend, this is again divided by the radix 3, and the remainder noted. Division is continued until the dividend is less than the divisor 3. The value of each remainder indicates the corresponding integer. If there is no remainder then the integer is zero.

The process is summed up by the progressive formula given below and illustrated by the following example.

$$N/3 = N' + F_0;\quad N'/3 = N'' + F_1;$$

Decimal Number $N_{10} = 41$

$N/3$; $41/3 = 13$	Remainder	$= 2$	$F_0 = 2$
$N'/3$; $13/3 = 4$	,,	$= 1$	$F_1 = 1$
$N''/3$; $4/3 = 1$	,,	$= 1$	$F_2 = 1$
$N'''/3$; $1/3 = 0$	,,	$= 1$	$F_3 = 1$

Ternary Number $N_3 = 1112$.

To convert back from ternary to decimal numbers it is only necessary to multiply each integer by the appropriate power of three and then add them together.

As most computers are in binary form it is important to be able to convert from binary into ternary and vice-versa. Two methods can be used, the first is to convert the binary into decimal and only then into ternary, while the second method is a direct conversion using only binary and ternary codes. The binary number is divided by the radix three, as before in conversion from decimal into ternary, only here the radix three is taken in the binary form of 11. The remainder again gives the successive integers F_n but as the radix of the binary is smaller than that of ternary the integers F_n formed are still in the binary code. The conversion of the integers F_n into ternary form can be effected using the following propositions:

If the remainder is 0 then $F_n = 0$
If the remainder is 1 then $F_n = 1$
If the remainder is 10 then $F_n = 2$

The following example demonstrates this conversion, taking the binary number 101001.

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Binary number $N_2 = 101001$.

$$N/3 = \frac{101001}{11} = 1101 \quad \text{Remainder } 10 \quad F_0 = 2$$

$$N'/3 = \frac{1101}{11} = 100 \quad ,, \quad 1 \quad F_1 = 1$$

$$N''/3 = \frac{100}{11} = 1 \quad ,, \quad 1 \quad F_2 = 1$$

$$N'''/3 = \frac{1}{11} = 0 \quad ,, \quad 1 \quad F_3 = 1$$

Ternary number $N_3 = 1112$.

To convert the ternary number into the binary code it is necessary to carry out the same operation as before but this time dividing the ternary number into successive powers of 2. The following example demonstrates this operation.

$$N_3 = 1112 \quad N/2 = \frac{1112}{2} = 202 \quad \text{Remainder } 1 \quad F_0 = 1$$

$$N'/2 = \frac{202}{2} = 101 \quad ,, \quad 0 \quad F_1 = 0$$

$$N''/2 = \frac{101}{2} = 12 \quad ,, \quad 0 \quad F_2 = 0$$

$$N'''/2 = \frac{12}{2} = 2 \quad ,, \quad 1 \quad F_3 = 1$$

$$N''''/2 = \frac{2}{2} = 1 \quad ,, \quad 0 \quad F_4 = 0$$

$$N'''''/2 = \frac{1}{2} = 0 \quad ,, \quad 1 \quad F_5 = 1$$

$$N_2 = 101001$$

These operations for conversion from one code system into any other can be summed up as follows. The number N of the system which is being converted, is divided by successive powers of the radix of the new system. The radix code system and thus the integers obtained are still in the old system. If the new radix is smaller than the old radix then the integers formed are identical with those of the new code system. If the new radix is larger than the old radix then the integers must be converted into the new system.

Arithmetic Operations

All the mathematical operations of ternary numbers are similar to those of decimal numbers. In the decimal system, where there are ten symbols, if the addition of two symbols is greater than 9 then a digit is 'carried" to a higher order of ten, e.g. $5 + 7 = 12$. The same applies in ternary numbers, where there are only three symbols. If on adding two ternary numbers one obtains a number greater than two a digit is carried to a higher order of three. This is summed up in the following table:

Augend Number		Addend Number		Sum	Carry
0	+	0	=	0	0
0	+	1	=	1	0
0	+	2	=	2	0
1	+	0	=	1	0
2	+	0	=	2	0
1	+	1	=	2	0
1	+	2	=	0	1
2	+	1	=	0	1
2	+	2	=	1	1

In general, let X be the augend number, Y the addend number and N the whole addition of $X + Y$. Then for for all values of $N < 3$ the sum will be equal to N and the carry will equal zero, where $N = 3$ the sum will equal zero and the carry equal to one, and for all values of $N > 3$ both sum and carry digits will be formed, except when $(X + Y)/3$ is a whole number giving $S = 0$. If more than two ternary numbers are added, then the sum is equal to $N - 3n$ (having $N - 3n$ the smallest real number), and the carry will be equal to n ($n = 0,1,2\ldots$).

Subtraction of ternary numbers is effected by using the same method as in decimal subtraction. If the difference of the subtrahend number from the menuend number is less than zero then a digit is borrowed from a higher order of the menuend number. The table for ternary subtraction is as follows:

Subtrahend Number		Menuend Number		Difference	Borrow
0	−	0	=	0	0
0	−	1	=	2	1
0	−	2	=	1	1
1	−	0	=	1	0
1	−	1	=	0	0
1	−	2	=	2	1
2	−	0	=	2	0
2	−	1	=	1	0
2	−	2	=	0	0

In general, let X be the subtrahend number, Y the menuend number and N the full subtraction of $X - Y$. Then for all values of $N \geqslant 0$ the difference is equal to N and the borrow equal to zero, and for all $N < 0$ the difference is equal to $N + 3 = (X - Y + 3)$ and the borrow is equal to one.

Mutiplication of two numbers is obtained by adding repetitively the multiplicand number as many times as is defined by the multiplier number, e.g $X.3 = X + X + X$. In any numerical system it is usual to accomplish multiplication with the aid of a table which defines all the primary multiplications. The ternary multiplication table is very simple, as shown below:

	0	1	2
0	0	0	0
1	0	1	2
2	0	2	11

If two ternary numbers of 2 are multiplied then the result should be equal to 4 but as there can be no code bigger than 2 a digit is carried to a higher order and the result is $2 \times 2 = 11$.

Multiplication of any ternary numbers larger than those given in the table is accomplished in the same way as is employed for decimal multiplication. The multiplicand number is multiplied separately by each power of three of the multiplier number and then all the partial products are added together. This is demonstrated by the following example:

```
Multiplicand      1020
Multiplier        2001
                  1020
                 0000
                0000
               2110
Product        2111020
```

Division of ternary numbers is performed as in decimal division only keeping to the laws of ternary addition and subtraction. This is demonstrated in the following example:

```
1020)2111020    (2001)
  (-)2110
       1020
    (-)1020
```

Fractional Numbers

In the same way that the ternary code is defined, fractional numbers may be defined by using negative powers of three. Any number N, where $1>N>0$, can be represented in the form:

$$N=E_1R^{-1}+E_2R^{-2}+E_3R^{-3}+\ \dots\ +E_nR^{-n}=\sum_{m=1}^{n} E_mR^{-m}$$

The integers E_m can only take the value of 0, 1 or 2 and the negative powers R^{-n} represent the fractional powers. Fractional ternary numbers take the following form:

$$N = E_1/3 + E_2/9 + E_3/27 + \ \dots + E_n/3^n.$$

Any number N, if so desired, can be described in the ternary system where N comprises both fractional and whole parts, $0 < N < \infty$, in the following form:

$$N=F_mR^m+\ \dots F_2R^2+F_1R^1+F_0R^0+E_1R^{-1}+E_2R^{-2}+\ \dots\ +E_nR^{-n}$$

In abbreviating it is common to write the integers without the powers of the radix, having a comma between the whole and fractional parts.

$$N = F_m \dots F_2F_1F_0,\ E_1E_2E_3 \dots + E_n.$$

Conversion of fractional numbers into the ternary code must be done separately from the whole numbers. The fractional number is multiplied by the radix of three. If the result is less than one then $E_1 = 0$. If the result is greater than one then E_1 will equal the whole number of the result. The fractional part of the result is multiplied again to give E_2. The process is continued to obtain all the other integers. If the process is discontinued before it terminates, the last integer is rounded off. The following example demonstrates this process, converting 0·2 in the decimal code into the ternary code.

$N_{10} = 0{\cdot}2$	$0{\cdot}2 \times 3 = 0{\cdot}6$	$E_1 = 0$
	$0{\cdot}6 \times 3 = 1{\cdot}8$	$E_2 = 1$
	$0{\cdot}8 \times 3 = 2{\cdot}4$	$E_3 = 2$
	$0{\cdot}4 \times 3 = 1{\cdot}2$	$E_4 = 1$
	$0{\cdot}2 \times 3 = 0{\cdot}6$	$E_5 = 0$

The ternary number will be $N_3 = 0{,}01210$; this process can be continued to give the same pattern indefinitely and should be rounded off according to the required accuracy.

To convert a ternary fractional number back into the decimal code, it is necessary to divide each integer by the associated power of three, as demonstrated below.

$$N_3 = 0{,}01210$$
$$N_{10} = 0/3 + 1/9 + 2/27 + 1/91 + 0/273$$
$$= 0 + 0{\cdot}1111 + 0{\cdot}0741 + 0{\cdot}0109 = 0{\cdot}1961 \simeq 0{\cdot}2$$

Other Ternary Codes

As explained before, the ternary codes 0, 1 and 2 can be represented in electronic circuits by the three states of the circuit minus, zero and plus $(-1;\ 0;\ +1)$. These three states can be used to indicate the ternary codes by replacing 0 by $-$, 1 by 0, and 2 by $+$, and all the mathematical operations can be done by using this new code.

For example, the ternary code of 1112, 01210 ($N_{10}=41{\cdot}2$) will be replaced by $000+,\ -0+0-$.

The addition table will then take the following form, using the sign $(+)$ for addition:

Augend		Addend		Sum	Carry
—	(+)	—	=	—	—
—	(+)	0	=	0	—
—	(+)	+	=	+	—
0	(+)	—	=	0	—
+	(+)	—	=	+	—
0	(+)	0	=	+	—
0	(+)	+	=	—	0
+	(+)	0	=	—	0
+	(+)	+	=	0	0

The subtraction table will take the following form, using the sign $(-)$ for subtraction:

Subtrahend		Menuend		Difference	Borrow
—	(—)	—	=	—	—
—	(—)	0	=	+	0
—	(—)	+	=	0	0
0	(—)	—	=	0	—
0	(—)	0	=	—	—
0	(—)	+	=	+	0
+	(—)	—	=	+	—
+	(—)	0	=	0	—
+	(—)	+	=	—	—

There are other ternary codes which can be used in practice. The most interesting method is the one described below which enables the representation of negative numbers. The integers F_m can only take the form of -1, 0 and $+1$ or abbreviated, $-$, 0 and $+$. The row of codes represents the row of powers of three as before. The zero represents the absence of a power of three, the positive sign represents the presence of a power of three and the negative sign also represents the presence of a power of three, but in a negative sense. The following example demonstrates this method.

$$N_3 = F_n3^n+\ \dots\ +F_23^2 + F_13^1 + F_03^0$$
$$N_{10} = -3 \quad N_3 = 0{,}3^2 + (-1)3^1 + 0{,}3^0 \equiv 0 - 0$$
$$N_{10} = -8 \quad N_3 = 1{,}3^2 + 0{,}3^1 + (-1)3^0 \equiv + 0 -$$

The mathematics of this method is much simpler than of those considered above and can easily be applied in computer networks by using differential circuits.

Conclusion

As shown in the appendix, by using two cores per digit it is possible to obtain a three state device which can be used as a ternary switch. Ternary codes have the advantage over binary codes in that less digits are required to store the same number. Using two cores per digit doubles the number of cores which are required for the ternary store, so this system will be extremely expensive. Other three state devices are being investigated and may well prove a better solution.

Appendix

Ternary Switch and Store

A ternary switch could use ferrite cores which have a rectangular hysteresis loop. These cores are in one of two remanent states, $+B_m$ or $-B_m$, when no external field is applied.

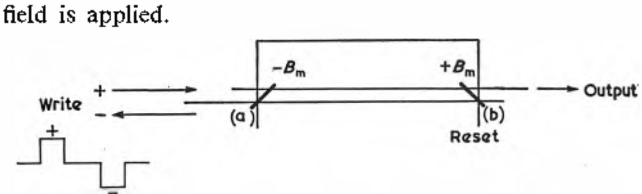
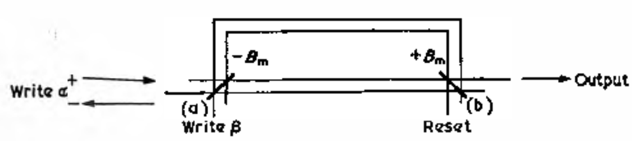


Fig. 1. Ternary core switch

Fig. 2. Ternary core switch using coincident currents



Two cores will be necessary for each ternary stage, one core being used to indicate the positive digit and the other to indicate the negative digit, while the initial state for both cores indicates the zero digit.

Both cores are wired in series as shown in Fig. 1.

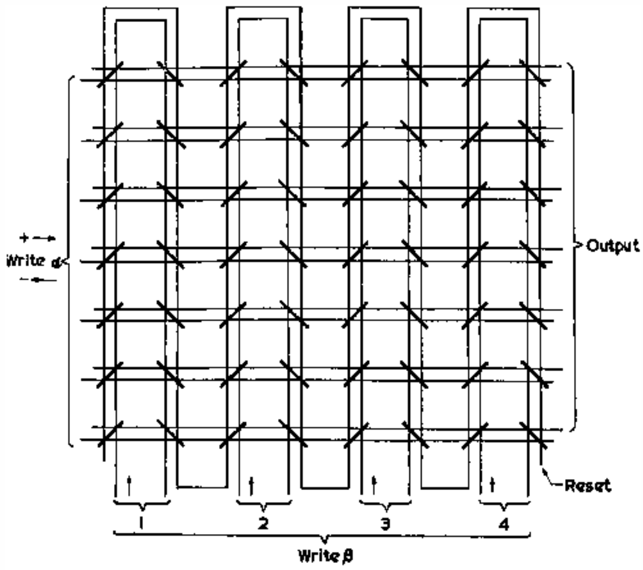


Fig. 3. Ternary core store

Initially, one core is in the $+B_m$ state while the second core is in the $-B_m$ state. The input pulses to write the digit '+' are of positive polarity and thus can only switch core (a). The input pulses to write the digit '−' are of negative polarity and thus can only switch core (b). The reset pulse, which occurs some time after the write pulse, is driven through both cores in such a way that it will return either of the cores back to its original state. An output pulse of a positive or negative polarity is obtained on the output coil, when the cores are reset, to indicate if a '+' or a '−' was previously put in the switch. If no output occurs at the reset time it will indicate the digit '0'. The magnitude of the write and reset currents in both cores will have to be of sufficient amplitude to switch the cores, viz. I_m.

Two coincident currents of magnitude $I_m/2$ can be used to switch the cores as shown in Fig. 2. This will enable the use of this circuit in a matrix form as shown in Fig. 3. The α write pulses will be of positive or negative polarity according to the digit which is to be stored. The β write pulses are always of positive polarity. Both write current pulses have an amplitude of $I_m/2$ thus only one chosen core will get the full switching field. Using the half-current selection method reduces the disturbance which is obtained on the output coil of the unswitched core. This is due to the fact that the α write pulse, on the core which is not switched, is in the opposite direction to the β write pulses. The selection of any core in the matrix is effected by selecting a row (write α) and a coloumn (write β).

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An Electronic Storekeeper

The Electronic Machine Co. Ltd. of Thornton Heath, Surrey, have recently introduced a specialized computer known as the 'Storekeeper'. It is a compact machine, little larger than an office desk and, as its name implies it is intended mainly for storekeeping applications.

Initiation of all records is done through the medium of a permanently incorporated electric typewriter which has been specially adapted to initiate the recording of information and is also available for the typing of all associated information on such documents as goods inwards or outwards advice notes, invoices and works issuing dockets etc., being adaptable for the handling of all suitable type continuous stationery or single sheets.

The console shown in the illustration is the basic model which is capable of storing 10 000 articles under each of 5 000 different catalogue numbers, thus giving a total record for up to 50 000 000 articles. Larger models are available according to article and catalogue requirements and are obtainable in multiples of 10 000 and 5 000 respectively. Operation is simple. By pressing the initiation tab and the plus or minus tab followed by the catalogue number, the recording drum is prepared for the acceptance of the quantity of articles to be added or subtracted, the manipulation of the following keys causing their number to be recorded.

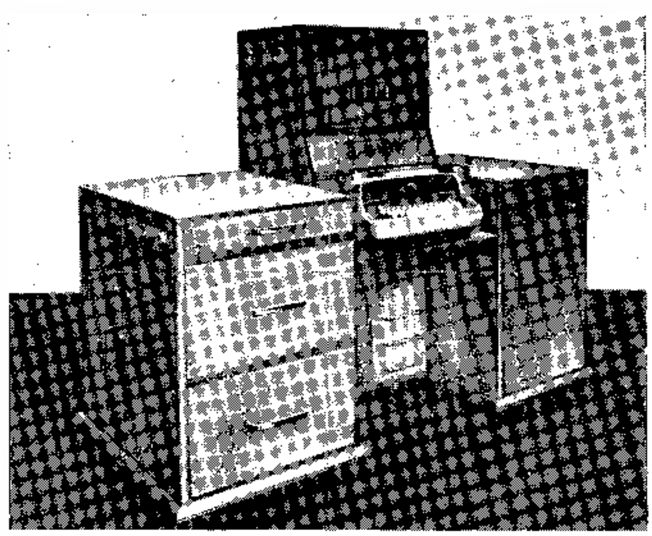
Two things then happen: The top row of illuminated digits will show the quantity of goods then being placed in or taken out of stock and the lower line of illuminated digits will show the number of items in stock under that particular catalogue heading *before* the new stock movement is recorded.

The 'Storekeeper' then calculates the new stock position and the electric typewriter then comes into operation and gives a typed record of these figures on the form that will have been inserted and for this purpose one half only of the platen is used, the second half being available for documentary typing.

In the event of requiring, without movement of stock, to know what stock is in any or all of the catalogue numbers, this can be done by the use of the appropriate tabs and keys, each answer coming up on the lower line of illuminated digits while being typed on the stock form.

The 'Storekeeper' can, of course, be used for other purposes and may find application in theatre and travel ticket agencies etc.

The basic 'Storekeeper'



A Hall Effect Multiplier for Use at Radio Frequencies

By E. Cohen*, M.Sc., A.Inst.P., A.M.I.E.E.

A short discussion is given of the limitations of available low frequency Hall effect when used at radio frequencies. It is shown that the determining factor in high frequency operation is carrier break-through caused by inductive pick-up in the Hall circuit (in the absence of a magnetic field) due to carrier current flowing in the plate. This may be minimized by certain circuit configurations and, in conjunction with plug-in type of broad band transformers, lead to carrier break-through greater than 50dB below maximum Hall signal.

(Voir page 592 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 600)

DETAILS have already been given[1,2] of the design and performance of a Hall effect multiplier for use as an analogue multiplier, as a modulator, or in control or power measuring equipment. These are low frequency applications and, for reasons to be described, the maximum frequency at which it is permissible to drive the Hall plate without loss of performance is of the order 100kc/s.

When a Hall effect field is established in a semiconductor a brief time elapses while the current carriers redistribute themselves. Welker[3,4] has discussed theoretically the frequency dependence of the Hall effect and Barlow *et al.*[5,6] have shown experimentally that substantial Hall voltages can be produced in germanium at frequencies up to several thousand megacycles per second. The frequency limitation in a practical multiplier is thus not in the Hall effect itself, but in the inevitable stray capacitances and inductances of the plate and its associated wiring.

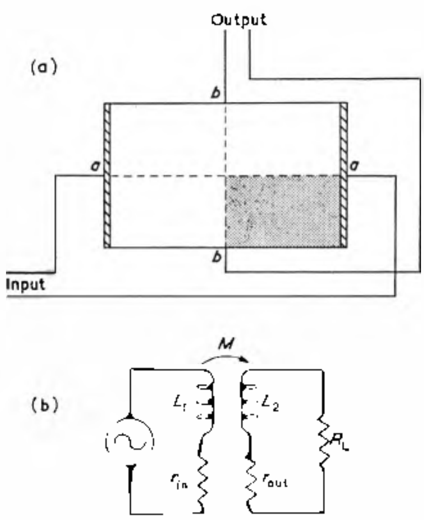


Fig. 1. (a) Cross hatched area one of greatest coupling (b) Equivalent circuit of r.f. injection process

L_1 - inductance associated with input circuit
L_2—inductance of Hall output circuit
r_{in}, r_{out}—resistances of semiconductor plate between a-a, b-b, respectively
R_L - load resistance

Consider a Hall plate, Fig. 1(a), whose Hall electrodes lie on an equipotential, i.e. zero Hall voltage output for zero applied magnetic field orthogonal to the plane of the plate. Ignoring capacitive effects, it is clear that the input and output circuits are inductively coupled together and a voltage will be injected into the Hall output loop. The equivalent circuit of Fig. 1(a) as regards r.f. injection is shown in Fig. 1(b). The configuration of Fig. 1(a) in any proposed device would be particularly bad as it would lead to a high coefficient of coupling and consequent large break-through of carrier signal into the output circuit. Also the rising impedance of L_1 with frequency will lead to serious mismatch of the input circuit.

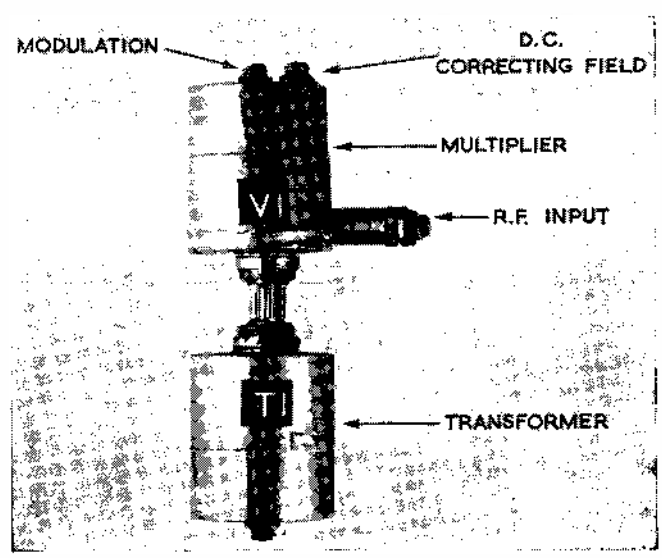


Fig. 2. Complete multiplier-transformer assembly

Starting from a ferrite assembly and indium arsenide Hall plate similar to that described previously[1], it has been found possible by careful design to reduce the carrier break-through to not less than 50dB below the Hall voltage in the frequency band up to 5Mc/s. The input impedance is also substantially resistive within this range.

Carrier break-through may also occur due to misalignment of the Hall connexions on the sample. In the low frequency multiplier this can be eliminated by means of a simple resistive network. In the high frequency case

* *Associated Electrical Industries (Manchester) Ltd.*

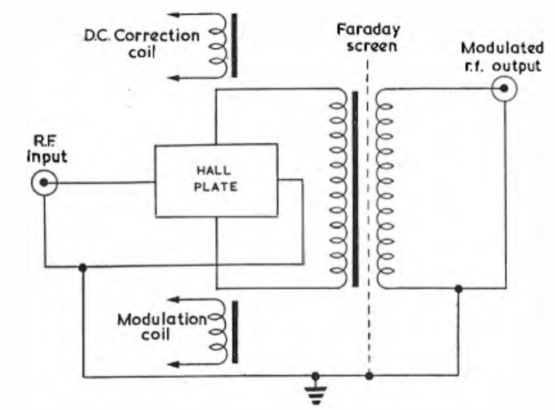


Fig. 3. Circuit representation of multiplier-transformer assembly

it is found more satisfactory to provide a small subsidiary coil on the ferrite core through which a direct current may be passed to provide a correcting magnetic field.

As a Hall plate is a four-pole network whose poles are internally connected, it is essential that the input source and output load should possess no common terminal. To accomplish this, wide band transformers have been constructed using a similar ferrite pot core assembly so that the transformer can easily be plugged into a corresponding socket on the multiplier. The photograph of Fig. 2 shows a complete multiplier and transformer assembly. Besides presenting a balanced input to the Hall electrodes one can make use of a step-up ratio in the transformer to obtain a larger output. A range of transformers has been developed possessing differing electrical characteristics as regards bandwidth, turns ratio, input impedance, etc., to provide optimum conditions for a given arrangement. The transformers have been designed in accordance with criteria laid down by Maddox and Storer[7], in which very low values of leakage inductance and stray secondary capacitance are ensured by the geometry of the windings and winding materials.

TABLE 1

Characteristics of typical indium arsenide multiplier—transformer assembly

MULTIPLIER PLATE	
Electrical conductivity	$200\Omega^{-1}$ cm^{-1}
Hall coefficient	$120cm^3$/coulomb
Plate dimensions	0·9cm × 0·4cm × 0·005cm
Resistance—current connection	3·2Ω
Resistance—Hall connexion	11·5Ω
Input inductance of plate circuit (L_1)	0·058μH
Output inductance of Hall circuit (L_2)	0·048μH
Mutual inductance (M, see Fig. 1(b))	$2·52 \times 10^{-10}$H
$k\,[=M/\sqrt{(L_1L_2)}]$	0·005
MULTIPLIER COIL	
Magnetic core gap	0·05cm
Coil winding data	2 × 1 000 turns, 40 s.w.g.
Coil capacitance	30pF
Coil inductance	206mH
Coil resistance	68Ω
Coil resonant frequency	64kc/s
TRANSFORMER CHARACTERISTICS	
Primary turns	4
Secondary turns	15
Primary inductance	10μH
Secondary inductance	139μH
Leakage inductance	0·1μH
Primary resistance	0·041Ω
Secondary resistance	0·83Ω
Open-circuit output at secondary of transformer with plate current 218mA, 5Mc/s and coil current 80mA d.c.	325mV

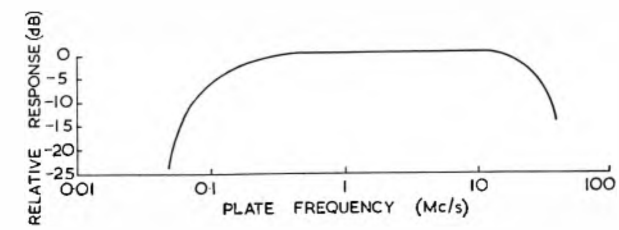


Fig. 4. Frequency characteristic of complete multiplier-transformer assembly on open circuit

Plate current 218mA
Coil current 80mA d.c.
Output voltage at 0dB 325mV

A circuit diagram of the compete multiplier-transformer assembly is given in Fig. 3 and performance data in Table 1. The main coil providing the modulating signal may be wound to permit operation up to several tens of kilocycles per second should this be desired.

The frequency characteristic of a typical multiplier-transformer device is given in Fig. 4, where the transformer has a step-up ratio of 3·75:1. The low frequency cut-off, which in practice is determined by the output impedance of the multiplier (~5Ω) and the transformer primary inductance (10μH) is approximately 100kc/s.

In practice it is likely that such a multiplier will be operated in one of two ways:

(a) Feeding into a transistor or the grid of a thermionic valve.

(b) Feeding a matched line of say 70Ω impedance.

When feeding into a valve or transistor, the transformer will be lightly loaded and the Hall plate operating into a high impedance. This is the condition for optimum linearity as described previously[2]. When feeding a 70Ω line a transformer of the type illustrated in Fig. 2 with a step-up ratio of 3·75:1 will be suitable for matching a 5Ω Hall plate. Some small loss of linearity will now occur due to magnetoresistive effects in the semiconductor.

Possible uses for such a multiplier may be as a suppressed carrier modulator[8], a contactless r.f. potentiometer[9], or a harmonic analyzer.

Acknowledgment

The author wishes to thank Sir Willis Jackson, F.R.S., Director of Research and Education, Associated Electrical Industries Ltd, for permission to publish this article.

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A 'DECIMAL' ATTENUATOR

A Constant-Resistance Voltage or Current-Dividing Network Indexed in Decimals

By E. R. Wigan*, B.Sc. (Eng.), A.M.I.E.E.

A network is described which combines the properties of a conventional voltage or current divider with those of a constant-resistance attenuator, retaining the most useful features of each while being free of their disadvantages. It will deliver at its output terminals precisely known decimal-fractions of its input voltage; for instance in a divider which has three decimal-dials the output/input ratio can be adjusted in 1 000-steps of 0·001 from zero up to unity The divider retains at all times a strictly constant output resistance; this means that current may be drawn from the output terminals without losing the ability to make precisely known adjustments of the output voltage or current, although the presence of the load will reduce the voltage or current available.

(Voir page 592 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 600)

EXCEPT for the fact that the output/input ratio is indicated in decimal fractions rather than decibels, the divider to be described has most of the properties of a conventional constant-resistance attenuator; the title given above emphasizes this point.

For instance, under the proper conditions, the effective output resistance is invariant with the dial-setting, and the input-resistance is similarly invariant. So far as the input resistance is concerned the divider behaves exactly as an attenuator, the constancy of resistance depending upon there being a load-resistance equal to the 'design-resistance' of the instrument; in the case of the output resistance the similarity is not exact, for the requirement is that the source-resistance feeding the divider shall be zero.

At first sight it would seem that to fulfil this condition would limit the field of use of the instrument drastically. However, as will be demonstrated, this is not the case, and it is found that the divider can, like an attenuator, be inserted into a transmission-path, without compromising its calibration, in order to produce a predetermined insertion-loss.

In other respects it behaves like a generator of constant internal resistance and continuously adjustable e.m.f.

The author's attention has been drawn to the publication by Chalmers University, Gotemborg, of a paper by Nils Bjork, entitled 'A Novel Class of Voltage-dividing Networks Inherently Free from Loading-errors.' Among the many networks he discusses, one of them embodies the conductance principle adopted here; however, it cannot be described or used as a constant-resistance 'attenuator' because the input-resistance is a function of the dividing-ratio even when correctly loaded.

The Circuit Principle

Although it can quite properly be described as a voltage-divider, the basic function of the instrument is to provide an output current, and its performance is best analysed on that basis.

The objective, then, is to provide, in a load-resistance (R_L)—see Fig. 1—a current (i_2), the value of which is to be adjustable with precision by means of a divider- network R_a, R_b, R_s.

The magnitude of i_2 will, of course, depend upon the value of R_L, and, writing E_o and R_o as the Thévenin parameters:—

$$i_2 = E_o/(R_o + R_L) \quad (1)$$

and, to obtain the equivalent output resistance (R_o), the input terminals are short-circuited (which eliminates R_s

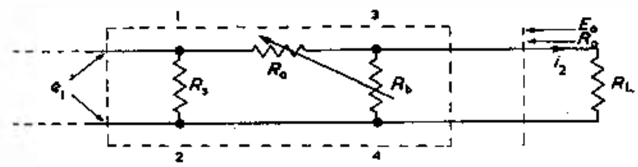

Fig. 1. Divider with generalized load resistance

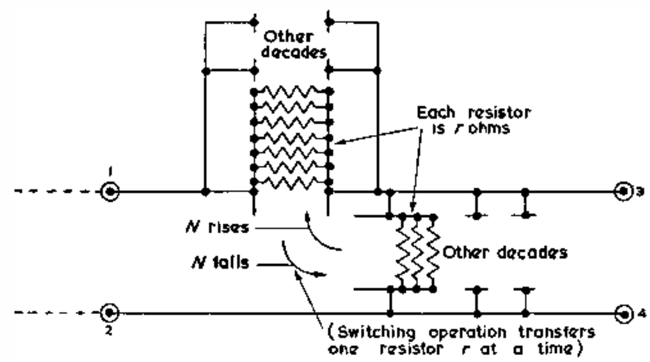

Fig. 2. Arrangement for varying the conductance of R_a and R_b

from the discussion), then:—

$$1/R_o = 1/R_a + 1/R_b \quad (2)$$

$$\text{and} \quad E_o = e_1 . R_b/(R_a + R_b) \quad (3)$$

which, on dividing through by ($R_a . R_b$), yields:—

$$E_o = e_1 . (1/R_a)/(1/R_o) \quad (4)$$

From the last equation it can be seen that, if R_o can be kept constant as the dividing-ratio is changed, the open-circuit output voltage E_o will vary directly as the conductance ($1/R_a$).

Fig. 2 shows diagrammatically how these conditions are fulfilled in the divider; details of one of the decade switches are shown in Fig 6. The effect is that conductance is transferred from the R_a to the R_b-branch of the network or *vice versa*, as the dial is turned. The figure shows that the output resistance is constant and equal to the inverse of the conductance of the complete array of resistors of which the divider is constructed. This resistance will be called R'; it is described as intrinsically constant because nothing but switch-contact resistance can modify it.

* *British Broadcasting Corporation.*

Writing the open-circuit output/input voltage ratio as:—

$E_o/e_1 = N$, (say);

then from equation (4):—

$$N = (1/R_a)/(1/R') \quad \ldots\ldots\ldots \ (5)$$

Again, from equation (1):—

$$i_2/e_1 = (1/R_a)/(1 + R_L/R') \quad \ldots\ldots \ (6)$$

Thus, once R' is established as a constant independent of N, the output current i_2 will always vary in proportion to N, which in turn varies directly as the conductance of the R_a-branch of the net, and therefore as the number of resistors-r transferred to that branch by the switching operation indicated in Fig. 2.

It follows that the value of N is indicated directly by the number of steps made by the switch in moving from its rest-position. For instance, in a divider consisting of one decade only, reading '0', '0·1' . . . . '1·0', the switch would have eleven positions, and there would be ten resistors-r. If r were made 10kΩ, R' would become 1kΩ — a value which will be adopted throughout this article.

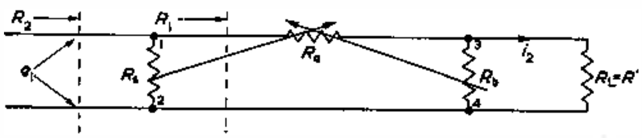

Fig. 3. Conditions when the load is equal to the design resistance

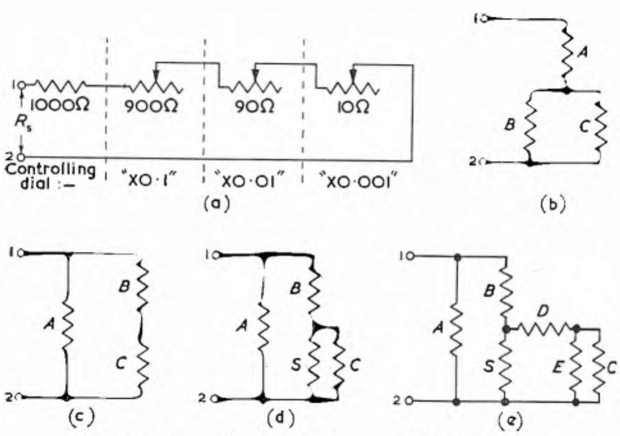

Fig. 4. Alternative forms of the variable shunt R_s

To build a subsidiary decade (×0·01) ten resistors of 100kΩ would be needed; these would be connected, as indicated in the figure, in parallel with the first set, but at the same time the eleventh position would be eliminated from the first decade. The same procedure is repeated for the next lower decade, using 1MΩ resistors, and so on. However, for reasons to be explained later, three decades suffice even for those applications of the divider which demand the highest possible precision.

Equation (6) shows that i_2 will vary in proportion to N, no matter what the value of R_L may be, provided e_1 remains constant; it is the function of the shunt-element R_s to ensure this. It has already been pointed out that none of the arguments up to this point involve R_s, and it follows that one is free to give it any desired value. In the network to be described R_s is controlled by the decade-dial switches, and varies as N varies. The relationship between R_s and N is discussed later; for the present it is sufficient to remark that R_s is so adjusted that the input resistance (R_2) in Fig. 3 is kept constant (or as nearly constant as will suffice) irrespective of the dial-setting, but before this can be done the value of the load R_L has to be specified. The designs dealt with here comply with:—

$$R_2 = R_L = R' = 1/(1/R_a + 1/R_b) \quad \ldots\ldots \ (7)$$

Having made the input resistance R_2 invariant with N, the internal resistance of the power source from which e_1 is derived becomes immaterial, and e_1 remains similarly invariant, as required.

Circuit Arrangements to Stabilize the Input Resistance

When R_L is made equal to R', the circuit takes the form of Fig. 3. The object is to choose a value for R_s which will make the total effective input resistance R_2 equal to R' for all values of N.

The arrangements of Fig. 2 ensure that the total conductance of R_a and R_b is always $1/R'$, and the switching operations lead to:—

$R_a = R'/N$

and $R_b = R'/(1 - N)$.

From Fig. 3 it is easily shown that when the load $R_L = R'$:—

$$R'/R_1 = (N - N^2/2) \quad \ldots\ldots\ldots\ldots\ldots \ (8)$$

Clearly also, if R_2 as well as R_L is to be made equal to R', it is necessary that:—

$$1/R_2 = 1/R'_s + 1/R_1 = 1/R' \quad \ldots\ldots\ldots\ldots \ (9)$$

Whence $R'/R'_s = 1 - N + N^2/2 \quad \ldots\ldots\ldots\ldots \ (10)$

In these equations R_s is written as R'_s to indicate that this is the ideal value called for to make $R_2 = R'$ for all values of N.

A Simple Form of R_s

In one very simple arrangement R_s is varied in sympathy with N so that it is equal to $(1 + N) . R'$; by this means R_s can be brought to within a small percentage of the ideal value R'_s. With R'_s = 1kΩ, each step of the auxiliary switch on the '× 0·1'-dial contributes 100Ω to the shunt path, and the auxiliary switches on the other dials contribute similarly, in proportion. All the auxiliary resistances are joined in series with 1kΩ and the total constitutes the R_s-branch (see Fig. 4(a)).

