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Commentary

ACCORDING to the Annual Report issued by the Electronic Engineering Association, 1964 was a most significant period for the Association which represents the major portion of the capital electronic equipment in this country.

During the year a number of governmental reports and White Papers appeared which are likely to produce long-term effects for the industry, and the interest by the present Government in the modernization of Great Britain and the various proposals for the advancement of science and technology leading to the formation of the new Ministry of Technology are welcomed by the Association since the Ministry is to become the sponsoring government department for industry as a whole.

At the same time the association feels that it has not yet been made clear what is the full extent of the Ministry's responsibilities and how these will be carried out, and, in addition, there is a great deal of anxiety concerning the Government's views on military defence, civil aviation, transport and communications, all of which are likely to have considerable effect on the electronic industry.

The aviation section of the industry, in particular, has been disturbed by the Government's actions over the Concord plan, the cancellation of certain military projects and its hesitancy over the TSR2 military aircraft (a hesitancy which has been ended since publication of the Report). This has therefore led to strong action by the Association firmly proposing that where it is necessary to purchase aircraft from abroad, British electronic equipment, should be specified wherever possible and if foreign equipment is to be installed in the foreign built aircraft it should at least be made in this country under licence.

But if there is some doubt in the Association's mind about the future the Association feels that the year under review has been a successful one for the total output of the capital goods sector of the electronic industry was provisionally valued at about £204M, an increase of 32 per cent on the 1963 figure, and of this about £60M worth of goods were exported representing an increase of 22 per cent on the 1963 figures.

At the same time however, there has been a considerable increase in the imports of electronic equipment and while no figure has been given for the value of these imports the capital goods sector of the electronic industry remains a substantial net exporter.

The increase of imported electronic equipment must be expected as the level of business rises but the Association feels that the Government, being the largest single importer of these goods, could do more to moderate this rate of increase by buying more equivalent British equipment, and this feeling cannot be better expressed than by quoting from a passage in the Annual Report itself.

"As far as electronic capital goods are concerned the factor which will have the greatest effect on the industry both in the short term and the long term will be the recognition by H.M. Government and all its departments that the purchase of British electronic equipment is of vital importance to the life blood of the country in this age of technology. Furthermore in the field of development it is not acceptable to contract out of the advanced areas of applied research in electronics and rely on other countries, hoping to catch up in a later phase. This is one of the fields of technology in which Great Britain just cannot afford to lag behind, and while it must be recognized that the investment involved must be very heavy necessitating the joint participation both by Government and Industry in many projects, the prize also will be great not only in technical leadership but in greater productivity at home.

"Above all it could restore our position in the export markets of the world. Electronic equipment itself, with its high content of native soft ware and minimum content of imported materials is particularly appropriate for export."

In short, unless the Government responds to these demands there is the great danger, according to the Chairman of the Association when presenting the Report that "we shall become the industrial serfs of other nations."

The Association has of course been active in the setting up of a number of specialized committees to co-operate with other associations.

For example space technology is a subject of considerable interest and the Association, in conjunction with the Telecommunication Engineering and Manufacturing Association (T.E.M.A.), are co-operating with the General Post Office to ensure that the maximum possible British contribution is made in satellite communication development with the provision of British equipment.

The Association is also giving support to the British independent space programme based on Black Knight for such a programme could undoubtedly stimulate the development of reliable equipment and components in the electronic industry and so enable the industry to occupy a prominent position in the international space programme.

Despite considerable increases in production costs and postal charges the price of ELECTRONIC ENGINEERING has remained 3/- per copy since 1956. In view of further recent increases however we are now compelled to revise our charges as follows:—

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The new charges will come into effect with the June issue, but all subscription orders already placed will be completed at the present rates.
It is regretted that these changes have become necessary.

The LADDIC
in All-Magnetic Logic Circuits

By D. J. Morris*, Dipl. Eng., M.Sc., A.M.I.E.E., A.M.I.E.R.E.

This article describes the application and performance of the 'laddic' in all-magnetic logic circuits where the connexion between the magnetic devices is of wire only. The laddic is used in these circuits as an input device which may perform the AND *operation and also may be used as a d.c. to pulse convertor, (having the input signals in d.c. form and the output in repetitive pulse form).*

An analysis of the flux switching patterns in the laddic is presented and a suggestion of an alternative flux pattern is also given.

(Voir page 354 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 361)

THE extreme reliability of the all-magnetic circuit, as compared with circuits incorporating transistors, diodes, resistors, etc., has directed much attention in recent years to the development of logic circuits consisting solely of ferrite magnetic elements and their connecting wires. The elements so far used for this purpose are the multi-aperture devices (m.a.d.). This article describes the incorporation of the laddic as an additional element in all-magnetic logic circuits.

The laddic, first reported by Gianola and Crowley[1], is a ladder shaped device of square loop ferrite material. This device has many attractive features as a switching element for digital circuits where as a single device it may perform a wide range of combinational logic operations. This device is essentially intended for combinational logic circuits however, for sequential logic circuits, intermediate buffers had to be added between the devices[2].

In all-magnetic circuits, the coupling between the devices, being of wire only, imposes an extremely low impedance load on the output arm of each device. The multiple switching paths within the multi-aperture device can provide the buffer isolation between the input and output circuits and hence eliminate any need for additional buffers. It is possible to heavily load the devices without reflecting the output load to the input circuit. Although the laddic is made of the same ferrite material as multi-aperture devices, it does not exhibit the same feature of isolation of output to the input circuits. This article describes means of obtaining this isolation of the laddic without affecting its basic performance. The low impedance load may be connected to the output leg of the laddic and thus it can be used in coupling it to other devices. However, the input to the laddic must still be applied from a relatively high impedance source and hence the laddic may be used in all-magnetic circuits only as an input element. It is hoped that future research will eliminate this restriction.

Various types of all-magnetic systems have been suggested, the one, described here for the employment of the laddic, is known as the m.a.d.-R system[3]. The name of this system comes from the fact that only multi-aperture devices and connecting wires are used, where the resistance value of the coupling loop is of extreme importance to the operation of the circuit.

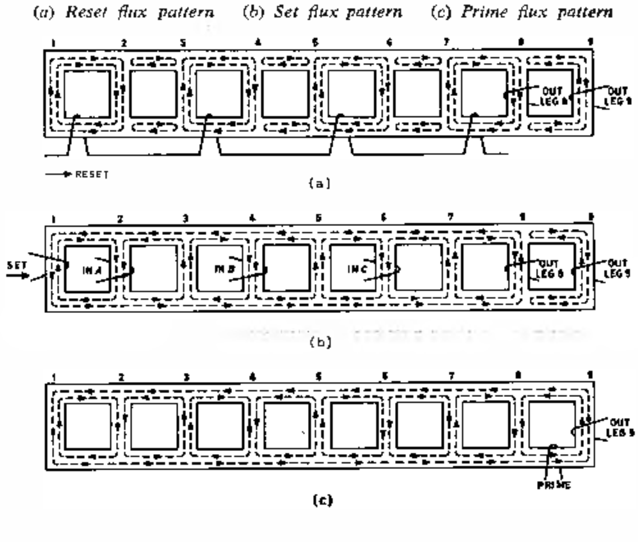
The Basic Laddic Operation

The laddic, although it may have many apertures and be made of the same ferrite material as the multi-aperture devices, is in a class of its own. In multi-aperture devices the input m.m.f. switches flux along selected paths, while in the laddic the input m.m.f. only 'holds' the flux in a specific direction, but does not switch flux.

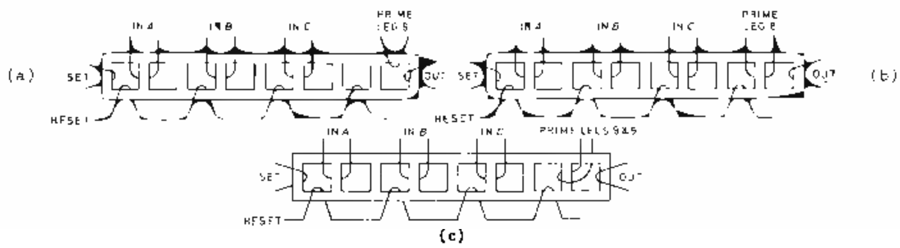
The basic operation of the laddic can best be followed with the aid of the flux patterns shown in Fig. 1(a) and 1(b). The initial state of the laddic is achieved by applying an m.m.f. to the 'reset' winding which switches the flux in a clockwise direction round each aperture, i.e. in an upward direction in the odd numbered legs and a downward direction in the even numbered legs. An m.m.f. applied to the 'set' winding on leg 1 of the laddic, after the laddic was switched to its reset state, will tend to switch the flux in a path along legs 1 and 2, which is the shortest available path. If an input m.m.f. is applied to leg 2, at the same time as the set m.m.f., it will prevent any switching in this leg. The input m.m.f. to the laddic will 'hold' the flux in a downward direction in the leg where it is applied, and redirect the induced 'set' field to switch along another path. Having leg 2 held by the input m.m.f., the induced set field may switch the flux round legs 1 and 4, since this provides the next available path. The flux may not switch in leg 3 since it is already saturated in the direction of the induced field and thus is not 'available' for switching.