That such an arrangement of the shunt element R_s is on right lines, although imperfect, is shown by writing:

$R_s/R' = (1 + N)$ which (since $N < 1$), is the same as:

$$R'/R_s = (1 - N + N^2 - N^3 + \ \ldots\ldots) \quad \ldots\ldots \ (11)$$

whereas the ideal is, from equation (10):—

$R'/R_s' = (1 - N + N^2/2)$.

This very simple arrangement gives precisely the correct value of R_s when N is zero or unity; but when N has intermediate values the conductance $1/(1 + N)R'$ of the shunt is slightly too large to bring $1/R_2$ to equality with $1/R'$. Writing the error in $1/R_2$ as $(1/R_2 - 1/R')R'$, or more simply $\epsilon . (1/R')$:

$$\epsilon = (N^2/2)(1 - N)/(1 + N) \quad \ldots\ldots\ldots\ldots \ (12)$$

which has a maximum value of

$$\epsilon_{max} \simeq 0{\cdot}045$$

when $N \simeq 0{\cdot}65$.

Precision Versions of the Shunt-Element R_s

It is not difficult to obtain a closer approximation between R_s and R_s'; for instance, in the network of Fig. 4(b) the error ϵ is reduced to less than 0·4 per cent for all values of N. Detailed analysis of this and other arrangements is made in the Appendix, and it will be sufficient to say here that in Fig. 4(b) the resistance C, varies linearly in proportion to the settings of the second and third dials, while appropriate pairs of values of A and B are selected by the switches on the first dial.

The value of ϵ can be reduced still further by the network of Fig. 4(c). Here a linear variation of the C-element is discarded, and it becomes possible to bring the error ϵ precisely to zero whenever N is a multiple of 0·1; however, ϵ is finite at intermediate values when the second dial reads 0·05, and reaches a maximum of about 0·2 per cent at $N = 0{\cdot}05$ and 0·95.

This defect can be removed by adding another circuit-element as in Fig. 4(d) which allows ϵ to be reduced to zero whenever N is zero or a multiple of 0·05. The shunt is then correct to about 20 parts in 10^6 provided the 0·001-dial is not operated (except to zero or maximum), for this can increase the error.

If it is thought necessary this restriction, too, can be removed by adding another element to the network as shown in Fig. 4(e); the maximum value of ϵ lying then between 10 and 20 parts in 10^6 for all possible settings of the dials. Of course, in this case the constituent resistors throughout the whole divider must be set with an equivalent precision and this includes the load-resistance R_L.

For many purposes the shunt of Fig. 4(c) will serve, in spite of an error of rather less than 0·2 per cent, as will be seen from the discussion of the next section. The reason for this can be inferred from the last sentence of the section 'The Circuit Principle': an error in R_s affects measurement only so far as the input voltage e_1 is caused to vary from its (presumed) fixed value, and this variation in its turn can be reduced without limit by lowering the internal impedance of the power-source which supplies e_1.

Indeed, on this basis, the extremely simple shunt of Fig. 4(a), with its error of 4·5 per cent or so, will serve in some situations. However, in the next section applications of the divider will be discussed in which the power-source has necessarily an internal resistance equal to R', and in such cases the influence of the shunt-error is reduced by only one half; this, nevertheless, is sufficient to make the 0·2 per cent error of the shunt of Fig. 4(c) acceptable in any test-assembly aiming at 0·1 per cent precision.

Some Applications Using a Single Power Source

Three of the more elementary applications of the divider are illustrated in Fig. 5. In Fig 5(a) the network is shown as being inserted into a transmission path of design-resistance Z_p at the one side, and Z_q at the other side of the network. If $Z_p = Z_q = R'$, the divider will act only to produce an insertion loss equal to $N/2$, N being the dial reading; but, further than that, if Z_p is not equal to Z_q, but one of them is equal to R', the dial-reading will still be proportional although not equal to the insertion-loss-ratio—this might happen, for instance, if P represented the output stage of an amplifier or oscillator and Q the load; the divider then delivers, at Q, voltages or currents which will vary in strict proportion to the dial-settings.

In this arrangement the very fine adjustments which can be made to the loss-ratio by means of the second and third dials of the instrument provide a facility missing from the conventional 'attenuator' or 'output-control' commonly used for the purpose; and in Figs. 5(b) and 5(c), the fine-adjustment is used to yield calibration-voltages or currents for the meters M_2, in terms of the meters M_1 as 'standards'; the ability to subdivide the input voltage or current into 1000 parts allows the scales of the instruments to be checked (or, if required, to be marked off) with a precision of 0·1 per cent.

When, in the case of Fig. 5(b), it is impossible to arrange that M_2 has a resistance which is equal to R', the variation of the dial-setting will cause changes in the input resistance of the divider, but, if M_1 is restored to its original reading, the output voltage e_2, or current I_2, will be in strict proportion to the divider setting.

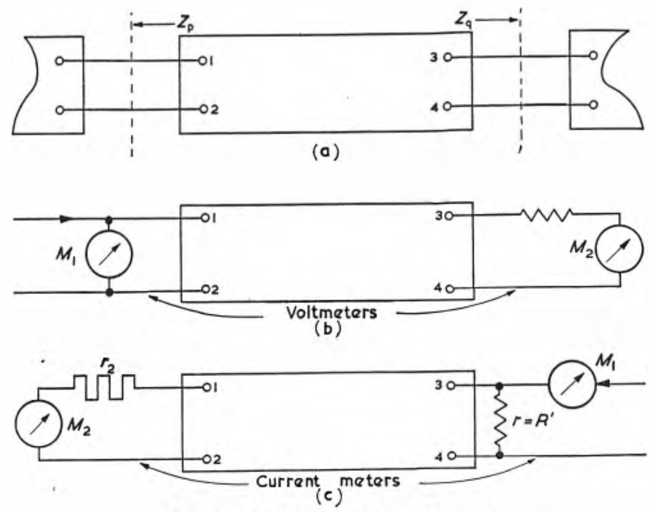

Fig. 5. Some elementary applications

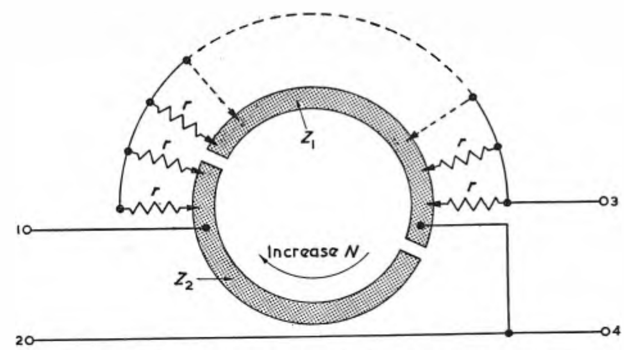

Fig. 6. Switch for controlling R_a and R_b elements

Fig. 5(c) is rather different; the meters are current-meters, and the divider is used in reverse with the 'input' current applied to terminals 3-4. It is to be noted that a shunt resistance r equal to R' has been connected across terminals 3-4; the purpose of this is to adjust the effective output-resistance (looking to the right from terminals 1-2) to the value R'. That R' is in fact the effective output resistance can be seen by treating the 'constant-current' feeding-circuit (the meter M_1 and the shunt R') as identically equivalent to a 'constant-voltage' feeding-circuit of e.m.f. $(I_1 . R')$ and internal resistance R'. It is then clear that R' acts to make constant the resistance seen from the 'output' terminals 1-2. The current I_2, which flows in meter M_2 is then given by:—

$$I_2 = I_1 . N/2 . (R')/(R' + r_2) \qquad \ldots\ldots \ (13)$$

(where r_2 is the resistance of meter M_2).

It will be clear that, whatever the value of r_2, the current fed to M_2 will vary in strict proportion to I_1, provided only that r_2 is constant. This means that alternating current meters which include rectifiers cannot be calibrated by this means because the resistance r_2 is a function of the meter reading; they can be dealt with by methods given later, however.

By these simple arrangements, therefore, the divider can be used to compare voltmeters with voltmeters, current meters with current meters, and voltmeters with current meters; if the resistances of the meters under test are unknown it is not possible to carry out a true calibration, but the scales may be subdivided with precision in terms of a fiducial mark—say at full-scale deflexion.

The output voltages or currents will necessarily be less than those at the input of the divider, and in many cases may be of the order of one half; this means that the field

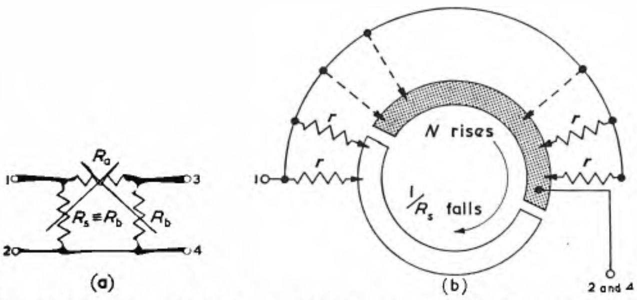

Fig. 7. (a) Divider which has constant input resistance R′ when load is zero
(b) Associated shunt-element switch

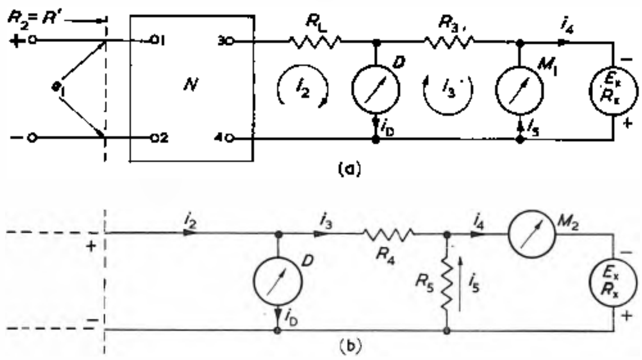

Fig. 8. Use of divider for precision voltage and current measurements

of use of these simple circuits is restricted, for the voltage which can be permitted at the output terminals of the divider will be limited to, say, 10V; the output currents will similarly be limited to about 10mA. For larger values a different technique is called for, but before dealing with this, a very simple form of the divider will be described which is particularly suited to that technique.

A Divider Which Operates into a Zero Resistance Load

In Fig. 7(a) the divider network has been re-drawn with the shunt-element R_s modified to be identical with the R_b-element. This means that the switching arrangements made to change R_b are duplicated in the R_s-element, except that there is, in the R_s-element no transfer of conductance from the shunt to the series branch of the network as was provided in the switch of Fig. 6, for controlling R_a. The switch-structure (Fig. 7(b)) is identical with that used for R_b, but the connexion to the ring Z_2 is broken, and the connexion to terminal 3 is transposed to terminal 1: then, when N is decreased, the conductance of R_s grows in sympathy with the increase of conductance of R_b, and both are at all times equal to $(1-N)/R'$. Element R_a remains as before, and has the conductance $(N)/R'$.

If now the total conductance of the divider is measured either from terminals 1-2 or from 3-4, with the opposite pair short-circuited, the value is the same, and independent of N (namely $(N + (1 - N))/R' = 1/R'$) the first term being provided by the series branch of the network, and the second by whichever of the two shunt branches is immediately adjacent to the measuring point; the other shunt element is connected across the terminals which, on the Thévenin basis, are considered to be 'short-circuited', and therefore does not contribute. It follows that when the R_s-element is arranged in this way, the divider has both its input and its output resistances equal to R', provided the load (R_L) is of zero resistance. Consequently, the short-circuit load-current will be strictly proportional to N, while the input resistance will remain fixed at the value R'. The short-circuit output-current will be:—

$$I_2 = N\,.e_1\,./R' \quad \ldots\ldots\ldots\ldots\ldots \quad (14)$$

Measurement Techniques Which Employ Two Power Sources

PROCEDURE

In Fig. 8 there are two sources of power, and they are so connected that both assist the flow of the currents i_2, i_3 and i_4. For simplicity d.c. conditions are assumed, but the equivalent a.c. circuit differs only in minor points, one of them concerns the 'detector', the null-reading instrument D, and will be dealt with later.

The secondary source is indicated as E_x, R_x and requires only to be stable over the period of the experiment and continuously adjustable. It supplies all the power necessary to operate the meters M_1 or M_2, the primary source at e_1 providing only such input current to the divider as will produce the current i_2 at the output.

In calibrating a voltmeter, connected at M_1 in Fig. 8(a), the primary input voltage e_1 is first standardized by one of the conventional procedures, the terminals of the null-meter D being short-circuited at this stage in order to make certain that the input resistance of the divider is precisely defined at the value R' while the test proceeds.

The reason for short-circuiting D at this stage is that when the test is made the current i_D in the detector D is brought to zero by feeding from (E_x, R_x) a current i_3 which is precisely equal to i_2; the divider must therefore be operating under conditions which cannot be distinguished from those shown in Fig. 3. When i_D is zero the output current i_2 is therefore given by equation (6) and is completely defined in terms of e_1, N and R'.

If the divider is of the type described in the last section, R_L is made zero, and $i_2 = e_1\,.N/R'$; if the alternative network is employed, i_2 has exactly half this value, R_L being set equal to R'.

In both cases, however, i_3 is precisely equal to i_2 when i_D has been brought to zero; thus, if the meter M_1 is to be calibrated at 100V, a resistance is provided at R_3 which is sufficient to bring the product $(i_3\,.R_3)$ to 100V. This can always be done even though the voltage e_1 available at the input of the divider is very small (for instance it will be 1·0185V if a direct comparison with a standard cell has been made). The current i_5 required to operate the meter is provided by the source E_x and does not play any part in the calibration procedure.

Current meters are connected at M_2 in Fig. 8(b). The precision resistors R_4 and R_5 act as current-shunts and allow heavy-duty ammeters to be dealt with, if need be, by reducing R_5/R_4 sufficiently; to obtain a calibration of M_2 in amperes relative to milliamperes at i_2, the ratio R_5/R_4 is set to 1/999; and so on.

Since the meter resistance cannot, in either case, affect the accuracy of calibration in any way, alternating current instruments which have non-linear resistance may be tested by this procedure.

Means for Increasing the Precision of Measurement

It will be noticed that the circuits illustrated have the considerable advantage that both power-sources, and also the 'detector' have one terminal at earth-potential—a feature which improves the precision of a.c. measurements very considerably. In addition the 'sharpness' of the null-reading can be improved very greatly by taking advantage of the pseudo-zero impedance existing across the terminals of D; in a.c. measurements D represents a series-resonant circuit the input terminals of an amplifier being connected across a small resistance in series, an arrangement which will not only remove unwanted distortion terms in the signal but also the greater part of the circuit noise.

There are many reasons for making the resistance at the point D as low as possible. One of them is concerned with a very useful technique for extending the decade-range of the divider: a three-decade instrument may by this means be used to make measurements at points which lie between the three-figure index-marks. It is, of course, a well-established procedure to make such interpolations, the 'null'-indicator being allowed to show a small reading which is then calibrated in terms of the available voltages or currents; in the present circuit, however, no such 'calibration' of the off-zero reading of D is required.

It is easily shown that when the resistance of D is small (e.g. $\not> R'/100$) a given change of instrument setting (e.g. the moving of the final dial by one step) leads to virtually the same value of off-zero current in D, irrespective of the dial-setting N and of the circuit arrangement to the right of D in the figure, except in the unlikely event that $R_3 \ll R'$. The current i_D is, of course, proportional to the voltage e_1, but once this has been fixed the reading of D for one step of the final dial can be measured and adjusted (e.g. by a shunt across D) to some convenient figure, say 100-divisions: the divider has now an effective extension of two decades. This extension cannot be justified of course unless the component resistors in the R_a and R_b branches are accurate to five-figures.

It is worth noting that the same precision is not essential in the R_s-element; if the adjustment of this element is imperfect the only effect is to modify e_1, and this voltage, it is to be presumed, will be under continuous observation, so that any changes induced by slight fluctuations in the input resistance of the divider will be observed and corrected before the measurements are recorded. It will be appreciated, however, that a fairly constant input resistance is desirable or else a long series of sequential adjustments of e_1 and i_D may be called for before the correct conditions are obtained.

It should be pointed out that, in allowing D to show a reading, the conditions for e_1 remaining constant are in any case violated, and that a minor adjustment of the source supplying e_1 will always be called for, and must be carried out, before the reading i_D becomes reliable.

Acknowledgments

A decimal-attenuator of 1kΩ design-resistance was built as part of an investigation carried out in the Research Department of the BBC at Kingswood Warren, and acknowledgments are made to the Director of Engineering for permission to publish this material.

A copy of this design* was made by Messrs Electronic Instruments Ltd, Richmond, and exhibited at the Physical Society Exhibition of Instruments in 1960.

Appendix

The Design of the Shunt-Element R_s

(1) *Shunt for the circuit of Fig. 3 when* $R_L = 0$

The shunt is a precise copy of the element-R_b, and therefore conforms to:—

$$1/R_s = (1 - N)/R' \qquad \ldots\ldots\ldots (15)$$

The equation implies that R_s has to vary in sympathy with changes in the dial-settings, the appropriate number of 10kΩ, 100kΩ 1MΩ resistors being connected in parallel, exactly as in the element R_b of the divider-network. The value of R_s which results will be in error only insofar as the individual resistors are incorrectly calibrated.

(2) *Shunts for the circuit of Fig. 3 for use when* $R_L = R'$

For this type of divider no shunt can be designed which will make $R_s = R_s'$ for all values of N. This section deals with 'errors' in the input resistance R_2 of the network (when correctly loaded) which arise from this fact, for, since all other elements in the divider are assumed ideal, errors in R_2 arise purely from errors in R_s. It is convenient to normalize all resistances to kΩ and conductances to milliohms, and to take R' as kΩ. The the ideal conductance of the shunt (R_s') is given by:

$$1/R_s' = 1 - N + N^2/2 \qquad \ldots\ldots\ldots (16)$$

and the error ϵ is a conductance-error and equals $(1/R_s - 1/R_s')$.

In view of the normalization, the ideal value of R_2 is unity.

The quantity N is the sum of the index-readings of the three dials, so $N = (N_a + N_b + N_c)$, where these are the settings of the 0·1, 0·01 and 0·001 dials.

The resistances shown in Fig. 4 all vary with changes in N_a, N_b and N_c but the similarity of lettering does not imply that B varies when N_b varies, etc.

In the equations, the resistance of element B (say) may appear as $B_{.9}$ to indicate that the first dial controls B, and has the setting 0·9; similarly the resistance of element-C may appear as $C_{.05}$, to show that the second dial controls C and has been set at 0·05; and so on.

In part of the discussion a symbol is needed to show the change of the conductance of the shunt when dial-settings are altered. The following definitions are used:

$$_{.8}\Delta_{.9} \quad (1/R_s) \equiv (1/R_s)_{.9} - (1/R_s')_{.8}$$

$$\text{Generalized:} \quad {}_{n}\Delta_{n+0.1}(1/R_s) \equiv (1/R_s)_{n+0.1} - (1/R_s')_{n} \qquad \ldots\ldots\ldots\ldots 17)$$

(2a) *The shunt of Fig.* 4(a)

The value of the error in $1/R_s$ is given by equation (12).

(2b) *The shunt of Fig.* 4(b)

In this circuit, elements A and B are both controlled by the N_a-dial, while element C varies linearly in proportion to $(N_b + N_c)$ and is partly controlled by the N_b and partly by the N_c-dials. It is found convenient to give C the maximum value of 0·1295. Thus whenever $(N_b + N_c)$ total 0·1, C has this resistance irrespective of the setting (N_a) of the 0·1-dial).

From equation (16) Table 1 may be constructed.

TABLE 1

N	0·0	0·1	0·2	0·3	0·4	0·5	0·6	0·7	0·8	0·9	1·0
$1/R's$	1·0	0·905	0·82	0·745	0·68	0·625	0·58	0·545	0·52	0·505	0.5

* *Patents Pending*

To illustrate the design-procedure take the case in which N happens to have the value 0·9, and it is presumed that this is provided by the N_a-dial alone.

Then $N_a = 0{\cdot}9$; $(N_b + N_c) = 0$.

The resistance at A (written $A_{.9}$) now constitutes the shunt resistance R_s, and must, from Table 1, have the value

$A_{.9} = 1/0{\cdot}505 = 1{\cdot}98019 \ldots$ to make $R_s = R_s'$ as required.

Suppose, then that the N_b and N_c dials are both set to maximum, so that $(N_b + N_c) = 0{\cdot}1$; then $N = 1{\cdot}0$, and from Table 1, the resistance $R_{s(1.0)}$ of the shunt should be 2·0. In fact the resistance of the shunt is the sum of $A_{.9}$ and B with C in parallel. Now C has its maximum value (chosen above) of 0·1295 and this is far too large; but it is clear that element-B can be used to shunt-down the value of C_{max} to such a value that B and C_{max} in parallel will have the correct resistance of $(2 - 1{\cdot}98019 \ldots)$.

The resistance at B must therefore be varied when element A is varied, and both of them by the N_a-dial; thus for every value of A there is a corresponding value of B. To find the value of $B_{.9}$ which has to be put into circuit simultaneously with $A_{.9}$ one can proceed formally as follows (the symbol // implies 'in parallel with'):—

$${}_{.9}\Delta_{1.0}(R_s) = 0{\cdot}019801 \ldots = B_{.9}//C_{max}$$

(C_{max} has already been set as 0·1295)

from which:—

$$1/B_{.9} = 1/({}_{.9}\Delta_{1.0}(R_s)) - 7{\cdot}73 \ldots$$

or, using generalized symbols:—

$$1/B_n = 1/({}_n\Delta_{n+0.1}(R_s)) - 7{\cdot}73 \ldots \qquad (18)$$

Equation (18) defines all the values of B; the A values are taken from the lower line of Table 1, inverted.

The design procedure ensures that this shunt has no errors when N has any of the values 0, 0·1, 0·2....0·9, 1·0; but when N is an odd multiple of 0·05 (and half of C_{max} is put into circuit) the linear variation of C fails to fit the true law of variation of R_s', and errors appear. They will not be computed here—(there is a local maximum of 0·5 per cent when $N = 0{\cdot}55$)—because the shunt next to be discussed, which has the same number of elements, has rather smaller errors.

(2c) *The shunt of Fig.* 4(c)

In this case again both A and B elements are controlled by the N_a-dial; C varies (but not quite linearly) with $(N_b + N_c)$.

Again an infinity of alternative designs are possible (provided only that $C_{max} > 0{\cdot}105$). The design discussed here takes C_{max} as 0·200 ($C_{max} = 200\Omega$).

Consider any specific value of N, say $N = 0{\cdot}4$: then with $N_a = 0{\cdot}4$ and $(N_b + N_c) = 0$:—

$$1/A_{.4} + 1/B_{.4} = (1/R_s')_{.4} = \text{(from Table 1), } 0{\cdot}68.$$

Then when N is raised to 0·5 by setting $(N_b + N_c)$ to maximum (= 0·1), and consequently $C = C_{max} = 0{\cdot}200$:—

$$1/A_{.4} + 1/B_{.4} + 0{\cdot}200) = (1/R_s')_{.5} = \text{(from Table 1) } 0{\cdot}625.$$

On subtracting these two equations, $B_{.4}$ is isolated:—

$${}_{.4}\Delta_{.5}(1/R_s') = ({\cdot}200)/(B_{.4} + 0{\cdot}200)) = 0{\cdot}055.$$

With $B_{.4}$ derived, $A_{.4}$ follows, for:—

$$1/A_{.4} + 1/B_{.4} = (1/R_s')_{.4}$$

Expressed in generalized terms:—

$$\left.\begin{aligned} &1/({}_n\Delta_{n+0.1}) \,.\, 1/R_s') = B_n + 5 \,.\, (B_n)^2 \\ &\text{and } 1/A_n = (1/R_s')_n - 1/B_n \end{aligned}\right\} \qquad (19)$$

(It is to be observed that the '5' appears only as the inverse of the arbitrarily chosen quantity $C_{max} = 0{\cdot}2$. Equation (19) applies only to this choice of C.)

For the design of this and other, later, shunts the quantity ${}_n\Delta_{n+0.1}1/R_s'$ is needed and has been tabulated in Table 2.

TABLE 2

Increment in N	0/·1	·1/·2	·2/·3	·3/·4	·4/·5	·5/·6	·6/·7	·7/·8	·8/·9	·9/1·0
Increment in shunt conductance $1/R's$	·095	·085	·075	·065	·055	·045	·035	·025	·015	·005

The resistors required at A and B in the shunt network are defined by equation (19), and, provided $(N_b + N_c)$ is set to zero or 0·1 the shunt will have no errors. But, to take care of settings of $(N_b + N_c) < 0{\cdot}1$, the value of C_n will have to be considered; a non-linear variation of C_n is found to be necessary to keep errors as small as possible. Rather less than half of C_{max} is called for when $(N_b + N_c) = 0{\cdot}05$.

In deciding the resistances required, another quite arbitrary decision has to be made. Here the shunt will be designed on the basis that when $N_a = 0$ the element $-C$ is arranged so that there are no errors when:—

$N_a = 0$; $N_b = 0{\cdot}01$, $0{\cdot}02 \ldots 0{\cdot}09$

and, provided $N_c = 0$, or maximum (= 0·1).

(It will be remembered that Equation (19) already ensures that there are no errors when $(N_b + N_c) = 0$, and N_a has any of its possible values.)

Table 3 shows the appropriate resistances for the C-element. They have been derived by first calculating from equation (16) the quantity $1/R_s'$ appropriate to $N = 0{\cdot}01 \ldots 0{\cdot}09$; from this obtaining the increments in $1/R_s'$ and then deriving the associated resistance $C_{.01}$, $C_{.02} \ldots C_{.09}$ by the process of writing into equation (19) the now known quantities A_0 and B_0; replacing the Δ term by the incremental conductance just derived and solving for the term, which in equation (19) appears as 5—in the new equation this term is $1/C_n$; whence the required C_n.

TABLE 3

N	N_a	N_b	N_c	Cn (resistance normalized)	Normalized increments in C_n
0·01	0	0·01	0	0·01850146...	0·01882023...
0·02	0	0·02	0	0·03732169...	0·01914430...
0·03	0	0·03	0	0·05646599...	0·01947373...
0·04	0	0·04	0	0·07573972...	0·01980851...
0·05	0	0·05	0	0·09574823...	0·02014866...
0·06	0	0·06	0	0·11589689...	0·0204942...
0·07	0	0·07	0	0·1363911 ...	0·0208450...
0·08	0	0·08	0	0·1572361 ...	0·0212012...
0·09	0	0·09	0	0·1784373 ...	0·021562...7
0·1	0	0·09	0·01	0·2000000 ...	

(The series has been completed by showing the effect of setting Nc = Nc (max)).

The last column shows that the resistance of element-C must not vary linearly with N_b, no increment being precisely equal to 0·02.

For the same reason, it is quite clear that there will always be an error in R_s when the N_c-dial is operated, except when it is set to maximum. The best that can be done is to make the part of the element-C controlled by this dial vary in equal steps of 0·002 (i.e. 2Ω); if then $N_c \neq 0{\cdot}01$ and $N_b = 0{\cdot}04$ to 0·05, there will be little effect, but for lower or higher values of N_b there will be an error ϵ which approaches 0·1 per cent.

(It must be remembered that 1Ω error in element-C does not imply 0·1 per cent error in $1/R_s$, for the resistance of the branch B-C already contains, at B_0, some 1 350Ω. Moreover B_n gets larger as N_a is increased towards unity, and errors in C_n are, by that, more deeply swamped.)

The error just mentioned is additive to another which arises from the arbitrary decision, made earlier, that C_n is to be calculated when $N_a = 0$. The data of Table 3 leads to values of R_s which are precisely correct when this is the case, but will necessarily lead to errors in R_s when $N_a \neq 0$. Such errors reach a maximum whenever N is an odd multiple of 0·05, falling to zero when N is a multiple of 0·1. The local maxima at $N = 0{\cdot}45$ and 0·65 are 0·064 per cent and 0·14 per cent respectively; to this must be added the error, if any, due to the use of the N_c-dial; this may be either positive or negative, depending upon whether $N_b >$ or $< 0{\cdot}05$.

(2d) *The shunt of Fig.* 4(d)

It was mentioned in the main text that for the majority of purposes shunt-errors of the order of 0·2 per cent can be tolerated, and this can be achieved with the three-element shunt just mentioned. However, the errors can be reduced to about 20 parts in 10^6 by the addition of one extra variable, and this may be justified as a small price to pay for a ten-to-one increase in precision.

Unfortunately, there will remain larger errors whenever the N-dial is operated, but it has already been pointed out that the use of the lower-order dials can be avoided in certain circumstances.

The shunt of Fig. 4(d) has the additional element S which is controlled by the N_a-dial. (Elements A and B are controlled as before also by the N_a-dial). Its purpose is so to modify the resistance contributed to the B-S-C branch as to bring it to precisely the correct value whenever N is an odd multiple of 0·05.

It is convenient to make use of the already computed data of Table 3 to establish the resistance C_n; then S will be unnecessary when $N_a = 0$, and thus $S_0 = \infty$. The design-procedure which establishes the values of S_n will be merely outlined here since it is a simple, though tedious, extension of the approach used in (2c), above.

First it must be noted that the values of A and B will differ from those used in (2c)—except for the case of A_0 and B_0. One has to deal with values of N which are odd multiples of 0·05 and consequently one needs:—

$$C_{05} = \text{(from Table 3) } 0{\cdot}09574823\ldots$$

and $1/C_{05} = 10{\cdot}444057\ldots$

The procedure is illustrated by taking as an example the case of $N = 0{\cdot}45$. The first step is to derive the values of $A_{.4}$ and $B_{.4}$:—

From Table 2:—${}_{.4}\Delta_{.5}(1/R_s') = 0{\cdot}68 - 0{\cdot}625 = 0{\cdot}055$.

The increment is derived from $1/A_{.4} + 1/B_{.4} = 0{\cdot}68$

and $1/A_{.4} + 1/(B_{.4} + S_{.4}//C_{max}) = 0{\cdot}625$; for $S_{.4}$ shunts C_{max}.

By subtraction:—$1/B_{.4} - 1/(B_{.4} + S_{.4}//C_{max}) = 0{\cdot}055$, which reduces to:—

$$\frac{(B_{.4})^2}{S_{.4}} + \frac{(B_{.4})^2}{C_{max}} + B_{.4} = 1/0{\cdot}055 = 18{\cdot}1818\ldots$$

(taking $C_{max} = 0{\cdot}2$).

This equation is insufficient to define either $S_{.4}$ or $B_{.4}$. But a second and similar equation can be set up using ${}_{.4}\Delta_{.45}(1/R'_s)$ in place of ${}_{.4}\Delta_{.5}$, and by replacing C_{max} by $C_{.05}$.

Again, by subtraction of equations, the terms in $S_{.4}$ can be eliminated and the value of $B_{.4}$ derived from the resulting quadratic equation; whence, finally $A_{.4}$, $B_{.4}$ and $S_{.4}$, as required.

Omitting the effect of using N_c at other than zero and maximum, the errors of this shunt are zero when N is any multiple of 0·05, and rise to local maxima when N lies between these values; the error varies with N_a, but does not exceed 20 parts in 10^6.

(2e) *The shunt of Fig.* 4(e)

This shunt structure has been included in order to round off the series, but it is unlikely ever to be needed. The design principle consists in controlling the element C with the N_c-dial only, C varying linearly with N_c; A, B and S are controlled by the N_a-dial; D and E are controlled by the N_b-dial.

The design proceeds by assigning to C a maximum resistance in excess of any of the increments shown in Table 3; 22Ω might suffice. D is given the values set out under C_n in Table 3, while E is adjusted, in sympathy with changes in N_b, so as to bring $C//E$ to such a resistance as would precisely cover the next increment of D.

Thus when $N_b = 0{\cdot}06$; $C//E_{06} = 0{\cdot}022//E_{05} =$ (from Table 3) 0·020492..

Whence E_{06} is calculable.

This shunt is estimated to show no errors in excess of 10 parts in 10^6 for any setting of the three dials whatever, except those which would arise from incorrect adjustment of component resistors. It is to be remembered that the 'load' R_L on the shunt must be set to to 1kΩ with an equal precision if the high accuracy of this shunt-structure is to be exploited to the full.

Improvements in Photocathodes *

It has been found that a photo-emissive surface can be formed by evaporating a thin film (250 to 400Å) of bismuth-antimonide upon a glass or other suitable substrate, and sensitizing this layer with the alkali metals sodium, potassium and caesium, under high vacuum.

Layers so formed can have high sensitivities, i.e. 160μA/L or more to white light (2 854°K colour temperature lamp).

The processing details are as follows:—

Sodium and potassium metals are distilled into side tubes by a well-known method. The three metals are derived by separate thermite reactions on the chromates.

The bismuth antimonide layer is deposited by evaporation from a single bead of the alloy. The amount deposited is measured by the increase in optical opacity of the layer during evaporation. Layers having optical transmissions between 60 and 35 per cent are used. The alloy is prepared by heating pure antimony and bismuth metal in a glass bulb under high vacuum. The constituents are mixed in a 6:1 ratio of weight, the major constituent being antimony.

The layer is first sensitized with potassium at a temperature of 160° to 175°C, the sensitivity achieved is about 1μA/L and the activitation time is approximately 4 to 8min. Next, sodium is allowed to react at 230°C temperature until a steady sensitivity is obtained (usually 10 to 14μA/L); the activation takes about 10 to 15min. Finally, the partly activated layer is sensitized with caesium at a temperature of 150°C; the process is continued until a drop of 25 per cent of the peak sensitivity is reached. After this, a post-caesiation bake is applied at 160°C until maximum sensitivity (about 80μA/L) is obtained. On cooling the layer, a further increase in sensitivity occurs and the final magnitude is of the order of 160μA/L. The total time is less than 55min.

* *A communication from E.M.I. Ltd.*

The Role of Space Charge Waves in Modern Microwave Devices

(Part 3)

The transmission line theory of tubes with continuous interaction between the electron beam and the circuit.

By G. D. Sims*, M.Sc., Ph.D., A.M.I.E.E., and I. M. Stephenson*, M.Sc., Ph.D.

(Voir page 460 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 465)

Transmission Line Theory

IN the earlier parts of this article, the elementary properties of space charge waves have been discussed together with the ways in which they can be launched. It has been shown that a space charge wave mode can be represented by a transmission line and thus the performance of say, a microwave valve can be discussed by considering the interaction of the appropriate valve circuits with the equivalent transmission line or lines representing the space charge waves launched within the valve. From impedance considerations alone, it is clear that as soon as a wave propagating circuit, with an impedance and phase velocity comparable with those of the space-charge-wave equivalent transmission line, is brought into proximity with the beam then mutual interaction will take place, which modifies the properties of the space charge waves as well as the properties of the characteristic modes of the circuit.

This Part will be largely concerned with the consequences of the interaction of space charge waves with circuits in modern electronic devices and an examination of the way in which the presence of space charge invalidates ballistic theories.

An instructive example to consider is that of the backward wave oscillator, for the properties of a number of other devices can be deduced from a similar argument. Gould[7] has considered this problem in terms of space charge wave interaction and his discussion will be essentially reproduced here. A backward wave oscillator consists basically of an electron beam travelling in proximity to a microwave circuit propagating a wave whose fundamental space harmonic has a velocity roughly equal to that of the beam, and which is by nature a 'backward' wave. (A backward wave in this sense being one in which the phase velocity is opposite to the direction of energy flow.) The fact that a backward wave circuit is being considered suggests that the interaction may be primarily with the slow space charge mode whose negative kinetic power flow has already been discussed. Gould considers the problem as one of coupled transmission lines—one representing the slow space charge wave and the other the circuit wave, the argument proceeding as follows:

Let the slow wave amplitude be P and the phase constant in the absence of the circuit be β_a. Let the amplitude of the circuit wave be S and the phase constant in the absence of the space charge wave be β_b, then if K is a constant representing the coupling between the waves, the equations for the coupled modes are:

$$\partial P/\partial z + j\beta_a P - KS = 0 \qquad (43)$$

$$\partial S/\partial z + j\beta_b S + KP = 0 \qquad (44)$$

and their solution is of the form:

$$P = C_1 \exp(-j\beta_1 z) + C_2 \exp(-j\beta_2 z) \qquad (45)$$

$$S = -jC_1\left(\frac{\beta_1-\beta_a}{K}\right)\exp -j\beta_1 z - jC_2\left(\frac{\beta_2-\beta_b}{K}\right)\exp(-j\beta_2 z) \qquad (46)$$

where $\beta_{1,2} = (\beta_a + \beta_b)/2 \pm \sqrt{[((\beta_a - \beta_b)/2)^2 + K^2]}$ (47)

Suppose the electron beam enters the interaction region at the plane $z = 0$. At this point it will be unmodulated and thus the amplitude of the space charge wave is zero, i.e. $P = 0$ at $z = 0$. If the tube is perfectly matched at the end of the structure, say at $z = l$ then, because all the power flows in the negative direction, the amplitude of the circuit wave will be zero there, i.e. $S = 0$ when $z = l$. Application of these conditions leads to a relation between the constants C_1 and C_2 and also gives a dispersion relation connecting β_a, β_b, K and l. Solution of the latter equation shows that for oscillation to take place $\beta_a = \beta_b$ i.e. the space charge wave velocity must equal the velocity of the circuit wave and also that $Kl = (2n + 1)\,\pi/2$. Using these results the amplitudes of the modes in the valve are found to be:

$$P = -2jC_1 \exp(-j\beta_a z)\sin KZ \qquad (48)$$

$$S = -2jC_1 \exp(-j\beta_a z)\cos KZ \qquad (49)$$

The propagation constants of the modified waves are $\beta_{1,2} = \beta_a \pm K$ and hence $K = (\beta_1 - \beta_2)/2$. The condition $Kl = (2n + 1)\,\pi/2$ thus gives that for $n = 0$ oscillation will only be possible when $l = \pi/2K$ i.e. when l is one quarter beat wavelength long (beat wavelength between the modified space charge and circuit waves). Fig. 9 indicates that after $\frac{1}{4}$ beat wavelength the current modulation impressed on the circuit has been completely changed to velocity/kinetic voltage modulation on the electron beam.

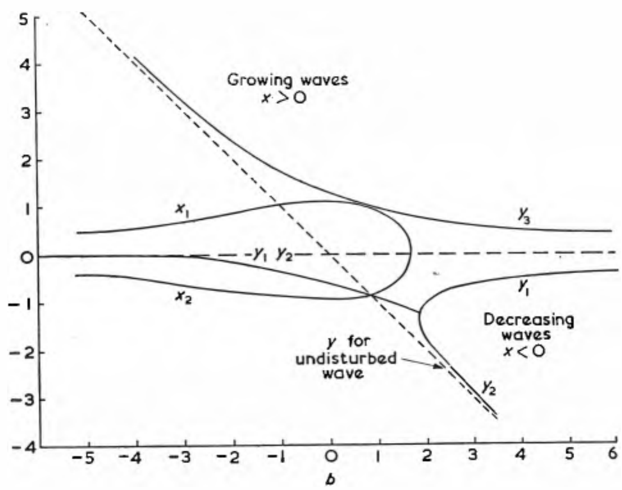

Fig. 7. Roots of travelling wave tube equation as a function of b
No space charge. QC = O

* *University College, London*

Comparison of the coupled mode dispersion relation with the general travelling wave tube dispersion relation with all four modes present shows that when only two modes are important:

$$K^2 = (\beta_e C)^3/2\beta_p = (Z_{cct}/Z_\omega)(\beta_e^2/4) \quad \ldots\ldots (50)$$

Whence for the start of oscillation:

$$(Z_{cct}/Z_\omega)(l/\lambda)^2 = \tfrac{1}{4} \text{ i.e. } N = 2\sqrt{(Z_{cct}/Z_\omega)} \quad \ldots. (51)$$

This corresponds to the normal expression for start of oscillation which states that:

$$CN = (\beta_q l/16\pi)^{1/3} \quad \ldots\ldots\ldots\ldots (52)$$

The Effect of Space Charge on Simple Travelling Wave Tube Theory

In the orthodox travelling wave tube (forward wave amplifier) the first, simple theories took no account of space

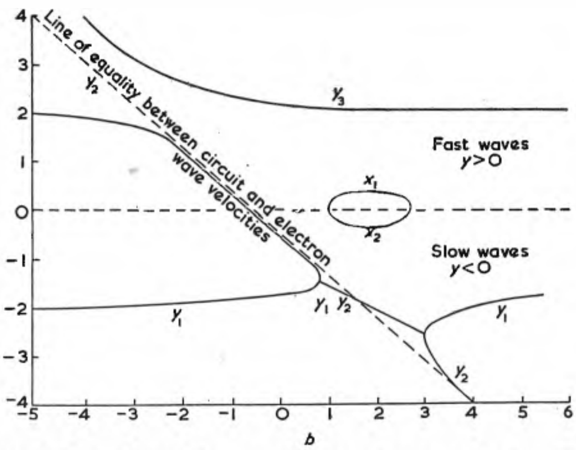

Fig. 8. Roots of travelling wave tube equation as a function of b
$QC = 1$

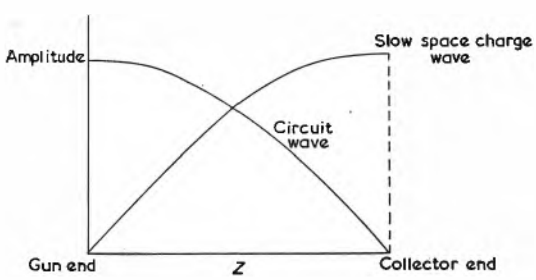

Fig. 9. Wave amplitudes in backward wave oscillator

charge. It was shown that if the propagation constant of the circuit wave was written as:

$$\Gamma_1 = \beta_e C d + j\beta_e(1 + Cb) \quad \ldots\ldots\ldots (53)$$

(where b effectively accounts for the amount by which the velocity of the circuit wave differs from that of the beam) and all waves are assumed to propagate as $e^{-\Gamma z}$, and if the propagation constant of the waves in the electron stream is written:

$$\Gamma = -\beta_e C x + j\beta_e(1 - Cy) \quad \ldots\ldots\ldots\ldots (54)$$

then x and y were related to b, d and β_e by the relation:

$$\delta^2 = 1/(-b + jd + j\delta) \quad \ldots\ldots\ldots (55)$$

where $\delta = x + jy$.

A plot of the roots of equation (55) is shown in Fig. 7. The amplitude of the wave on the beam is proportional to:

$$\exp(\beta_e C x z) \exp - j\beta_e(1 - Cy)z \quad \ldots\ldots (56)$$

and hence in those regions where x has a maximum positive value, maximum amplification will take place. It is noted that this value corresponds to the point $b=0$ i.e. a condition such that the velocity of the beam is equal to the velocity of the wave. It is seen however, that the disturbance induced in the beam by the circuit travels more slowly than the undisturbed wave under all conditions. The other roots shown in Fig. 7 are not of particular relevance to the present argument and are only included for the sake of completeness.

When the beam current is such that space charge ceases to be negligible the picture alters substantially. As soon as the propagation of true space charge waves becomes possible these modify the disturbance induced on the beam by the slow wave circuit and must be taken into account in the calculations. Pierce[8] shows that the effect of coupling of the space charge waves with all non-synchronous modes on the circuit can be considered in terms of an additional positive or negative distributed capacitance between the beam and the circuit. The dispersion relation is modified to the form:

$$\delta^2 = [1/(-b + jd + j\delta)] - 4QC \quad \ldots\ldots (57)$$

where Q is related to this distributed capacitance C_0 through the relation:

$$Q = (\beta_e/\omega C_0)\ (E^2/\beta^2 P) \quad \ldots\ldots\ldots\ldots (58)$$

and is referred to as the *space charge parameter*. For a fairly high value of QC,—in practice about 1·0 say, the roots of the dispersion relation take the form shown in Fig. 8. It is seen that now the maximum amplification no longer occurs at a velocity corresponding to that of the beam, but to a positive value of b denoting a circuit wave travelling more slowly than the beam. This indicates that the travelling wave tube is predominantly dependent on the slow space charge for its operation. Amplification can now only be obtained over a relatively narrow range of circuit wave velocities. In distinction from the case of no space charge it is apparent that for large values of b, both positive and negative, the three roots of the equation approach a pair of asymptotes. These are given by the expression:

$$\delta = \pm 2j\sqrt{(QC)} \quad \ldots\ldots\ldots\ldots (59)$$

When the circuit wave velocity is comparable with the electron velocity, interaction occurs modifying both circuit and electron waves. However, when the velocity of the circuit wave is very different from that of the electron beam ($|b|$ large) no interaction occurs and the disturbances on the beam become the pure space charge waves, previously discussed. This argument amounts to the statement that the electrons do not 'see' the circuit wave unless the two have similar velocities.

Since for large $|b|$, $\delta = \pm 2j\sqrt{(QC)}$ one may equate the resulting phase constant $\beta_e(1 \pm C\delta)$ with the value obtained from the analysis of Part 1, viz: $j\beta_e(1 \pm \omega/\omega_p)$

$$\text{Thus } 2C\sqrt{(QC)} = \omega/\omega_p \quad \ldots\ldots\ldots\ldots (60)$$

$$\text{or } Q = (\omega_p/\omega)^2/4C_3 = \tfrac{1}{4}\ Z_\omega^2/Z_B Z_{cct} \quad \ldots\ldots (61)$$

The role of the fast space charge wave in travelling wave tube operation has been discussed by Gould[7] in a manner analogous to his discussion of the backward wave oscillator. In this case both the electron wave and the circuit wave are forward waves and flow of power takes place in the positive z direction. If in equations (43) and (44) P now stands for the amplitude of the fast wave then the same general solution as before applies but the boundary conditions are now stated as: at $z = 0$ there is no modulation on the beam, i.e. $P = 0$, while the circuit wave will have some amplitude S_0. Applying these conditions one finds similarly to the argument before that:

$$P = S_o \exp - j\beta_e z \sin KZ \quad \ldots\ldots\ldots\ldots \quad (62)$$

$$S = S_o \exp - j\beta_e z \cos KZ \quad \ldots\ldots\ldots\ldots \quad (63)$$

This indicates that at a distance of $2\pi/K$ (or one quarter beat wavelength between the modified circuit and fast space charge waves), from the input, all the signal initially on the circuit has been transferred to the beam. Similarly, had there been any signal on the beam this would have been transferred to the circuit. It is clear then that in a tube of this type, one quarter beat wavelength long, the interaction of the beam and circuit can be compared with a 1:1 lossless directional coupler. Such a tube would have no output from its circuit at all and is in effect operating in the Kompfner dip condition. The variation of both the beam and circuit wave under the Kompfner dip condition is shown in Fig. 9.

The Use of Fast Waves in Low Noise Microwave Valves

A basic limitation of the travelling wave devices discussed so far is that the noise figure is of the order of several decibels. Shot noise at the cathode sets up pairs of space charge waves on the beam at each frequency. These pairs of waves interfere to give maxima and minima noise along the length of the beam[3]. For maximum signal-to-noise ratio any external signals with which the beam is to interact must be coupled at points corresponding to noise minima. In conventional tubes these noise minima are never zero and consequently the noise figure will be non-zero too. The reason for this is implicit in the discussion of kinetic power flow given in Part 1. For the slow wave the power flow is negative and thus in order to launch the wave energy must be extracted from the beam. Once a noise stimulated slow wave is launched it cannot be removed from the beam, as in order to do so an amount of energy equal to the noise energy must be added to the beam, see Fig. 1, possessing a spectrum (both amplitude and phase) identical with that of the noise measured on the beam at that instant. Since the noise is random in nature this cannot be achieved. The reverse applies in the case of the fast wave. The power flow is positive and a positive quantity of energy is added to the beam in launching it. If the beam is passed through suitable filters this noise can be removed—consequently it is possible, in a narrow band at least, to obtain a noise free fast wave. Any tube in which both slow and fast waves are launched will be noisy because of the noise which cannot be extracted from the slow wave. However, if a tube can be designed which utilizes the fast wave only it should be possible to make it noise free by extraction of the noise, as described.

The problem of fast wave launching has received a fair amount of attention recently and a summary of one method is given below. The problem is to design a coupler which ensures that an outside signal is impressed on the beam in the form of a pure fast wave while at the same time also removing all the noise from the beam.

Use is first made of a short length of helix circuit through which the beam is passed. The external signal is fed on to the helix which is designed to be one quarter beat wavelength long (beat wavelength between helix mode and fast wave). The previous argument shows that at the end of the helical section all signal on the helix will be converted into velocity modulation of the beam while all fast wave noise on the beam will have been transferred to the helix, thus a noise free fast wave modulated as desired has been produced. Such a device will also of course launch a slow wave too but as has been argued previously the amplitude ratio of slow to fast wave can be made fairly small by making the space charge density high. Clearly then a high value of QC or $Z_{\omega}J$ is required. This will ensure that the minimum slow wave is launched and the greater part of the residual slow wave can then be removed by a suitably placed velocity jump. Such a jump is bound to result in some of the noise on the residual slow wave being converted into fast wave noise but although this means that the final fast wave will not be completely noise free, the noise level will be apprecialy lower than in any conventional device yet described.

Having once launched a fast wave signal, energy can be fed into it by 'pumping' at twice the signal frequency. This increases the signal amplitude and in theory adds no additional noise to the wave at the signal frequency. The amplified signal is then extracted from the beam in a further coupler which operates on identical principles to the input coupler.

Another promising way of producing a fast wave amplifier has been described by Adler[9]. The Adler tube is a form of parametric amplifier that uses a fast wave on an electron beam as the operating medium. The fast wave is launched by a Cuccia coupler, shown in Fig. 10. It is seen that any modulating voltage applied to the coupler causes a transverse force to be exerted on the electrons and if the frequency of the voltage applied to the coupler is equal to the cyclotron frequency then the electrons will follow a spiral path in the coupler. If the coupler is long compared with $2\pi/\beta_e$ then all the electrons leaving it will travel on

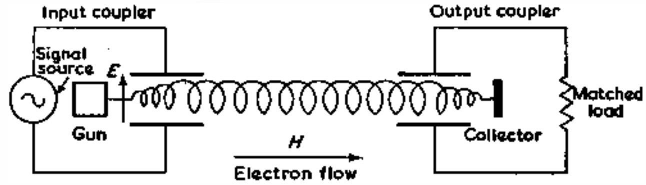

Fig. 10. Cuccia coupler

helical paths with approximately the same radius. The input signal to the coupler will have to supply the power necessary to impart the radial motion to the electrons. The longitudinal velocity of the electrons is unchanged by the coupler.

This type of coupler launches a pair of cyclotron waves whose phase velocities are given by $v_p = u_o/(1 \pm \omega/\omega_c)$ but as $\omega = \omega_c$ the fast wave will have infinite phase velocity and the slow wave a phase velocity of $u_o/2$. The difference between these two phase velocities is so huge that any practical device will only be affected by one wave or the other and in the Adler tube it is the fast wave that is used. The Cuccia coupler can also be used to extract any existing fast wave noise on the beam as it has already been shown that fast wave noise can be completely removed from a beam.

If a second coupler were placed at some point along the beam the power coupled to the beam by the input coupler could be extracted, but the device would not amplify the original signal. Adler places a quadrupole device between the input and output couplers which pumps the beam at some frequency, usually twice the input frequency. By examining Fig. 11 it is seen that the effect on the electron of the field in the quadrupole depends upon the position of entry of the electron and this in turn depends upon the phase of the input signal when the electron enters the input coupler. Thus some electrons have their radius of gyration increased while that of others is decreased. The increase in radius is greater than the decrease so that some of the pumping power is transferred to the electron beam. The electrons leaving the quadrupole region are still all travelling on helical paths but the path radius of individual electrons is not the same. The variation of radius with longitudinal distance or time is not sinusoidal as can be seen from Fig. 12, and contains terms at both the original frequency and the difference between the pump and input frequency. The beam leaving the quadrupole then passes through a second Cuccia coupler which extracts the signal, now amplified, from the beam. Unlike most electron beam amplifiers there is no conversion of d.c. beam power to r.f. power, the electrons leaving the ouput coupler all have exactly the same velocity as when they left the electron gun. All the increased power comes from the 'pump.'

Wave Propagation in Plasmas

A final example of space charge wave phenomena of current interest occurs in the gas discharge plasma. Propagation of electromagnetic waves in plasmas is of considerable importance in view of its use in determining ion and electron densities. The method is employed as a diagnostic tool in both thermonuclear devices using plasma as well as for a number of other more conventional purposes.

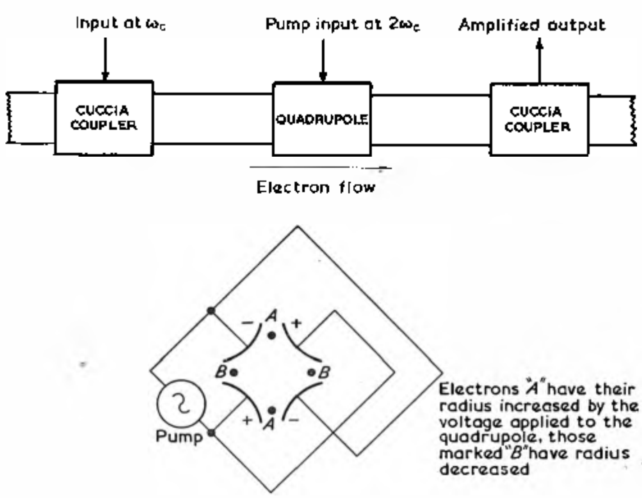

Fig. 11. Quadrupole and complete tube

Hitherto only waves in pure electron gases have been considered. These gases are frequently considered to be neutralized in theoretical treatments in order to overcome space charge divergence troubles and this requires the presence of positive ions which are assumed to play no part in the interaction with the waves being discussed. A plasma is an ionized gas which is electrically neutral—containing equal numbers of positive and negative charges. The positive charges being heavy compared with the electrons can move only relatively slowly and at high frequencies the assumption that their absence or presence is immaterial, is often largely justified. At lower frequencies however, this is not the case. The plasma is then a homogeneous mixture of an electron gas and a positive ion gas each of which under certain assumptions may be treated as an independent medium. The cases of most interest however, are those in which the two gases function collectively, that is, in which both positive ion and electron phenomena are coupled.

Some coupling is always present as a result of collisions. Normally electron-electron collisions are unimportant as in a normal gas discharge the collision frequency is so small as to be negligible. However, once gas molecules permeate the electron gas, electron-molecule collisions give rise to energy loss from the electrons and propagating waves suffer attenuation. This attenuation reaches a peak where the collision frequency is comparable with the wave frequency and, in the presence of a magnetic field, where the wave frequency is comparable with the cyclotron frequency (c.f. gyromagnetic effects on wave propagation in the ionosphere). In general the collisions between charged particles are of less importance than collisions between neutral molecules and charged particles. Such collisions can give rise to energy losses and clearly the higher the particle density the more important will collisional effects be. However for normal discharge densities the critical frequencies are low and the effects of collisions will not be taken into account here.

In a waveguide completely filled with plasma[10] in a longitudinal magnetic field a variety of types of space charge wave may propagate. The normal waveguide boundary conditions must be satisfied and this leads to the conclusion that a given waveguide mode will be characterized by some transverse wave-number β_{mn} leading to the same value of cut-off wavelength as for an empty guide. For each mode, i.e. for each discrete value of β_{mn} six principal waves can propagate in the plasma filled guide, they are:

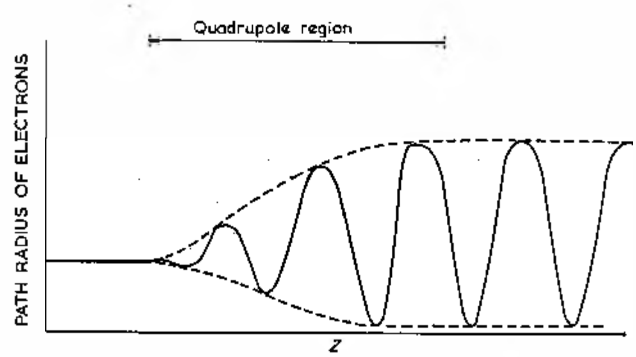

Fig. 12. Electron trajectory in Adler tube

(1) A pair of modes resembling the simple electron space charge waves previously discussed for which $\beta \simeq \beta_e = (\beta_{pe}/\beta_{mn})$. (Note that in the absence of a waveguide boundary or drift tube around the plasma β_{mn} would assume the value 1.)

(2) A pair of waves corresponding to the electron cyclotron waves—similar to those discussed for the Adler tube and for which $\beta \simeq \beta_e \pm \beta_{ce}$. This pair corresponds to both a slow and a fast wave and it should be noted that for values of magnetic field over 100 gauss β_e will probably be less than β_{ce} and hence for the fast cyclotron wave the phase velocity will be negative. This differs from the pure space charge wave case where although the kinetic power flow for the slow wave is negative the actual absolute phase velocity is positive.

(3) A pair of waves corresponding to propagation in the ion gas alone. For an ion gas of finite transverse dimensions a pair of waves with positive and negative values of β can propagate, the dispersion curves being shown in Fig. 13. The region between ω_{pi} and ω_{ci} is a stop-band, but a low frequency pass-band from $\omega = 0$ to ω_{pi} or ω_{ci} whichever is the smaller exists, as well as a higher frequency pass-band between ω_{pi} or ω_{ci} (whichever is greater) and $\sqrt{(\omega_{pc}^2+\omega_{ci}^2)}$. This latter band is such that $\partial\omega/\partial\beta$ is negative and hence the propagation will be of the backward wave type. The dispersion relation is given by:

$$\beta_i^2 = -\left\{\frac{1-(\omega_{pi}^2+\omega_{ci}^2)/\omega^2}{[1-(\omega_{pi}/\omega)^2][1-(\omega_{ci}/\omega)^2]}\right\}\beta_{mn}^2$$

As indicated previously when coupling occurs between two modes, e.g. a circuit mode and a space charge wave mode the characteristics of both modes change. When the phase velocity of a circuit mode approaches that of a space charge wave, strong coupling occurs between them and growing waves or instability may occur. It has already been stated how this explains the behaviour of the travelling wave tube or backward wave oscillator. In a similar way coupling between the fast and slow space charge waves arising in a multi-beam, multi-velocity system can be discussed in terms of the transmission line model and an explanation of the growing waves observed in the Haeff double beam tube[11] can be obtained. This type of argument is capable of extension to the present case of the plasma guide and the following couplings may occur giving rise to basic modifications of the wave types:

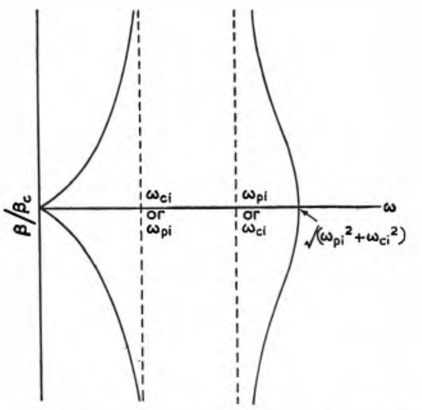

Fig. 13. Dispersion curve for ion filled waveguide

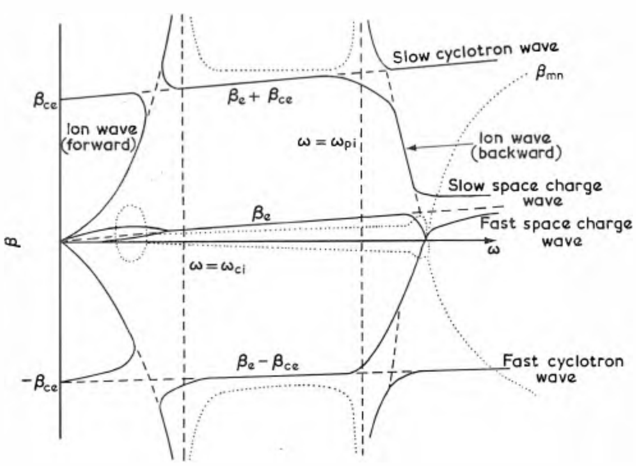

Fig. 14. Dispersion curve for plasma filled waveguide

(1) Near $\omega = \omega_{ci}$ the ion space charge waves couple to the electron cyclotron waves to produce a pair of complex propagation constants. These represent both forward and backward growing and decaying waves. If β_1 is taken to correspond to the usual slow circuit wave in the travelling wave tube and the slow cyclotron wave to the slow space charge wave, the above interaction can be seen to give rise to a similar forward growing wave to that of the t.w.t.

(2) Near $\omega = \omega_{pi}$—in a similar manner to (1) there is a phenomenon similar to that in the b.w.o. The ion wave is here a backward wave corresponding to the circuit wave in a b.w.o. and this interacts again with the slow cyclotron wave to give a new pair of real roots which represent a potential time instability.

(3) Between ω_{ci} or ω_{pi} and $\sqrt{(\omega_{pi}^2 + \omega_{ci}^2)}$ coupling also occurs between the electron space charge waves and the ion waves giving again a conjugate pair of growing and attenuating waves.

Fig. 14 shows the complete dispersion curve, indicating the way in which it evolves from the component waves taken separately.

Conclusion

The properties of simple space charge waves have been discussed and it has been shown how such waves arise under various conditions of excitation. When circumstances are such that the propagation constants of a pair of waves are comparable, interaction occurs between them and may produce instability or amplification within the device. In most modern devices for microwave generation, and in gaseous devices also, this type of interaction is fundamental. Any space charge wave, cyclotron wave or ion wave can be made to interact with a circuit wave producing such a phenomenon and similarly any pair of degenerate ion, cyclotron or space charge waves may also interact and bring about mutual modification. Whether such interaction results in a useful usable device often depends upon other considerations however. Although for instance many interesting cases exist in the case of the plasma waveguide, devices depending upon these properties would probably be too noisy for use as satisfactory microwave generators. Similarly high power devices depending entirely on the interaction of particle waves for their operation may well be limited by focusing troubles.

No attempt has been made to cover all cases in the discussion of devices in the last two Parts of this article but it should be clear that the phenomena in the devices not mentioned are broadly similar. For instance in the magnetron a rotating space charge cloud exists in which space charge waves can propagate in an azimuthal direction. The dispersion relation for these waves will be modified by the crossed magnetic field and some sort of cumulative interaction will be apparent even in the absence of an external circuit. Such waves have been often observed[12] and their importance in determining the noise characteristics and build up of oscillations when a circuit is present are well appreciated. This particular case is further complicated by the fact that the electrons move on a closed path indicating that some sort of electronic feed-back condition will modify the spectrum of the free waves which would otherwise propagate.

This and other crossed field devices would yield to a similar study but the principle that a space charge, or similar wave can be represented by a transmission line, and its interaction with a circuit can be discussed in terms of coupled transmission line theory, is fundamental and of importance in the analysis of any space charge device. This importance arises because of the fundamental difficulty inherent in solving Poisson's equation subject to arbitrary boundary conditions and, particularly for predicting the initial characteristics of newly conceived devices, the simplicity of the method is too apparent to need further comment.

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Short News Items

Sir Willis Jackson, F.R.S., Director of Research and Education of Associated Electrical Industries (Manchester) Ltd, is to return in October 1961 to academic life as Professor of Electrical Engineering at Imperial College, University of London. Sir Willis, who is a former apprentice of the old Metropolitan-Vickers Electrical Company, now Associated Electrical Industries (Manchester) Limited, resigned his Chair at Imperial College some seven years ago to become Director of Research and Education of the Manchester Company. In 1938 he was appointed Professor Electrotechnics at Manchester and Professor of Electrical Engineering at the Imperial College of Science and Technology, London, in 1946. He returned to the company as Director of Research and Education in 1953. Sir Willis has wide experience in both education and research. Among other activities he has been a member of the Scientific Advisory Council of the Ministry of Supply and the British Broadcasting Corporation and has been President of both the Engineering and Education Sections of the British Association for the Advancement of Science. Sir Willis is at present a member of the University Grants Committee, the National Council for Technological Awards, the Research Council of the Department of Scientific and Industrial Research, the National Council for the Supply and Training of Teachers, the Committee on the Organization and Control of Government Research Expenditure, and the Federation of British Industries Research Committee. He is also a member of Council and this year's President of the Institution of Electrical Engineers. His main research interests are in the properties of electrical insulating materials and in microwave measurements.

A new transatlantic telephone cable will be laid direct between the United Kingdom and the United States under an agreement signed by the British Post Office and the American Telephone and Telegraph Company, which will come into operation in 1963. This cable, known as TAT3, will be 3 400 nautical miles in length and will carry 128 circuits for telephony or telegraphy as required. The cable will also transmit radio programme material. It will not be used for live television as the bandwidth is not sufficient. It will be of the lightweight type, using polythene insulation without external armouring. This form of cable will find its first application in the cable between the United Kingdom and Canada which is to be laid next year. The overall diameter of the deep sea cable will be about 1¼in. About 180 rigid type repeaters of a new design developed by the Bell Telephone Laboratories will be required to boost the speech at intervals of about 20 miles along the length of the cable. The cost of the project will be £12M.

The First International Congress of the International Federation of Automatic Control (IFAC) was held in Moscow from 27 June to 7 July this year. Arrangements are now being made by the Institution of Mechanical Engineers for discussion of the British papers presented at the Congress at a Symposium in London on 27 and 28 September 1960 for the benefit of those who are interested in this important subject but were not able to attend the Moscow Congress. In addition, at the discussion of the papers it is planned to report briefly the more important contributions to the Moscow Congress. Preprints of the British papers are available and will be supplied to those registered to attend the Symposium. A list of the papers, together with a registration form to attend the Symposium may be obtained on application to the Secretary of The Institution of Mechanical Engineers.

The Institution of Electrical Engineers has announced that the following have been elected to fill the offices of the Institution for the session 1960/61: President: Sir Hamish D. MacLaren, K.B.E., C.B., D.F.C. Ordinary Members of Council: Professor H. E. M. Barlow, D. A. Barron, C. O. Boyse, F. H. S. Brown, Professor M. W. Humphrey-Davies, L. Drucquer, J. M. Ferguson, R. J. Halsey, Professor J. M. Meek, The Hon. H. G. Nelson, H. V. Pugh, R. L. Smith-Rose, J. R. Rylands, G. A. V. Sowter, H. G. Taylor. Chairman Electronics and Communications Section: T. B. D. Terroui. Ordinary Members of Committee: W. H. Aldous, W. J. Perkins, K. F. Sander, J. A. Saxton, T. R. Scott, F. J. D. Taylor.

Full time and part time day and evening courses in Telecommunications, Electronics and Television, are to be held during the coming session at the Northern Polytechnic. A course of 24 evening lectures on colour television is to be held for the third year running and visiting lecturers for this course will be drawn from the major research organizations, the BBC, D.S.I.R., and Radio Research Board. Full particulars of the courses can be obtained from the Head of the Department of Electronics and Telecommunications, Northern Polytechnic, Holloway, London N.7.

The Council of The Institution of Electrical Engineers has recently agreed to the establishment of an annual prize for the best Higher National Certificate student who takes electronics or telecommunications subjects in his final examination. The award, which will be known as the Goldup Prize, has been endowed by Mullard Ltd as a memorial to the life and work of the late T. E. Goldup, a past president of the Institution. As a consequence, on the establishment of the Goldup Prize the existing Page Prize will in future be awarded to the corresponding Higher National Certificate student who has taken electrical power subjects in the final examination.

The French and Algerian television networks have been connected since 14 July by an experimental 4 000Mc/s tropospheric scatter link between the Pyrenees and Algiers. A single repeater station only is required to cover this distance of nearly 450 miles which is a world record. This equipment has been developed and manufactured by Télécommunications Radioélectroniques et Téléphoniques, Paris, for the French Broadcasting and Television Company (R.T.F.).

The General Electric Co. Ltd is now producing a broadband radio equipment in the 6 000Mc/s frequency band. This equipment will convey either a television circuit or 960 speech circuits, and conforms to the latest CCIR recommendations. The first order for this system has already been received from the General Post Office. It covers the supply and installation of equipment for two television channels between Carlisle and Kirk O'Shotts, near Glasgow. There will be two intermediate repeater stations at Riddingshill and Green Lowther.

The Bluebird car which is to make an attempt this month at setting up a new world land speed record at Bonneville Salt Flats, Utah, is to be fitted with a radio telephone system for communication purposes between the car and pit areas and control positions. Pye Telecommunications have fitted a specially built f.m. duplex radiotelephone transmitter and receiver into one of the wheel fairings of Bluebird, with a microphone in the driver's oxygen mask. A completely separate r.t. network provides continuous contact between the main and turnround pits, the course marshall and the control point. A radio telemetry system* is being supplied by Armstrong Whitworth Aircraft Ltd which will transmit to a recording station data regarding temperatures and stresses set up in selected points of the car during its run.

* *Electronic Engineering 32, 230 (1960).*

Thorn Electrical Industries Ltd and Standard Telephones and Cables Ltd announce the signing of an agreement under which the BRIMAR Valve and Cathode Ray Tube Division of Standard Telephones and Cables Ltd has been acquired by Thorn. A new company is being formed under a name including the trade mark 'BRIMAR' which will continue all the activities previously carried out by Standard Telephones in this field. The 'Special' Valve Division of Standard Telephones and Cables Ltd is not included in the agreement and will continue as hitherto. Thorn Electrical Industries Ltd is the largest manufacturer of domestic television receivers in the United Kingdom, marketing under various brand names, including Ferguson, Philco and (in association with E.M.I. Ltd) 'His Master's Voice' and Marconiphone.

The Fourth Summer School was held by A.E.I. (Rugby) Ltd at Rugby from 11-15 July, which was attended by some 30 Professors of Electrical Engineering from British Universities and Technical Colleges including visiting Professors from Montreal, Witwatersrand and California. A series of lectures was arranged covering the latest developments in electrical and electronic engineering and visits were made to laboratories and works. The opening address was given by Sir Willis Jackson, F.R.S., President of The Institution of Electrical Engineers, who spoke on the relation between Industry and the Universities.

A course of 20 lectures on pulse techniques is to be given at the Borough Polytechnic on Monday evenings at 7 p.m. commencing on 3 October. A 12-week laboratory course on the same subject is to be held concurrently. Further particulars can be obtained from The Secretary, Borough Polytechnic, London S.E.1.

The Technical Publications Association have recently established a City and Guilds of London Institute syllabus for Technical Authorship. A list of colleges registered for the Technical Authorship Courses for the 1960/61 session is now prepared. Full details of the Technical Authorship Syllabus or the courses may be obtained from The City and Guilds of London Institute, 76 Portland Place, London W.1. Inquiries regarding membership of the Technical Publications Association should be addressed to the Secretary, Technical Publications Association, 86-88 Edgware Road, London W.2.

The Institution of Electrical Engineers held its Summer Meeting at Glasgow from 27 June to 1 July, with a total attendance of some 530. A civic reception was held in Glasgow by the Lord Provost and Magistrates of the City of Glasgow and visits were arranged to local centres of electrical and electronic interest.

De Havilland Propellers Ltd have now set up at their factory at Stevenage, Hertfordshire, an organization which can perform data processing on a commercial basis. This is believed to be the first such service which is generally available to industry in the South of England. Data from strip chart records on film or paper can be rapidly and accurately transferred to standard punched cards, punched paper tape, or into tables. The punched cards or tape may then be fed into a suitable computer elsewhere. Alternatively, if the information is presented on punched cards or tape it can be automatically displayed, up to six variables being plotted together on a common base scale. Handling rates in either case are normally in excess of 800 points per hour. Work has already been undertaken for the Central Electricity Generating Board, the United Kingdom Atomic Energy Authority, and several missile contractors.

An agreement for technical and commercial co-operation has now been concluded between E.M.I. Electronics Ltd and the Swedish SAAB Aircraft Company. In future SAAB will sell E.M.I.'s equipment in the industrial, automation, broadcasting, instrument and special tube fields throughout Sweden, Norway, and Denmark, through their sales organization SAAB Electronics, Stockholm. Likewise, E.M.I. Electronics will market SAAB products in the British Commonwealth and other countries. SAAB Aircraft Company was incorporated in April 1937 to develop and manufacture military aircraft. Today the company employs more than 8 000 and operates major factories at Linköping, Trolhättan, Jönköping, Gothenburg, and several other towns. Its range of products includes civil aircraft, helicopters, small cars and electronic equipment. SAAB is at present engaged in a major expansion programme for its automobile production and new facilities at Trollhättan will make possible an annual production of more than 24 000 cars. Other commercial developments include digital and analogue computers, numerically controlled machine tools, electronic instruments and special components. Full implementation of the agreement will begin in a few months' time. In addition, the two companies have agreed to exchange technical information and licences in the future.

An advanced inter-island and aeronautical telecommunications system was officially inaugurated in Barbados on 10 June. This £500 000 multi-channel v.h.f. scheme was engineered in conjunction with and on behalf of International Aeradio Limited by Pye Telecommunications Limited and Ericsson Telephones Limited. It provides modern automatic telephone and teleprinter services for islands operating in the Caribbean and gives instant, round-the-clock contact between any airport, airline office and aircraft, irrespective of weather conditions. Twelve separate and simultaneous voice channels have been provided, along with 12 to 18 teleprinter channels. The main route runs from Trinidad via Grenada, St. Vincent, St. Lucia, Dominica, Antique and St. Kitts and by a link, through All-American Cable and Radio Company, to St. Croix, Puerto Rica and on by coaxial cable to North America. There are spur circuits from Grenada to St. Vincent, from St. Lucia to Barbados and Martinque and from Antigua to Guadaloupe. Facilities provided by the system include automatic dialling to any position on the network, air-to-ground circuits and automatic routing and selection of all telegraph messages. A survey of the system and suitable sites was made four years ago by I.A.L. when it was found that to make the project possible, roads had to be built to mountain-top sites, power supplies provided and hurricane-proof buildings erected. All radio units in the system are duplicated and provision made for automatic changeover of any unit should performance fall below a predetermined level. Indication of such changeover is immediately relayed back to Trinidad where engineers are able to check every unit in the system and changeover at will.

Standard Telephones and Cables Ltd have installed a comprehensive system for theatre sound reproduction and communication in the newly opened Royalty Theatre, Kingsway, London. The system comprises in addition to extensive sound amplification from the stage, special loudspeaker calling and rehearsal equipment, thus providing facilities both front-of-house and rear-of-house. The equipment required for this include 19 amplifiers, 12 high quality microphones, 41 loudspeakers and over 2 000 yards of cable.

The Sectional Meeting of the World Power Conference was held in Madrid on 5-9 June 1960 when over 2 100 participants attended, a record number for a Sectional Meeting. The number of participants from the United Kingdom was 154. It is proposed to hold the Sixth (Plenary) World Power Conference at Melbourne, Australia, on 20-26 October 1962, when the theme of the Conference will be 'The Changing Pattern of Power'. Arrangements for the preparation of the papers to be presented by the British National Committee are in the hands of a special Papers Committee (Chairman: Sir Josiah Eccles, Deputy Chairman, The Electricity Council). Any suggestions as to papers should be sent to: The Secretary, British National Committee, World Power Conference, 201-2 Grand Buildings, Trafalgar Square, London W.C.2. Further information concerning the Melbourne Conference can be obtained from the Secretary of the British National Committee or from the Australian National Committee, 45 Little Lonsdale Street (P.O. Box 1823Q), Melbourne C.1, Victoria, Australia.