If similar to leg 2, legs 4 and 6 are held, the switching due to the set m.m.f., will occur in a path along the two side rails and along legs 1 and 8, giving an output pulse from leg 8. The flux pattern of the laddic, after it

Fig. 1. Flux patterns of the laddic in different switching modes
(*a*) *Reset flux pattern* (*b*) *Set flux pattern* (*c*) *Prime flux pattern*



* *English Electric-Leo-Marconi Computers Ltd.*



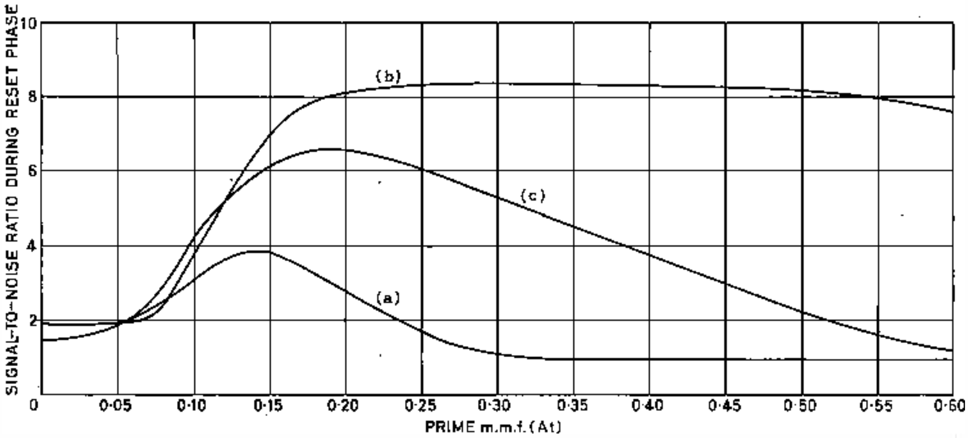


Fig. 2. Output from spare aperture for different prime winding

(a) Prime on leg 9 (b) Prime on leg 8 (c) Prime on legs 8 and 9

has been ' set ', and with legs 2, 4 and 6 held, is shown in Fig. 1(b). The absence of any of the input m.m.f.s will cause switching along a shorter path and no output will be induced from leg 8. Hence, the laddic in this wiring configuration operates as an AND gate.

The output load, imposed on leg 8 of the laddic, must be of a relatively high impedance. A low impedance load on leg 8, will increase the reluctance of this leg and will prevent any flux switching in it. For this reason, transistors must be used at the output of the laddic in order to act as buffers between the following stages and the laddic. In all-magnetic circuits, the load is of the order of a fraction of an ohm, and thus means must be found to prevent this load from damping the operation of the laddic.

Spare Aperture

To achieve input-output buffering-isolation, which is required for all-magnetic circuits, an extra aperture of the laddic is used.

In most of the laddics which are produced, there is a spare aperture which is not required for the AND operation, which provides an extra leg to the laddic. This aperture is added to enable logic operations other than the AND function to be used[1]. When using the laddic for AND operation the spare aperture may be the first or the last aperture. If, for example, the spare aperture is at the beginning of the laddic, in a 9 leg laddic, the set m.m.f. would be applied on leg 2 and the output would be taken from leg 9, which is the final leg of the laddic. Where the spare aperture is at the end of the laddic, as given in Fig. 1, the set m.m.f. is applied on leg 1 and the output is taken from leg 8. The input-hold and reset windings must be moved according to the position of the spare aperture.

The spare aperture takes no part in the switching mechanism of the laddic; and there will be no flux change round the aperture, either during the set or reset switching phases. Although the spare aperture provides an extra flux path, the switching will always be along the shortest available path, (along leg 8 if the input is on leg 1), which is not along the extra leg (leg 9) of the spare aperture. Fig. 1, (a) and (b), show the flux patterns of a laddic with a spare aperture at the final end of the laddic, for set and reset remanent states. An output winding on the extra leg (leg 9) will reveal no output voltage either when the set or reset pulses are applied, and it is presumed that since the flux lines must always be continuous, that the flux pattern in the spare aperture must be similar to that shown in Fig. 1 (a) and (b), which has a net flux of zero in leg 9.. This flux pattern will not change when the laddic is operated in the AND mode and when the output is taken from leg 8.

Initially, the flux in leg 9 may have had a different pattern due to a previous setting. Once the first reset pulse is applied, there may be some flux reorientation in the final leg to preserve flux continuity. Any pulse after the first pulse, set or reset, will not change the flux in the final leg.

The flux pattern in leg 9 of the spare aperture, having the flux in half up and half down direction suggests that it is possible to ' prime ' this aperture; that is, transferring the input information (of upward direction flux) from leg 8 to leg 9 after the laddic had been set. The prime could be by means of pulses or d.c. m.m.f.s applied round the spare aperture, in a direction to switch the flux in a counter clockwise direction round that aperture. This will enable the output to be taken from leg 9 instead of leg 8, the latter being part of the laddic which performs the AND function; and thus it will be possible to obtain the desired isolation of the input circuit which consists of legs 1 to 8, from the output circuit, which consists of the spare aperture. Priming with a d.c., m.m.f. is preferred, thus reducing the need for an extra clock pulse.

The application of the prime m.m.f. on the spare aperture will reverse the flux round that aperture, after the set m.m.f. has been applied, giving the flux pattern shown in Fig. 1(c). After the prime operation, the next reset pulse will switch the laddic back to its initial state, shown in Fig. 1(a), giving an output pulse from leg 9. No switching will occur in leg 8 during this operation, since all the flux in that leg is already in a downward direction.

The prime m.m.f. may be applied to leg 8 or leg 9 since they both form part of the spare aperture. The signal-to-noise ratio of the output pulses from leg 9, during the reset phase, as a function of the prime m.m.f., is shown in Fig. 2. This test was repeated for different wiring patterns of the prime winding; (a) only on leg 8, (b) only on leg 9, (c) shared equally on legs 8 and 9. The prime m.m.f. when applied only on leg 9 must be limited to switch the flux round the spare aperture only, and so only along legs 8 and 9. There is a danger that when the laddic is in its reset remanent state, a large prime field may tend to switch along legs 7 and 9. This can be seen from the narrow range of curve (a) in Fig. 2. With no load on leg 9 and with the prime m.m.f. applied on leg 8, there is no apparent upper limit for the prime m.m.f., as can be seen from curve (b).

The Basic All-Magnetic Circuit

Since, as explained in the previous section, the spare aperture takes no part in the switching function of the laddic, it is possible to have a short turn winding on leg 9 without affecting the basic operation of the laddic. This feature of the spare aperture provides the link required for the application of the laddic in all-magnetic circuits; namely input-output isolation. The output can now be taken from leg 9, which is isolated from the input, instead of leg 8, which is part of the input circuit, while the prime m.m.f. is transferring the information from the input part of the laddic, which performs the AND function, to the output part.

The basic all-magnetic circuit which includes the laddic is shown in Fig. 3. The principle of this circuit is similar to that of the m.a.d.-R shift register[3,4]. The logical functions are performed by the laddic while the second device, shown in Fig. 3, acts as a one clock delay stage.

Two clock pulses, designated drive 1 and drive 2, must be used; drive 1 to set the laddic and drive 2 to reset it, where the input pulses must be coincident with the drive 1 pulses. The input pulses do not switch any flux in the laddic, but only hold the flux in the legs in a given direction and redirect the set field (drive 1) to other legs; hence, the input may also be in a d.c. form. Whatever the form of the input m.m.f., it will have an operational effect only during drive 1 time.

The logic operation of the laddic, shown in Fig. 3, is intended for a three-input AND gate, but it could also be wired to perform other logical functions. Initially, the flux pattern in the laddic is as shown in Fig. 1(a) which results from drive 2 m.m.f. (reset). If all three inputs are applied to legs 2, 4 and 6, then drive 1 (set) applied to leg 1 will be able to switch the flux in leg 8. That is, the flux switching path for a field applied to leg 1, will be along the two side rails of the laddic and also along legs 2 and 8. If only some of the inputs are applied, the return flux switching path due to field induced by drive 1, will be along a shorter path. For example, if no input is applied to leg 4 and inputs are applied to legs 2 and 6, then the flux switching path will be along legs 1 and 4 and along the two side rails between these two legs, with leg 8 being left undisturbed. The flux switching after drive 1 pulse was applied and with all the three inputs holding the three legs, is shown in Fig. 1(b).

During the input phase (drive 1), the low impedance load imposed on the output winding on leg 9, is isolated from the rest of the laddic. After the laddic is in its 'set' state, the prime m.m.f. switches the flux round the output spare aperture, thus transferring the upward direction flux from leg 8 to leg 9, as shown in Fig. 1(c). Since leg 9 is heavily loaded by the coupling loop, the prime operation will be relatively slow. The duration of the prime operation will depend on the loop wire resistance. The maximum recorded speed of operation for a full logic cycle is 10kc/s.