LETTERS TO THE EDITOR

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

A Transistor-Thermistor Feedback Quadrature Suppressor

DEAR SIR,—I was most interested in the article by Hutcheon and Harrison (February 1960) and the subsequent letter by Winter in your June issue. While at the Bristol Aircraft Company, I originated the self-balancing thermistor bridge[1] described by Winter and should like to draw attention to the following somewhat similar circuit for direct voltage measurement.

The voltage to be measured is compared with the voltage appearing at the junction of the resistor R and the bead of an indirectly heated thermistor Th (see Fig. 1). Initially the thermistor re-

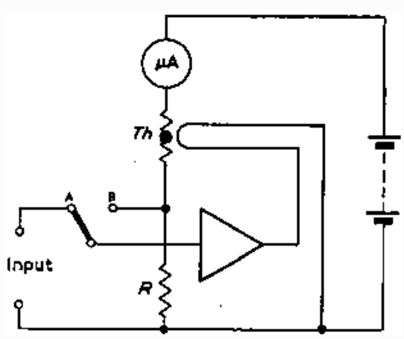


Fig. 1. Thermistor feedback voltmeter

sistance is high and a voltage difference will exist between A and B. This voltage is amplified, preferably by a chopper amplifier, and the output applied to the heater of the thermistor. The resistance of the thermistor bead will fall until the voltage difference between A and B is just sufficient, when amplified, to maintain the thermistor resistance at this equilibrium value. By making the amplifier gain very large the voltage difference between A and B can be made negligibly small and the input resistance presented to the unknown voltage at balance will become very large. This voltage will be very nearly equal to iR and i may be read directly on a microammeter.

The chief advantage of the system described is the virtual independence of the accuracy on all circuit parameters with the exception of R and meter sensitivity.

Yours faithfully,

B. J. WHITE,
Dawe Instruments Limited,
London, W.5.

REFERENCE

1. Bristol Aircraft Electronic and Vibration Note No. 40.

The Authors' Reply:

DEAR SIR,—Mr. White's use of a thermistor to smooth the output of a chopper-type d.c. amplifier is ingenious, and emphasizes the value of the indirectly heated thermistor as a circuit-element. It seems useful to regard the combination of a transistor a.c. amplifier, demodulator and thermistor potentiometer as a 'solid-state servo' which can be used for many purposes and is inferior to its mechanical equivalent only in linearity and the absence of a position output. Quite often these factors do not matter.

Yet another application of such a device is as part of an analogue multiplier, as shown in Fig. A. Although the response is not fast the accuracy can be made very high. Two mechanical choppers working synchronously allow the thermistor potentiometer to be shared between two circuits alternately, i.e.:—

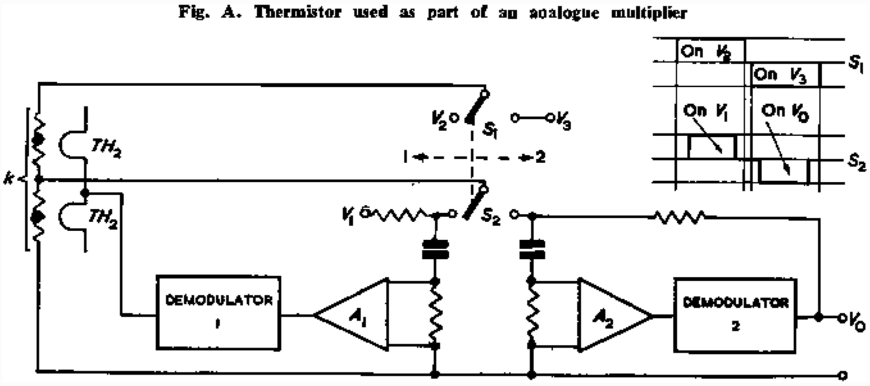
(1) Inputs V_1 and V_2.

(2) Input V_3 and output V_o.

During the periods when the choppers are in position (1) the gain K of the thermistor potentiometer is automatically adjusted by the error between KV_2 and V_1 at the input of the a.c. amplifier A_1. With sufficient gain in this amplifier, K can be made equal to V_1/V_2 very nearly, the ratio of two d.c. input voltages.

During alternate half cycles the choppers are in position (2), and the output voltage V_0 is adjusted so as to equal $K\ V_3$, i.e. $V_0 = V_1 . V_3/V_2$.

Fig. A. Thermistor used as part of an analogue multiplier



The device relies on the fact that the value of K remains constant over a complete cycle of operations due to the thermal lag in the thermistors. By using 50c/s choppers a high accuracy can be obtained, and a.c. amplifiers only need be used.

There must be many more applications of a similar nature for thermistor servos.

Yours faithfully,

I. C. HUTCHEON and D. N. HARRISON
Research and Development,
George Kent Limited,
Luton, Beds.

Explicit Solutions for the Inductive Circuit With Non-Linear Resistance

DEAR SIR,—Let a d.c. voltage E be applied to a circuit consisting of an inductance L and a non-linear resistance characterized by the current-voltage relation $i = kv^n$. The differential equation of the circuit is

$$L\ di/dt + (i/k)^{1/n} = E$$

or

$$L\ k\ n\ v^{n-1}\ dv/dt + v = E \quad \dots (1)$$

the solution of which is

$$t(v) = nkL \int_0^v v^{n-1}/(E-v)\ dv \quad \dots (2)$$

In another communication[1] a recurrence relation was established between the solution of equation (2) for the exponent n and that for the exponent $n+1$, so that the integral in equation (2) need be evaluated only for n in the range $0 < n \leqslant 1$.

In a recent text[2] it is pointed out that equation (2) can be solved explicitly (in terms of elementary functions) if n is a positive integer or an odd multiple of $\frac{1}{2}$. The purpose of this note is to show that equation (2) can be solved in terms of elementary functions if n is any rational number. This result broadens the scope of the previous treatment.

Consider the integral in equation (2):

$$I = \int_0^v v^{n-1}/(E-v)\ dv$$

Suppose that $n - 1 = p/q$, where p and q are integers in their lowest terms. The substitution $v = z^q$ leads to

$$I = q \int_0^{v^{1/q}} z^{p+q-1}(E - z^q)dz$$

so that equation (2) becomes

$$t(v) = (p + q)\ kL \int_0^{v^{1/q}} z^{p+q-1}/(E - z^q)dz \quad (3)$$

The integrand is a rational function of z and can thus be integrated by the method of partial fraction decomposition. The special cases given[2] are easily checked using equation (3).

By way of another example, if $n = \frac{1}{4}$, $p = -3$, $q = 4$, so that

$$t(v) = kL \int_0^{v^{1/4}} dz/(E - z^4)$$

$$= (kL/2E^{1/2}\left\{(1/2E^{1/4})\ ln\left(\frac{E^{1/4} + v^{1/4}}{E^{1/4} - v^{1/4}}\right) + (1/E^{1/4})\tan^{-1}(v/E)^{1/4}\right\}$$

By the recurrence formula previously given[1], the solution for $n = 5/4$ is readily obtained.

Yours faithfully,

H. KAUFMAN,
Mathematics Department,
McGill University,
Montreal.

REFERENCES

1. KAUFMAN, H. Recurrence Relation for Inductive Circuit with Non-Linear Resistance. *Electronic Engng.* 28, 273 (1956).
2. Electrical Engineering Staff, MIT Electric Circuits p. 689 (John Wiley & Sons, Inc., New York, 1940).

The Anode-Cathode Follower

DEAR SIR,—With reference to the article by Vincent and Kaine in the April issue of ELECTRONIC ENGINEERING, I believe the circuit described is better known as an 'amplified feedback cathode-follower' or just as a multiple valve cathode-follower. The circuit has been used in the way described in many applications where high performance cathode-followers are required and circuits using up to nine valves per cathode-follower unit have been described elsewhere[1].

While this application of the circuit may not be widely known, the actual circuit is. Redrawing the authors' circuit,

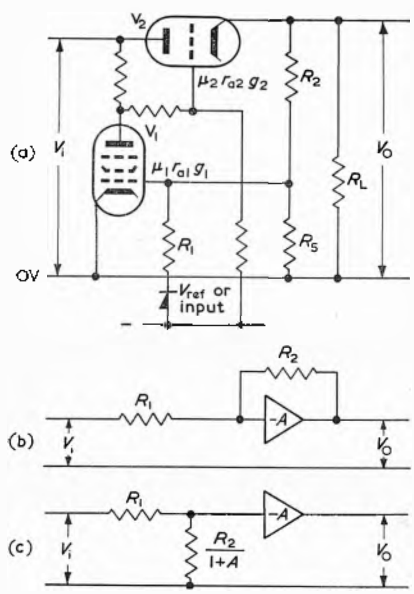


Fig. 1. A common application of the anode-cathode-follower

as shown in Fig. 1, illustrates a common application. The circuit is the normal series stabilizer used in regulated power supplies, one is in fact used by the authors in their power unit. The only difference between the output amplifier and the stabilizer is that the 'input' signal is replaced by a reference voltage. The output impedance of the circuit may be found in practically any work on regulated supplies and applied directly to the output amplifier. Unfortunately, for people using the circuit as an amplifier, the gain normally calculated is V_i/V_o and not V_o/V_{ref}. This may be found by solving the mesh equations, but a better approach is often found in the use of standard feedback equations. A big advantage of this method is that simple expressions are quickly derived and these may then be expanded as required. For example, consider the circuit gain V_o/V_{ref} or V_o/V_i. The circuit may be represented as in Fig. 1(b), which may be replaced by Fig. 1(c). The gain is then given by:

$$A' = (V_o/V_i) = \frac{A}{1+A}\left(\frac{R_2}{R_1+(R_2/(1+A))}\right) = \frac{R_2}{R_1(1+(1/A)+(R_2/R_1A))} \quad \text{.................} \; (1)$$

The difference of A' from the required value of R_2/R_1 is hence quickly given in terms of open loop gain. If an exact expression for A' is required, equation (1) may be expanded by substitution for A. Similarly the standard expression for output impedance of an amplifier with negative voltage feedback is:

$$Z_o' = \frac{Z_o}{1 + A\beta} \quad \text{...........} \; (2)$$

where Z_o is impedance with no feedback.

Again, exact expressions if required may be derived from this by substitution.

The circuit described by the authors may be limited in its applications due to its fairly low input impedance, approximately R_1. If a high input impedance is required the circuit may be modified by replacing V_1 by a long-tailed pair. Feedback is applied to one of the pair while the other provides a high impedance input terminal.

Yours faithfully,

R. THOMPSON,
Havant, Hants.

REFERENCE

1. THOMASON J. G. Multi-Valve Cathode Follower Circuits. *Wireless World.* 63, 310 & 373 (1957).

The Authors' Reply:

DEAR SIR,—Mr. Thompson has made a valid and interesting comparison between the anode-cathode follower and the circuit formed by isolating the series voltage-dropping valve and its control amplifier in a stabilized power supply of the type such as we used. However, the applications and detailed development of the circuit in the two cases are so different as to merit separate description, in our opinion.

With regard to the reference given, Mr. Thomason's Fig. 8 is clearly a prior publication of the basic circuit from which our development proceeded and we willingly acknowledge this. We greatly regret that we were not aware of this article, until it was brought to our attention by Mr. Thompson's letter, and trust that our article may nevertheless have served a useful purpose by drawing further attention to the possibilities of this circuit, particularly to its high peak-current handling capacity. The double cathode-follower, referred to in our article, is an example of a multi-valve cathode-follower which has the necessary low output impedance but is nevertheless not suitable for transmitting our large positive pulses.

The standard methods of deriving the gain and output impedance of any given feedback circuit such as this, whether by circuit analysis or by reference to the general feedback amplifier theory, are so well-known and so much a matter of personal preference that we decided to minimize discussion on this point. However, if a quick derivation of the results is desired, one may take the usual formula for the gain of a negative feedback amplifier and multiply this by a factor γ to allow for any attenuation (or gain) between the input terminal and the junction at the input end of the loop. (This apparently trivial modification is actually of quite general utility, because of the inevitable practical requirement that some form of mixing circuit must be provided for the input and feedback voltages.) This gives:

$$G = (\gamma/\beta) \cdot \frac{1}{1 + 1/\alpha\beta}$$

Our equation (1), which gives the precise value of the gain, is in this form and may be written down at once. The output impedance follows immediately, from the variations of G with R_L. Approximations can be made as desired, with the advantage that the magnitude of the terms being dropped is known.

Yours faithfully,

C. H. VINCENT and D. KAINE
United Kingdom Atomic Energy Authority,
Aldermaston, Berks.

Regulated Power Supplies

DEAR SIR,—With reference to "Letters to the Editor" February issue, I note with interest Professor Middlebrook's comment on stabilization of the anode load return voltage.

The problem exists whenever the series element is driven from a high source impedance device, e.g. a transistor (grounded emitter) or a pentode (grounded cathode), and is basically a problem of endeavouring to retain the effective 'μ' of the series element.

In general the greater the 'μ', the greater the benefit to be derived from stabilization of the return point, e.g. if the 'μ' is low (valve type 6080, $\mu = 2$) there is very little gain.

The method mentioned by Middlebrook can produce an additional benefit, in that, if the series element is a pentode type (e.g. 12E1, CV4060) its screen can be returned to the stabilized point, thus increasing the effective 'μ' and allowing a lower minimum V_a for the element.

Although the advantages of this well-known method cannot be denied, one feels that more elegant, and economic methods may eventually be achieved.

Yours faithfully,

D. J. COLLINS,
Solartron Research & Development Ltd.,
Dorking, Surrey.

BOOK REVIEWS

Mechanization of Thought Processes, Volume 1 and 2

958 pp. 84 Figs. Foolscap 4to.. H.M.S.O. 1960. Price 50s.

AS early as the seventeenth century Leibnitz proposed to investigate the way in which computing machines could be used to simulate or even to replace human intellectual effort. With the coming of automatic digital calculators after the Second World War the newspapers and popular press seized upon the same idea with avidity, and almost daily announcements appeared about the near-human performance of the latest 'giant brain'. Machine designers themselves were far more cautious in their approach, and the scientific literature contains numerous refutations of any anthropomorphic abilities on the part of the machines. Now that the place of computers in pure and applied science has become well established and their capabilities, considered purely as calculators, are known, the attention of the more forward-looking workers in the computer field has naturally turned to an examination of the question of how nearly a machine can be made to simulate animal, or, more hopefully, human thought processes. A number of more or less informal conferences on intelligence simulation by machines has from time to time been held, but the Conference which was held at the National Physical Laboratory in November 1958 is hitherto the most ambitious and the most international in character.

The volumes under review contain an account of the proceedings of this Conference. In a short review like this it would be impossible to mention by name even the titles of the three dozen papers which were presented at the Conference, and all that can be done is to give some indication of the way in which the proceedings occurred.

The first session of the Conference was devoted to general principles within the field and probably the most interesting contribution in this session was that by Macarthy who discussed programmes which have, in some respects, the attributes of common sense. A concrete example of a computer programme exhibiting some characteristcs of common sense is given, in which the machine is invited to describe the way in which a man seated at his desk can go to an airport.

The second session on automatic programming of computing machines is perhaps the least interesting of the whole series, since most of the papers are well known and at least one of them appears to the reviewer to be a 'rehash' of material which has appeared some four times in other places. The only exception to this regrettable lack of novelty is the paper by Ershov of the Academy of Sciences of the U.S.S.R., whose work, although well known in the Soviet Union, has not been particularly accessible in this country, so that the present reprint will be of interest and utility to western workers. Rather curiously, the session on automatic programming ran into a session of the machine translation of language. Here again, the most interesting feature is the paper describing work in the Soviet Union.

The third session of the Conference dealt with speech recognition and included a discussion of learning in machines. The most controversial paper in this section is probably that by Rosenblat on the Perceptron, a device for which apparently exaggerated claims have been made in the Press The reviewer must confess that his reading of Dr. Rosenblat's paper does little to clarify the matter, since, so far as can be gathered from this paper, the Perceptron is no different from many devices that have been considered over the past decade and that have been rejected for various reasons, some of which are given in the discussion of Rosenblat's paper.

The fourth session, divided into two parts, consists of implications for biology and for industry respectively. Among the biological implications the most notable is that to medical diagnosis discussed in the paper by Dr. Paycha. The French workers in this field seem to have had more co-operation from their medical colleagues than has been available in this country, but even so the day when a computing machine will automatically diagnose the ailments of a patient or, if this is impossible, call for further tests, still seems somewhat remote. In the section on industrial implications a provocative paper, which, so far as can be judged from the discussion, failed to provoke, was that by representatives of the Treasury on the extent to which administration can be mechanized. Clearly this subject is capable of considerable elaboration, but the proposals which are presented in the paper seem to fall short of the ideal, which would be to replace almost all civil servants by machines which are more intelligent, more adaptable and probably better mannered.

During the course of the meeting there were lecture-demonstrations of many kinds, including a self-reproducing machine, a conditional probability computer, automatic pattern recognition by means of a simulated neuron net, and a library retrieval procedure. Most of these demonstrations are well known to those workers who specialize in the particular fields concerned, but probably never before have these various fields and demonstrations been brought together in one place.

There is no doubt whatever that these two volumes form an extremely valuable addition to the rather meagre literature in this field. They will certainly be acquired by all specialist workers, but because they are extremely clearly written in most cases, they will also be of great interest to the intelligent layman who wishes to know what to expect from the computing machines of the next decade.

A. D. BOOTH

Materials and Techniques for Electron Tubes

By W. H. Kohl. 618 pp. 237 Figs. Medium 8vo. Reinhold Publishing Corpn., New York. Chapman & Hall Ltd, London. 1960. Price 132s.

KOHL'S new guide to materials and technologies in the electronics industry is even more comprehensive than its predecessor.

Despite the wide scope, each subject is covered in detail: the approach is essentially practical, sensibly directing the reader elsewhere for more fundamental knowledge. The book is packed with information and especially thorough with references (at least, those in English), although belated additions demand a conscientious reader.

Defects stem from the apparently hasty assembly, leading to errors and inconsistencies and occasional unsystematic presentation so that valuable information may be lost in unsuitable contexts. The inexpert reader may suffer because much information is not critically assessed but collected, without comment, in long passages of direct quotation. Some of these are of doubtful applicability and the reader must himself discriminate when he is most in need of guidance.

The materials—glasses, ceramics, the metals, etc.—are well covered. Their properties and those of associated materials are fully tabulated, the applications well described and suitable tests are frequently given, but not always well presented (e.g. micrographs of copper). At times, perhaps, more emphasis could be given to the limitations and defects of the materials (e.g. rhenium is surely not so attractive; rarely has full advantage been taken of the potentials of chrome-copper and internally oxidized metals; arsenical glasses are hazardous; tungsten often suffers from embrittlement and recrystallization—a particular problem in dispenser cathodes; silver crucibles must not be used at 1 000°C).

Chapters 12-16 on technologies are welcome additions, and are very thorough and generally excellent, with few omissions. There are, unfortunately, a number of slips and more serious mistakes (e.g. p. 577 the sequence given for absorptive behaviour of gases is unacceptable; some values in Tables 15.3, 16.2 are suspect and misleading; Wagener is misrepresented, p. 592).

This is a necessary reference book, which will be even more valuable when fully corrected.

G. L. DAVIS

Beam and Wave Electronics in Microwave Tubes

By R. G. E. Hutter. 367 pp. 158 Figs. Medium 8vo. D. Van Nostrand Company Ltd. 1960. Price 73s. 6d.

THERE are perhaps less than half a dozen books in English devoted to the detailed theory of microwave tubes. Over the last decade a vast amount of knowledge has been accrued concerning the interaction between space charge waves on electron beams and circuit waves on slow wave structures. Most of this information resides in original papers in a rather variegated form of presentation. In this book Dr. Hutter has succeeded in presenting an elegant and unified treatment of the small signal theory of microwave valves. The emphasis is completely analytical with no reference to constructional aspect nor to the problems of electron beam formation and focusing.

The book opens with a short phenomenological description of microwave tubes and this is followed with a chapter on Maxwell's equations, wavefunctions, phase and group velocity and the Poynting vector. These ideas are illustrated by considering propagation along rectangular and circular waveguides. Resonant cavities are then described and the calculation of resonant frequencies and Q factors are made for some special examples.

An excellent chapter follows on slow wave structures. Homogeneous delay lines such as the sheath helix, are first considered and these lead on to inhomogeneous periodic lines and the concept of space harmonics and Brillouin diagrams. Several representative slow wave structures are analyzed to give the reader an idea of the manner in which these problems are attacked.

The next two chapters are devoted to the small signal behaviour of an electron beam when acted upon by a localized field or allowed to drift through a field free region. Generalized electronic equations for the a.c. current density and a.c. velocity modulation are obtained in terms of the impressed field and the beam parameters, and are then applied to a number of examples. These include the microwave diode, the modulation produced by a klystron gap, velocity jump regions, space charge waves and some methods of space charge wave amplification. The transmission line analogue of a modulated beam is also derived.

The small signal O type interaction between a slow wave and an electron beam is then considered. Travelling wave tubes and backward wave oscillators are covered in fair detail. The author then goes on to discuss crossed field devices. One chapter is reserved for thin beam models and a further chapter is devoted to analysis of M type devices with thick beams.

Chapter 13 is concerned with energy conversion in microwave tubes. After defining the concepts of kinetic and potential beam power, several theorems are proved and examples are worked for both O and M type interactions.

In the penultimate chapter the concepts of coupling between circuit modes and beam modes are developed and the matrix formalism is discussed. The final chapter is devoted to a rather abbreviated discussion of noise in longitudinal field tubes.

Considering the large amonnt of analysis in the text there are very few typographical errors and the book is well produced. It is a pity perhaps that there is no mention of the recent exciting ideas of parametric amplification of space charge waves, but then the book is already almost 400 pages long. It is to be highly recommended to anyone with interests in the microwave tube field.

P. N. ROBSON

Electrical Noise: Fundamentals and Physical Mechanism

By D. A. Bell. 342 pp. 98 Figs. Medium 8vo. D. Van Nostrand Company Ltd. 1960. Price 50s.

MANY aspects of the theoretical interpretation of electrical noise, in particular of that arising in thermionic valves, have been the subject of controversy since the early days of experimental verification of the Schottky formula in 1930. In writing a text book on noise, therefore, the choice is open between giving the present position, as is usually done, or covering in addition the background history, in order to give the reader a greater physical insight into the problems. The present work, written by one who has been in the forefront of the controversy, follows the second course, and could well have been called a collection of critical essays or reviews of the literature on noise, tempered, as one would expect, by the personal views of the author.

Following an opening on the necessary mathematical techniques, a very complete discussion is given of the basis of Nyquist's theorem and Johnson noise, including the derivation from the ideas of statistical mechanics, and the application to non-linear elements. Three chapters are devoted to noise in thermionic diodes, the one on space charge control diodes being controversial in that it is particularly concerned with the limiting value of noise (at high anode voltage and small anode current) where the author's result differs from that of most other workers. Noise in grid controlled valves includes a good discussion of the possibilities of correlation between grid and anode noise. The chapter on beam and parametric amplifiers is very up to date, but could with advantage have been made more extensive.

Excess noise at low frequencies in semiconductors and film resistors is well covered, as is the author's contention that a solution of many of the associated theoretical problems can be obtained by treating the duration of stay of charge carriers as a queuing problem.

The last two chapters deal with Barkhausen noise and with standards and measuring techniques.

This is a stimulating book, in which the author does not hesitate to show the difficulties associated with different approaches to noise problems. This mode of presentation makes the book more suitable for engineers and physicists at the post-graduate level than for student study. Comprehensive bibliographies are given for each chapter and the whole book is agreeably free from errors.

W. H. ALDOUS

Structure and Properties of Thin Films

Edited by C. A. Neugebauer, J. B. Newkirk and D. A. Vermilyea. 546 pp. 322 Figs. Demy 4to. John Wiley & Sons Inc., New York, Chapman & Hall Ltd, London. 1960. Price 120s.

THIS book presents the complete proceedings of an international conference on thin films held at Bolton Landing on Lake George, New York, in September 1959. Review papers, contributed by recognized experts in their specialized fields, discuss problems which have been recently solved and those still awaiting solution. Topics include the preparation, growth and structure of thin films, and the mechanical, electrical, optical, and magnetic properties of thin films. In addition, a review of processes occurring at the metal surface is given, as a large fraction of thin film atoms resides on or near the surface.

A complete transcript of the conference discussions is included.

THE INTERNATIONAL MEDICAL ELECTRONICS EXHIBITION

A description, compiled from information supplied by the manufacturers, of some of the exhibits at the Medical Electronics Exhibition held at Olympia, London, from 21 to 27 July

(Voir page 585 pour la traduction en français: Deutsche Übersetzung Seite 593)

ALLEN AND HANBURYS LTD

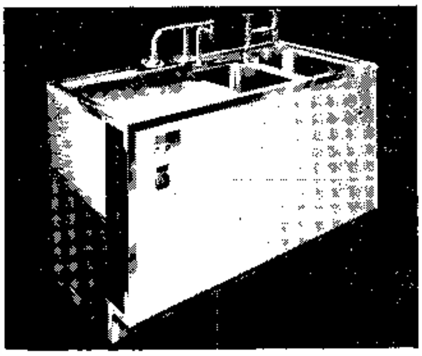
Bethnal Green, London E.2

ULTRASONIC CLEANER

(Illustrated below)

This ultrasonic cleaner was designed for installation in the theatre suites and central sterile supply department of the new surgical building at Guy's Hospital. The cleaner is now manufactured commercially by Kerry (Ultrasonics) Ltd and distributed by Allen and Hanburys Ltd.

The unit has twin stainless steel tanks and drainage board of one piece construction mounted on an angle frame-



work, with removable stainless steel panels. The ultrasonic generating equipment is built-in to the unit and there is side access to the generator. A suitable stainless steel basket to contain instruments in the tank is provided with each unit. A specially designed automatic double rinsing head is fitted over the rinse section.

A suitable hot/cold water mixing tap is fitted to allow a controllable water temperature. The maximum usable water temperature is 60°C, which is also the optimum temperature for efficient cleaning. Should there be danger of water at a higher temperature entering the cleaner, the fitting of an external thermostatically controlled mixer is recommended.

A standing overflow drain governs the amount of water to be used and at the same time allows for a continuous flow to waste of gross contamination. The design of the tank and transducer system eliminates the necessity for variable tuning or complex controls; operation is by a simple push-button on/off switch.

The ultrasonic power is 2kW pulsed at 40kc/s and is supplied by a Mullard generator.

EE 25 751 *for further details*

PACEMAKER DEFIBRILLATOR

(Illustrated on right)

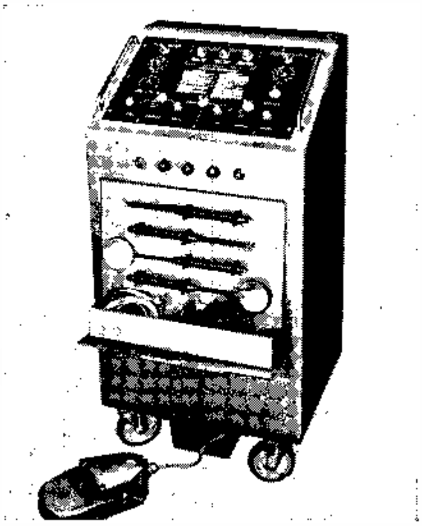
This comprehensive instrument (type T.P.D.3) which is manufactured by Sealey's provides the electrical outputs required for pacemaking and defibrillation in both medical and surgical emergencies. It can be used to apply the electrical power either directly to the open heart or indirectly through the chest wall.

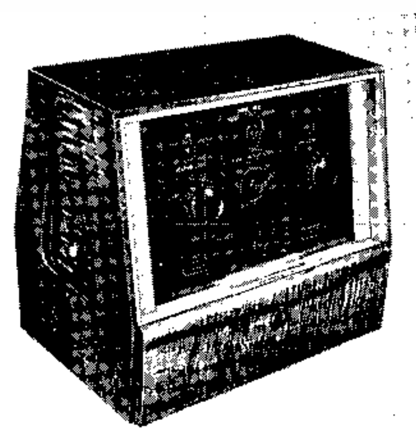
The layout of the control panel is such as to provide optimum safety and simplicity of operation and the whole unit is constructed so that the weight is concentrated at the base of the console to provide maximum stability. Ball bearing castor wheels with anti-static rubber tyres are fitted to the console while in front of the instrument is a drawer unit containing the electrodes and leads and this is designed to pull out and drop down when the apparatus is in use.

The setting of the voltage range and the selector switches determines the level of the pacemaker output power that is periodically connected to the output terminals. The outputs available are discreetly variable between 1 and 17V on the low range and 10 and 170V on the high range, each successive voltage level being approximately 30 per cent greater than the preceding level. The output pulse frequency is continuously variable from 30 to 120 pulses per minute. A neon lamp flashes in synchronism with operation of the pulse circuit and should the mains supply fail while the unit is in operation the battery powered buzzer alarm sounds immediately and the warning light is switched on.

The defibrillator output provided for each operation of the initiating switch is a 0·1sec burst of approximately sinusoidally alternating current. The amplitude of the output voltage is determined by the selector switches on the control and the voltages provided are 60, 80, 106, 140, 190, 250, 330, 440, 590, 780V r.m.s. each voltage level being approximately 30 per cent greater than the preceding value. To reduce the risk of very high voltages being applied direct to the heart the voltages up to 250V appear on one pair of terminals and those above this level on another and it is necessary to have both the switch in the wrong position and the leads on the wrong terminals before these high voltages can be applied accidentally. The output can be initiated by means of a toggle switch on the control panel or by the foot switch provided.

EE 25 752 *for further details*





AMPLIVOX LTD

Beresford Avenue, Wembley, Middlesex

CLINICAL AUDIOMETER

(Illustrated above)

The model 82 Audiometer has 10 fixed frequencies for air conduction and 7 for bone conduction, with common zero, to British or American Standard as desired. An advanced type of masking is incorporated and enables accurate bone conduction measurements to be made, even in cases of severe unilateral deafness. A white noise generator is followed by a filter network, automatically selecting narrow bands of noise centred about the pure tone frequency in use. Narrow band masking excludes all unwanted noise except the critical band necessary, and results in low subjective loudness with maximum efficiency. Applied via an insert earphone, having the minimum of contact with the head, this narrow band masking can be used without the fear of cross-stimulation, and makes it possible to confine the pure tone positively to the tested ear, with resultant increase in accuracy of audiograms.

A second pure tone channel from the main oscillator provides facilities for loudness balance tests and is presented

(Pages 585 to 600 French and German Sections)

in a most elegant form, without the complications of additional controls. The second channel is calibrated in increments of 2dB for additional accuracy in plotting loudness balance curves.

Automatic check of control settings ensures simplicity in use, and a speech communication channel is incorporated.

EE 25 753 *for further details*

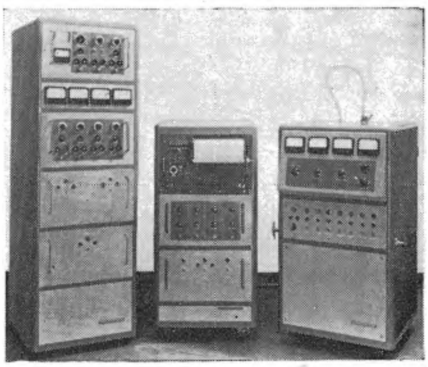
ASSOCIATED ELECTRICAL INDUSTRIES LTD

Crown House, Aldwych, London W.C.2

RESPIRATORY MASS SPECTROMETER

(Illustrated below)

This instrument is produced specifically for clinical practice and research associated with lung diseases. It provides for the continuous analysis of inspired and expired gases, either as passing the lips of a patient breathing through a facepiece, or as extracted from within the bronchial tree by means of a heated



hypodermic sampling probe. Analyses are performed quickly and accurately, and the results are permanently recorded.

The complete apparatus is housed in three mobile cubicles.

The tube unit contains the analyser tube and associated pumping equipment, and the electron-beam supplies. The vacuum pumps which are used to evacuate the tube are a 'Metrovac' type A022 air-cooled oil-diffusion pump and a 'Metrovac' type DR1 rotary vacuum-pump. The pump used on the sampling system is also a type DR1.

The recorder unit is a Sanborn, four-channel, heated-stylus recorder. Each channel is 50mm wide and the total chart width is 250mm. The four recorder amplifiers are mounted below the recorder.

There are nine paper speeds from 0·25mm/sec to 100mm/sec but the usual speed is 5mm/sec. This speed allows nearly three hours' recording with a 200ft roll of waxed recording paper. The recorder has a marker pen which will indicate 1sec intervals, or it can be used as an event-marker. Signals from other instruments (e.g. from an integrating flowmeter) can be applied to any of the channels.

The control unit incorporates all the controls needed to operate the mass spectrometer directly from the mains supply. In the top chassis of the control cubicle are mounted a cathode-ray tube, the main electrometer amplifier, scan unit and metering sub-chassis.

The cathode-ray tube trace shows the full mass range being scanned and the four meters mounted below measure the amplitude of selected peaks in this range. The peaks required are selected by a control mounted on the sub-chassis below each meter, and are indicated by a bright spot on the trace.

Although the main amplifier's individual stages are d.c. it is so connected to the circuits following that the overall effect is that of an a.c. amplifier. This effect makes unnecessary the highly-stable power supplies essential with d.c. amplifiers. The output to the recorder amplifier is 20V full scale.

EE 25 754 *for further details*

EYE MAGNET

(Illustrated below)

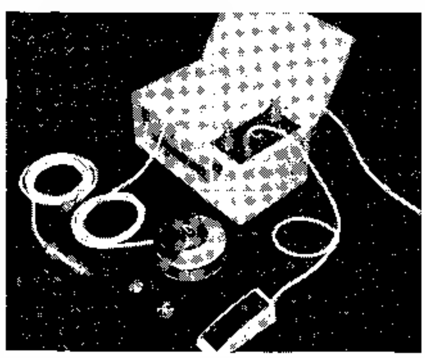
The PMX magnet, for the removal of ferromagnetic bodies from the eye, is capable of exerting an attractive force of over 100 times the weight of the attracted particle.

This high efficiency, combined with the small size and weight of the PMX magnet, provides the ophthalmic surgeon with an instrument in advance of anything previously available.

The weight of the magnet is only 15 lb, enabling the surgeon to apply a really powerful attractive force to the eye of the patient with a high degree of flexibility and convenience.

The equipment includes a small 'pencil' magnet with three interchangeable pole tips. This pencil magnet can be energized continuously without damage. It is particularly useful when the foreign body is on or near the surface of the eye and large attractive forces are unnecessary.

The complete equipment is contained in two carrying cases. One case holds the magnet, mains connecting cable, foot switch, interchangeable cobalt-steel pole tips and the pencil magnet. The other contains the rectifying and control gear, enabling the magnet to be used on a.c. mains.



E 25 755 *for further details*

ATELIERS DE CONSTRUCTION BEAUDOUIN

57, rue de Paris, Bagneux (Seine), France

PORTABLE ELECTROCARDIAGRAPH

This portable electrocardiagraph has been developed to meet the demand for a direct writing instrument for use by the general practitioner either in his surgery or at the home of the patient. This unit is easy to use and is small and light in weight. It uses 6cm wide heat sensitive paper and has overall dimensions of 15cm × 22cm × 28cm and weighs 8·5kg.

This firm is represented in the United Kingdom by SFIM (Great Britain) Ltd, 218-221 Bedford Avenue, Slough, Buckinghamshire.

EE 25 756 *for further details*

I. G. BAILEY AND CO LTD

51 Bearton Green, Hitchin, Hertfordshire

MONITOR-DEFIBRILLATOR-PACEMAKER

This machine is designed primarily for use in the theatre and will provide defibrillation and cardiac stimulation with facilities for monitoring the patient's acceptance or otherwise of this stimulation.

For defibrillation a positive and negative-going square wave pulse at a frequency of 125c/s is applied for a duration of 100msec. In the 'normal' position the pulse voltage is 175V and in the 'high' position 250V. These conditions will apply with a normal adult heart. The machine may be safely used on children as in the case of small hearts the voltage applied will automatically be reduced without the necessity of any external adjustment by the operator. A safety factor exists with this machine as it is impossible to discharge it inadvertently as two depressions of the foot-switch are necessary to produce the defibrillation pulse.

The apparatus is operated by a waterproof and explosion proof foot-switch which will produce one pulse when depressed.

In the stimulation condition a 12msec square wave pulse fully variable from zero to 15V at a pulse rate of 60 to 150 pulses per minute may be applied at the discretion of the operator. An audible click will be heard each time the pulse is applied and provided the patient accepts this pulse and a contraction is produced, a neon light on the panel will flash for each pulse. If no reaction is present in the patient due either to misplacement of the electrodes, inadvertently disconnected leads, etc, this will be indicated by a cessation of the flashing light. A remote audible alarm system to indicate this condition can also be fitted if required.

The instrument is housed in a gas-tight steel case—operates on its own internal batteries, and is completely earth free. The batteries are permanent, and may be recharged from the mains, a safety device is incorporated which renders the instrument inoperative when connected to the mains.

The standard probes provided are cup-shaped electrodes 3in in diameter with insulated handles, but special shapes and sizes of probe can be made to order.

EE 25 757 *for further details*

ELECTRONIC THERMOMETER

This thermometer is a device for measuring cavity and skin temperatures quickly and accurately. It has been designed primarily for use in hypothermia, but is also of considerable value in normal anaesthesia. The thermometer, which is simple and quick to operate, is direct reading and has four independent channels allowing four independent readings to be taken. The four probes consist of an unbreakable thermistor element mounted in a 12 E.G. rubber tube 10ft long. Special probes are also obtainable for other uses (e.g. mounted in gas tubing or hypodermic needles).

The instrument is housed in a small steel case. It has a linear scale from 0°C to 40°C. The accuracy of the instrument in this range is plus or minus 1 per cent. As a bridge circuit is used the accuracy is not affected by ambient changes, nor is any 'settling down' or balancing required. The battery used has a normal life of 12 months, and again the accuracy is not affected by its condition, the instrument remaining within the stated limits up to the useful life of the battery, without adjustment other than the initial zero setting before use. The instrument is completely safe in explosive atmospheres. As it is extremely simple and reliable it may be used with advantage for post-operative temperature measurement in the ward.

EE 25 758 *for further details*

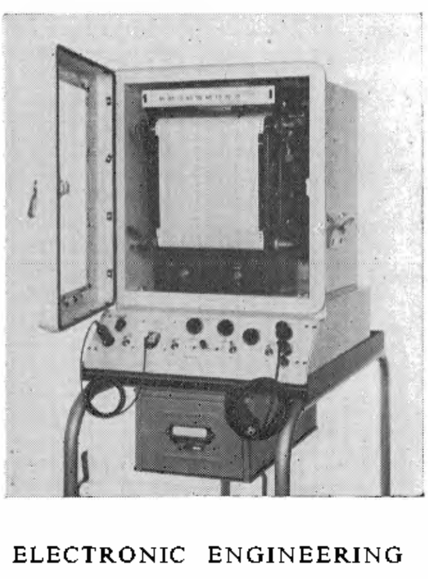
CAMBRIDGE INSTRUMENT CO LTD

13 Grosvenor Place, London S.W.1

DYE-DILUTION CURVE RECORDING APPARATUS

(Illustrated below)

The Cambridge dye-dilution apparatus is designed to record the change of concentration of a dye injected at selected sites in the venuous circulation. This change is detected by a photo-electric attachment on the ear or by a photo-electric cuvette through which blood—from the radial artery, for example—is permitted to flow. The output from either of these units is fed into a special Cambridge electronic recorder which incorporates a highly sensitive and stable



d.c. amplifier. The instant of dye injection is marked on the recorder chart by momentarily depressing an external push-button switch. When the dye passes through the ear or the cuvette, the recorder pen moves across the chart and a time concentration curve is traced.

The mark II version of this equipment, which was exhibited incorporates many features designed to increase versatility, to simplify operation, to improve baseline stability, to allow linear interpretation of the curves, even when large doses of Coomassie Blue dye are used and to allow a series of curves to be drawn, each having exactly the same sensitivity as the last, even in the presence of an increasing background of dye.

A switch is provided on the new control panel (which is brought outside the instrument and sloped for convenience) to enable blue and green dyes to be used with a single earpiece or cuvette.

No external accumulator is required as a highly stabilized built-in supply is provided for the lamp in the earpiece or cuvette. The stability of this unit is such that mains voltage fluctuations, from under 190V to over 250V, produce deflexions of less than two millimetres on the record, and normal mains variations have no effect.

Provision is made for preliminary warming of the earpiece so that when it is fixed to the ear there will be a minimum of delay (5 to 10min) before complete thermal equilibrium is achieved.

The photocell which normally contributes 90 per cent of the thermal drift encountered with this type of apparatus is completely excluded from the circuit as far as drift is concerned, and a very high degree of stability has therefore been achieved without recourse to any special arrangements for maintaining thermal equilibrium. The miniature lamp dissipates less than one watt. Its heating effect on the ear is, therefore, negligible, and cooling is unnecessary.

The pulse rejection circuit positively eliminates arterial pulsations from the record when blue dyes are used, without introducing any form of damping, which would affect the speed of response to dye signals, and without varying the red-cell load resistance, which would cause the instrument to have an unpredictable sensitivity to dye.

EE 25 759 *for further details*

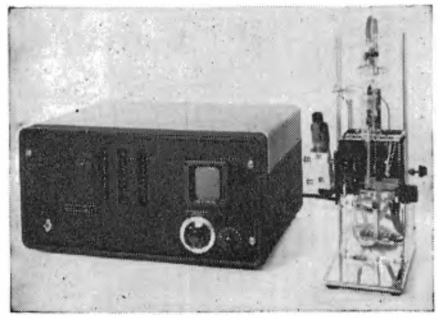
COULTER ELECTRONICS LTD

4 Auriol Mansions, Edith Road, London W.14

BLOOD CELL COUNTER AND SIZE ANALYSER

(Illustrated above right)

The instrument employs a non-optical scanning system providing a counting rate in excess of 6 000 individual cells per second with a counting interval of 15sec. A suspension of blood cells is passed through a small orifice simultaneously with an electric current. The individual blood cells passing through the orifice introduce an impedance



change in the orifice determined by the size of the cell. The system counts the individual cells and provides cell size distribution. The number of cells counted per sample is approximately 100 times greater than the usual microscope count to reduce the statistical error by a factor of approximately 10 times.

The instrument is operated as follows:

(1) A small blood sample is taken with a self-filling pipette. 2mm^3 in a typical case.

(2) Sample is mixed in a beaker with diluent. (Diluent is delivered by automatic burette).

(3) The beaker is positioned on a spring platform so that orifice tube is immersed in the sample.

(4) Count is automatically made by actuating a stopcock and resetting a count switch.

(5) Instrument count is noted, beaker is removed and next sample positioned on stand. No rinsing or cleaning of orifice tube is required between samples except when extreme accuracy is needed.

The principle involved being briefly:

(a) When the stopcock is opened, an external vacuum initiates sample flow from beaker through 0·1mm orifice and 'unbalances' mercury in the manometer.

(b) An electrode in the sample beaker and one in the orifice tube cause an electric current to flow through the orifice. (c) An individual cell as it is carried through the orifice displaces some of the conductive fluid to raise the electrical resistance of the orifice contents. Resistance increase produces a voltage pulse of short duration which has an amplitude proportional to cell size.

(d) Pulses are amplified and displayed on the oscilloscope screen as distinct vertical 'spikes'. Relative cell size is indicated by relative height of the spikes. Pulses are also fed to a threshold circuit allowing selection of a level which if reached by a pulse causes the pulse to be counted. Theshold level is indicated by a brightening of pulse segments above the threshold level.

(e) Closing the stopcock isolates the system from the external vacuum. Syphoning action of the manometer continues sample flow. Rising mercury column contacts start and stop electrodes to activate the digital counter as 0·5mm is drawn through the orifice.

EE 25 760 *for further details*

STANLEY COX LTD

Electrin House, 93-97 New Cavendish Street, London W.1

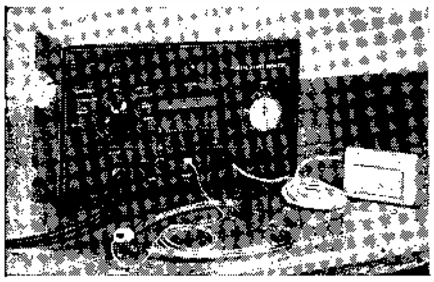
FOETAL HEART MONITOR

(Illustrated below)

This instrument which is manufactured by Bruce Peebles and Co. Ltd and distributed by Stanley Cox Ltd is based on the work of the Medical Physics Unit of the University of Edinburgh.

The pick-up is essentially a mechanical device, tuned to resonate with maximum amplitude at 57c/s, the nominal most pronounced component in the heart beat of the foetus. The attenuation to frequencies only a few cycles from 57c/s is very high indeed, such that very little unwanted noise is fed to the amplifier which, in turn, provides a higher toned signal within the range of easy audibility.

This equipment is fully transistorized, and makes use of printed circuits. The output level is adequate for use in most locations, and can be heard even in a noisy ward, if the speaker is used. The headset may, of course, be used in areas of very high ambient noise. Operation



is from mains supplies, or from built-in dry batteries.

The pick-up is positioned on the mother's abdomen in the best position to pick up the vibrations from the foetus. Visual indication of the heart beat of the foetus is given by the flashing light on the front of the instrument. Audible indications may be had from the loudspeaker, the loudspeaker and headphones or the headphones only. If desired, the light and speaker indications may be cut out, leaving only the headset output.

The sensitivity of the amplifier is adjustable, but would not frequently require to be moved. Audible or visual signals may be monitored against a stop watch, which may easily be fitted to a special design of security mounting on the front panel of the instrument.

EE 25 761 *for further details*

ELECTRONIC MACHINE CO. LTD

Mayday Road, Thornton Heath, Surrey

ELECTRO-CARDIAC MONITOR

(Illustrated on right)

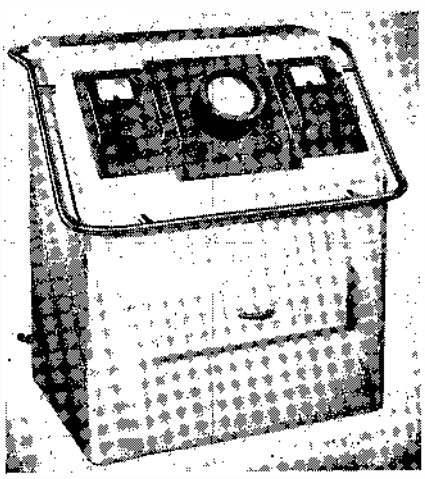
The complete console consists of a mobile case in which are housed a cardiac monitor, pacemaker and defibrillator, each of which can be easily removed and used wherever required. The main case has a large drawer to keep accessory leads and electrodes and is mounted on anti-static rubber castors.

The cardiac monitor displays a cardiogram on a 5in long persistence cathode-ray tube. A lead switch is incorporated for switching the leads 1, 2, 3 and 4 for correct working of leg, arms and chest cardiography. The unit is complete with its own high gain amplifier and does not need a separate electrocardiograph. A meter is fitted to indicate the heart rate (i.e. cardiotachometer) which can be controlled entirely independently from the cardiac-monitor. A socket has been provided for plugging in a standard pen recorder. The amplifier is of a balanced push-pull type with a very low hum level and noise output, it has a continually variable gain control and has ample gain for a 1mV signal input. Power supplies to the cardiac monitor are fully stabilized and will accommodate a mains voltage variation of plus or minus 20V.

The defibrillator was designed by the Physics Department of the Westminster Hospital, London, and has been in regular clinical use at the Westminster Hospital and elsewhere for approximately two years. Controlled pulses are derived from an isolating transformer actuated by a push-button switch and internally set to give a shock of 0·2sec duration. The output from the Westminster defibrillator is a fixed voltage of 120V and the current is variable in a range of 0 to 5A. A memory circuit is incorporated which measures the peak value of current during the pulses and this is indicated on the meter. Provision is made for automatic cut-out in the event of electrodes accidentally shorting or the instrument becoming overloaded. The reset indicator lamp then lights and the unit can then only again be used after the 'reset' button has been pressed.

The pacemaker was designed with the help of the Post-Graduate School of Medicine at Hammersmith, for internal or external application, it has various safety features incorporated.

The pulse rate per minute is from 30 to 160, with voltage for internal use 1 to 20V, for external use the voltage is from 20 to 200V. A meter is incorporated to indicate the current passed by the heart and which also gives warning of any electrode polarization during prolonged use. A neon lamp indicates that the instrument is working.



A feature of the 'pacemaker' is that it is battery or mains operated, it being an automatic action for the batteries to immediately take over on mains disconnexion or failure. This feature makes it possible to use the equipment from its batteries, if desired, while used in the theatre, thus obviating trailing lead hazards.

Connexion to the mains automatically switches the batteries off. There is also a 'battery-check-switch' which gives an immediate indication on the meter of the 'charge' in the battery.

EE 25 762 *for further details*

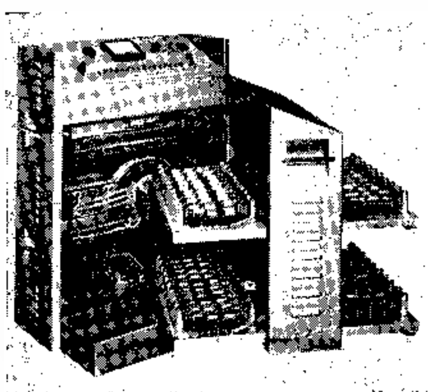
FARADAY ELECTRONIC INSTRUMENTS LTD

Blue Town, Sheerness, Kent

LOW FREQUENCY WAVE ANALYSER

(Illustrated below)

Information about the components of complex functions of time may be



obtained by analysis of the variable in terms of the various frequencies in the original data-record.

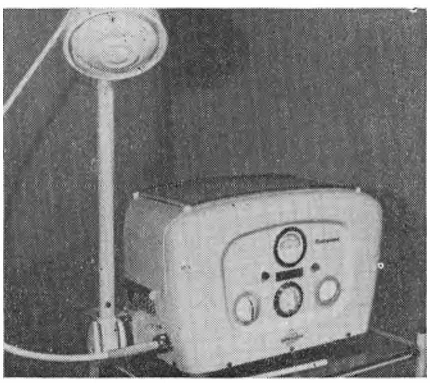
This method has been applied successfully to several problems in industry and medicine, particularly in the analysis of records of electroencephalograms.

The essential feature of this instrument is that it provides a continuous analysis which is immediately visible on the primary record as a frequency histogram. The signal to be analysed is taken from the output of a conventional amplifier and is applied simultaneously to a set of suitably tuned frequency-selective circuits. The outputs of these are integrated and stored in a storage-register, the contents of which are then written out by a special wide-span pen on the original record. The integration is carried out over a period of 10sec (called an epoch) and during the next ten seconds the stored data are written-out sequentially to form a frequency histogram. This is done by a rotating switch which selects each storage capacitor in turn and allows it to discharge through the ballistic pen. This write-out is usually displayed on the original record of the raw data with a 10sec delay (by staggering the analyser pen and the data recording pens). This display is a series of spikes whose amplitude denotes the 'abundance' of a particular frequency component of the data in the epoch analysed. Abundance here means the summation in time of the voltage amplitude of a particular frequency component of the original data. The gains of the filter circuits are set so that the amplitude of the write-out is proportional to the voltage at the analyser input and should be the same for all frequencies, i.e. a two volt input at 3c/s should give the same amplitude of write-out of the 3c/s spike as a two volt input at 30c/s gives the 30c/s spike write-out.

During the write-out of an analysed epoch, analysis of the incoming data continues and the integration during this time is done by using another set of storage capacitors. At the end of ten seconds the role of these capacitors changes from storing to writing and the set of capacitors which have been discharged during the write-out are then used as integrators. Thus, a continuous analysis of the incoming data is achieved.

A further set of storage capacitors is used to store a fraction of each component to compute and record the average of a number of epochs (either 4 or 9) and this average is written out (in a similar manner as the 10 second analyses) at the end of the averaging epoch (40 or 90sec).

EE 25 763 *for further details*



C. J. FOX AND SONS (Chemical) LTD

117 Victoria Street, London S.W.1

MICROWAVE DIATHERMY EQUIPMENT

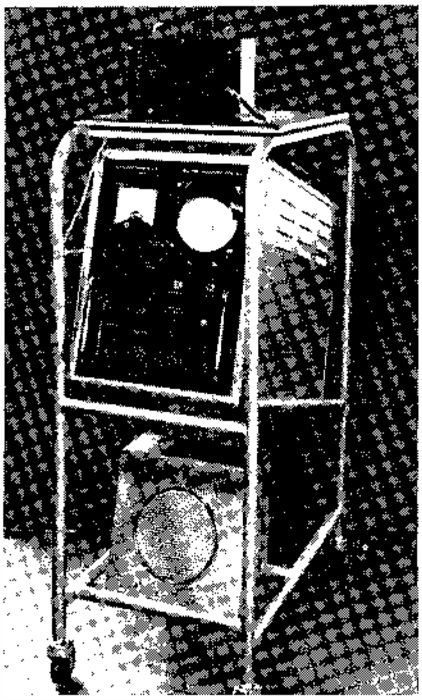
(Illustrated above)

This microwave diathermy equipment, which is manufactured by Deutsche Elektronik of Berlin and distributed by C. J. Fox and Sons, is available in either console or table mounting models. The output of these units is 200W maximum at 2 425Mc/s ±25Mc/s. The energy is fed via a coaxial cable to a dipole aerial where it is 'bunched' by convex mirrors and thus directed to the treatment area.

Focusing electrodes of various dimensions for concentrating the energy on very small areas and body cavity electrodes are available.

The main advantages of microwave diathermy is the ease of focusing the radiation on to the selected area and the absence of electrodes attached to the patient as in h.f. or v.h.f. therapy.

EE 25 764 *for further details*



G.H.S. ELECTRONICS LTD

Severalls Avenue, Chesham, Buckinghamshire

ELECTROMYOGRAPH

(Illustrated above)

The type IIIc is a development from the type III R.A.F. model. A special feature is the addition of a cathode follower input unit on an extension cable, which allows the use of much shorter electrode leads reducing interference pick-up to a minimum.

The head amplifier has inputs with switching facilities for surface electrodes, needle electrodes, connected as either co-axial or bi-polar. Vacuum electrodes may be used in the surface electrodes connexions and facilities are provided for connecting a vacuum pump.

The main amplifier consists of a four stage differential amplifier having a gain sufficient to give a maximum sensitivity of 30μV input for a deflexion on the c.r.t. of 1cm. A sensitivity selection switch allows the gain to be adjusted to deflexion sensitivities of 100, 300, 1 000, 3 000μV for 1cm.

An extra stage in the main amplifier provides for the connexion of a magnetic tape recorder. A relay is provided to switch the amplifier from the normal recording position to playback, in which condition the recorded signals can be displayed on the c.r.t. or heard on the loudspeakers.

Also associated with the main amplifier is the valve-voltmeter which gives visual indication of the peak-to-peak amplitude of the input signal on a 2½in meter mounted adjacent to the c.r.t. The circuit is arranged so the meter rises rapidly to the maximum signal, and returns slowly, allowing time for the readings to be recorded.

EE 25 765 *for further details*

THE INFRA RED DEVELOPMENT CO. LTD

11 Broadwater Road, Welwyn Garden City, Hertfordshire

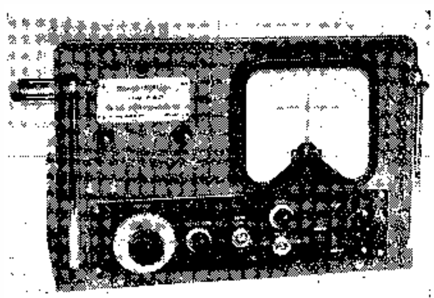
MICROMANOMETER

(Illustrated below)

This instrument is a sensitive pressure measuring device capable of measuring differential gas pressures down to ±10dyn/cm^2 for full-scale deflexion. This pressure is equivalent to ±0·004in or ±0·1mm water gauge. The scale is linear so that pressures down to ±·00016in or ±·004mm water gauge can be readily determined.

The measuring head consists of two symmetrically arranged cavities separated by a metal diaphragm. This diaphragm, together with a fixed electrode on either side, forms two capacitors which are part of the tuning capacitances of two tuned circuits, both equally coupled to an r.f. oscillator. Movement of the diaphragm causes a variation in the capacitance between it and each adjacent electrode, thus unbalancing the voltages across the tuned circuits. These voltages are compared by a differential rectifier voltmeter, and the difference is shown on a meter calibrated directly in pressure.

The pressures to be measured are taken to the measuring head via ¼in



connexions on the front panel of the instrument. Manometers can be produced to give full-scale readings for pressures ranging from ±0·004in water gauge to ±10in w.g. Static working pressures can be up to 1 000lb/in^2 and a differential overload up to at least ten times full-scale pressure can be applied without damage. Multi-range instruments are not supplied, but it is a relatively simple matter to change the measuring head, and provision is made to enable the sensitivity of the electronic circuits to be adjusted so that alternative heads will measure exact multiples of the pressures shown on the meter.

It is possible to use considerable lengths of external connexions without sacrificing the accuracy of calibration; for example, a steady pressure difference between two points can be measured through tubing no more than 1mm diameter bore, up to 20ft in length on each side. The sensitive element in the measuring head will respond to sine wave pressure variations up to 200p/s or more, but long lengths of connexions will form attenuators at these speeds.

A jack is provided to enable an external recorder or oscillograph to be fed in addition to the built-in meter, so that advantage can be taken of the inherently high speed of the measuring system. The output from this jack is approximately $\pm 0{\cdot}5$V across 10kΩ. A switch is provided to increase the time constant of the measuring circuit to 0·3 sec, so that, when required, relatively steady readings can be obtained for pressures fluctuating rapidly about a mean value.

EE 25 766 *for further details*

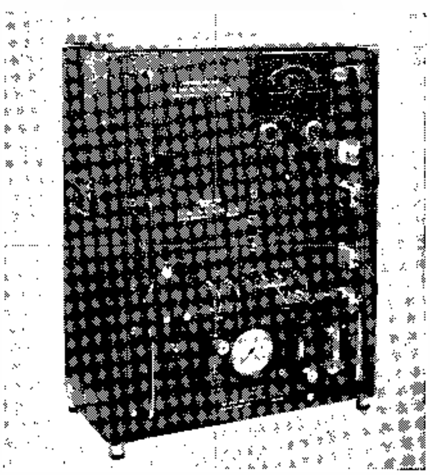
Gas Analyser

(Illustrated below)

This instrument is designed to provide a gas analyser, complete with all necessary gas sampling and flow control equipment, which can be assembled in a single case or supplied as separate units.

The analyser normally consists of three units, of similar overall dimensions which can either be mounted one above the other in a standard 19in cubicle type rack, or in separate cases containing either one or two units. In the latter instances, the units can be separated by distances up to 12ft. If required, a suitable recorder can be included in the assembly.

The three basic units are as follows: An optical system, including the infrared source, rotating shutter, analysis and reference tubes, detector unit and preamplifier. A type 408 low frequency amplifier supplied from a built-in constant voltage transformer, and complete



with all necessary gain controls, a meter calibrated in gas concentration, switched time-constant circuits, and output jacks to feed suitable recording systems. The third unit comprises the gas handling system and its design will depend on the application. In certain cases, particularly when the apparatus is required to measure CO_2 or H_2O, it will include a flushing system to ensure that certain sections of the optical system are maintained free of this gas or vapour.

EE 25 767 *for further details*

LIGHT LABORATORIES

10 Ship Street Gardens, Brighton, Sussex

Thermometers

The model 1GPD is specially designed for hypothermia. The temperature range is 26°C to 38°C with an accuracy of 0·2°C; presentation is on a 6in scale meter. Oesophageal, rectal, skin and pharangeal probes are provided; these are individually connected by plug and socket and can be instantaneously switched. The unit operates from the a.c. mains, the power unit being sealed and spark proof.

A wide range of electric thermometers, both mains and portable battery-operated types, are available for a wide range of applications.

EE 25 768 *for further details*

MEDINTRON LTD

Yeo Street, Bow Common, London E.3

Endoradio Sonde

The Heidelberg probe (endoradio sonde for measuring pH values in the intestinal track) is a capsule measuring about 10mm by 5mm covered with a plastic material which is unaffected by the gastric juices. The capsule, which is swallowed by the patient, contains an oscillator, operating with a transistor, the frequency of which is approximately 1·8Mc/s. Power for the oscillator is supplied by a small battery within the radiosonde. The battery is activated shortly before use of the device, by filling it with an electrolyte.

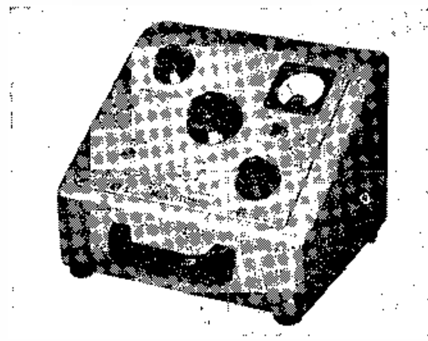
For the purpose of measuring the pH values of gastric juices the radiosonde comprises a pH-measuring cell with one external annular electrode which comes into direct contact with the gastric juice during the measurement. The other measuring electrode is located on the inside of the sonde, being separated from the external electrode by a semi-permeable membrane.

Any change of the pH value of the gastric juice also alters the frequency of the oscillator. The shift in frequency is nearly proportional to that of the pH value and amounts to approximately 120kc/s for the pH 1 to pH 8 range.

It is therefore possible, after calibrating the frequency of the endoradiosonde by immersion in standardized solutions of exactly pre-ascertained pH values, to determine the pH value of the gastric juice by the frequency.

The transmissions from the endoradiosonde are received on an aerial placed on the patient's body, fed to a heterodyne receiver, detected and amplified and fed to a meter calibrated directly in pH values. The output may also be fed to a pen-recorder.

EE 25 769 *for further details*



MULTITONE ELECTRIC CO. LTD

12-20 Underwood Street, London N.1

Cardiac Pacemaker

(Illustrated above)

This cardiac pacemaker is built to a design evolved by the Thoracic Research Unit of Guy's Hospital. It is completely portable, and being fully transistorized will run for several hundred hours from self-contained batteries. It is entirely independent of the mains, and a reserve battery is contained in the instrument. There is a built-in meter for monitoring the battery voltage, and when this falls below a predetermined level, the instrument is switched over to the reserve battery without interrupting the stimulus. The used battery is then replaced by a new one which is ready as a reserve.

The rectangular wave output is of 2msec duration, it may be varied between 1 and 15V and is applied directly to the ventricular myocardium through a flexible wire electrode. The stimulus rate may be set between 60 and 150 per minute and a separate output provides a synchronizing pulse for use in conjunction with an electrocardioscope.

EE 25 770 *fro further details*

NEW ELECTRONIC PRODUCTS LTD

360 Kennington Road, London S.E.11

Physiological Recorder

The type 1195 rack unit provides accommodation for a maximum of six amplifiers, selected in any combination from those described below. Amplifiers are of identical physical dimensions, and each has its own inbuilt stabilized power supply. Units are therefore immediately interchangeable for service purposes.

Provision is made for the mounting of two 6in meters on the top of the rack, which are coupled to the electromanometer amplifiers, and display the mean pressure levels on scales calibrated in mm Hg.

Electrocardiograph Amplifier

This unit is designed especially for electrocardiography, but provision is made for recording other phenomena such as respiration which may need a d.c. response. A low frequency phonocardiograph may be recorded, and filters are provided for eliminating unwanted a.c. interference.

Phonocardiograph Amplifier

This amplifier provides the three stages of filtration described in the British Medical Bulletin 1952 (Vol. viii, No. 4, pp. 333-342). It uses high quality crystal microphones with easily replaceable inserts, and the sounds may be visualized on an oscilloscope and monitored acoustically by means of the stethoscope transducer provided.

Pulse Recorder Amplifier

A two channel unit which makes possible the recording of two simultaneous indirect arterial or venous pulse waveforms. The output may be taken either directly to a 130c/s galvanometer, or to the d.c. input sockets of an e.c.g. amplifier.

Pressure Measurement

The A644 carrier amplifier is designed to work in conjunction with the NEP differential or single ended pressure head. The output is normally taken directly to a 130c/s galvanometer for recording while the mean pressure is displayed on one of the large meters mounted on the rack unit. Facilities for flushing and calibrating are provided on a separate platform.

E.E.G. (Single Channel)

Space is provided in the rack for a battery power supply and charging unit for use with the e.e.g. pre-amplifier type A649.

The e.e.g. pre-amplifier is normally situated close to the patient and is connected by a single cable to the rack mounted power unit.

EE 25 771 *for further details*

OFFICINE GALILEO S.p.A

c/o O.T.E., Via di Caciolle 15, Firenze, Italy or Leland Instruments Ltd, 145 Grosvenor Road, London S.W.1

MULTI-RECORDING APPARATUS

(Illustrated below)

This equipment has been designed specifically for obstetric use.

The built-in oscilloscope can be used to display up to four traces with an electronic switch and can be supplied with a recording camera. The front panel and table carry all the main controls and also a pen recorder.

The pre-amplifiers are contained in single metal cases and are easily interchangeable.

The complete apparatus is designed for the insertion of the following pre-amplifiers:

One amplifying channel for maternal electrocardiography

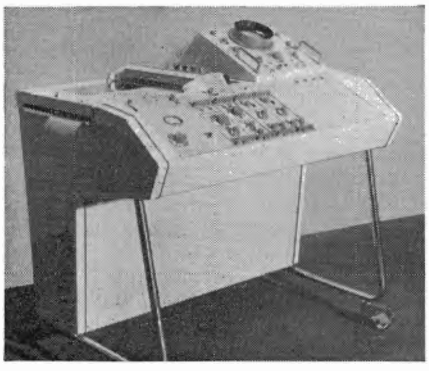
One amplifying channel for foetal electrocardiography

One amplifying channel for maternal tonography

One amplifying channel for foetal tonography

One electromyographic amplifying channel for uterine contraction examination.

The available channels for the insertion of other particular amplifiers, viz, those used for encephalography sphygmoscillometry, pneumography, pressure measurements, plethysmography, rheography, etc.



EE 25 772 *for further details*

OFFNER ELECTRONICS INC

5320 N. Kedyie Avenue, Chicago, 25, U.S.A.

TRANSISTORIZED ELECTROENCEPHALOGRAPH

This completely transistorized electroencephalograph consists of two major units. The first contains the input lead-selector-calibrator panel, the pre-amplifiers, and the small battery supply.

The second unit contains the power amplifiers, oscillator and Dynograph recorder. The two units are identical in size and outward appearance.

The type T employs a new system of amplification which makes possible its sensitivity, stability and reliability. The input signal is converted directly to a high frequency a.c. signal by a vibrating chopper switch. This chopper is of special design for low noise and long life—it is guaranteed for five years of normal operation. The a.c. signal is amplified by the six-stage transistor pre-amplifier, then electronically demodulated and amplified by the two-stage transistor power amplifier, providing the current to operate the low resistance Dynograph units.

The choppers are operated by a transistor oscillator unit, providing the necessary 400c/s power for their operation. All the major units of the type T are of plug-in construction; the pre-amplifiers, the power amplifiers, the Dynographs, and the oscillator unit. All amplifier components are contained in the plug-in units, excepting only the input transformers, the power supply transformer, and components associated with panel controls.

Because of the very low voltage and low current consumption of the pre-amplifier stages, it is more convenient to operate the pre-amplifiers from batteries. Only six small flashlight cells are required for the complete eight channels, and these will give more than one thousand hours of normal operation. The battery voltage is monitored by the panel meter.

Any desired combination of twenty-four electrodes may be selected by duocentric selector switches. In addition, up to nine predetermined runs may be chosen by the preset selector switch.

Seven calibration microvoltages, from 10 to 1 000μV, accurate to within 2 per cent are provided. Electrode resistance is measured by the panel meter to ensure electrodes are properly affixed.

EE 25 773 *for further details*

S.S. ELECTRONICS LTD

Severalls Avenue, Chesham, Buckinghamshire

STIMULATORS

These stimulators were designed by the R.A.F. for diagnosis and therapy of nerve and muscle injuries in their own hospitals and they are now generally available.

They are available in two types. The first is a combined treatment and diagnostic unit giving, in addition to a wide range of square topped pulses, facilities for surging the shorter pulses, galvanic surge and plain galvanic outputs.

The second type is a precise diagnostic stimulator providing the wider range of accurately defined square topped pulses needed for the complete diagnosis of nerve and muscle injuries.

EE 25 774 *for further details*

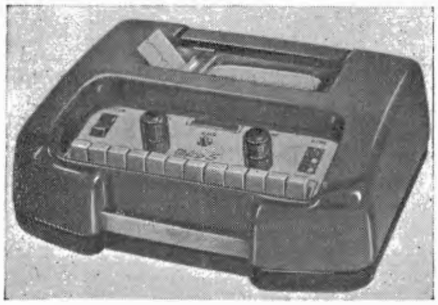
TELCO

45-47, rue de la Division du General Leclerc, Gentilly (Seine), France

PORTABLE ELECTROCARDIOGRAPH

(Illustrated below)

The Telco 5 is a light weight electrocardiograph for use at a patient's home. A voltage indicator and a switch are



used to match any mains voltage and to obtain correct electronic regulation. The inkless record on heat-sensitive chart paper gives a sharp and contrasted tracing. The new Telco pen-motor, moving-coil type, using the most recent magnetic alloys, gives high frequency response, correct damping independently of the pen pressure, and complete absence of hysteresis. The amplitude of the tracing can reach 40mm on a 50mm wide paper chart.

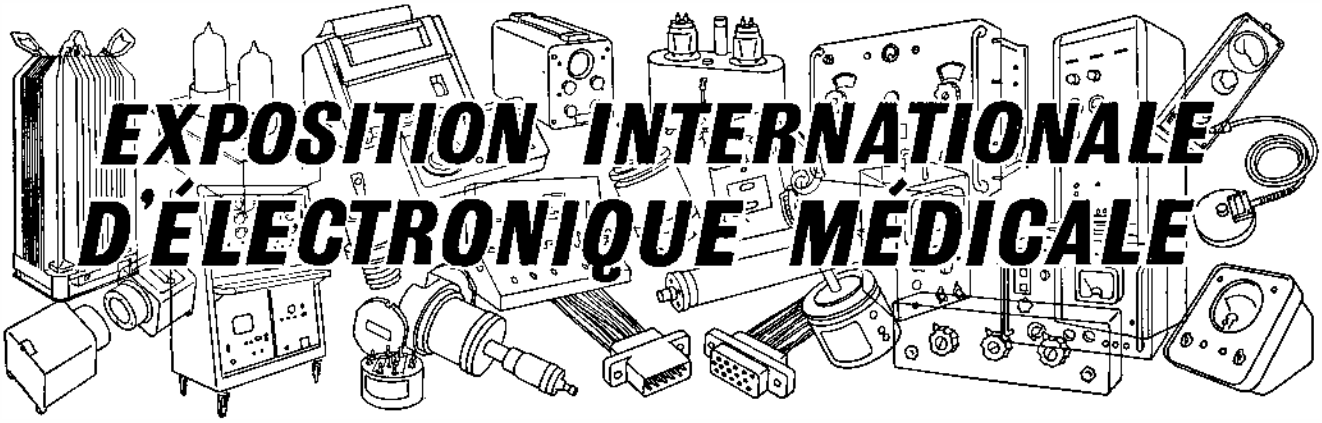
The amplifier uses valves, transistors and printed circuits.

A push-button switch is used for starting, lead selection (with automatic stabilizing of the base line) and calibration (millivolt).

Two speeds are available 25mm/sec (normal) and 2·5mm/sec (survey).

The drive mechanism is a very elaborate block which affords constant paper speed (by a synchronous motor). Loading is quite simple; a push on a lever lifts up the viewing table and gives full access to the place for the paper.

EE 25 775 *for further details*



EXPOSITION INTERNATIONALE D'ÉLECTRONIQUE MÉDICALE

Une description, basée sur des renseignements fournis par les fabricants, de certains des appareils exposés au Salon de l'Electronique Médicale qui s'est tenu à l'Olympia de Londres du 21 au 27 juillet 1960.

Traduction des pages 578 à 584

ALLEN AND HANBURY'S LTD

Bethnal Green, London E.2

DÉGRAISSEUR À ULTRA-SONS

(*Illustration à la page 578*)

Ce dégraisseur à ultra-sons a été prévu pour être installé dans les salles d'opération et dans le département central de fournitures stérilisées du nouveau pavillon chirurgical du Guy's Hospital de Londres. Il est maintenant fabriqué commercialement par Kerry (Ultrasonics) Ltd. et distribué par Allen and Hanbury's Ltd.

Le dégraisseur comprend des cuves jumelées en acier inoxydable et un drain construit en une pièce et monté sur une charpente angulaire, avec panneaux amovibles en acier inoxydable. Le générateur d'ultra-sons est encastré dans l'appareil et on peut y avoir accès latéralement. Chaque appareil est muni d'une corbeille en acier inoxydable pour les instruments. La section de rinçage est dotée d'une double tête de rinçage automatique d'un dessin spécial.

Un robinet mélangeur d'eau chaude et d'eau froide permet de contrôler la température de l'eau. La température d'eau maximum utilisable est de 60°C., ce qui est aussi la meilleure température pour un nettoyage efficace. S'il y a risque de pénétration d'eau à une température élevée dans le dégraisseur, il est recommandé de fixer un mélangeur extérieur thermoréglé.

Un drain fixe de trop-plein règle la quantité d'eau à employer et permet en même temps un écoulement continu des matières contaminées grossières. La conception de la cuve et du système transducteur évitent la nécessité d'un accord variable ou de commandes compliquées. L'actionnement se fait par simple interrupteur arrêt/marche à bouton-poussoir.

La puissance en ultra-sons est de 2kW, à impulsions de 40kHz, et elle est fournie par un générateur Mullard.

EE 25 751 *pour plus amples renseignements*

DÉFIBRILLATEUR STIMULATEUR DU NOEUD SINUSAL

(*Illustration à la page 578*)

Cet instrument complet (type T.P.D.3), fabriqué par Sealey's, fournit les sorties de puissance électrique voulues pour défibriller et stimuler, tant dans les cas d'urgence chirurgicaux que dans les cas médicaux. Il peut être employé pour appliquer la puissance électrique, soit directement au coeur ouvert, soit indirectement à travers la paroi de la poitrine.

Le panneau de commande est conçu de façon à assurer la plus grande sécurité et simplicité de fonctionnement et l'ensemble de l'appareil a été construit de telle sorte que le poids est concentré à la base de la console pour garantir le maximum de stabilité. La console est muni de roues pivotantes sur roulements à billes, avec pneus antistatiques, cependant qu'à l'avant de l'instrument se trouve un élément à tiroir, contenant les électrodes et les fils, qui a été étudié de manière à pouvoir être tiré abaissé lorsque l'appareil est utilisé.