Drive 2 (reset), when applied after the prime operation in order to reset the laddic, will switch back the flux in leg 9 to a downward direction, thus causing a voltage pulse to be induced in the coupling loop. This causes a current pulse to flow in the coupling loop in a direction to set the following device. In order to achieve efficient energy transfer between the laddic and the following device, the coupling loop resistance must be very small.

An output voltage pulse is induced in the coupling loop also during the prime operation (of opposite polarity) when switching the flux in leg 9 to an upward direction. Since this output voltage will cause a current to flow, in the coupling loop, in a direction to block the following device (which is already in its blocked state) it will have no effect on the operation of the circuit.

The coupling loop wire resistance which must be of the order of a fraction of an ohm, is obtained from the resistance of the wire itself with no external resistance elements required. Full considerations of the loop wire resistance has been previously discussed for the m.a.d.-R shift register[4].

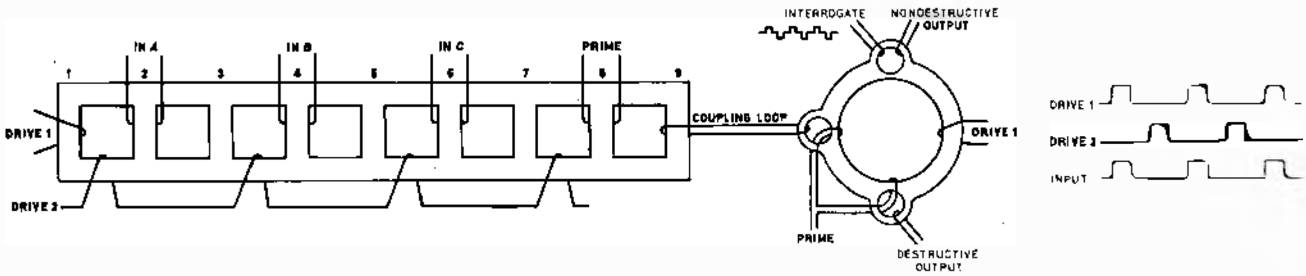


Fig. 3. The basic all-magnetic circuit

The second device, shown in Fig. 3, may be interrogated by an alternating m.m.f., thus giving non-destructive read-out of the stored information. This device may be connected to other devices, using wire only, for further manipulation of the logic information.

Means have to be provided to ensure that information, previously transferred to the second device, does not flow back to the laddic when the second device is reset by drive 1. Two methods could be used to prevent back-flow of information in all-magnetic circuits[4]; one method is to 'hold' the output leg of the device in question and prevent it from switching, while the following device is driven to its 'reset' state. The other method is to prime the input aperture of the second device after it has switched to its 'set' state.

In the basic all-magnetic circuit, shown in Fig. 3, the prime method must be used. That is, after the information is transferred from the laddic, the input aperture of the second device must be primed. It is impossible to use the 'hold' method in preventing back flow of information, since the flux in leg 9 is in both directions, thus an m.m.f. applied to leg 9 during drive 1 will switch the flux round the output aperture into a flux pattern other than that required.

Although the transfer of information from the laddic

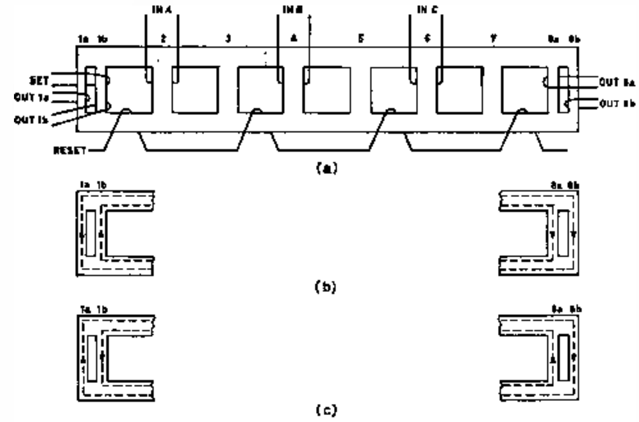


Fig. 4. Flux patterns in input and output legs
(a) Wiring diagram of test (b) Reset flux pattern (c) Set flux pattern

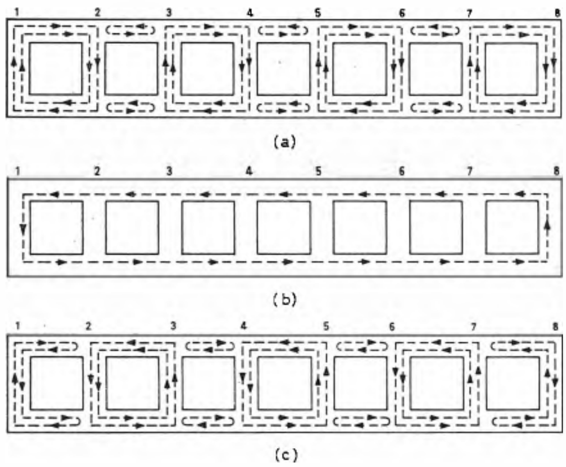


Fig. 5. Suggested flux patterns of the laddic
(a) Reset flux pattern (b) Flux switching path during set phase with all the inputs applied (c) Set flux pattern

to other devices is possible, the use of the prime method in preventing back-flow of information will reduce the efficiency of this circuit. (It has been shown elesewhere[4] that priming two apertures of one device reduces the range of operation).

The disadvantage of the laddic in the basic all-magnetic circuit, i.e. the prime method used for preventing back-flow of information, stimulated thoughts to be directed into finding alternative means of using the laddic in all-magnetic circuits. Furthermore, the laddics available at the initial stages of the investigation gave poor signal-to-noise ratio of the output pulses and had narrow operating tolerances. A new approach to the all-magnetic circuits was made possible by the introduction of a new type laddic, having split input and output legs. Before the modified all-magnetic circuit may be discussed, the flux patterns in the laddic must be reconsidered.

The Laddic Flux Patterns

The original flux pattern model of the laddic was suggested by Gianola and Crowley[1], which in principle is similar to that shown in Fig. 1. These flux patterns seemed to be confirmed in all the initial investigations. But further investigation, made possible by the availability of new laddic shapes, having split input and output legs, indicated that these flux pattern models need reversing. There are still a number of unknown factors in the laddic which make it difficult to assess fully the actual flux pattern within the device. Nevertheless, it is intended here to suggest a modified model of flux patterns which could assist in understanding the performance of the modified laddic.

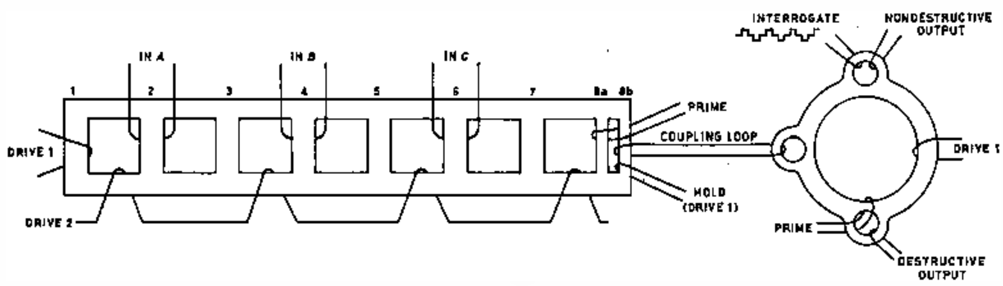
In ordinary AND operations, the output is taken from leg 8 of the laddic as shown in Fig. 1. There are two flux paths in each leg and in the side rails of the laddic, which suggest the possibility of split legs. In leg 8 only one switching path takes an active part in the performance of the laddic. According to the pattern given in Fig. 1, the active path is the outer half of leg 8; but further tests carried with a split output leg prove this flux pattern to be incorrect.

The wiring diagram of the modified shaped laddic which has split input and output legs, is shown in Fig. 4(a). The circuit was wired in a similar manner to that given in Fig. 3, having the set, reset and input windings on the appropriate legs. The set winding was round both halves of leg 1 to give the same operational effect as before. Output windings were assembled on each half of the input and output legs.

An output voltage pulse was obtained, when setting and resetting the laddic, from leg 8a and not from leg 8b. This result could have been anticipated, on the basis that leg 8a provides the shortest available switching path for the set field induced in leg 1. In the same manner, the flux switched in leg 1b and not leg 1a. Although it may be claimed that leg 1b provides the shortest available path, this selected path is not so obvious, since at the same time as the 'set', the input to leg 2 is also applied. When the set field is induced in leg 1 and the input field in leg 2, the field in leg 2 should have some effect on leg 1. According to Fig. 1(b), the field in leg 2 holds the non-active flux path in leg 1 in an upward direction. The results obtained with a split leg 1 have proved this incorrect. In order to eliminate any side effects which might influence the switching of flux in this experiment, the set and input m.m.f.s were applied in series, with equal numbers of turns in both windings. From these results it is possible to postulate the flux patterns within the input and output legs, and these are given in Fig. 4(b) and (c). It can be seen that the flux switches in the inner half of legs 1 and 8.

The flux patterns in the centre of the laddic are of a complex nature, since two opposing fields operate at the same time; the set and input fields. If it

Fig. 6. The modified all-magnetic circuit



is accepted that the set field switches the flux along the shortest path as shown in Fig. 5(b), it may then be asked what part the input fields play in the selection of switching paths. The induced input fields do not cause any flux switching within the laddic, but hold the flux in legs 2, 4 and 6 in a downward direction and redirect the set field to other legs. At the junction of legs 2, 4 and 6 with the side rails, two opposing forces are in operation. The influence of these two forces on the flux switching performance is yet unknown, and could only be experimentally investigated if each leg and side rail were split to enable the assembly of additional windings. Since no such laddic is available, one can only conjecture the actual switching paths. A suggestion of the flux patterns in the whole of a laddic, having 8 legs and performing a three input AND operation, is given in Fig. 5.

The Modified All-Magnetic Circuit

The two flux switching paths in the output leg could be utilized for all-magnetic applications. By splitting leg 8, the output circuit of the laddic will be similar to other multi-aperture devices since it can provide input to output isolation, and thus a low impedance load may be applied to the output circuit without affecting the laddic operation. Furthermore, there is no need for the 'spare' aperture or the use of the prime method in preventing back-flow of information.

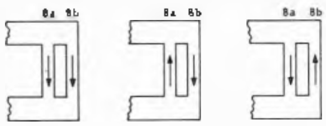


Fig. 7. Flux patterns in the split output leg
(a) Reset state (b) Set state (c) Primed state

The wiring diagram of the modified all-magnetic circuit is shown in Fig. 6. In order to understand the operation of this circuit, only the flux direction in the split output legs are relevant, and these are given in Fig. 7. Initially, in the reset state, the flux in both halves of leg 8 are in the downward direction. With the three inputs present, the set m.m.f. (drive 1) will switch the flux in leg 8a in an upward direction. Since the load is applied to leg 8b, this load is isolated from the rest of the laddic and will not affect the required magnitude of the set m.m.f. The prime m.m.f. will switch the flux round both halves of leg 8, thus transferring the upward direction flux from leg 8a to leg 8b; and since leg 8b is heavily loaded this operation will be relatively slow. The performance of the split leg is similar to the minor output aperture of the multi-aperture device[4] and hence the best range of operation is obtained with the prime winding shared on both halves of leg 8. The prime m.m.f. may be in d.c. form, since it can only cause switching after the laddic is fully set (Fig. 7).

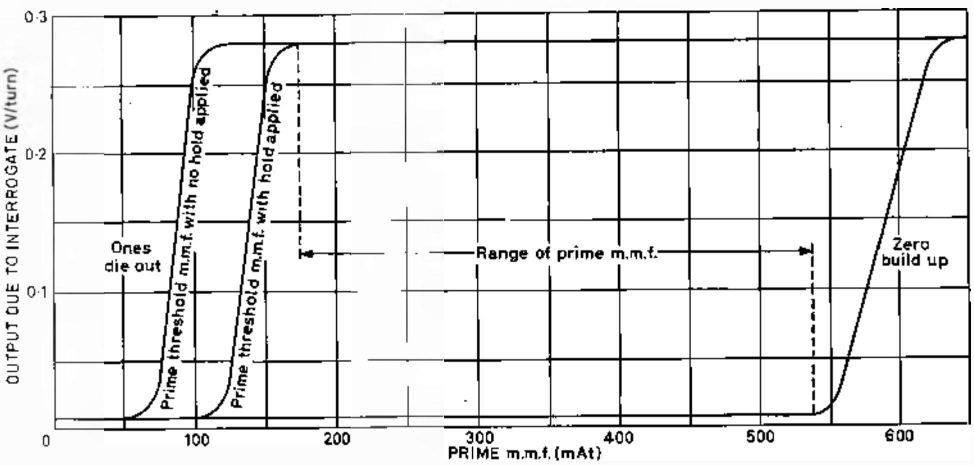


Fig. 8. Range of prime m.m.f. in the laddic

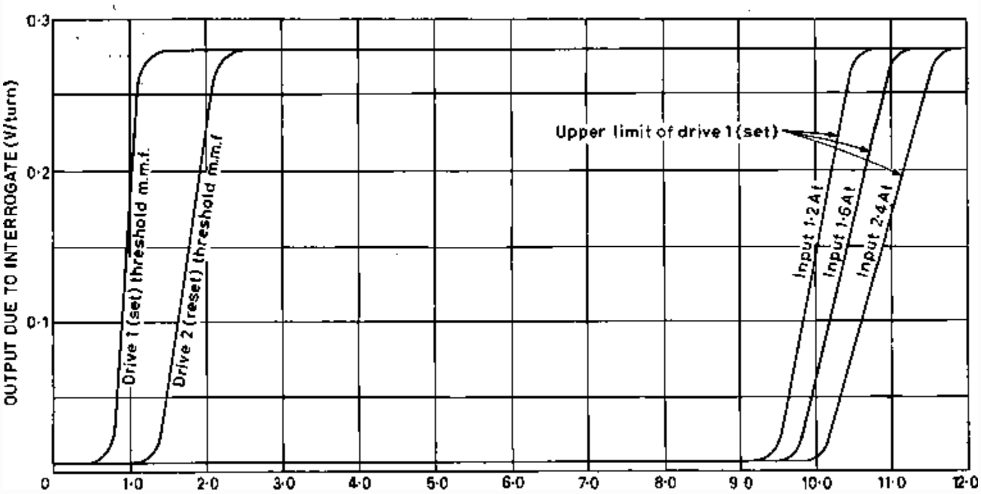


Fig. 9. Range of drive m.m.f.'s in the laddic

In order to prevent backflow of information, the flux in leg 8b may be held in a downward direction while the following device is driven. In the circuit shown in Fig. 6, the laddic is held at drive 1 time, which is also the time the laddic is set. This will not affect the direction of flux in leg 8b, since both in the set and reset states the flux is already in a downward direction, as shown in Fig. 4. Because of this, the prime operation cannot start functioning while the hold is applied, and thus it is delayed to occur between drive 1 and drive 2 pulses.

The range of the prime m.m.f. shown in Fig. 8, is wider than in a multi-aperture device, since there are longer switching paths in the laddic and therefore less danger of a large prime m.m.f. switching in unwanted flux paths. The curves of Fig. 8 were obtained by measuring the amplitude of the non-destructive output pulses from the second device shown in Fig. 6. All the applied m.m.f.'s were kept constant while varying the prime m.m.f. If the prime m.m.f. is below the threshold m.m.f. required for switching the material in the split leg 8, a one installed in the laddic will be transmitted as a zero. If, on the other hand, the prime m.m.f. is too large, it may switch flux between leg 8b and leg 7, during the reset state, and thus may cause a zero to be transmitted as a one.

The lower limit of the prime m.m.f. will depend on whether the 'hold' field is applied, as shown in Fig. 8, since the hold field will delay the prime operation.

The range of the drive m.m.f.'s applied to the laddic, is shown in Fig. 9, which was obtained in a similar manner to that of Fig. 8. The upper limit of the set m.m.f.

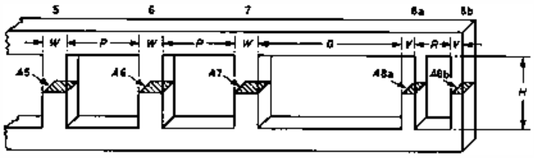


Fig. 10. The laddic dimensions

depends on the magnitude of the input m.m.f.'s. (The relation of the set and input m.m.f.'s was discussed by Gianola and Crowley).

The Laddic Dimensions and Configuration

It was shown in the previous section that splitting the output leg of the laddic enables it to be used in all-magnetic circuits. This split of the output leg also improves the performance of the laddic, since each half now accommodates only one switching path.

In order to ensure that there is only one switching path situated in each half of the split leg, it is essential for the cross-sectional area of each half to be equal to half the cross-sectional area of any other leg; and the cross-sectional area of all the other legs must be equal. These cross-sectional areas are shown schematically in Fig. 10.

$$A_{8a} = A_{8b}$$

$$A_5 = A_6 = A_7$$

$$A_7 = 2A_{8a}$$

Further improvements of the laddic performance could be achieved by increasing the distance between the two final legs (7 and 8), as shown in Fig. 10. The worst noise output is created by an input signal on the leg nearest to the output leg; i.e. by the input applied to leg 6. By increasing the distance between the two final legs the noise in the output leg is reduced. Furthermore, by splitting the output leg, the noise output pulses will be concentrated in the inner half of the split leg (8a) with little noise output from the outer half of this leg (8b). The distance between legs 7 and 8 must not be increased indiscriminately, since increasing this distance also increases the magnitude of the set and reset m.m.f.'s.

The distance between the two halves of the split leg should be kept short so that the prime m.m.f. will be as small as possible. On the other hand, this distance must be large enough physically to accommodate comfortably all the windings.