Le réglage de la gamme de tensions et les interrupteurs sélecteurs déterminent le niveau de la puissance de sortie du stimulateur qui est reliée de façon permanente aux bornes de sortie. Les sorties disponibles sont modérément variables entre 1 et 17V, dans la gamme inférieure, et 10 et 170V dans la gamme supérieure, chaque niveau successif de tension étant environ 30% plus élevé que le niveau précédent. La fréquence d'impulsions de sortie est à variation continue, de 30 à 120 impulsions par minute. Une lampe au néon s'éclaire en synchronisme avec le fonctionnement du circuit d'impulsions et, en cas de panne de l'alimentation secteur pendant le fonctionnement de l'appareil, l'avertisseur vibreur à batterie sonnera immédiatement et le voyant d'alarme s'allumera.

La puissance de sortie du défibrillateur, pour chaque opération de l'interrupteur primaire, est d'une rafale de 0,1sec. de courant alternatif à peu près sinusoïdal. L'amplitude de la tension de sortie est déterminée par les interrupteurs sélecteurs sur le panneau de contrôle et les tensions prévues sont de 60, 80, 106, 140, 190, 250, 330, 440, 590, 780V efficaces, chaque niveau de tension étant environ 30% supérieur à la valeur précédente. Afin de réduire le risque d'application directe sur le coeur de très hautes tensions, les tensions allant jusqu'à 250V apparaîssent sur une paire de bornes et celles au-dessus de ce niveau sur une autre paire, de sorte qu'il faudrait que l'interrupteur soit sur la position incorrecte et que les fils soient sur la fausse borne avant que ces hautes tensions ne puissent être appliquées accidentellement. La sortie pourra être déclenchée au moyen d'un tumbler sur le panneau de commande ou de l'interrupteur à pédale prévu à cet effet.

EE 25 752 *pour plus amples renseignements*

AMPLIVOX LTD

Beresford Avenue, Wembley, Middlesex

AUDIOMÈTRE CLINIQUE

(*Illustration à la page 578*)

L'audiomètre modèle 82 a dix fréquences fixes pour la conduction par l'air et sept fréquences pour la conduction par les os, avec un zéro commun, selon la norme britannique ou américaine, au choix. Un type perfectionné de masque est incorporé et permet d'effectuer des mesures précises de conduction des os, même dans les cas de surdité unilatérale grave. Un générateur de "bruit blanc" (ou bruit à spectre continu et uniforme) est suivi par un réseau de filtrage, choisissant automatiquement des bandes étroites de bruit centrées autour de la fréquence de tonalité pure utilisée. Un masque de bande étroite exclut tous bruits indésirables, sauf la bande critique nécessaire, et produisant une intensité sonore subjective réduite, avec le maximum d'efficacité. Ce masque de bande étroite, qui est appliqué à travers un écouteur intercalé ayant un contact minimum avec la tête, peut être utilisé sans crainte d'interstimulation et permet de limiter la tonalité pure positivement à l'oreille soumise à l'examen, résultant en un accroissement de la précision des audiogrammes.

Un second canal de tonalité pure partant de l'oscillateur principal, permet les essais d'equilibre d'intensité sonore et il est présenté sous une forme des plus élégantes, sans les complications de commandes supplémentaires. Le second canal est étalonné en incréments de 2dB, pour plus de précision, dans le tracé des courbes d'équilibre d'intensité sonore.

Le contrôle automatique des réglages de commande assure la simplicité d'emploi de l'appareil, auquel est incorporé un canal de communication vocale.

EE 25 753 *pour plus amples renseignements*

ASSOCIATED ELECTRICAL INDUSTRIES LTD

Crown House, Aldwych, London W.C.2

SPECTROGRAPHE RESPIRATOIRE DE MASSE

(*Illustration à la page 579*)

Cet instrument a été spécifiquement réalisé pour la pratique clinique et les recherches se rapportant aux maladies des poumons. Il permet l'analyse continue des gaz aspirés et expirés, soit lorsqu'ils passent à travers les lèvres du malade respirant au moyen d'un masque, soit lorsqu'ils sont extraits de l'intérieur des bronches à l'aide d'une sonde d'échantillonnage hypodermique chauffée. Les analyses se font rapidement et avec précision et les résultats sont enregistrés de façon permanente.

L'appareil complet est logé dans trois coffrets mobiles.

L'élément à tube contient le tube analyseur ainsi que l'équipement de pompage s'y rapportant, de même que le matériel de faisceau électronique. Les pompes de vide utilisées pour faire le vide dans le tube sont une pompe METROVAC type A022, à refroidissement par air et à diffusion par huile, et une pompe à vide rotative METROVAC type DRI. La pompe employée pour le système d'échantillonnage est également du type DRI.

L'élément d'enregistrement est un appareil Sanborn à quatre voies et style chauffé. Chaque voie a 50 mm de large et la largeur totale du papier est de 250 mm. Les quatre amplificateurs d'enregistrement sont montés sous l'enregistreur.

Neuf vitesses d'avance du papier sont prévues, de 0,25 mm/sec., à 100 mm/sec. mais la vitesse habituelle est de 5 mm/sec. Cette vitesse permet un enregistrement de près de trois heures avec un rouleau de papier ciré de 200 pieds de longueur. L'enregistreur a une plume marqueuse indiquant les intervalles de 1 sec. ou pouvant être employée comme marqueuse "d'événements". Des signaux d'autres instruments (comme, par exemple, un débitmètre intégrateur) peuvent être appliqués à n'importe lesquels des canaux.

Le bloc de contrôle comprend toutes les commandes nécessaires au fonctionnement direct du spectromètre de masse à partir de alimentation secteur. Le châssis supérieur du coffret de contrôle renferme un tube cathodique, le principal amplificateur électromètre, l'élément de balayage et le sous-châssis de mesure.

La trace du tube cathodique montre la gamme entière de masse balayée et les quatre appareils de mesure fixés au dessous mesurent l'amplitude de crêtes choisies dans cette gamme. Les crêtes requises sont choisies au moyen d'une commande montée sur le sous-châssis, placé sous chaque appareil de mesure, et elles sont indiquées par un point brillant sur la trace.

Quoique les étages individuels de l'amplificateur principal soient à courant continu, il est relié aux circuits suivants de telle façon que l'effet total est celui d'un amplificateur de courant alternatif. Cet effet rend inutile les blocs d'alimentations à grande stabilité qui sont essentiels avec les amplificateurs de courant continu. La puissance de sortie à l'amplificateur de l'enregistreur est de 20 V de l'échelle totale.

EE 25 754 *pour plus amples renseignements*

AIMANT OCULAIRE

(*Illustration à la page 579*)

L'aimant PMX, pour extraire les corps ferromagnétiques de l'oeil, peut exercer une force d'attraction de plus de 100 fois le poids de la particule attirée.

Cette grande efficacité, alliée aux dimensions et au poids réduits de l'aimant PMX, offrent à l'ophtalmologue un instrument bien en avance sur tout ce qui a été fait à ce jour.

Le poids de l'aimant n'est que de 7 Kg., permettant au chirurgien d'appliquer une force d'attraction réellement puissante sur l'oeil du malade, avec un degré élevé de souplesse et de commodité.

L'appareil comprend un petit aimant "crayon" avec trois cornes polaires interchangeables. Ce "crayon" magnétique peut être excité continûment sans dommage. Il est particulièrement utile lorsque le corps étranger est à la surface ou près de la surface de l'oeil, rendant ainsi superflues de grandes forces d'attraction.

L'appareil complet est contenu dans deux coffrets portatifs. L'un de ceux-ci renferme l'aimant, le câble de branchement au secteur, l'interrupteur à pédale, les cornes polaires interchangeables en cobalt-acier et le crayon magnétique. L'autre contient le mécanisme redresseur et de commande, permettant d'employer l'aimant sur secteur alternatif.

EE 25 755 *pour plus amples renseignements*

ATELIERS DE CONSTRUCTION BEAUDOIN

57, rue de Paris, Bagneux (Seine), France

ELECTROCARDIOGRAPHE PORTATIF

Cet électrocardiographe portatif a été réalise pour répondre à la demande d'un instrument d'enregistrement direct à l'usage du médecin ordinaire, soit dans sa clinique, soit au domicile du malade. Il est petit, léger et d'un emploi facile. Il utilise un papier sensible à la chaleur, de 6 cm de large, et ses dimensions hors-tout sont de 15 cm × 22 cm × 28 cm, son poids n'étant que de 8,5 Kg.

Les fabricants de cet appareil sont représentés dans le Royaume-Uni par S.F.I.M. (Great Britain) Ltd., Bedford Avenue, Slough, Buckinghamshire.

EE 25 756 *pour plus amples renseignements*

I. C. BAILEY & CO. LTD

51 Bearton Green, Hitchin, Hertfordshire

CONTRÔLEUR DÉFIBRILLATEUR STIMULATEUR

Cet appareil a été principalement conçu pour être employé dans la salle d'opération. Il assure la défibrillation et la stimulation cardiaque et permet en outre de contrôler l'acquiescement du malade à cette stimulation ou son refus.

Pour la défibrillation, on applique une impulsion d'onde carrée positive et négative à une fréquence de 125 Hz. pendant une durée de 100 msec. Dans la position "normale", la tension d'impulsion est de 175 V, et dans la position "élevée" elle est de 250 V. Ces conditions s'appliquent à un coeur adulte. L'appareil peut être employé en toute sécurité pour des enfants car, dans le cas de petits coeurs, la tension appliquée sera réduite automatiquement sans nécessité d'un quelconque réglage extérieur par l'opérateur. Il comporte un facteur de sécurité vu qu'il est impossible de le décharger par mégarde, étant donné qu'il faut presser à deux reprises sur l'interrupteur à pédale pour produire l'impulsion de défibrillation.

Il est actionné par un interrupteur à pédale à l'épreuve des explosions et étanche à l'eau et produit une impulsion par pression de la pédale.

Pour la stimulation, une impulsion d'onde carrée de 12 msec, pleinement variable de 0 à 15 V à un taux d'impulsions de 60 à 150 impulsions par minute peut être appliquée au choix de l'opérateur. Un déclic distinct peut être entendu chaque fois que l'impulsion est appliquée et, à condition que le malade accepte cette impulsion et qu'une contraction se produise, une lumière au néon s'éclairera sur le panneau à chaque impulsion. Si aucune réaction ne se produit de la part du malade par suite du déplacement des électrodes, de fils disjoints par inadvertance ou pour d'autres motifs, ceci sera indiqué par l'arrêt du feu clignotant. Un système de téléalarme audible, indiquant cette absence de réaction, pourra être fixé au besoin.

L'appareil est logé dans un coffret d'acier à l'épreuve des gaz, fonctionnant sur ses propres batteries intérieures, et il est complètement exempt de conducteurs de terre. Les batteries sont permanentes et peuvent être rechargées à partir du secteur. Un dispositif de sécurité incorporé à l'appareil rend ce dernier inopérant lorsqu'il est relié au secteur.

Les palpeurs couramment fournis sont des électrodes en forme de tasses, de 7,5cm de diamètre, avec poignées isolées, mais des palpeurs de grandeurs et de formes spéciales peuvent être fournis sur commande.

EE 25 757 *pour plus amples renseignements*

THERMOMÈTRE ÉLECTRONIQUE

Ce thermomètre permet de mesurer rapidement et avec précision les températures de cavités et de la peau. Il a été principalement prévu pour l'hypothermie, mais il est également d'une très grande utilité pour l'anesthésie normale. Il est à lecture directe et d'un fonctionnement simple et rapide. Etant à quatre voies indépendantes, quatre lectures indépendantes peuvent être effectuées. Les quatre palpeurs consistent en un élément à thermistance incassable, monté dans un tube de caoutchouc de 330cm de long. Des palpeurs spéciaux peuvent également être obtenus pour d'autres usages (par exemple, montés dans des tuyaux à gaz ou des aiguilles hypodermiques).

L'instrument est logé dans un petit coffret en acier. Il est muni d'une échelle linéaire de 0°C à 40°C. La précision de l'instrument dans cette gamme est de plus ou moins 1%. Vu qu'un circuit en pont est employé, la précision n'est pas affectée par les changements ambiants et aucun "tassement" ou équilibrage n'est nécessaire. La batterie employé a une durée normale de 12 mois et, une fois encore, sa précision n'est pas affectée par son état, l'instrument demeurant dans les limites indiquées pendant toute la durée utile de la batterie, sans autre réglage que celui du zéro initial avant l'emploi. L'instrument est parfaitement sûr dans des atmosphères explosives. En raison de son extrême simplicité et de sa sûreté de fonctionnement, il peut être utilisé avantageusement pour les mesures de température postopératoire dans les salles d'hôpitaux.

EE 25 758 *pour plus amples renseignements*

CAMBRIDGE INSTRUMENT CO. LTD

13 Grosvenor Place, London S.W.1

ENREGISTREUR DE DILUTION DE COLORANTS

(*Illustration à la page 580*)

L'appareil à dilution de colorants Cambridge a été étudié pour l'enregistrement des changements de concentration de colorants, injectés à des points déterminés de la circulation veineuse. Ce changement est détecté par un accessoire photoélectrique posé sur l'oreille ou par une cuvette photoélectrique à travers laquelle on fait passer le sang, de l'artère radiale, par exemple. Les données de l'un ou l'autre de ces éléments sont transmises à un enregistreur électronique spécial Cambridge, comportant un amplificateur de tension continue, extrêmement sensible et stable. Le moment d'injection du colorant est marqué sur le papier enregistreur en pressant momentanément un interrupteur à bouton-poussoir extérieur. Lorsque le colorant passe par le récepteur ou la cuvette, la plume enregistreuse se déplace sur le papier et trace une courbe de concentration de temps.

La version Mark II de cet instrument, qu'on a pu voir à l'Exposition, comporte de nombreux perfectionnements qui augmentent la souplesse d'emploi, simplifient le fonctionnement, améliorent la stabilité de la ligne de base, permettent une interprétation linéaire des courbes, même lorsque de grandes doses de colorant bleu sont employées, ainsi que le tracé d'une série de courbes, dont chacune a exactement la même sensibilité que la précédente, même lorsqu'il y a un fond croissant de colorant.

Un interrupteur est prévu sur le nouveau panneau de commande (placé à l'extérieur de l'instrument et de forme inclinée, pour plus de facilité) qui permet d'employer des colorants bleus et verts avec un seul récepteur ou cuvette.

Aucun accu extérieur n'est nécessaire car une source d'alimentation incorporée de grande stabilité est fournie pour la lampe dans le récepteur ou cuvette. La stabilité de cet élément est telle que les fluctuations de tension secteur, inférieures à 190 V, produisent des déviations de moins de 2mm sur l'enregistrement, tandis que les variations de secteur normales sont sans effet.

Le récepteur peut être réchauffé à l'avance, de sorte que lorsqu'il est fixé à l'oreille, il y aura un minimum de retard (5 à 10 min.) avant que l'équilibre thermal complet ne soit réalisé.

La cellule photoélectrique, qui provoque normalement le 90% de la dérive thermale produite par ce type d'appareil, est complètement exclue du circuit en ce qui concerne la dérive, et un degré très élevé de stabilité a donc été atteint sans recourir à des dispositifs spéciaux, quels qu'ils soient, pour le maintien de l'équilibre thermal. La lampe miniature dissipe moins d'un watt. Son effet de chauffage sur l'oreille est, donc, insignifiant, et le refroidissement n'est pas nécessaire.

Le circuit de rejet d'impulsions élimine positivement les pulsations artérielles de l'enregistrement lorsque des colorants bleus sont employés, sans introduire une forme quelconque d'atténuation, ce qui affecterait la vitesse de réponse aux signaux de colorants, et sans varier la résistance de charge à cellule rouge, ce qui donnerait à l'instrument une sensibilité imprévisible aux colorants.

Etant donné que la sécurité du fonctionnement est de la plus grande importance avec des appareils de ce type, il est indispensable que l'opérateur puisse vérifier son comportement à n'importe quel moment et qu'il puisse le comparer avec celui de l'instrument à l'état de neuf.

La stabilité de l'enregistreur peut être vérifiée. Les cellules rouges et infrarouges peuvent être vérifiées indépendamment en ce qui a trait à leur sensibilité et à leur stabilité, et de nombreux autres essais peuvent également être effectués.

EE 25 759 *pour plus amples renseignements*

COULTER ELECTRONICS LTD

4 Auriol Mansions, Edith Road, London W.14

COMPTEUR DE CELLULES SANGUINES ET ANALYSEUR DE GRANDEUR

(*Illustration à la page 580*)

Cet instrument utilise un système de balayage non optique donnant un taux de comptage supérieur à 6000 cellules individuelles par seconde, avec un intervalle de comptage de 15 secondes. Une suspension de cellules sanguines passe à travers un petit orifice en même temps qu'un courant électrique. Les cellules sanguines individuelles passant par l'orifice apportent un changement d'impédance dans l'orifice, déterminé par la grandeur de la cellule. Le système compte les cellules individuelles et assure une distribution selon la grandeur des cellules. Le nombre de cellules comptées par échantillon est environ 100 fois plus grand que le nombre compté au microscope, de façon à réduire l'erreur statistique par un facteur d'approximativement 10 fois.

L'instrument fonctionne comme suit:

(1) Un petit échantillon de sang est prélevé au moyen d'une pipette autoremplissable (2 mm^3 dans un cas typique).

(2) L'échantillon est mélangé dans un gobelet avec un diluant (ce dernier étant fourni par une éprouvette graduée automatique).

(3) Le gobelet est placé sur une plaque à ressort, de manière à ce que le tube orifice soit plongé dans l'échantillon.

(4) Le comptage se fait automatiquement en actionnant un robinet d'arrêt et en réenclenchant un interrupteur de comptage.

(5) Le comptage de l'instrument est relevé, le gobelet est enlevé et l'enchantillon suivant est placé sur la plaque. Aucun rinçage ou nettoyage du tube orifice n'est nécessaire entre les prélèvements d'échantillons, sauf lorsqu'une précision extrême est requise.

Le principe appliqué est brièvement le suivant :

(a) Lorsqu'on ouvre le robinet d'arrêt, un vide extérieur fait passer le flot de l'échantillon par un orifice de 0,1 mm et "déséquilibre" le mercure dans le manomètre.

(b) Une électrode dans le gobelet d'échantillonnage et une autre électrode dans le tube orifice provoquent le passage d'un courant électrique à travers l'orifice.

(c) Au cours du passage de chaque cellule individuelle par l'orifice, elle déplace une partie du fluide conducteur afin d'augmenter la résistance électrique du contenu de l'orifice. L'augmentation de résistance produit une impulsion de tension de courte durée dont l'amplitude est proportionelle à la grandeur de la cellule.

(d) Les impulsions sont amplifiées et reproduites sur l'écran oscilloscopique sous forme "d'épis" verticaux distincts. La grandeur relative des cellules est indiquée par la hauteur relative des "épis". Les impulsions attaquent également un circuit de seuil, facilitant le choix d'un niveau qui, une fois atteint par une impulsion, permettra de compter cette dernière. Le niveau de seuil est indiqué par l'éclat des segments d'impulsions au-dessus du niveau de seuil.

(e) La fermeture du robinet d'arrêt isole le système du vide extérieur. L'effet de siphon du manomètre continue le flot de l'échantillon. La colonne montante de mercure touche les électrodes du marche et d'arrêt afin d'activer le compteur numérique tandis que 0,5 mm sont tirés à travers l'orifice.

EE 25 760 *pour plus amples renseignements*

STANLEY COX LTD

Electrin House, 93-97 New Cavendish Street, London W.1

CONTRÔLEUR CARDIAQUE FOETAL

(*Illustration à la page 581*)

Cet instrument, fabriqué par Bruce Peebles & Co. Ltd. et distribué par Stanley Cox Ltd., est basé sur les travaux de la section de Physique Médicale de l'Université d'Edimbourg.

Le capteur est essentiellement un dispositif mécanique, accordé pour résonner avec le maximum d'amplitude à 57 Hz, la composante nominale la plus prononcée dans les pulsations du coeur du foetus. L'atténuation à des fréquences à quelques cycles seulement de 57 Hz est extrêmement élevée, de telle sorte que très peu de bruit indésirable est transmis à l'amplificateur, lequel, à son tour, fournit un signal d'une plus grande tonalité dans la gamme d'audibilité aisée.

L'appareil est entièrement transistorisé et utilise des circuits imprimés. Le niveau de sortie est suffisant pour l'emploi dans la plupart des endroits, et il peut être entendu même dans une salle bruyante si l'on emploie le haut-parleur. Le casque peut, evidemment, être employé dans des lieux à très forts bruits ambiants. Le fonctionnement se fait sur alimentations secteur ou sur piles sèches incorporées.

Le capteur est placé sur l'abdomen de la mère, dans la meilleure position possible pour capter les vibrations du foetus. L'indication visuelle des pulsations du coeur du foetus est donnée par la lumière à éclats à l'avant de l'appareil. Des indications auditives peuvent être obtenues au moyen du haut-parleur, du haut-parleur et des écouteurs ou des écouteurs seulment. Au besoin, les indications lumineuses et celles du haut-parleur peuvent être supprimées, en ne laissant que la sortie des écouteurs. La sensibilité de l'amplificateur est réglable, mais ne devrait pas être fréquemmment touchée. Les signaux visuels ou auditifs peuvent être contrôlés au moyen d'un chronomètre, qui peut être facilement fixé à l'appareil, selon un montage spécial de sécurité, sur le panneau frontal.

EE 25 761 *pour plus amples renseignements*

ELECTRONIC MACHINE CO. LTD

Mayday Road, Thornton Heath, Surrey

CONTRÔLEUR ÉLECTROCARDIAQUE

(*Illustration à la page 581*)

L'installation complète se compose d'un coffret mobile dans lequel sont logés un contrôleur cardiaque, un stimulateur du noeud sinusal et un défibrillateur, pouvant tous être facilement extraits et utilisés dans le lieu particulier où ils sont requis. Le coffret principal a un grand tiroir pour les électrodes et les conducteurs accessoires, et il est monté sur roulettes en caoutchouc antistatiques.

Le contrôleur cardaque présente un cardiogramme sur un tube cathodique de 12·75 cm à grande persistance. Un interrupteur à conducteurs est incorporé au contrôleur pour commuter les conducteurs 1, 2, 3 et 4, afin d'assurer le fonctionnement correct de la cardiographie des jambes, des bras et de la poitrine. L'appareil est complet avec son propre amplificateur à grand gain et ne nécessite pas d'électrocardiographe à part. Il est muni d'un instrument de mesure pour indiquer le train des pulsations du coeur (c.à.d. un électrocardiographe) qui peut être commandé complètement indépendamment du contrôleur cardiaque. Une prise permet d'enficher un enregistreur à plume ordinaire. L'amplificateur est d'un type push-pull équilibré avec un niveau de ronflement et de bruit très réduits. Il a une commande de gain à variation continue et un gain fort élevé pour une entrée de signal de 1 mV. L'alimentation au contrôleur cardiaque est totalement stabilisée et elle peut s'accomoder à une variation de tension de secteur de plus ou moins 20 V.

Le défibrillateur a été conçu par le Département de Physique du Westminster Hospital, Londres, et il est en service clinique régulier dans le susdit hôpital, ainsi que dans d'autres hôpitaux depuis deux ans environ. Des impulsions contrôlées s'obtiennent d'un transformateur isolant, actionné par un interrupteur à bouton-poussoir et réglé intérieurement pour produire un choc d'une durée de 0,2 secondes. La puissance de sortie du défibrillateur Westminster est une tension fixe de 120 V et le courant est variable dans une gamme de 0 à 5 A. Un circuit à mémoire incorporé à l'instrument mesure la valeur de crête du courant pendant les impulsions et l'indique sur l'organe de mesure. La mise hors circuit se fait automatiquement en cas de court-circuit accidentel des électrodes ou de surcharge de l'instrument. L'indicateur lumineux à réenclenchement s'allume alors et l'instrument ne peut être employé ensuite que lorsque le bouton de réenclenchement a été pressé.

Le stimulateur du noeud sinusal a été conçu avec le concours de l'Ecole de Médecine Post-Universitaire de Hammersmith, pour des applications internes ou externes et il comporte plusieurs dispositifs de sûreté.

Le stimulateur du noeud sinusal a été coups par minute, avec une tension de 1 à 120 V pour l'usage externe. Un instrument de mesure incorporé indique le courant passant par le coeur et signale également toute polarisation d'électrode pendant l'usage prolongé.

Le stimulateur se distingue par le fait qu'il marche sur batterie ou secteur, les batteries entrant automatiquement en fonction en cas de panne ou interruption de l'alimentation secteur. Cet avantage permet de faire marcher l'appareil sur batteries, si nécessaire, pour l'utiliser dans la salle d'opération, évitant ainsi les risques de fils traînant sur le sol.

Le branchement sur secteur romp automatiquement le contact avec les batteries. Il y a aussi un commutateur de contrôle des batteries qui donne une indication immédiate, sur l'appareil de mesure, de l'état de charge de la batterie.

EE 25 762 *pour plus amples renseignements*

FARADAY ELECTRONIC INSTRUMENTS LTD

Blue Town, Sheerness, Kent

ANALYSEUR D'ONDES BASSE FRÉQUENCE

(*Illustration à la page 581*)

Des informations sur les composantes de fonctions compliquées de temps peuvent être obtenues par l'analyse de la variable en fonction des diverses fréquences de l'enregistrement de données original.

Cette méthode a été appliquée avec succès à plusieurs problèmes industriels et médicaux, particulièrement dans l'analyse d'enregistrements d'encéphalogrammes.

Cet instrument se caractérise essentiellement par le fait qu'il fournit une analyse continue, rendue immédiatement visible sur l'enregistrement primaire sous forme d'histogramme de fréquence. Le signal soumis à l'analyse est pris de la sortie d'un amplificateur classique et il est appliqué simultanément à un ensemble de circuits sélecteurs de fréquence accordés de façon adéquate. Les sorties de ces circuits sont intégrées et emmagasinées dans un registre à accumulation, dont le contenu est ensuite annoté par une plume spéciale à grande portée sur l'enregistrement original. L'intégration s'effectue pendant une période de 10 secondes, appelée "époque", et durant les dix secondes suivantes, les données enregistrées sont écrites en séquence pour former un histogramme de fréquence. Ceci se fait par un commutateur rotatif qui choisit tour à tour chacune des capacités d'emmagasinage et leur permet de se décharger à travers la plume balistique. Cette écriture est habituellement présentée sur l'enregistrement original des données brutes avec un retard de 10 secondes (en échelonnant la plume d'analyse et les plumes enregistreuses de données). L'image présentée est une série "d'épis" dont l'amplitude dénote "l'abondance" d'une composante de fréquences particulière des données de l'époque analysée. Abondance signifie ici la sommation dans le temps de l'amplitude de tension d'une composante de fréquence particulière des données originales. Les gains des circuits de filtrage sont réglés de manière à ce que l'amplitude de l'écriture soit proportionelle à la tension, à l'entrée de l'analyseur, et qu'elle soit la même pour toutes les fréquences, c'est à dire qu'une puissance d'entrée de 2 volts à 3 Hz doit donner la même amplitude d'écriture de l'épi de 3 Hz qu'une entrée de 2 volts à 30 Hz donne pour l'écriture de l'épi de 30 Hz.

Durant l'écriture d'une époque analysée, l'analyse des données reçues se poursuit et l'intégration se fait pendant ce temps par l'utilisation d'une autre série de capacités d'emmagasinage. Au bout de dix secondes, le rôle de ces capacités passe de l'emmagasinage à l'écriture et les capacités qui se sont déchargées pendant l'écriture sont ensuite employées comme intégrateurs. On obtient ainsi une analyse continue des données reçues.

Une série supplémentaire de capacités de stockage est employée pour emmagasiner une fraction de chaque composante, afin de calculer et d'enregistrer la moyenne d'un nombre d'époques (soit 4 ou 9) et cette moyenne est écrite (d'une façon semblable à l'analyse de 10 secondes) à la fin de l'époque moyenne (40 ou 90 secondes).

EE 25 763 *pour plus amples renseignements*

C. J. FOX & SONS (CHEMICAL) LTD

117 Victoria Street, London S.W.1

EQUIPEMENT DE DIATHERMIE POUR MICRO-ONDES

(Illustration à la page 582)

Cet équipment de diathemie pour microondes, fabriqué par Deutsche Elektronik de Berlin et distribué par C. J. Fox & Sons, est livrable en modèles pour montage sur console ou sur table. La puissance de sortie de ces appareils est de 200 W au maximum à 2425 MHz ± 25 MHz. L'énergie attaque, à travers un câble coaxial, un aérien dipolaire, où elle est recueillie par des miroirs convexes et dirigée sur la zone de traitement.

Des électrodes de focalisation de diverses dimensions, pour concentrer l'énergie sur de très petites régions ainsi que des électrodes de cavités du corps peuvent être fournies.

Les principaux avantages de la diathermie microondes sont la facilité de focalisation de la radiation sur la région choisie et l'absence d'électrodes attachées au malade, comme dans le cas de la thérapie haute fréquence ou hyperfréquence.

EE 25 764 *pour plus amples renseignements*

G.H.S. ELECTRONICS LTD

Severalls Avenue, Chesham, Buckinghamshire

ELECTROMYOGRAPHE

(Illustration à la page 582)

L'électromyographe type IIIc est un modèle dérivé du type III de la R.A.F. Il se distingue, en particulier, par l'adjonction d'une unité d'entrée à cathode asservie sur câble d'extension, ce qui permet l'emploi de fils d'électrodes beaucoup plus courts, réduisant ainsi le captage de parasites au minimum.

L'amplificateur à tête a des entrées avec facilités de commutation pour électrodes de surface et électrodes à aiguilles, reliées de façon coaxiale ou bipolaire. Des électrodes à vide peuvent être employées dans les connexions d'électrodes de surface et on peut également relier une pompe à vide.

L'amplificateur principal se compose d'un amplificateur différentiel à quatre étages, ayant un gain suffisant pour fournir une sensibilité maximum de 30 μV d'entrée pour une déviation de 1 cm sur le tube cathodique. Un interrupteur de sélection de sensibilité permet d'ajuster le gain à des déviations de sensibilité de 100, 300, 1000, 3000 μV pour 1 cm.

Un étage supplémentaire dans l'amplificateur principal permet de brancher un magnétophone. Un relais est prévu pour commuter l'amplificateur de la position normale à celle de réenregistrement, ce qui autorise la présentation des signaux enregistrés sur tube cathodique ou leur audition sur haut-parleurs.

Un voltmètre électronique est également relié à l'amplificateur principal, ce qui permet une indication visuelle de l'amplitude de crête à crête du signal d'entrée sur un appareil de mesure de 8 cm, monté près du tube cathodique. Le circuit est disposé de telle sorte que l'appareil de mesure monte rapidement au signal maximum, et retourne lentement, donnant le temps nécessaire à l'enregistrement des lectures.

EE 25 765 *pour plus amples renseignements*

THE INFRA RED DEVELOPMENT CO. LTD

11 Broadwater Road, Welwyn Garden City, Hertfordshire

(Illustration à la page 582)

MANOMÈTRE

Cet instrument est un dispositif sensible de mesure de pression pouvant mesurer des pressions de gaz différentielles jusqu'à ±10 dyn/cm^2 pour une déviation totale. Cette pression est équivalente à ±0,004 pouces ou ±0,1 mm de pression hydraulique. L'échelle est linéaire, de sorte que des pressions de ±0,00016 pouces ou ±0,004 mm de pression hydraulique peuvent être facilement déterminées.

La tête de mesure se compose de deux cavités symétriquement disposées et séparées par un diaphragme métallique. Ce dernier, ainsi qu'une électrode fixe des deux côtés, forme deux capacités d'accord de deux circuits accordés, tous deux également accouplés à un oscillateur haute fréquence. Le mouvement du diaphragme cause une variation de la capacitance entre celui-ci et chacune des électrodes adjacentes, déséquilibrant ainsi les tensions parcourant les circuits accordés. Ces tensions sont comparées par un voltmètre redresseur différentiel et la différence est indiquée sur un appareil de mesure étalonne directement en pression.

Les pressions devant être mesurées sont dirigées vers la tête de mesure au moyen de connexions de ¼ pouce sur le panneau frontal de l'instrument. Des manomètres peuvent être produits pour donner des lectures d'échelle totale de pressions s'étendant de ±0,004 pouces à ±10 pouces de pression hydraulique. Les pressions de régime statiques peuvent aller jusqu'à 1 000 lbs/pouce^2 et une surcharge différentielle atteignant au moins dix fois la pression totale peut être appliquée sans crainte. Des instruments multigammes ne sont pas fournis, mais il est relativement simple de changer la tête de mesure. L'instrument permet donc de régler la sensibilité des circuits électroniques de manière à ce que des têtes alternantes mesurent des multiples exacts des pressions indiquées sur l'appareil de mesure.

On peut employer des longueurs considérables de connexions extérieures, sans pour autant sacrifier la précision de l'étalonnage. Par exemple, une différence régulière de pression entre deux points peut être mesurée au moyen de tubes d'un diamètre ne dépassant pas 1 mm et ayant jusqu'à 20 pieds de long de chaque côté. L'élément sensible de la tête de mesure réagit à des variations de pression d'ondes sinusoïdales de 200 impulsions par seconde, mais les longues connexions constituent des atténuateurs à ces vitesses.

Un jack est fourni pour permettre de relier supplémentairement un enregistreur ou un oscillographe extérieur à l'appareil de mesure incorporé, de manière à pouvoir tirer profit de la grande vitesse inhérente du système de mesure. Le rendement de ce jack est environ ±0,5 V à travers 10 kΩ. La constante de temps du circuit de mesure peut être augmentée jusqu'à 0,3 sec à l'aide d'un commutateur, pour que des lectures relativement stables puissent être obtenues au besoin pour des pressions fluctuant rapidement autour d'une valeur moyenne.

EE 25 766 *pour plus amples renseignements*

ANALYSEUR DE GAZ

(*Illustration à la page 583*)

Cet analyseur de gaz est complet avec tous les éléments nécessaires d'échantillonnage de gaz et de contrôle de débit, pouvant être assemblés en un seul coffret ou fournis en tant qu'organes séparés.

L'analyseur se compose normalement de trois organes, de même grandeur, pouvant être soit superposés dans un châssis standard de 19 pouces, du type cabine, soit logés dans des coffrets séparés contenant chacun un ou deux organes. Dans ce dernier cas, les organes peuvent être séparés par des distances allant jusqu'à 12 pieds. Un enregistreur approprié peut être inclus, si nécessaire, dans l'assemblage.

Les trois organes de base sont un système optique, comprenant la source infra rouge, un obturateur rotatif et des tubes d'analyse et de référence, un détecteur et un préamplificateur; un amplicateur basse fréquence type 408, alimenté par un transformateur incorporé de tension constante type 408, complet avec toutes commandes de gain nécessaires, un appareil de mesure étalonné en concentration de gaz, des circuits à constante de temps commutés et des jacks de sortie pour alimenter des systèmes d'enregistrement appropriés; le troisième organe comprend le système de traitement des gaz et sa construction dépend de son application particulière. Dans certains cas, en particulier lorsque l'appareil est nécessité pour mesurer le contenu en CO_2 ou H_2O, il pourra être doté d'un système de chasse pour veiller à ce que certaines sections du système optique soient maintenues exemptes de gaz ou de vapeurs.

EE 25 767 *pour plus amples renseignements*

LIGHT LABORATORIES

10 Ship Street Gardens, Brighton, Sussex

THERMOMÈTRES

Le modèle ICPD a été spécialement prévu pour l'hypothermie. La gamme de températures s'étend de 26°C. à 38°C. avec un degré de précision de 0,2°C. La présentation se fait sur un appareil de mesure à échelle de 6 pouces. Des sondes oesophagiennes, rectales et pharingiennes sont fournies. Elles sont reliées individuellement par fiche et prise et elles peuvent être commutées instantanément. L'appareil est mis en oeuvre à partir du secteur alternatif, le bloc d'alimentation étant scellé et à l'épreuve des étincelles.

Une série étendue de thermomètres électriques, fonctionnant sur secteur ou sur batteries portatives, sont livrables pour une grande gamme d'applications.

EE 25 768 *pour plus amples renseignements*

MEDINTRON LTD

Yes Street, Bow Common, London E.3

SONDE ENDORADIO

Le palpeur Heidelberg (sonde endoradio pour mesurer les valeurs pH du canal intestinal) est une capsule mesurant environ 10 mm × 5 mm, recouverte d'une matière plastique non affectée par les sucs gastriques. Cette capsule, qui est avalée par le malade, contient un oscillateur, fonctionnant avec un transistor, dont la fréquence est d'environ 1,8 MHz. La puissance nécessaire à l'oscillateur est fournie par une petite batterie à l'intérieure de la radiosonde. La batterie est activée un peu avant l'emploi de l'appareil, en la remplissant avec un électrolyte.

Pour mesurer les valeurs pH des sucs gastriques, la radiosonde comprend une cellule de mesure pH avec une électrode annulaire extérieure qui est en contact direct avec le suc gastrique durant la mesure. La seconde électrode de mesure est logée à l'intérieur de la sonde et elle est séparée de l'électrode extérieure par une membrane semi-perméable.

Tout changement dans la valeur pH du suc gastrique modifie également la fréquence de l'oscillateur. Le déplacement de fréquence est presque proportionnel à celui de la valeur pH, et s'élève approximativement à 120 kHz pour la gamme pH à pH8.

Il est donc possible, après avoir étalonné la fréquence de l'endoradiosonde, en l'immergeant dans des solutions normalisées de valeurs pH précontrôlées avec précision, de déterminer la valeur pH du suc gastrique par la fréquence.