The range of the prime m.m.f. will depend on the ratio of distances of the final legs 7 and 8. The flux switches along the shortest available path due to applied m.m.f., $F = \oint H_c dl$. Since the coercive force H_c is a constant for the material, the applied m.m.f. will be a direct function of the path length $\oint dl$. The minimum path length for the prime field to cause switching round the split leg will be $2(H + R)$, as can be seen from Fig. 10. The path length required to saturate fully the material in the split leg will be $2(H + R + 2V)$. The prime m.m.f. must be limited so as not to switch flux between leg 7 during the reset state. Thus the danger path length will be $2(H + R + V + Q)$. From this it is possible to define the operational range of the prime m.m.f. which is the ratio of the last two path lengths $(H+R+V+Q)/(H+R+2V)$. Since the prime winding is shared equally between the two halves of leg 8, leg 8b is subjected (when the laddic is in its reset state) to only half the prime m.m.f. This will reduce, by a factor of 2, the danger of unwanted switching and the range of the prime m.m.f. is therefore a function of $2(H+R+V+Q)/(H+R+2V)$.

The use of the input split leg where the set m.m.f. is applied has very little advantages over the unsplit input leg for the applications described here. For that reason, the split input leg is not shown in the circuit given in Fig. 6. Furthermore, the use of the split input leg imposes practical difficulties when not all the legs are required. For example, for a two input AND gate, two dummy inputs would have to be used if only a 10 leg laddic is available.

The information processed in the laddic may be read non-destructively by interrogating the following device as is shown in Fig. 6. It is possible to interrogate the laddic itself by switching the flux between the two halves of leg 8 in a manner similar to the prime operation. However, the split output leg (leg 8) may be used for one function only, i.e. either for destructive read-out, as in all-magnetic circuits, or for non-destructive read-out. In order to achieve both functions from the same laddic, the split leg must be modified. This is done by dividing the space between the two halves of the split output legs into two small windows, as shown in Fig. 11, having the cross-sectional area between the two windows at least equal to twice the cross-sectional area of the side rungs. This is in order to enable each one of these windows to function separately. However, the operational range of the prime and the interrogate m.m.f.'s is thereby considerably reduced; and for this reason, the configuration is not very practicable, although future modifications may improve it.

The Applications of the Laddic in All-Magnetic Circuits

The inputs to the laddic must come from a relatively high impedance source. The input m.m.f.'s do not switch any flux in their respective legs, but hold the flux in these

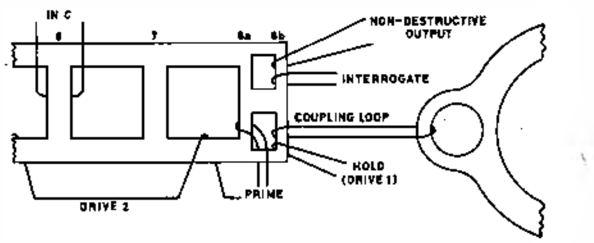


Fig. 11. Non-destructive read-out from laddic in the all-magnetic circuit

legs in a given direction and prevent any switching in them. If the inputs were applied from a low impedance source it would prevent flux switching occurring in the legs even when no pulse currents flow in the input windings. Given sufficient time, switching can take place even when a low impedance load is imposed on these legs; but, in this specific case, the time in question is the pulse duration of the set pulse, which is relatively short. For this reason, it is difficult to use the logical inputs to the laddic in all-magnetic circuits. The only place where a

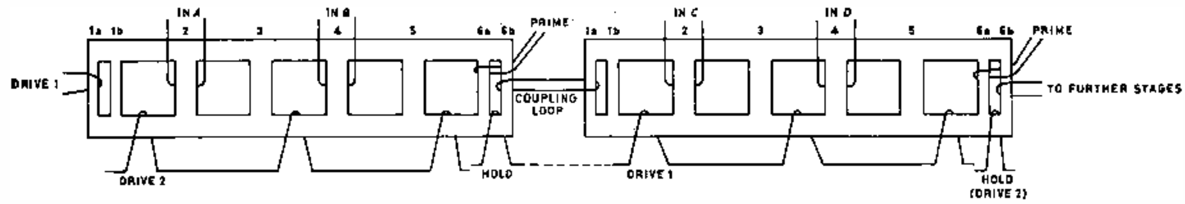


Fig. 12. Cascading laddics in all-magnetic circuits

low impedance load will not affect the operation of the laddic is when it is connected to the output rung of the final leg.

To date, the main function of the laddic when used in an all-magnetic circuit is to act as an input device. That is, where the information is fed into the laddic from a relatively high impedance source and then transferred from the laddic to other devices, using only the connecting wires, for further processing. The laddic itself may perform some logical operations, but only combinational functions of the first sequential stage. In the circuits shown in Figs. 3 and 6 the laddic performs the AND function, although many other combinational functions may also be performed by the laddic, such as OR-AND-INHIBIT.

One of the most important applications of the laddic when it operates as an input device for an all-magnetic circuit, is as a d.c. to pulse convertor; i.e. as an AND function where one of its inputs is a clock pulse. In this application, the inputs are applied as d.c. and the state of these inputs is checked at every reset pulse time. This information, obtained from the laddic, is then transmitted for further processing. An example of such an application is a fail-safe circuit, where the inputs are supplied from various measuring equipments, giving a d.c. input for a safe reading and no input for an unsafe reading. The laddic performs the operation of a scanner, checking the state of these inputs at the frequency of the clock pulses.

Where the inputs to the laddic are in the form of direct currents, it is possible to use the prime as an extra input source (for the AND operation), since the absence of the prime current prevents any output pulse from being fed out.

One possible application of the split input leg could be in cascading laddics in all-magnetic circuits, as shown in Fig. 12. The input split leg isolates the set and the input switching circuits. This circuit may be used when the number of combinational variables is larger than can be accomplished on a single laddic.

Conclusions

The two main incentives for the work on all-magnetic circuits are the high-reliability factor of the magnetic devices and the absence of any associated components, other than connecting wires. The laddic, being a magnetic device, may be used in all-magnetic circuits as an input element having its output connected, by wire only, to other devices.

One of the main applications of the all-magnetic circuits is in 'fail-safe' usage. An example of such an application may be in reactor trip control in atomic power stations.

The tolerances of the driving currents, applied to these transistorless circuits, are extremely wide. The main limitation for this device, when used in all-magnetic circuits, is the slow speed of operation, the maximum speed being of the order of 10kc/s. The slow speed is due to the low impedance imposed by the coupling loop on the laddic, without it, the laddic may switch at much faster speeds.

At this stage, only the laddic output may be connected by the low impedance coupling loop. It is hoped that future development will enable the laddic inputs to be also coupled by the wire system, which will allow the laddic to be fully exploited.

Acknowledgments

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A Single Sideband Automatic Communications System

A fully automatic remotely controlled, high frequency single sideband communications system for operation in the frequency range of 2Mc/s to 30Mc/s has recently been developed by Racal Electronics Ltd of Bracknell, Berkshire.

Known as 'Speedrace', this communications system is available in static, transportable, or mobile form with power output up to 10kW.

The heart of this system is the type MA250 synthesizer which accepts a standard signal frequency, either internal or external, and produces 280 000 possible operating frequencies separated by 100c/s, each operating frequency being controlled by a frequency standard with an accuracy of better than 1 part in 10^9 and a stability of better than 5 parts in 10^{10} per day.

All tuning, frequency-setting and mode-setting is completely automatic and can be remotely controlled so as to permit the operation of unattended stations.

The average tuning time for a transmitter system is about 20 seconds.

Various types of emission can be employed, including single sideband or independent sidebands with either 3kc/s or 6kc/s bandwidths giving one or two speech channels per sideband. Compatible amplitude modulation, c.w., and frequency shift telegraphy can also be used.

A Tachometer Using Circular Diffraction Gratings

By B. V. Jayawant*, Ph.D., B.Sc., A.M.I.E.E., and D. P. Rea*, B.Sc.

The instrument described utilizes two circular diffraction gratings with their axes accurately centred to provide a signal, the frequency of which is proportional to the speed of the machine to which the gratings are coupled. By a sample measurement of frequency over a small period of time, a continuous trace may be displayed on an oscilloscope, the level of which is proportional to frequency, and hence to speed. Since the measurement of speed is rapid, the instrument can follow and record the transient behaviour of a machine, enabling transients to be studied directly.

(Voir page 354 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 361)

THE application of diffraction gratings making use of Moiré fringes[1,2,3] provides an extremely versatile measuring method of great accuracy and is well known especially in the control of machine tools. Figs. 1(a) and 1(b) illustrate the production of Moiré fringes by linear and circular diffraction gratings. The subject of Moiré fringes in machine tool control is dealt with quite comprehensively elsewhere[1]. The advantage of the use of Moiré fringes is that since they are produced by very many accurately spaced grating lines great accuracy of measurement is obtainable. Also in the case of measurement of linear or angular displacements the effect of any remaining errors of spacing of individual lines is greatly reduced by the averaging process which takes place in the production of the fringes.

The circular diffraction gratings are readily adaptable

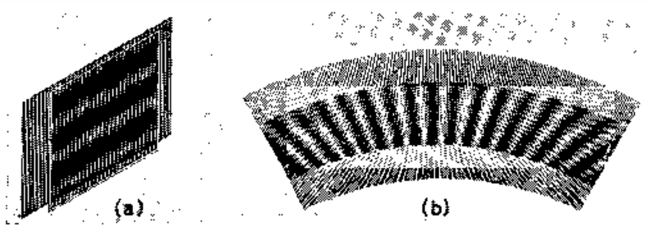

Fig. 1(a). Moiré fringes with linear gratings
(b). Moiré fringes with radial gratings

to a continuous measurement of angular velocity in the same manner as a gear wheel with a tooth-sensing device. Only the inaccuracy of individual lines of circular gratings will not be reflected to the same extent as the spacing accuracy of the gear teeth. The problem of adapting a commercially made tachometer worked from a gear tooth sensing device to circular diffraction gratings would probably be a trivial one. All these tachometers appear to require, however, for a steady state measurement of speed a certain period of constant speed running before the measurement of the speed can be displayed. It was, therefore, decided to construct a fast acting tachometer using circular diffraction gratings for measurement of speed transients in the range from 0 to 4 000 rev/min and this instrument is described here.

Principle and Specification of the Instrument

Continuous measurement of angular velocity can be carried out in two ways:

(a) Measuring the angle turned through in fixed time intervals, i.e. if the angle turned through in a time Δt is $\Delta\theta$ then over the n^{th} time interval of measurement the angular velocity is given by $\omega_n = \Delta\theta_n/\Delta t$.

(b) Measuring the time intervals for fixed angular displacements so that over the n^{th} angular displacement the angular velocity is $\omega_n = \Delta\theta/\Delta t_n$.

In the above cases $\Delta\theta_n/\Delta t_n \neq \Delta\theta_{n-1}/\Delta t_{n-1}$ only if the speed is not constant. Also in both cases angular displacements can be measured by, say, a pulse for each $\Delta\theta_n$. In the second case, therefore, measurement of angular velocity would consist of measurement of time elapsed between a given number of pulses and displaying the time intervals as a voltage from one sampling period to the next. In theory this sounds relatively simple but the difficulties of

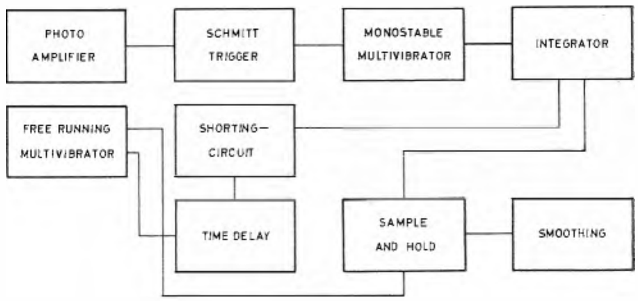

Fig. 2. Arrangement of the tachometer

converting varying time intervals to voltages rapidly and obtaining voltages from these proportional to speed did not prove to be entirely surmountable. The instrument described here is therefore based on the first method, i.e. measurement of angular displacements, i.e. the number of pulses generated in a given interval of time. If there is a reasonable length of time over which the pulse repetition frequency remains constant then again the problem becomes trivial. Where, however, the speed transients are so rapid as to not only affect the p.r.f. but successive pulse widths as well, the problem starts to become interesting.

The criteria set for the instrument were, (a) linearity between 0 to 4 000 rev/min, (b) sensitivity up to changes in speed of 20 rev/min and (c) ability to follow a speed transient of 0 to 3 000 rev/min in 0·1sec. This last criterion necessitates that the cut-off frequency of the instrument must not be less than 2·5c/s.

Layout

The layout of the instrument is shown in the diagram of Fig. 2. The initiating pulse train is generated by the circular diffraction gratings, one fixed and the other coupled to the drive (each diffraction grating having 120

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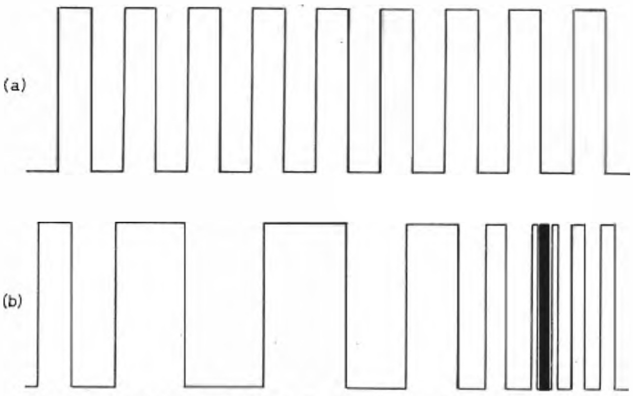


Fig. 3(a). Pulse train at constant speed
(b). Pulse train at varying speed

dark and 120 bright sectors) and a photocell recording the dark and bright periods as the drive shaft turns the movable grating through every $\Delta\theta = 1{\cdot}5°$. The photocell pulses for a fluctuating speed drive and a constant speed are shown in Fig. 3. It may be noted that regardless of the width of individual pulses they represent the same angle turned through by the movable grating. These photocell pulses are amplified and used to actuate a Schmitt trigger. The pulses from the Schmitt trigger are converted to equi-length pulses by the monostable multivibrator and are fed into an integrator. The integrating capacitor is shorted out every $1/75^{th}$ of a second and consequently the voltage (staircase) waveform on the integrating capacitor at the end of two $1/75^{th}$ second intervals would be as illustrated in Fig. 4. Just before the capacitor is shorted out the voltage on it is sampled and held. This necessitates a time delay between sample and hold and shorting the integrating capacitor. The free running multivibrator determines the sampling period and could be a more refined oscillator. The output from the sample and hold circuit is smoothed out to eliminate the 75c/s ripple, which of course introduces attenuation at the top end of the range of the instrument.

Complete Circuit

The complete circuit is given in Fig. 5. The circuit is designed to operate between −9V and +9V.

Photo-amplifier

This is a conventional amplifier incorporating a photo transistor OCP71.

Schmitt Trigger

The trigger level is set at approximately −5V so as to eliminate the effect of spurious signals.

Monostable Multivibrator

The time duration of the equi-length pulses has to be substantially less than the duration of pulses actuating it at the top end of the speed range. For the given diffraction gratings and intended speed range it was decided to make it of 75μsec duration which allows interference free operation up to 13·3kc/s which corresponds to a rotational speed greater than 6 000 rev/min.

Integrator

This is an emitter-follower preceding a high gain stage incorporating an npn transistor. The 50kΩ potentiometer is included to offset drift and the slight negative bias due to the monostable multivibrator. It also acts as a zero setting control. A certain amount of leakage of the feed-back capacitor through the shorting transistor is inevitable and though this does not affect the required linear operation it puts a limit on the output voltage range. The value of the capacitor is a compromise between making it as large as possible to counteract this and small enough to prevent the feed-in resistance being too low or introducing unwanted lags in the sample and hold circuit.

Free Running Multivibrator

This acts as the master oscillator and determines the pulse counting intervals. The frequency of 75c/s was arrived at from two considerations: (a) definition, i.e. the number of steps required to show a trace with sufficient accuracy, and (b) sensitivity of the instrument, i.e. reaction to small speed changes. Obviously the first consideration would dictate a large number of steps such as those of Fig. 4(c), i.e. small time interval. The criterion of sensitivity, however, requires as equal a number of pulses per counting interval when running at constant speed as possible or if not the effect of any departure from an averaged mean should be insignificant to avoid the output displayed on an oscilloscope as a trace from 'dithering'. This condition is satisfied by a large counting interval in relation to the p.r.f. of the astable multivibrator actuated by Schmitt trigger. The choice of $1/75^{th}$ of a second interval allows a sensitivity to a speed change of 36 rev/min and with smoothing it can be made as sensitive as 18 rev/min.

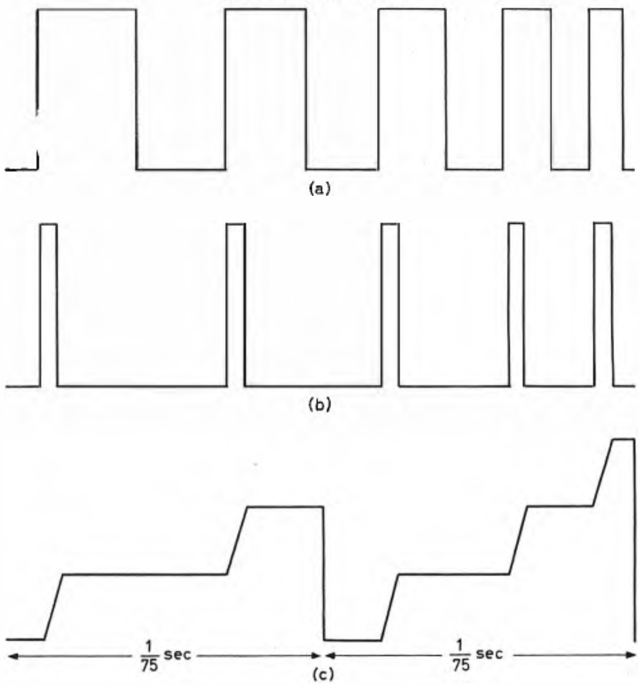
Sample and Hold

This is a straightforward circuit relying on the pulses from the multivibrator acting as switching pulses to enable the sample and hold transistor to acquire the voltage on the integrating capacitor.