Les émissions de l'endoradiosonde sont reçues par une antenne placée sur le corps du malade, transmises à un récepteur hétérodyne, détectées et amplifiées, puis retransmises à un appareil de mesure étalonné directement en valeurs pH. Les données de sortie peuvent également être "injectées" à un préenregistreur.

EE 25 769 *pour plus amples renseignements*

MULTITONE ELECTRIC CO. LTD

12-20 Underwood Street, London N.1

STIMULATEUR DU NOEUD SINO-AURICULAIRE

(*Illustration à la page 584*)

Ce stimulateur cardiaque portatif est fabriqué selon un dessin réalisé par le service de Recherches Thoraciques du Guy's Hospital. Etant entièrement transistorisé, il peut fonctionner pendant plusieurs centaines d'heures à l'aide de batteries autonomes. Il est entièrement indépendant du secteur et comprend une batterie de réserve incorporée. Il comporte également un compteur incorporé pour contrôler la tension de la batterie, et lorsque cette dernière tombe au dessous d'un niveau prédéterminé, l'instrument est commuté sur la batterie de réserve sans interrompre l'incitation motrice. La batterie usée est ensuite remplacée par une nouvelle batterie prête à servir de réserve.

La sortie d'ondes rectangulaires est d'une durée de 2 msec. On peut la faire varier entre 1 et 15 V et elle est appliquée directement au myocarde ventriculaire au moyen d'une électrode de fil métallique flexible. Le rythme du stimulus peut être réglé entre 60 et 150 pulsations par minute et une sortie séparée fournit une pulsation synchronisante pouvant être utilisée avec un électrocardioscope.

EE 25 770 *pour plus amples renseignements*

NEW ELECTRONIC PRODUCTS LTD

360 Kennington Road, London S.E.11

ENREGISTREUR PHYSIOLOGIQUE

L'enregistreur du type armoire 1195 permet de loger un maximum de six amplificateurs, choisis d'une combinaison formée par ceux décrits ci-après. Les amplificateurs sont de dimensions identiques et chacun d'eux a sa propre alimentation stabilisée incorporée. Les éléments sont donc immédiatement interchangeables en cas de besoin.

Deux compteurs de 6 pouces peuvent être montés sur le haut de l'armoire. Ils sont accouplés aux amplificateurs à électromanomètres et indiquent les niveaux de pression moyenne sur des échelles étalonnées en mm Hg.

Amplificateur électrocardiographe

Cet instrument a été spécialement prévu pour l'électrocardiographie, mais il peut enregistrer d'autres phénomènes, tels que la respiration, pouvant nécessiter une réponse du courant continu. Un phonocardiographe basse fréquence peut être enregistré et des filtres sont fournis pour éliminer toute interférence de courant alternatif indésirable.

Amplificateur phonocardiographe

Cet amplificateur assure les trois étages de filtrage décrits dans le British Medical Bulletin 1952 (vol. 3, No. 4, pp. 333-342). Il utilise des microphones piézoélectriques de haute qualité avec pièces intercalées facilement remplaçables. Les sons peuvent être rendus visibles sur un oscilloscope et contrôlées acoustiquement au moyen du transducteur stéthoscopique fourni à ce effet.

Enregistreur de pulsations

Il s'agit ici d'un instrument à deux canaux autorisant l'enregistrement de deux formes d'ondes de pulsations artérielles ou veineuses. La sortie peut être transmise directement à un galvanomètre de 130 Hz ou aux prises d'entrée de courant continu d'un amplificateur d'électrocardiogrammes.

Mesure de pression

L'amplificateur de la porteuse A644 est prévu pour l'utilisation conjointe avec la tête de pression NEP à extrémité unique ou différentielle. La sortie est normalement dirigée vers un galvanomètre de 130 Hz, aux fins d'enregistrement, tandis que la pression moyenne est indiquée sur l'un des trois grands compteurs montés sur le bâti. L'étalonnage et les chasses peuvent se faire sur une plateforme à part.

E.E.G. (Canal Unique)

Le bâti est pourvu de l'espace nécessaire pour une alimentation par batterie et un appareil de charge, pouvant être utilisés avec le préamplificateur e.e.g. type A.649.

Le préamplificateur e.e.g. est normalement placé près du malade et il est relié par un seul câble au bloc d'alimentation monté sur bâti.

EE 25 771 *pour plus amples renseignements*

OFFICINE GALILEO S.p.A.

c/o O.T.E., Via di Caciolle 15, Florence, Italie ou Leland Instruments Ltd, 145 Grosvenor Road, London S.W.1

Appareil multi-enregistreur

(Illustration à la page 584)

Cet appareil a été spécialement conçu pour l'usage obstétrique.

L'oscilloscope incorporé peut être utilisé pour présenter jusqu'à quatre traces au moyen d'un interrupteur électronique, et il peut être fourni avec une caméra enregistreuse. Le tableau et la table frontaux portent toutes les commandes principales, ainsi qu'un enregistreur à plume.

Les préamplificateurs sont contenus dans des coffrets métalliques indépendants, et ils sont facilement interchangeables.

L'appareil complet a été étudié pour y insérer les préamplificateurs suivants:

Un canal amplificateur pour l'électrocardiographie maternelle.

Un canal amplificateur pour l'électrocardiographie foetale.

Un canal amplificateur pour la tonographie maternelle.

Un canal amplificateur pour la tonographie foetale.

Un canal amplificateur électromyographique pour l'examen des contractions utérines.

Les canaux disponibles peuvent servir à l'insertion d'autres amplificateurs particuliers, comme ceux utilisés pour l'encéphalographie, la sphygmoscillométrie, la pneumographie, les mesures de tension, la pléthysmographie, la rhéographie, etc.

EE 25 772 *pour plus amples renseignements*

OFFNER ELECTRONICS LTD

5320 N. Kedyie Avenue, Chicago, 25, U.S.A.

Electroencéphalographe transistorise

Cet électroencéphalographe entièrement transistorisé se compose de deux éléments principaux. Le premier contient le panneau à conducteur d'entrée/sélecteur/étalonneur, les préamplificateurs et l'alimentation à petite batterie.

Le second élément contient les amplificateurs de puissance, l'oscillateur et l'enregistreur Dynograph. Les deux éléments sont identiques quant à la grandeur et à l'aspect extérieur.

Le type T utilise un nouveau système d'amplification qui rend possible son degré élevé de sensibilité, de stabilité et de sécurité de fonctionnement. Le signal d'entrée est converti directement en un signal alternatif de haute fréquence par un interrupteur vibrant. Ce dernier a été spécialement conçu pour une longue durée et un bruit réduit; il est garanti pour cinq ans de fonctionnement normal. Le signal c.a. est amplifié par le préamplificateur à transistors à six étages, puis démodulé électroniquement et amplifié par l'amplificateur de puissance à transistors à deux étages, qui fournit le courant actionnant les éléments Dynograph à faible résistance.

Les écrêteurs sont actionnés par un organe oscillateur à transistors, fournissant la puissance de 400 Hz nécessaire à leur mise en oeuvre. Tous les principaux organes du Type T sont à fiches, c.à.d. les préamplificateurs, les amplificateurs de puissance, les Dynograph et l'organe oscillant. Les pièces d'amplificateurs sont renfermées dans les éléments à fiches, à l'exception des transformateurs d'entrée, du transformateur d'alimentation et des pièces rattachées aux commandes de panneau.

En raison de la consommation très réduite de tension et de courant des étages préamplificateurs, il convient mieux de faire fonctionner les préamplificateurs à partir de batteries. Seules six petites piles du type lampe de poche sont nécessaires pour tous les huit canaux, et elles assurent plus de mille heures de fonctionnement normal. La tension de batterie est contrôlée par le compteur du panneau.

N'importe quelle combinaison voulue de vingt-quatre électrodes peut être choisie par deux interrupteurs sélectifs duocentriques. En outre, jusqu'à neuf parcours prédéterminés peuvent être choisis par l'interrupteur sélectif préréglé.

Sept microtensions d'étalonnage, de 10 à 1000μV, précises à 2% près, sont prévues. La résistance des électrodes est mesurée sur le compteur du panneau afin de s'assurer qu'elles sont convenablement fixées.

EE 25 773 *pour plus amples renseignements*

S.S. ELECTRONICS LTD

Severalls Avenue, Chesham, Buckinghamshire

Stimulateurs

Ces stimulateurs ont été conçus par la R.A.F. pour le diagnostic et la thérapie des blessures aux muscles et aux nerfs dans leurs propres hôpitaux, et ils sont maintenant en vente générale.

Ils sont livrables en deux modèles. Le premier de ces modèles est un appareil combiné de diagnostic et de traitement offrant, en plus d'une gamme étendue d'impulsions de forme rectangulaire, la possibilité de contrôler les impulsions courtes, la surtension galvanique et les sorties galvaniques ordinaires.

Le second type est un stimulateur de diagnostic précis fournissant une gamme étendue d'impulsions de forme rectangulaire exactement définies, qui sont nécessaires au diagnostic complet des blessures nerveuses et musculaires.

EE 25 774 *pour plus amples renseignements*

TELCO

45-47, rue de la Division du Général Leclerc, Gentilly (Seine), France

Electrocardiographe portatif

(Illustration à la page 584)

L'électrocardiographe Telco 5 est un appareil poids léger pour l'emploi au domicile du malade. Il est muni d'un indicateur de tension et d'un commutateur pour l'adapter à n'importe quelle tension de secteur, ainsi que pour obtenir une régulation électronique correcte. L'enregistrement sans encre sur papier sensible à la chaleur donne un tracé net et contrasté. Le nouveau moteur à plume Elco, du type à bobine mobile, utilisant les plus récents alliages magnétiques, donne une réponse de haute fréquence, un amortissement correct, indépendamment de la pression de la plume, et une absence complète d'hystérésis. L'amplitude du tracé peut atteindre 44 mm sur un papier de 50 mm de large.

L'amplificateur utilise des tubes, des transistors et des circuits imprimés.

Un commutateur à bouton-poussoir est employé pour la mise en marche, le choix du conducteur (avec stabilisation automatique de la ligne de base) et l'étalonnage (millivolts).

Deux vitesses sont prévues: 25 mm/sec (normale) et 2,5 mm/sec (étude).

Le mécanisme d'entraînement est un organe très perfectionné qui assure une vitresse constante d'avance du papier (par moteur synchrone), la charge de papier s'effectuant de façon très simple: une pression sur un levier soulève la table de vision et donne libre accès à l'emplacement du rouleau de papier.

EE 25 775 *pour plus amples renseignements*

Résumés des Principaux Articles

Le systéme de commutation statique "à coup" Elliott Sheffer par P. Kellett

Résumé de l'article
aux pages 534 à 539

L'auteur décrit un système dans lequel les interconnexions entre un nombre d'éléments logiques, tous du même type (le "Coup Sheffer"), permettent n'importe quel montage logique ou séquence de commutation. Il explique les régles simples qui régissent les interconnexions, décrit un nombre d'interconnexions fondamentales d'usage géneral et compare le "Coup Sheffer", le "ni" ainsi que les éléments logiques plus connus; "et", "ou" et "non".

Ceci est suivi d'une description des interrupteurs de puissance existants. Ces derniers peuvent être commandés par la puissance de sortie d'un élément logique. Les charges peuvent être reliées, au besoin, dans une matrice, avec une économie considérable.

Circuits diviseurs de fréquence pour instruments de musique par Alan Douglas

Résumé de l'article
aux pages 546 à 549

Les particularités marquantes de cette catégorie de circuits sont le fait que le facteur de division est toujours de deux, leur simplicité maximum pour des raisons économiques et la condition exigeant que la forme d'onde convienne à une synthèse de tonalité stable. Toutes ces conditions peuvent difficilement être réunies dans un circuit simple.

La réduction des effets de glissement dans les calculatrices analogiques à courant continu par E. T. Emms et K. H. Brinkmann

Résumé de l'article
aux pages 550 à 553

Les auteurs expliquent les effets de glissement d'amplificateur à courant continu sur les calculatrices analogiques à courant continu, et indiquent des méthodes pour choisir des facteurs d'intégration et répartir les gains afin de réduire ces effets au minimum.

Ils formulent un nombre de règles de gouverne qui permettront au réalisateur de systèmes de tirer le meilleur parti possible de son équipement.

Introduction au système à chiffre de code ternaire par D. J. Morris et W. Alexander

Résumé de l'article
aux pages 554 à 557

Cet article examine les connaissances concernant l'arithmétique ternaire et traite du système ternaire et de sa conversion en systèmes numériques binaires et décimaux et vice versa.

Un multiplicateur d'effets Hall pour l'utilisation aux hautes fréquences par E. Cohen

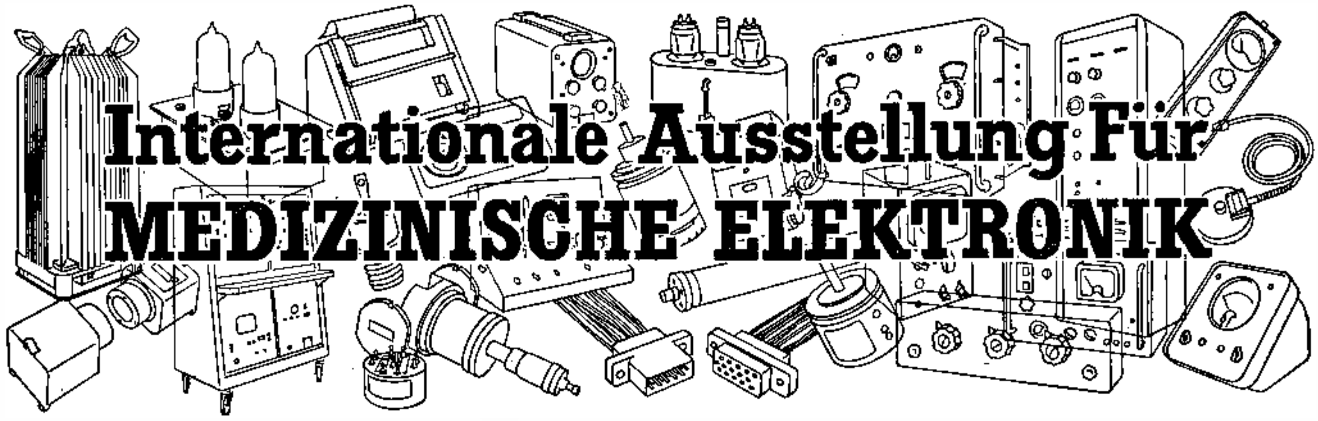
Résumé de l'article
aux pages 558 à 559

Cet article traite sommairement des limitations de l'effet Hall basse fréquence lorsqu'il est utilisé aux hautes fréquences. Il y est démontré que le facteur déterminant dans le fonctionnement haute fréquence est le passage du courant porteur amené par le captage inductif dans le circuit Hall (en l'absence d'un champ magnétique), par suite de la circulation de courant porteur dans l'anode. Ceci peut être réduit par certains montages de circuits et, grâce également à des transformateurs à large bande du type à fiches, on peut produire un passage de courant porteur supérieur à 50 dB au-dessous du signal Hall maximum.

Un atténuateur "décimal"; un réseau de tension à résistance constante, ou diviseur de courant à indices décimaux
par E. R. Wigan

Résumé de l'article
aux pages 560 à 566

L'auteur décrit un réseau qui allie les propriétés d'un diviseur classique de tension ou de courant à celles d'un atténuateur de résistance constante, tout en gardant les caractéristiques les plus utiles de ces deux éléments, sans en avoir, toutefois, les désavantages. Le réseau fournit à ses bornes de sortie des fractions décimales, connues avec précision, de sa tension d'entrée. Ainsi, le rapport entrée/sortie d'um diviseur à trois cadrans décimaux pourra être réglé en 1000 plots de 0,001, de zéro à l'unité. Le diviseur garde toujours une résistance de sortie rigoureusement constante ce, qui veut dire que le courant peut être pris des bornes de sortie sans qu'on ne perde la faculté de pouvoir effectuer des réglages de tension ou de courant de sortie connus avec précision, malgré que la charge réduise le courant ou la tension disponibles.



Beschreibung einer Auswahl der auf der Ausstellung für Medizinische Elektronik vom 21. bis 27. Juli in den Olympia-Hallen in London gezeigten Geräte nach Angaben der Hersteller.

Übersetzung der Seiten 578 bis 584

ALLEN AND HANBURYS LTD

Bethnal Green, London E.2

ULTRASCHALL-REINIGUNGSANLAGE

(Abbildung Seite 578)

Diese Ultraschall - Reinigungsanlage wurde besonders für Einbau in Operationssälen und Hauptverteilungsstellen für keimfreie Materialien im neuen Gebäude des Guy's Hospital entwickelt. Die Anlage wird nunmehr von Kerry (Ultrasonics) Ltd hergestellt und von Allen and Hanburys Ltd vertrieben.

Die Anlage hat rostfreie Doppelbecken und Tropfbretter, die in einem Stück hergestellt und in einem Rahmen aus Winkelprofilen mit abnehmbaren Feldern aus rostfreiem Stahl eingebaut sind. Der Ultraschall Generator ist eingebaut und von der Seite zugänglich. Mit jeder Anlage wird ein rostfreier Korb zur Aufnahme chirurgischer Instrumente geliefert. Über dem Spülteil ist ein automatischer Kopf für Doppelspülung angebracht.

Für Wasserzufuhr geregelter Temperatur ist ein Mischhahn für kaltes und warmes Wasser vorgesehen. Höchsttemperatur ist 60°C, was auch die beste Temperatur für wirksame Reinigung darstellt. Sollte Gefahr bestehen, dass Wasser höherer Temperatur in die Anlage gelangt, so wird eine thermostatisch geregelte Mischung ausserhalb des Zuflusses empfohlen.

Ein Überflussablauf regelt die verbrauchte Wassermenge und ermöglicht gleichzeitig einen dauernden Durchfluss zur Entfernung schwerer Verschmutzungen. Konstruktion von Becken und Gebersystem machen Nachstimmung und komplizierte Bedienung unnötig; die Bedienung erfolgt über einen einfachen Ein-Aus Druckschalter.

Die von einem Mullard-Generator gelieferte Überschall-Leistung ist 3 kW bei 40 kHz Impulsfrequenz.

EE 25 751 *für weitere Einzelheiten*

SCHRITTMACHER-DEFIBRILLATOR

(Abbildung Seite 578)

Dieses von Sealey hergestellte Gerät (Type T.P.D.3) liefert die elektrische Leistung, die für Schrittmachen und Defibrillation sowohl in medizinischen als auch chirurgischen Notfällen benötigt werden. Elektrische Leistung kann entweder direkt an das offene Herz oder indirekt durch den Brustkasten zugeführt werden.

Die Bedienungsplatte ist für höchste Betriebssicherheit und Bedienungsvereinfachung ausgelegt, und das Gewicht des Gerätes ist im Fuss des Standgehäuses konzentriert, so dass das Gerät äusserst kippsicher ist. Die Laufrollen sind mit Kugellagern und antistatischen Gummireifen ausgerüstet. Vorn im Gerät ist eine Schublade angebracht, die die Elektroden und Zuführungen enthält. Wenn das Gerät in Gebrauch ist, kann die Schublade herausgezogen und heruntergeklappt werden.

Einstellung des Spannungsbereiches und eines Wählerschalters bestimmen die Ausgangsleistung des Schrittmachers, die periodisch an die Ausgangsklemmen gelegt wird. Der Ausgang kann im niedrigen Bereich zwischen 1 und 17 V und im hohen Bereich zwischen 10 und 170 V diskret geregelt werden, wobei jeder folgende Spannungspegel ungefähr 30% höher als der vorhergehende Pegel liegt. Die Ausgangstaktfrequenz ist stufenlos von 30...120 Impuls/min regelbar. Eine Glimmlampe leuchtet synchron mit der in Betrieb befindlichen Impulsschaltung auf. Sollte das Stromnetz während des Gebrauchs ausfallen, so ertönt sofort ein batteriebetätigter Summer und eine Alarmlampe wird eingeschaltet.

Der Defibrillator-Ausgang gibt für jede Bedienung des Schalters zur Inbetriebsetzung einen etwa sinusförmigen Wechselstromstoss von 0,1 s Dauer. Die Amplitude der Ausgangsspannung wird durch den Wählerschalter der Bedienung eingestellt. Die einstellbaren Spannungen sind 60, 80, 106, 140, 190, 250, 330, 440, 590 und 780 V_{eff}, und jeder Spannungspegel liegt ungefähr 30% höher als der vorhergehende. Um das Risiko zu vermeiden, dass hohe Spannungen direkt ans Herz gelegt werden, werden die Spannungen bis zu 250 V einem Elektrodenpaar und die über diesem Pegel einem anderen Elektrodenpaar zugeführt. Beide Schalter müssen falsch eingestellt und die Leitungen an die falschen Klemmen angeschlossen werden, bevor diese hohen Spannungen zufälligerweise angelegt werden können. Zur Inbetriebsetzung kann ein Kippschalter am Gerät oder ein Fussschalter benutzt werden.

EE 25 752 *für weitere Einzelheiten*

AMPLIVOX LTD

Beresford Avenue, Wembley, Middlesex

KLINISCHES AUDIOMETER

(Abbildung Seite 578)

Das Audiometer Modell 82 hat 10 feste Frequenzen für Fortpflanzung in Luft und 7 für Fortpflanzung in Knochen mit gemeinsamem Nullpunkt, auf Wunsch nach Britischen oder Amerikanischen Normen. Eine fortschrittliche Abdeckungsmethode ermöglicht genaue Knochenfortpflanzungsmessungen selbst in Fällen schwerer, einseitiger Taubheit. Einem Weissgeräuschgenerator folgt ein Filter-Netzwerk, das automatisch ein enges Geräuschband wählt, das um die benutzte reine Tonfrequenz konzentriert ist. Engbandabdeckung schliesst alle unerwünschten Geräusche ausserhalb des notwendigen kritischen Bandes aus und ergibt eine niedrige subjektive Lautstärke mit grösstem Wirkungsgrad. Über einen Einsatz-Kleinhörer zugeleitet, also in geringster Berührung mit dem Kopf, kann die Engbandabdeckung ohne Gefahr der Kreuzreizung benutzt werden und beschränkt den reinen Ton zwangsläufig auf das zu prüfende Ohr, wodurch die Genauigkeit der Audiogramme erhöht wird.

Ein zweiter reiner Tonkanal vom Hauptoszillator gibt die Möglichkeit, das Lautstärkengleichgewicht zu prüfen; er ist in elegantester Form ohne zusätzliche Regler ausgeführt. Für grössere Genauigkeit in der Aufzeichnung dieser Gleichgewichtskurven ist der zweite Kanal in 2 dB Stufen geeicht.

Automatische Prüfung der Einstellungen erlaubt einfache Bedienung. Ein Kanal für Sprechverbindung ist auch vorgesehen.

EE 25 753 *für weitere Einzelheiten*

ASSOCIATED ELECTRICAL INDUSTRIES LTD

Crown House, Aldwych, London W.C.2

ATEMMASSENSPEKTROGRAPH

(Abbildung Seite 579)

Das Gerät wird besonders für klinische Praxis und Forschung im Zusammenhang mit Lungenkrankheiten hergestellt. Es gibt eine fortlaufende Analyse der ein- und ausgeatmeten Gase, entweder wenn sie über die Lippen eines durch eine Maske atmenden Patienten gehen oder wenn sie innerhalb der Bronchien durch beheizte Probenkanülen entnommen werden. Die Analyse wird schnell und genau mit Daueraufzeichnung ausgeführt.

Die komplette Anlage ist in drei Schränken untergebracht. Der Schlauchteil besteht aus Analysatorschlauch und zugehörigen Pumpen sowie der Stromversorgung für den Elektronenstrahl. Zum Auspumpen des Schlauches wird eine luftgekühlte Öldiffusorpumpe Type Metrovac A 022 und eine Umlaufvakuumpumpe Type Metrovac DR 1 benutzt. Dieselbe Type wird auch zum Probenehmen eingesetzt.

Als Schreiber wird ein vierkanäliges Sanborn Registriergerät mit beheiztem Stylus verwendet. Jeder Kanal ist 50 mm breit und die Gesamtpapierbreite ist 250 mm. Die vier Verstärker sind unter dem Schreiber untergebracht.

Neun Papiervorschübe von 0,25 mm/s...100 mm/s sind vorgesehen,5 mm/s ist jedoch der übliche. Mit einer 60 m-Rolle wachsbeschichteten Papiers können bei dieser Geschwindigkeit fast 3 Stunden lang Aufzeichnungen gemacht werden. Der Schreiber hat einen Stylus, der entweder 1 s Zeitintervalle oder Ereignisse markiert. An jeden der Kanäle können Signale von anderen Instrumenten (z.B. einem eingebauten Strömungsmesser) angelegt werden.

Das Bedienungsgerät enthält alle Bedienungselemente für Spektrograph und Netzanschluss. Im oberen Chassis des Bedienungsschrankes sind eine Elektronenstrahlröhre, der Elektrometer-Hauptverstärker, der Abtastteil und der Messbaustein untergebracht.

Die Spur der Elektronenstrahlröhre zeigt den vollen Massenbereich, der abgetastet wird. Vier darunter angebrachte Messgeräte zeigen die Amplitude vorgewählter Spitzen in diesem Bereich. Die gewünschten Spitzen werden an einer Bedienung vorgewählt, die unter jedem Messgerät am Baustein angebracht ist und werden auf der Kurve durch helle Punkte angezeigt.

Trotzdem die Einzelstufen des Hauptamplifiers galvanisch gekoppelt sind, sind sie so mit den nachfolgenden Kreisen zusammengeschaltet, dass die Gesamtwirkung der eines Wechselstromverstärkers entspricht. Dadurch werden die sonst für Gleichstromverstärker erforderlichen hochstabilen Netzanschlüsse unnötig. Die Ausgangsspannung zum Schreiberverstärker ist 20 V für Vollausschlag.

EE 25 754 *für weitere Einzelheiten*

AUGENMAGNET

(Abbildung Seite 579)

Der PMX-Magnet zur Entfernung ferromagnetischer Körper aus dem Auge kann eine Anziehungskraft ausüben, die 100 × grösser als das Gewicht des angezogenen Teilchens ist.

Die hohe Wirksamkeit verbunden mit kleinen Abmessungen und geringem Gewicht gibt dem Augenarzt ein Instrument, das alle bisher vorhandenen übertrifft.

Der Magnet wiegt nur 6,8 kg und ermöglicht dem Chirurgen, auf das Auge des Patienten eine wirklich starke Anziehungskraft mit grösster Anpassungsfähigkeit und Zweckdienlichkeit auszuüben.

Mit dem Instrument wird ein kleiner "Bleistift"-Magnet mit drei auswechselbaren Polspitzen geliefert. Dieser "Bleistift"-Magnet kann laufend ohneSchaden erregt werden. Er ist besonders nützlich, wo der Fremdkörper auf oder nahe der Augenoberfläche sitzt und keine grosse Anziehungskraft erforderlich ist.

Die gesamte Ausrüstung ist in zwei Tragkoffern untergebracht. Ein Koffer enthält den Magneten, Anschlusskabel, Fussschalter, auswechselbare Kobaltstahl-Polspitzen und den Bleistift-Magneten. Der andere Koffer enthält einen Gleichrichter und Steuerapparate für Benutzung des Magneten an Wechselstrom.

EE 25 755 *für weitere Einzelheiten*

ATELIERS DE CONSTRUCTION BEAUDOUIN

57, rue de Paris, Bagneux (Seine), France

TRAGBARER ELEKTRO-KARDIOGRAPH

Dieser tragbare Elektro-Kardiograph wurde entwickelt, um den Forderungen praktischer Ärzte gerecht zu werden, die entweder in der Sprechstunde oder im Hause des Patienten ein direkt schreibendes Gerät benötigen. Das Gerät ist einfach zu bedienen, klein und leicht. 6 cm breites wärmeempfindliches Papier wird benutzt, und die Gesamtabmessungen sind nur 15 cm × 22 cm × 28 cm. Das Gewicht ist 8,5 kg.

In England wird das Gerät durch SFIM (Great Britain) Ltd, 218-221 Bedford Avenue, Slough, Buckinghamshire vertrieben.

EE 25 756 *für weitere Einzelheiten*

I. G. BAILEY AND CO LTD

51 Bearton Green, Hitchin, Hertfordshire

MONITOR-DEFIBRILLATOR-SCHRITTMACHER

Dieses Gerät wurde in erster Linie für Defibrillation und Herzanregung in Operationssälen entwickelt und ermöglicht auch, durch Überwachung festzustellen, ob der Patient reagiert oder nicht.

Zur Defibrillation wird ein positiv- und negativgehender Rechteckwellenimpuls von 125 Hz für 100 ms angelegt. In der "normal"-Schalterstellung beträgt die Impulsspannung 175 V, in der "hoch"-Stellung 250 V. Das trifft für das normale Herz eines Erwachsenen zu. Die Maschine kann jedoch ohne Gefahr auch für Kinder benutzt werden, da für kleinere Herzen die angelegte Spannung automatisch herabgesetzt wird, ohne dass durch den Benutzer irgendwelche Nachstellungen vorgenommen zu werden brauchen. Als Sicherheitsmassnahme muss der Fussschalter zweimal bedient werden, um den Defibrillationsimpuls hervorzurufen, so dass es unmöglich ist, zufälligerweise Entladungen auszulösen.

Das Gerät wird durch einen wasser- und explosionsfesten Fussschalter bedient, der einen Impuls auslöst, wenn er heruntergedrückt wird.

Für die Herzanregung wird ein 12 ms Rechteckwellenimpuls benutzt, der stufenlos von 0...15 V regelbar ist und nach Gutdünken des Benutzers mit einer Impulsfrequenz von 60...150 Impulsen pro Minute arbeitet. Jedesmal wenn der Impuls abgegeben wird, wird auch ein hörbares Knacken hervorgerufen. Wenn der Patient auf den Impuls reagiert und eine Zusammenziehung stattfindet, wird eine Glimmlampe für jeden Impuls am Bedienungspult aufleuchten. Das Aufleuchten hört auf, wenn der Patient durch Verschiebung der Elektroden oder Unterbrechung der Leitungen nicht reagiert. Eine hörbare Fernwarnung kann auf Verlangen für diesen Fall auch geliefert werden.

Das Gerät ist mit seinen eigenen Batterien in einem gasdichten Gehäuse eingebaut und ist völlig erdfrei. Die Batterien sind Dauerbatterien, die von Zeit zu Zeit am Netz aufgeladen werden können. Eine eingebaute Schutzvorrichtung verhindert Inbetriebnahme während des Ladens.

Die gelieferten Standard-Sonden sind Becherelektroden mit 76 mm Durchmesser und isoliertem Griff; Sonderelektroden verschiedener Grösse und Form können gegen Sonderauftrag jedoch auch geliefert werden.

EE 25 757 *für weitere Einzelheiten*

ELEKTRISCHES THERMOMETER

Das Thermometer ist für schnelle und genaue Messungen von Höhlen- und Hauttemperaturen bestimmt. Es wurde hauptsächlich für Hypothermie entwickelt, ist aber auch für normale Betäubung sehr wertvoll. Das Thermometer kann einfach und schnell bedient werden, wird direkt abgelesen und hat vier unabhängige Kanäle, die vier unabhängige Ablesungen ermöglichen. Die vier Messköpfe bestehen aus unzerbrechlichen Thermistor-Elementen und sind in einen 3 Meter langen 12 E.G. Gummischlauch eingebaut. Für andere Zwecke sind auch besondere Messköpfe lieferbar (z.B. in Gasschläuchen oder Injektionskanälen).

Das Gerät wird in einem kleinen Stahlgehäuse geliefert und hat eine lineare Skala von 0...40°C. In diesem Bereich ist die Genauigkeit des Gerätes ±1%. Durch Benutzung einer Brückenschaltung wird die Genauigkeit durch Umgebungsänderungen nicht beeinflusst, ebenso ist weder "Einstellzeit" noch Ausgleichen erforderlich. Die verwendete Batterie hat normalerweise eine Lebensdauer von 12 Monaten, und auch hier wird die Genauigkeit nicht durch den Zustand beeinflusst und bleibt über die nutzbare Batterielebensdauer ohne Nachregelung oder Nullstellung in den genannten Grenzen. Das Instrument ist in explosiver Atmosphäre völlig ungefährlich. Es ist äusserst einfach und betriebssicher und kann vorteilhaft nach einer Operation zur Temperaturmessung auf der Station Verwendung finden.

EE 25 758 *für weitere Einzelheiten*

CAMBRIDGE INSTRUMENT CO LTD

13 Grosvenor Place, London S.W.1

FARBVERDÜNNUNGS-KURVENSCHREIBER

(Abbildung Seite 580)

Das Cambridge Farbverdünnungsgerät wurde zur Aufzeichnung der Konzentrationsänderung eines an bestimmten Stellen in den Blutumlauf eingespritzten Farbstoffs entwickelt. Dieser Wechsel wird entweder durch einen fotoelektrischen Zusatz am Ohr oder durch eine fotoelektrische Küvette entdeckt, durch die z.B. das Blut einer Arterie fliesst. Der Ausgang eines dieser Fühler wird einer Sonderausführung des elektronischen Cambridge-Schreibers zugeführt, in den ein hochempfindlicher, stabiler Gleichstromverstärker eingebaut ist. Durch Betätigung eines Druckknopfschalters wird der Zeitpunkt der Farbstoffeinspritzung auf dem Registrierpapier vermerkt. Sobald der Farbstoff durch das Ohr oder die Küvette geht, bewegt sich der Schreibstylus über das Papier und zeichnet eine Zeitkonzentrationskurve auf.

Das ausgestellte Baumuster II dieses Gerätes hat viele neue Konstruktionsmerkmale, die seine Vielseitigkeit erhöhen, die Bedienung vereinfachen, die Stabilität der Messbasis verbessern, lineare Deutung der Kurven erlauben, selbst bei Verwendung grosser Dosen Coomassic-Blau, und Aufzeichnung einer Reihe von Kurven ermöglichen, von denen jede genau dieselbe Empfindlichkeit wie die vorhergehende hat, trotzdem ein ansteigender Farbenhintergrund vorhanden ist.

Die neue Bedienungstafel ist aus dem Gerät herausgebracht und schräggestellt worden, um die Bedienung zu erleichtern. Ein Schalter, der Verwendung grüner und blauer Farbstoffe mit einem einzelnen Ohransatz oder Küvette ermöglicht, ist darauf vorgesehen.

Da ein hochstabiles Netzgerät die Spannungen für die Lampe im Ohransatz oder der Küvette liefert, wird kein zusätzlicher Akkumulator benötigt. Die Stabilität der Stromquelle ist so hoch, dass Netzschwankungen von unter 190 V bis über 250 V im Schreiber weniger als 2 mm Ablenkung hervorrufen und normale Netzschwankungen keinen Einfluss haben.

Der Ohransatz kann vorgewärmt werden, so dass nach Befestigung am Ohr die kleinstmögliche Verzögerund (5...10 min) auftritt, bevor thermisches Gleichgewicht hergestellt ist.

Die Fotozelle, die in Geräten dieser Art normalerweise 90% der Wärmedrift hervorruft, befindet sich, soweit Drift infrage kommt, nicht in der Schaltung, und dadurch wird eine sehr hohe Stabilität erzielt, ohne dass besondere Vorkehrungen zur Aufrechterhaltung des thermischen Gleichgewichts erforderlich sind. Die Miniaturlampe verbraucht weniger als 1 W, wodurch der Wärmeeinfluss auf das Ohr vernachlässigt werden kann, so dass keine Kühlung nötig ist.

Wenn blauer Farbstoff benutzt wird, schliesst ein Pulssperrkreis zwangsläufig Blutpuls von der Kurve aus, ohne dabei Dämpfung in irgendwelcher Form einzuführen, die die Ansprechgeschwindigkeit auf Farbsignale beeinflussen würde, und ohne Änderung der Belastung durch rote Blutkörperchen, durch die das Gerät eine nicht vorherbestimmbare Empfindlichkeit für Farbstoffe haben würde.

Betriebssicherheit ist für Geräte dieser Art äusserst wichtig, und der Benutzer muss in der Lage sein, jederzeit das Arbeiten des Gerätes zu prüfen und mit der Arbeitsweise des Gerätes, wenn neu, zu vergleichen.

Tests für Schreiberstabilität sind vorgesehen. Die roten und infraroten Zellen können unabhängig voneinander auf Empfindlichkeit und Stabilität geprüft werden. Für viele andere Tests ist Vorkehrung getroffen.

EE 25 759 *für weitere Einzelheiten*

COULTER ELECTRONICS LTD

4 Auriol Mansions, Edith Road, London W.14

BLUTZELLEN-ZÄHLER UND GRÖSSEN-ANALYSATOR

(Abbildung Seite 580)

Das Gerät benutzt ein nicht-optisches Abtastsystem mit einer Zählgeschwindigkeit von über 6000 Einzelzellen pro Sekunde und einem Zählintervall von 15 s. Eine Blutzellensuspension wird gleichzeitig mit einem elektrischen Strom durch eine kleine Öffnung geleitet. Die einzelnen durch die Öffnung gehenden Blutzellen rufen in der Öffnung einen Impedanzwechsel hervor, der von der Zellengrösse bestimmt wird. Das System zählt die einzelnen Zellen und gibt ihre Grössenverteilung. Die Anzahl der pro Probe gezählten Zellen ist ungefähr 100 × grösser als mit der üblichen Mikroskopzählung, so dass der statistische Fehler um einen 10 fachen Faktor verringert wird.

Das Gerät arbeitet wie folgt:—

(1) Mittels einer selbstfüllenden Pipette wird eine kleine Blutprobe entnommen; in einem typischen Fall 2 mm^3.