Time Delay and Shorting Circuit

These are simple extensions of the sample and hold circuit. The delay between sampling the integrator capacitor and the shorting is about 25μsec. The time for which the capacitor is shorted out is 2msec. If it were

Fig. 4(a). Pulse train
(b). Output from monostable multivibrator
(c). Integrator output



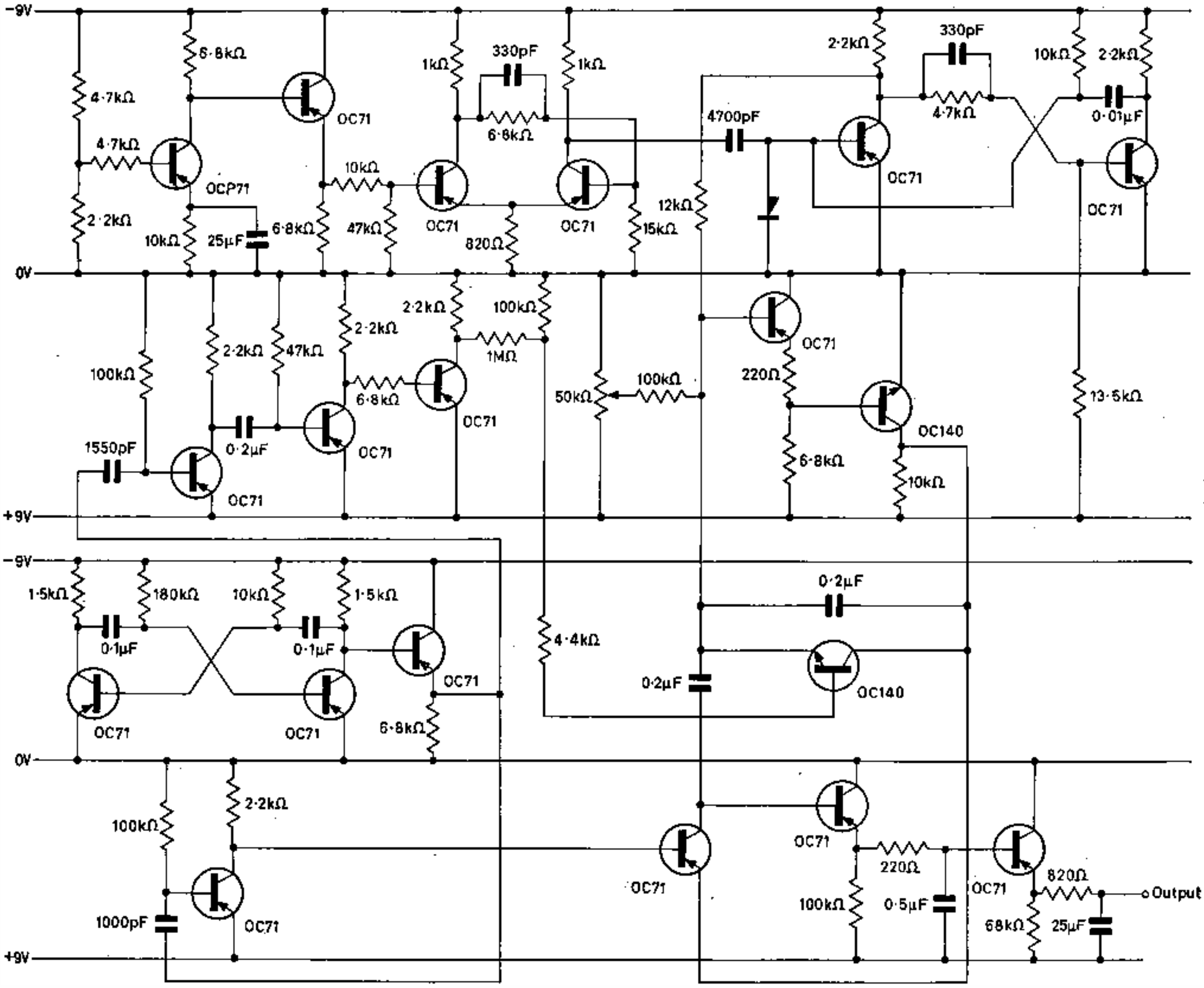


Fig. 5. Complete circuit diagram

possible to decrease the size of the capacitor ($0{\cdot}2\mu F$) the shorting time could be further reduced.

SMOOTHING CIRCUIT

This consists of two low-pass filters alternated with emitter-followers and preceded by the output of the sample and hold capacitor. The cut-off frequency is 10c/s and a slope asymptotic to 12dB/octave at higher frequencies.

Effect of Leakage on the Integrator

The effect of leakage associated with the integrator capacitor is to modify the transfer function to $1/(\alpha + pT)$ where $\alpha = R/R'$. The term $\alpha = R/R'$ may be neglected at high frequencies (Fig. 6). An estimate of the actual value of R' made from an experimental frequency response (Fig. 7(a)) where the curve cuts the zero axis and the fact that at this point $1/(\alpha + j\omega T) = 1$ and from the curve $\omega = 100\pi$ rad/sec gives that $\alpha \simeq 0{\cdot}655$, i.e. $R' = 12 \times 0{\cdot}655 = 18{\cdot}3\text{k}\Omega$ which appears to be rather low.

It is, however, possible to show, as follows, that in spite of the exponential decay (with a time-constant CR') of the output voltage the mean d.c. level in a given time is still proportional to the number of pulses arriving at the input to the integrator in this time.

If the input e_i has a mean level of V (d.c.) then the output is given by:

$$e_o(p) = \frac{1}{\alpha + pT} \, .(V/p)$$

$$= \frac{1}{1 + pT/\alpha} \, .(1/p) \, .(V/\alpha)$$

$$= \frac{1}{1 + pT'} \, .(1/p) \, .(V/\alpha) \text{ where } T' = T/\alpha$$

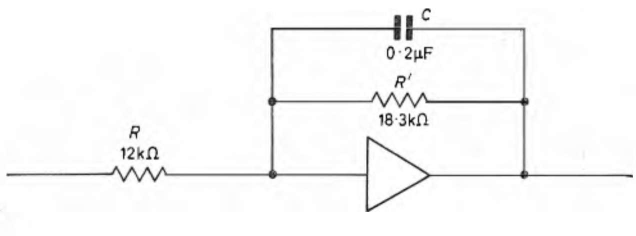
$$= V/\alpha \left\{ \frac{\alpha'}{p + \alpha'} \, .(1/p) \right\}; \; \alpha' = 1/T' = \alpha/T$$

$$= V/\alpha \left\{ (1/p) - \frac{1}{p + \alpha'} \right\}$$

$$e_o(t) = V/\alpha \, \{U_{-1}(t) - e^{-\alpha' t}\}$$

$$= R'/R \; V \, [1 - e^{-t/CR'}]$$

Fig. 6. Integrator



So that for a fixed time interval t as in the present application (the counting time), the output is proportional to the mean input d.c. level V. So although the leakage is an undesirable feature in reducing the output voltage it does not introduce any non-linearity. Though the actual mechanism of producing the final output voltage at the end of time t is due to each pulse being integrated separately and then discharging slightly, the result is the same as if a mean voltage level has been applied. This may be seen in Fig. 7(b).

Calibration and Sensitivity Tests

The curve of output voltage against speed in revolutions per second is shown in Fig. 8 and it shows that the curve begins to flatten slightly after 50 rev/sec. This is due to integrator shorting time not being long enough to discharge the capacitor completely, thus accumulating a charge and finally being driven into saturation.

A check on sensitivity made by feeding square pulses into the monostable multivibrator gave the required change in frequency of about 80c/s for a definite shift in the output trace on an oscilloscope. The sensitivity appeared to be better than this figure at higher frequencies (80c/s corresponds to about 18 rev/min). At the low frequency end as already mentioned the trace showed some dither but was limited to 0·03V peak-to-peak.