(2) Die Probe wird in einem Becher mit einem von einer automatischen

Bürette gelieferten Verdünnungsmittel gemischt.

(3) Der Becher wird so auf eine Federplatte gestellt, dass die Röhre mit der Öffnung in die Probe eintaucht.

(4) Die Zählung wird automatisch durch Betätigung eines Sperrhahns und Rückstellung des Zählers eingeleitet.

(5) Die Zählung wird notiert, der Becher entfernt und die nächste Probe auf die Platte gestellt. Falls nicht äusserst grosse Genauigkeit erforderlich ist, braucht die mit der Öffnung versehene Röhre zwischen Zählungen weder ausgewaschen noch gereinigt zu werden.

Das Verfahren kann folgendermassen kurz beschrieben werden:

(a) Wenn der Hahn geöffnet wird, ruft ein externes Vakuum einen Fluss der Probe im Becher durch die 0.1 mm Öffnung hervor und "stört" die Manometereinstellung.

(b) Eine Elektrode im Probenbecher und eine in der mit der Öffnung versehenen Röhre werden benutzt, um einen Strom durch die Öffnung zu leiten.

(c) Eine durch die Öffnung getragene Einzelzelle verdrängt leitende Flüssigkeit und erhöht den elektrischen Widerstand in der Öffnung. Die Widerstandserhöhung ruft einen Spannungsimpuls kurzer Dauer hervor, der im Verhältnis zur Zellengrösse verstärkt wird.

(d) Impulse werden verstärkt und auf einem Oszillographen als klare vertikale Spitzen dargestellt, wobei die relative Höhe der Spitze eine Anzeige der relativen Zellengrösse ist. Die Impulse werden einer Empfindlichkeitsschwellenschaltung zugeführt, die nach Wahl eines Pegels die Impulse zählt, die den Pegel erreichen. Die Empfindlichkeitsschwelle wird durch Aufhellung des Impulsteiles angezeigt, der über der Schwelle liegt.

(e) Schliessen des Hahnes isoliert das System vom internen Vakuum. Der Durchfluss der Probe wird durch die Ansaugwirkung des Manometers aufrecht erhalten. Die steigende Quecksilbersäule berührt "Start"- und "Halt"-Elektroden, die den Digitalzähler in Betrieb bringen, wenn 0.5 mm durch die Öffnung gezogen worden sind.

EE 25 760 *für weitere Einzelheiten*

STANLEY COX LTD

Electrin House, 93-97 New Cavendish Street, London W.1

MONITOR FÜR DAS FÖTALE HERZ

(Abbildung Seite 581)

Das von Bruce Peebles and Co Ltd hergestellte und von Stanley Cox Ltd vertriebene Gerät stützt sich auf Arbeiten des Medical Physics Unit der Universität Edinburg.

Der Geber ist im wesentlichen mechanischer Art und auf Resonanz mit Höchstamplitude bei 57 Hz, der nominell deutlichsten Komponente des Herzschlags des Fötus abgestimmt. Nur wenige Hz von 75 Hz ist die Dämpfung ausserordentlich hoch, so dass sehr wenig unerwünschte Geräusche an den Verstärker gelangen, der seinerseits einen höheren Ton im Bereich leichter Hörbarkeit abgibt.

Das Gerät ist volltransistorisiert mit gedruckten Schaltungen. Die Ausgangsleistung ist für Gebrauch an den meisten Stellen ausreichend und kann selbst auf geräuschvollen Stationen über den Lautsprecher gehört werden. Selbstverständlich kann bei hohen Umgebungsgeräuschen auch der Kopfhörer benutzt werden. Betrieb vom Netz oder eingebauten Trockenbatterien.

Der Geber wird in der besten Position zur Aufnahme der Schwingungen vom Fötus auf den Unterleib der Mutter aufgelegt. Eine aufleuchtende Lampe gibt auf der Vorderseite des Gerätes eine Sichtanzeige für den Herzschlag des Fötus. Höranzeige kann über den Lautsprecher, Lautsprecher und Kopfhörer, oder den Kopfhörer allein erfolgen. Auf Wunsch können Sichtanzeige und Lautsprecher abgeschaltet werden, so dass nur der Kopfhörer angeschlossen bleibt.

Die Empfindlichkeit des Verstärkers ist regelbar, braucht aber nicht oft nachgestellt zu werden. Hör- und Sichtsignale können mit einer Stoppuhr überwacht werden, die bequem in einem besonderen Sicherheitshalter auf der Vorderplatte des Gerätes angebracht werden kann.

EE 25 761 *für weitere Einzelheiten*

ELECTRONIC MACHINE CO LTD

Mayday Road, Thornton Heath, Surrey

ELEKTRO-HERZMONITOR

(Abbildung Seite 581)

In dem kompletten fahrbaren Standgehäuse sind ein Herzmonitor, Schrittmacher und Defibrillator untergebracht, die für Gebrauch auf Wunsch einfach herausgenommen werden können. Das Hauptgehäuse hat eine grosse Schublade zur Aufbewahrung der Zubehörkabel und Elektroden; es ist mit antistatischen Gummilaufrollen ausgerüstet.

Der Herz-Monitor stellt ein Kardiogramm auf dem langnachleuchtenden 127 mm Bildschirm einer Elektronenstrahlröhre dar. Ein eingebauter Schalter stellt die Verbindung mit Zuleitungen 1, 2, 3 und 4 zum einwandfreien Arbeiten für Bein-, Arm- und Brustkardiographie her. Das Gerät ist komplett mit eigenem Hochleistungsverstärker und erfordert keinen getrennten Elektrokardiograph. Ein eingebautes Instrument (z.B. ein Kardiotachometer) zeigt die Herzrate an und kann völlig unabhängig vom Herz-Monitor gesteuert werden. Ein Anschluss für einen Standard- Schreiber ist vorgesehen. Der symmetrische Gegentaktverstärker hat sehr geringe Restwelligkeit und ist geräuscharm; er hat stufenlos regelbare Verstärkung, die für 1 mV Eingangssignal reichlich bemessen ist. Das Netzanschlussgerät ist für Netzschwankungen von ±20 V vollstabilisiert.

Der Defibrillator wurde im Physikalischen Laboratorium des Westminster Hospital in London entwickelt und ist in diesem Krankenhaus und anderen Stellen seit über zwei Jahren regelmässig in klinischem Gebrauch. Von einem durch Druckknopfschalter gesteuerten Schutztransformator werden geregelte Impulse abgegeben, deren Zeitdauer auf 0,2 s eingeregelt ist. Der Westminster Defibrillator hat eine feststehende Ausgangsspannung von 120 V über einen regelbaren Strombereich von 0 ... 5 A. Eine Speicherschaltung misst den Stromspitzenwert während des Impulses und zeigt ihn auf einem Instrument an. Eine Schütze schaltet das Gerät ab, falls durch Zufall Elektroden kurzgeschlossen werden oder das Gerät überlastet wird. Eine Wiedereinstellungslampe leuchtet auf, und das Gerät kann erst wieder nach Betätigung des „Rückstellknopfes" benutzt werden.

Der Schrittmacher wurde mit Hilfe der Post-Graduate School of Medicine in Hammersmith für interne und externe Anwendung entwickelt und hat verschiedene eingebaute Schutzvorrichtungen.

Die Impulsfrequenz liegt zwischen 30...160 Impulse/min. Für interne Anwendung liegt der Spannungsbereich zwischen 20 ... 200 V. Ein Instrument zeigt den durch das Herz gehenden Strom an und gibt auch Warnung, falls während längerem Gebrauch Elektrodenpolarisation auftritt. Eine Glimmlampe zeigt an, dass das Gerät in Betrieb ist.

Der Schrittmacher ist entweder batterie- oder netzbetrieben, d.h. die Batterien werden automatisch angeschlossen, wenn die Verbindung zum Netz oder das Netz selbst ausfällt. Somit kann das Gerät auch im Operationssaal mit Batterien betrieben werden, wodurch die Gefahren freiliegender Kabel vermieden werden.

Anschluss ans Netz schaltet die Batterien automatisch ab. Ein „Batterie-Prüfschalter" gibt am Messgerät eine sofortige Anzeige der Batteriespannung.

EE 25 762 *für weitere Einzelheiten*

FARADAY ELECTRONIC INSTRUMENTS LTD

Blue Town, Sheerness, Kent

N.F.-WELLENANALYSATOR

(Abbildung Seite 581)

Information über die Komponenten

komplexer zeitabhängiger Funktionen kann man durch Aufgliederung der Analyse der verschiedenen Variablen der ursprünglichen Datenaufzeichnungen in verschiedene Frequenzen erhalten.

Die Methode wurde erfolgreich für mehrere Probleme in Industrie und Medizin eingesetzt, besonders zur Analyse der Aufzeichnungen von Elektroencephalogrammen.

Die grundlegende Eigenschaft dieses Gerätes ist die fortlaufende Analyse, die sofort sichtbar auf der ursprünglichen Aufzeichnung in Frequenz-Rechteckdarstellung erscheint. Das zu analysierende Signal wird dem Ausgang eines normalen Verstärkers entnommen und gleichzeitig an eine Reihe abgestimmter Filterkreise gelegt. Die Ausgänge dieser Filter werden integriert und Speicherzellen zugeführt, deren Inhalt durch Schreiber mit grosser Stylusablenkung auf die ursprüngliche Aufzeichnung ausgeschrieben wird. Die Integration wird für Zeitspannen von 10 Sekunden (Epoche genannt) ausgeführt, und während der nächsten 10 Sekunden werden die gespeicherten Daten folgerichtig ausgeschrieben und bilden eine Frequenz-Rechteckdarstellung. Das geschieht mit Hilfe eines Umlaufschalters, der die Speicherkondensatoren hintereinander einschaltet und über einen ungedämpften Stylus entlädt. Üblicherweise wird diese Darstellung auf der ursprünglichen Aufzeichnung der Rohdaten mit 10 s Verzögerung eingeschrieben; zu diesem Zweck werden die Analysator- und Datenaufzeichnungsstyli gestaffelt. Diese Darstellung zeigt eine Anzahl von Spitzen, deren Amplitude die „Ausgiebigkeit" der besonderen Frequenzkomponente in den während der Epoche analysierten Daten angibt. Ausgiebigkeit bezeichnet hier die Summierung der Spannungsamplituden einer bestimmten Frequenzkomponente der ursprünglichen Daten über einen Zeitabschnitt. Der Verstärkungsfaktor der Filterkreise ist so eingestellt, dass die ausgeschriebene Amplitude der Eingangsspannung des Analysators proportional und für alle Frequenzen gleich ist, d.h. 2 V Eingangsspannung für 3 Hz soll dieselbe Aufzeichnungsamplitude für die 3 Hz-Spitze hervorrufen wie 2 V am Eingang für 30 Hz für die 30 Hz-Aufzeichnungsspitze.

Während der Aufzeichnung einer analysierten Epoche läuft die Analyse der eingehenden Daten weiter, und die Integration erfolgt durch Einsatz einer weiteren Bank von Speicherkondensatoren. Nach 10 s ändert sich die Rolle dieser Kondensatoren von Speichern zum Ausschreiben, und die während der Aufzeichnungsperiode benutzten Kondensatoren werden als Integratoren benutzt. Dadurch wird eine laufende Analyse eingehender Daten erreicht.

Eine weitere Bank von Speicherkondensatoren wird zur Speicherung eines Bruchteils jeder Komponente benutzt, um den Durchschnittswert einer Epochenreihe (entweder 4 oder 9) zu berechnen und aufzuzeichnen. Diese Aufzeichnung erfolgt ähnlich wie die der 10 s Analyse am Ende der 40 s oder 90 s Epoche, für die der Durchschnitt ermittelt wird.

EE 25 763 *für weitere Einzelheiten*

C. J. FOX AND SONS (CHEMICAL) LTD

117 Victoria Street, London S.W.1

MIKROWELLEN-DIATHERMIEGERÄT

(*Abbildung Seite 582*)

Das von der Deutschen Elektronik, Berlin hergestellte und von C. J. Fox and Sons vertriebene Mikrowellen-Diathermiegerät ist als Standgerät oder Tischmodell lieferbar. Die max. Ausgangsleistung dieses Gerätes ist 200 W bei 2425 MHz ± MHz. Die Energie wird über ein koaxiales Kabel einer Diploantenne zugeführt, wo sie durch konvexe Spiegel gebündelt und dadurch auf den zu behandelnden Körperteil gerichtet wird.

Fokussierelektroden verschiedener Abmessungen zur Konzentration der Energie auf sehr kleine Flächen und Elektroden für Körperhöhlen sind lieferbar.

Der Hauptvorteil der Mikrowellen-Diathermie ist die Leichtigkeit, mit der die Strahlung auf die gewünschte Stelle fokussiert werden kann und die Abwesenheit von Elektroden, die für K.W.- und U.K.W.-Behandlung am Patienten befestigt werden müssen.

EE 25 764 *für weitere Einzelheiten*

G.H.S. ELECTRONICS LTD

Severalls Avenue, Chesham, Buckinghamshire

ELEKTROMYOGRAPH

(*Abbildung Seite 582*)

Type IIIc ist eine Weiterentwicklung des R.A.F.-Modells Type III. Ein besonderes Konstruktionsmerkmal ist der Kathodenverstärker an einem Verlängerungskabel, mit dessen Hilfe die Elektrodenleitungen gekürzt und Störsignalaufnahme auf ein Minimum verringert werden.

Der Kopfverstärker hat Eingänge mit Umschaltung auf Oberflächen- und Nadelelektroden, die entweder koaxial oder zweipolig zugeführt werden können. In der Schaltung für Oberflächenelektroden können Vakuumelektroden benutzt werden, und ein Anschluss für eine Vakuumpumpe ist vorgesehen.

Der Hauptverstärker besteht aus einem vierstufigen Differential-Verstärker mit ausreichender Verstärkung für eine Höchstempfindlichkeit von 30 μV/cm Ablenkung der Elektronenstrahlröhre. Ein Empfindlichkeitsschalter ermöglicht Einregelung der Verstärkung für Ablenkungsempfindlichkeiten von 100, 300, 1000 und 3000 μV pro cm.

Über eine Extrastufe im Hauptverstärker kann ein Magnetbandgerät angeschlossen werden. Ein Relais kann das Gerät von Aufnahme auf Wiedergabe umschalten, und die aufgezeichneten Signale können entweder auf der Elektronenstrahlröhre sichtbar gemacht oder im Lautsprecher abgehört werden.

Ein ebenfalls mit dem Hauptverstärker verbundenes Röhrenvoltmeter gibt auf einem 63,5 mm Instrument, das neben der Elektronenstrahlröhre eingebaut ist, eine Sichtanzeige des doppelten Scheitelwertes des Eingangssignals. Die Schaltung ist so angeordnet, dass die Anzeige schnell ansteigt und langsam abfällt, um genügend Zeit zum Ablesen zu geben.

EE 25 765 *für weitere Einzelheiten*

THE INFRA RED DEVELOPMENT CO LTD

11 Broadwater Road, Welwyn Garden City, Hertfordshire

MIKROMANOMETER

(*Abbildung Seite 582*)

Dieses Gerät ist ein empfindliches Druckmessgerät, das Differentialgasdrücke bis zu ± 10 Dyn/cm² Vollausschlag herunter messen kann. Dieser Druck entspricht $\pm 0{,}1$ mm Wasserstand. Die Skala ist linear, so dass Drücke von nicht mehr als $\pm 0{,}004$ mm Wassersäule leicht bestimmt werden können.

Der Messkopf besteht aus zwei symmetrisch angeordneten Kammern, die durch eine Metallmembran getrennt sind. Diese Membran bildet mit einer feststehenden Elektrode an jeder Seite zwei Kondensatoren, die einen Teil der Abstimmkapazität für zwei Schwingkreise bilden, die beide auf gleiche Weise zu einem h.f. Oszillator gekoppelt sind. Die Membranbewegung erzeugt gegenüber jeder der benachbarten Elektroden eine Kapazitätsänderung und bringt die Spannungen an den abgestimmten Kreisen aus dem Gleichgewicht. Diese Spannungen werden mittels eines differentiellen Gleichrichtervoltmeters verglichen und die Differenz auf dem direkt in Druck geeichten Messgerät abgelesen.

Die zu messenden Drücke werden durch 3,2 mm Verbindungen über die Frontplatte des Gerätes zum Messkopf geführt. Manometer können für Vollausschlagablesung von Drücken zwischen $\pm 0{,}1$ mm ... ± 245 mm Wassersäule hergestellt werden. Die statischen Betriebsdrücke können bis zu 700 kg/cm² gehen, und Differentialüberlastungen bis zu 10 mal Vollausschlag können ohne Schaden auftreten. Mehrbereichgeräte werden nicht geliefert, es ist aber verhältnismässig einfach, den Messkopf auszuwechseln, woraufhin dann die Empfindlichkeit der elektronischen Schaltung so eingeregelt werden kann, dass die alternativen Köpfe genaue Mehrfache des am Instrument abgelesenen Wertes messen.

Wesentliche externe Verbindungen können ohne Einbusse der Eichgenauigkeit benutzt werden; z.B. können beständige Druckdifferenzen zwischen zwei Punkten durch Schläuche mit 1 mm Innendurchmesser gemessen werden, die auf jeder Seite 6 m lang sind. Das empfindliche Element im Messkopf spricht auf sinusförmige Druckänderungen bis zu 200 Drücken pro Sekunde an, jedoch rufen bei diesen Geschwindigkeiten lange Verbindungen Dämpfung hervor.

Über eine Klinke kann ausser dem eingebauten Messgerät ein Schreiber oder Oszillograph gespeist werden, so dass aus der inhärent hohen Geschwindigkeit des Messystems Nutzen gezogen werden kann. Die Ausgangsspannung über die Klinke ist ungefähr $\pm 0{,}5$ V bei 10 kΩ Belastung. Die Zeitkonstante der Messschaltung kann durch einen Schalter auf 0,3 s erhöht werden, so dass auf Wunsch relativ beständige Ablesungen für um einen Mittelwert schwankende Drücke möglich sind.

EE 25 766 *für weitere Einzelheiten*

GASANALYSATOR

(Abbildung Seite 583)

Dieses Gerät wurde entwickelt, um einen Gasanalysator komplett mit Ausrüstung für Probenehmen und Durchströmungsregelung zu schaffen, der entweder in einem Gehäuse oder als getrennter Baustein geliefert werden kann.

Der Analysator besteht aus drei Bausteinen mit ähnlichen Aussenabmessungen, die entweder übereinander in einem genormten 19 Zoll Gestellgehäuse oder in getrennten Gehäusen für ein oder zwei Bausteine eingebaut sind. Im letzteren Falle können die Bausteine bis zu 3,65 m voneinander getrennt aufgestellt werden. Auf Wunsch kann ein Schreiber eingebaut werden.

Die drei Grundbausteine sind: Ein optisches System einschliesslich infrarote Quelle, Umlaufblende, Analysen- und Bezugsröhren, Nachweisgerät und Vorverstärker. Ein N.F.-Verstärker, der von einem Transformator mit konstanter Spannung gespeist wird, komplett mit allen erforderlichen Verstärkungsreglern, ein in Gaskonzentration geeichtes Anzeigegerät, umschaltbare Zeitkonstantenkreise und Ausgangsklinken zur Verbindung zu geeigneten Registriersystemen. Der dritte Baustein umfasst die Gasbehandlung, die entsprechend dem Verwendungszweck konstruiert wird. In bestimmten Fällen, besonders wenn CO_2 oder H_2O gemessen werden sollen, muss eine Spülanlage einbegriffen werden, um sicherzustellen, dass bestimmte Teile des optischen Systems von diesen Gasen und Dämpfen frei gehalten werden.

EE 25 767 *für weitere Einzelheiten*

LIGHT LABORATORIES

10 Ship Street Gardens, Brighton, Sussex

THERMOMETER

Modell 1 GPD wurde besonders für Hypothermie entwickelt. Der Temperaturbereich ist 26°C...38°C mit einer Genauigkeit von 0,2°C und Anzeige auf einer 152 mm Skala. Messköpfe für Kehlkopf, Rektalhaut und Speiseröhre werden mitgeliefert; diese können einzeln durch Steckverbindungen angeschlossen und momentan umgeschaltet werden. Das Gerät wird vom Wechselstromnetz betrieben. Das Netzgerät ist abgedichtet und funkenfest.

Ein grosses Fertigungsprogramm von elektrischen Thermometern sowohl für Netzanschluss als auch für tragbare, batteriebetriebene Typen steht für einen weiten Verwendungsbereich zur Verfügung.

EE 25 768 *für weitere Einzelheiten*

MEDINTRON LTD

Yeo Street, Bow Common, London E.3

ENDORADIO-SONDE

Die Heidelberger Sonde (Endoradio Sonde zur Messung der P_H-Werte in den Därmen) ist eine Kapsel von ungefähr 10 mm × 5 mm Grösse, die mit einem Kunststoff überzogen ist, der von Magensäften nicht angegriffen wird. Die Kapsel, die vom Patienten verschluckt wird, enthält einen mit einem Transistor bestückten Oszillator, der auf ungefähr 1,8 MHz arbeitet. Die Stromversorgung erfolgt durch eine kleine Batterie in der Radiosonde. Die Batterie wird kurz vor Gebrauch der Kapsel durch Auffüllen mit einem Elektrolyten geladen.

Für die Messung der p_H-Werte des Magensaftes besteht die Radiosonde aus einer p_H-Messzelle mit einer äusseren ringförmigen Elektrode, die während der Messung direkt mit dem Magensaft in Berührung kommt. Die andere Messelektrode liegt in der Sonde und ist von der äusseren Elektrode durch eine halbdurchlässige Membran getrennt.

Jede Änderung des p_H-Wertes des Magensaftes ändert auch die Oszillatorfrequenz. Die Frequenzverschiebung ist der des p_H-Wertes fast proportional und ist für den Bereich $p_H1...p_H8$ ungefähr 120 kHz.

Es ist daher möglich, durch Frequenzmessung den p_H-Wert des Magensaftes zu bestimmen, nachdem die Frequenz der Endoradiosonde durch Eintauchen in normalisierte Lösungen mit genau festgelegtem p_H-Wert geeicht wurde.

Die Ausstrahlungen der Endoradiosonde werden von einer auf den Körper des Patienten gelegten Antenne aufgenommen, einem Überlagerungsempfänger zugeleitet, gleichgerichtet, verstärkt und dann einem direkt in p_H-Werten geeichten Messgerät zugeführt. Der Ausgang kann einem Schreiber zugeführt werden.

EE 25 769 *für weitere Einzelheiten*

MULTITONE ELECTRIC CO LTD

12-20 Underwood Street, London N.1

HERZ-SCHRITTMACHER

(Abbildung Seite 583)

Der Herz-Schrittmacher wird nach einer von der Thoraric Research Unit in Guy's Hospital entwickelten Konstruktion hergestellt. Das Gerät ist tragbar und arbeitet, da es volltransistorisiert ist, mehrere hundert Stunden mit den eingebauten Batterien. Es ist vollständig unabhängig vom Netz, und Reservebatterien sind im Gerät vorgesehen. Ein eingebautes Messgerät überwacht die Batteriespannung, und wenn diese unter einen vorgewählten Wert fällt, wird automatisch und ohne Unterbrechung der Anregung auf die Reservebatterie umgeschaltet. Die verbrauchte Batterie wird dann durch eine neue ersetzt, die wiederum in Reserve bleibt.

Die Rechteckwellen-Ausgangsspannung dauert 2 ms, ist zwischen 1...15 V regelbar und wird direkt durch eine biegsame Drahtelektrode dem Herzkammermuskel zugeführt. Die Reizgeschwindigkeit ist zwischen 60...150 V pro Minute einstellbar, und eine getrennte Spannung ist als Synchronisierungsimpuls für einen Elektrocardiograph vorhanden.

EE 25 770 *für weitere Einzelheiten*

NEW ELECTRONIC PRODUCTS LTD

360 Kennington Road, London S.E.11

PHYSIOLOGISCHER SCHREIBER

Der Gestellschrank Type 1195 kann bis zu sechs Verstärker in jeder Kombination der unten beschriebenen Geräte aufnehmen. Die Verstärker haben identische Abmessungen und eingebaute stabilisierte Netzanschlüsse. Die Geräte sind also für Service austauschbar.

Über dem Gestell können zwei 152 mm Messinstrumente angebracht werden, die den Mitteldruck anzeigen, wenn sie mit den Elektromanometer-Verstärkern verbunden werden. Die Skalen sind in mm Hg geeicht.

Elektrokardiograph-Verstärker

Das Gerät wurde besonders für Elektrokardiographie entwickelt, kann aber auch zur Aufzeichnung anderer Vorgänge wie z.B. Atmung verwendet werden, die Ansprechen auf Gleichstrom erfordern. NF-Phonokardiographen können aufgezeichnet werden und vorgesehene Filter entfernen unerwünschte Wechselstromstörungen.

Phonokardiograph-Verstärker

In diesem Verstärker wird die dreistufige Siebung durchgeführt, die im British Medical Bulletin 1952 (Band 8 Nr. 4 Seite 333-342) beschrieben wurde. Hochwertige Kristallmikrophone mit leicht austauschbaren Kapseln finden Verwendung, und der Ton kann entweder auf einem Oszillographenschirm sichtbar gemacht oder mittels eines Stethoskopwandlers abgehört werden.

Impulsschreiber-Verstärker

Das Zweikanalgerät ermöglicht Aufzeichnung von zwei gleichzeitigen indirekten arteriellen oder venösen Impulswellenformen. Der Ausgang kann entweder direkt in ein 130 Hz-Galvanometer oder in den Gleichstromeingang eines E.C.G. Verstärkers gespeist werden.

Druckmessung

Der Trägerfrequenzverstärker A 644 wurde für Arbeiten mit dem NEP Differential- oder Einfach-Druckgeber entwickelt. Der Ausgang wird üblicherweise direkt zur Aufzeichnung an ein 130 Hz Galvanometer gelegt, während der Mittelwert des Druckes auf einem der grossen Instrumente im Gestell angezeigt wird. Ausspül- und Eichvorrichtungen sind auf einem getrennten Untersatz vorgesehen.

E.E.G. (Einzelkanal)

Im Gestell ist Raum für Batterie-Stromversorgung und Ladegerät vorgesehen, die für den E.E.C.-Vorverstärker A 649 benötigt werden.

Der E.E.G.-Vorverstärker wird in der Nähe des Patienten aufgestellt und ist mit dem Gestell durch ein einziges Kabel verbunden.

EE 25 771 *für weitere Einzelheiten*

OFFICINE GALILEO S.p.A.

c/o O.T.E., Via di Caciolle 15, Firenze, Italy or Leland Instruments Ltd, 145 Grosvenor Road, London S.W.1

VIELFACH-AUFZEICHNUNGSGERÄT

(Abbildung Seite 584)

Dieses Gerät wurde besonders für Geburtenhilfe entwickelt.

Auf dem eingebauten Oszillographen, für den auch eine Kamera geliefert werden kann, können über einen elektronischen Schalter vier Spuren dargestellt werden. Alle Bedienungselemente befinden sich auf der Frontplatte und dem Tisch, auf dem auch ein Schreiber untergebacht ist.

Die Vorverstärker sind in einzelne Metallgehäuse eingebaut und können leicht gegeneinander ausgetauscht werden.

Die gesamte Anlage ist für Einbau folgender Vorverstärker konstruiert:

Ein Verstärkerkanal für Elektrokardiographie der Mutter.

Ein Verstärkerkanal für Elektrokardiographie des Fötus.

Ein Verstärkerkanal für Tonographie der Mutter.

Ein Verstärkerkanal für Tonographie des Fötus.

Ein elektromyographischer Verstärkerkanal zur Untersuchung der Gebärmutterzusammenziehung.

Kanäle sind vorhanden für Einsatz anderer Sonderverstärker, z.B. Encephalographie, Blutdruckuntersuchungen, Lungenuntersuchungen, Druckmessungen, hohen Blutdruck und Rheographie.

EE 25 772 *für weitere Einzelheiten*

OFFNER ELECTRONICS INC.

5320 N. Kedyie Avenue, Chicago, 25, U.S.A.

TRANSISTORISIERTER ELEKTROENCEPHALOGRAPH

Der volltransistorisierte Elektroencephalograph besteht aus zwei Haupteinheiten. Die erste enthält das Bedienungsfeld für Eingangsleitungen, Wählschalter und Eichung, die Vorverstärker und kleine Batteriestromversorgung.

Der zweite Teil enthält die Leistungsverstärker, den Oszillator und Dynograph-Schreiber. In Abmessungen und äusserem Erscheinen sind die beiden Haupteinheiten identisch.

Im Modell I findet ein neues Verstärkungssystem Verwendung, das seine Empfindlichkeit, Stabilität und Betriebssicherheit möglich macht. Das Eingangssignal wird direkt durch einen Zerhacker in ein hochfrequentes Wechselstromsignal umgewandelt. Dieser Zerhacker ist eine geräuscharme Sonderkonstruktion mit langer Lebensdauer: unter normalen Betriebsverhältnissen ist er für 5 Jahre garantiert. Das Wechselstromsignal geht durch einen sechsstufigen Transistor-Vorverstärker, wird elektronisch demoduliert und in einen zweistufigen Transistor-Leistungsverstärker gespeist, der genug Strom zum Betrieb von niederohmigen Dynograph-Schreibern abgibt.

Ein Transistor-Oszillator liefert die für den Betrieb des Zerhackers notwendige 400 Hz-Energie. Alle Hauptbausteine des Modell I sind in Einschubkonstruktion ausgeführt: die Vorverstärker, Leistungsverstärker, Dynographen und der Oszillator. Alle Verstärkerbauelemente ausser Eingangsübertrager, Netztransformatoren und zum Bedienungsfeld gehörige Elemente befinden sich in Einsteckbausteinen.

Da die Vorverstärkerstufen sehr geringe Spannung und Strom benötigen, ist es einfacher, die Vorverstärker aus Batterien zu speisen. Nur sechs kleine Taschenlampenzellen sind für acht komplette Kanäle erforderlich und geben bei normalem Betrieb Strom für über 1000 Stunden. Die Batteriespannung wird durch ein Schalttafelinstrument überwacht.

Jede gewünschte Kombination kann durch Doppeldrehschalter eingestellt werden. Ausserdem können neun vorbestimmte Geschwindigkeiten durch einen Drehschalter gewählt werden.

Sieben geeichte Mikrospannungen sind im Bereich 10...1000 μV mit 2% Genauigkeit vorgesehen. Elektrodenwiderstand wird mittels eines Schalttafelinstruments gemessen, um ihre einwandfreie Befestigung zu prüfen.

EE 25 773 *für weitere Einzelheiten*

S.S. ELECTRONICS LTD

Severalls Avenue, Chesham, Buckinghamshire

REIZGERÄTE

Diese Geräte wurden von der Britischen Luftwaffe für Diagnose und Therapie von Nerven- und Muskelverletzungen in eigenen Krankenhäusern entwickelt und sind jetzt für Zivilbedarf freigegeben.

Zwei Typen sind lieferbar. Die erste ist ein Kombinationsgerät für Behandlung und Diagnose und hat ausser einer grossen Auswahl rechteckwellenförmiger Impulse kurzer Dauer Einrichtungen für Stromstösse kürzerer Impulsdauer, galvanische Stromstösse und einfache galvanische Ausgangsspannung.

Die zweite Type ist ein genaues diagnostisches Reizgerät und liefert eine grössere Auswahl genau bestimmter Rechteckwellenimpulse, die für die umfassende Diagnose von Nerven- und Muskelverletzungen erforderlich ist.

EE 25 774 *für weitere Einzelheiten*

TELCO

45-47, rue de la Division du General Leclerc, Gentilly (Seine), France

TRAGBARER ELEKTROKARDIOGRAPH

(Abbildung Seite 584)

Der Telco 5 ist ein leichter Elektrokardiograph zum Einsatz in der Wohnung des Patienten. Eine Spannungsanzeige und ein Schalter sind zur Anpassung an jedes Netz vorgesehen, um die richtige elektronische Regelung zu bekommen. Die tintenlose Aufzeichnung auf wärmeempfindliches Papier gibt eine scharfe und kontrastreiche Kurve. Der neue Telco Stylusantrieb arbeitet nach dem Tauchspulenprinzip mit den neusten Magnetlegierungen und spricht auf hohe Frequenzen an, hat die richtige Dämpfung unabhängig vom Stylusdruck und ist vollständig frei von Hysterese. Die Kurvenamplitude auf dem 50 mm breiten Registrierpapier kann bis zu 40 mm gehen.

Der Verstärker ist mit Röhren, Transistoren und gedruckten Schaltungen bestückt.

Für Anlauf, Zuleitungswahl (mit automatischer Stabilisierung der Messbasis) und Eichung (Millivolt) wird Druckknopfsteuerung benutzt.

Zwei Vorschübe stehen zur Verfügung mit 25 mm/s (normal) und 2,5 mm/s für Überblicksarbeiten.

Der Antrieb besteht aus einem sehr sorgfältig ausgearbeiteten Baustein, der über einen Synchronmotor konstante Papiergeschwindigkeit gibt. Laden ist sehr einfach: der Beobachtungstisch wird durch Herunterdrücken eines Hebels abgehoben, sodass der Platz für die Papierrolle zugänglich wird.

EE 25 775 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Das Elliott-Sheffer-Stroke ruhende Schaltsystem von P. Kellett

Zusammenfassung des Beitrages auf Seite 534-539

Ein System wird beschrieben, in dem Verbindungen zwischen einer Anzahl logischer Elemente derselben Type (Sheffer-Stroke) jede logische Anordung oder Schaltfolge erlauben. Die einfachen Regeln, die die Verbindungen begrenzen, werden angegeben. Eine Anzahl grundsätzlicher Verbindungen für Allgemeinverwendung werden beschrieben und ein Vergleich zwischen den "Sheffer-Stroke", den "NOR" und den bekannteren "und", "oder", "nein" logischen Elementen angestellt.

Zur Verfügung stehende Leistungsschalter werden beschrieben, die durch den Ausgang eines logischen Elementes gesteuert werden können. Durch die vorgesehene Zusammenschaltung der Verbraucher in eine Matrize können wesentliche Ersparnisse erzielt werden.

Frequenzteilerschaltungen für Musikinstrumente von Alan Douglas

Zusammenfassung des Beitrages auf Seite 546-549

Als Unterscheidungskennzeichen dieser Schaltungen gelten, dass der Teiler immer zwei ist, aus wirtschaftlichen Gründen grösste Einfachheit erforderlich ist, und dass die Wellenform für Dauertonsynthese geeignet sein soll. Die Befriedigung aller dieser Forderungen in einer einfachen Schaltung ist schwierig.

Herabsetzung des Drifteinflusses in einem Gleichspannungs-Analogrechner auf ein Minimum von E. T. Emms und K. H. Brinkmann

Zusammenfassung des Beitrages auf Seite 550-553

Der Einfluss der Drift in Gleichspannungs-Verstärkern auf Gleichspannungs-Analogrechner wird erläutert. Methoden zur Wahl von Zählfaktoren und Verteilung der Verstärkung zwecks Herabsetzung dieser Einflüsse auf ein Minimum werden gegeben.

Dem Systemingenieur ermöglichen formulierte Leitregeln die beste Ausnutzung seiner Geräte.

Einführung in die ternäre Ziffernverschlüsselung von D. J. Morris und W. Alexander

Zusammenfassung des Beitrages auf Seite 554-557

Dieser Beitrag bespricht die über ternäre Arithmetik vorliegende Information und gibt eine Darstellung des ternären Ziffernsystems und seiner Umwandlung in das binäre und dezimale System.

Ein Hall-Multiplikator für HF Verwendung von E. Cohen

Zusammenfassung des Beitrages auf Seite 558-559

Die Beschränkungen des bei niedrigen Frequenzen vorhandenen Hall-Effektes für Verwendung im HF-Gebiet werden kurz diskutiert. Es wird gezeigt, dass für HF-Betrieb der Trägerdurchbruch der entscheidende Faktor ist, der (bei Abwesenheit eines magnetischen Feldes) durch induktive Einstreuung in der Hall-Schaltung hervorgerufen wird, die vom in der Probe fliessenden Trägerstrom verursacht wird. Das kann durch gewisse Schaltungsanordnungen auf ein Minimum reduziert werden und führt in Verbindung mit Einsteck-Breitbandübertragern zu einem grösseren Trägerdurchbruch als 50 dB unter dem Hall-Spitzensignal.

Ein "Dezimal"-Dämpfungsregler; Ein Spannungs- oder Stromteilernetzwerk mit konstantem Widerstand und Dezimaleinstellung von E. R. Wigan

Zusammenfassung des Beitrages auf Seite 560-566

In einem beschriebenen Netzwerk werden die Eigenschaften eines herkömmlichen Strom- oder Spannungsteilers mit denen eines Dämpfungsreglers mit konstantem Widerstand verbunden und die nützlichsten Kennzeichen beibehalten, während es frei von ihren Nachteilen ist. An den Ausgangsklemmen werden genau bekannte Dezimalbruchteile der Eingangsspannung abgegeben; z.B. kann in einem Teiler mit drei Dezimalskalen das Eingangs-/Ausgangsverhältnis zwischen 0 . . . 1 in 1000 Stufen von 0,001 eingestellt werden. Der Teiler hat jederzeit einen absolut konstanten Ausgangswiderstand, d.h. den Ausgangsklemmen kann Strom entnommen werden, ohne dass die Fähigkeit verloren geht, genau bekannte Nachstellungen von Ausgangsspannung oder -strom vorzunehmen, obwohl die Gegenwart der Belastung Spannung oder Strom reduziert, die zur Verfügung stehen.