Frequency Response Test

The frequency response test was carried out by feeding a frequency modulated signal into the Schmitt trigger. The frequency response plot (Fig. 9) with all gains normalized with respect to the low frequency gain shows that the cut-off frequency (i.e. the frequency at which the gain is 3dB down on the very small frequency gain) is

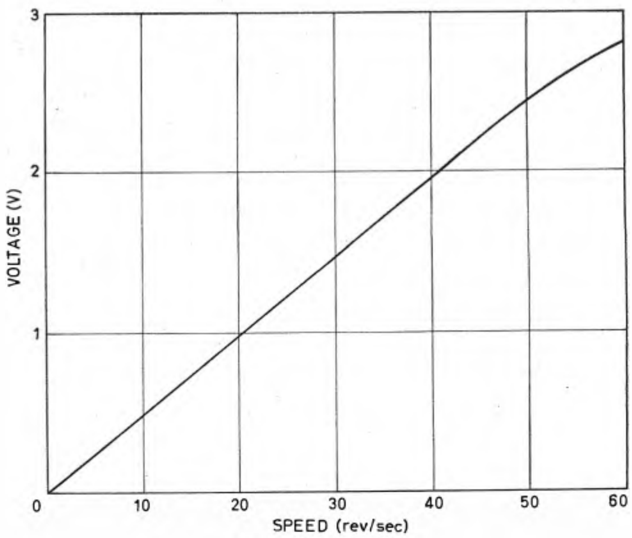


Fig. 8. Calibration

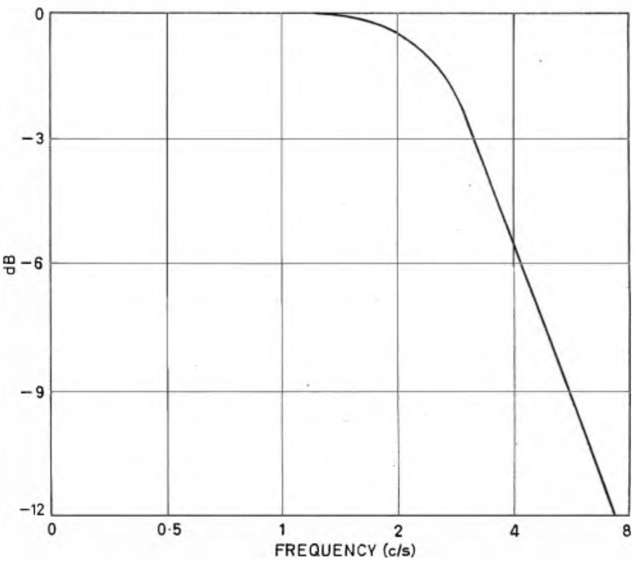
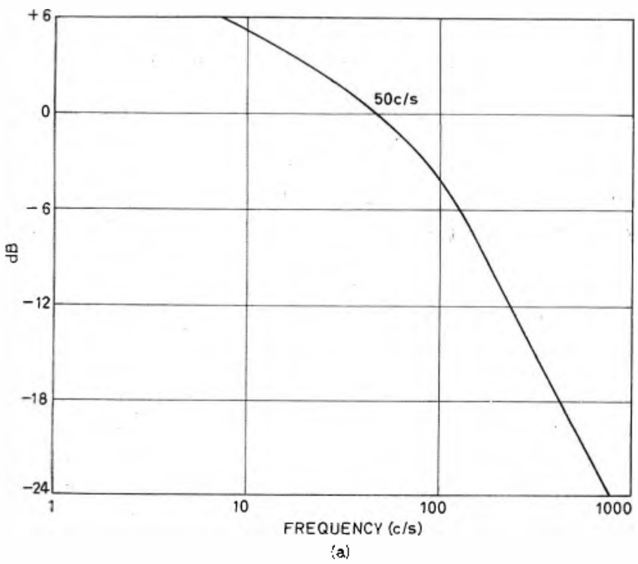
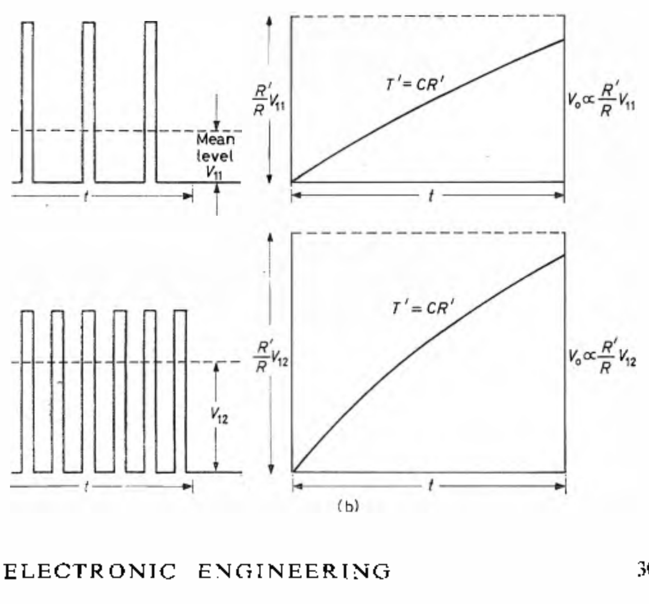


Fig. 9. Frequency response

Fig. 7(a). Integrator frequency response
(b). Low and high number of pulses





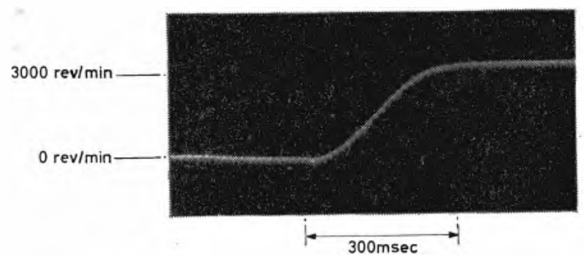


Fig. 10. An induction motor speed transient oscillogram obtained using the tachometer

just greater than 3c/s. After this the gain falls off sharply. Since the device is intended to follow acceleration from zero to full speed (50 rev/sec) in 0·1sec, the cut-off frequency must not be less than 2·5c/s, and this criterion is fulfilled.

Comments and Conclusions

The instrument was coupled to an induction motor and Fig. 10 shows one of the oscilloscope traces obtained from it for the switching on speed transients. (The top speed of the induction motor is 3 000 rev/min).

A device capable of many modifications and considerably greater accuracies than those obtained here has been described. Some may object to the use of the word diffraction gratings applied to elements having only 120 bright and 120 dark sectors but basically there is no reason for using what may correctly be termed diffraction gratings in optical terminology.

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A New Secondary Radar System

A new secondary radar system known as SECAR has been developed jointly by the Marconi Co. Ltd and the Compagnie Française Thomson Houston (CFTH) to meet all present and future requirements of secondary radar for air traffic control.

The Marconi Company was primarily responsible for the development of the interrogator/responser, and CFTH for the complex decoding and control system.

The complete system is fully transistorized with the exception of the high power output stage of the interrogator transmitter and a new aerial switching system has been incorporated.

The ground signal consists of one control pulse and one or two interrogate pulses, depending on the type of interrogation. In both cases the control and interrogate pulses are radiated by separate aerial configurations to produce different radiation patterns. A high speed switching system aerial enables the radiation patterns to be switched rapidly during the short time interval between pulses, and so permits the use of a single transmitter for both pulses as against other systems that require two transmitters.

The SECAR system employs both methods recommended by the ICAO, known as 'two-pulse' and 'three-pulse' respectively for interrogation on all civil and military codes. By a unique high-speed phase shifter, special radiation patterns are produced which enhance the effectiveness of the sidelobe suppression and the figures specified by ICAO are considerably exceeded.

Special mode interlace facilities enable up to three different modes to be transmitted sequentially during each aerial revolution. All modes can be interrogated during successive revolutions. Within this sequence, each mode can be interrogated with both two and three pulse signals.

The decoding equipment is extremely flexible and provides active, passive and height decoding facilities, together with emergency decoding.

Passive decoding consists simply of pre-setting any expected reply into the decoding equipment, and comparing the incoming replies with it. When the appropriate aircraft come within range, a signal is generated for display upon the operator's p.p.i. display.

Active decoding is used when the reply code is unknown to the operator. To decipher this code the operator isolates the response with a strobe marker. The active decoding circuits then examine this reply and provide the operator with an alpha-numeric display of the mode and code structure.

Passive height decoding is performed in the same manner as passive function decoding, and it is therefore possible to display aircraft responses that are above, below or between certain height levels which can be set in by the operator.

Active height decoding is achieved in exactly the same manner as active function decoding, except that code convertors translate the reply into decimal form, enabling the operator to read the height directly to the nearest 100ft.

Special decoding circuits are also included which cater for emergency replies initiated by the pilot. An emergency reply will give audio and visual alarms together with an automatic response on the p.p.i. display.

All aircraft producing replies to interrogations, can be displayed on a p.p.i. display of the SECAR system as single bar response coincident with their associated primary radar signals or on a separate p.p.i. display. Once a response has been passively decoded it can be displayed as a double bar response, distinguishing it from all other replies on the p.p.i. As a further aid to discriminate between passively decoded replies, the double bar response can be filled in to give a 'bloomed' effect.

Contained within the operator's control unit are two types of alpha-numeric display. One is used for active decoding and shows the mode and code structure of the signal, together with a figure from 0 to 5 depending on the validity of the reply. The other shows the height information in 10 000's, 1 000's and 100's of feet, together with the same validity indication.

Techniques are available to extract the information from decoded replies and to display it in alpha-numeric form on the p.p.i. display itself. This is done by programming a symbol character generator, using the SECAR decoder outputs together with information describing the position of the targets.

Two unique monitoring systems are available with the SECAR equipment.

One of these provides an absolute check of the interrogator/responser and aerial system performance. This is achieved by siting a special unit some 1 000ft away from the aerial, which re-radiates all transmitted interrogator pulses. This is done after translating the interrogator signals to the receiver frequency in a linear mixer, which is the only active circuit in the unit. The re-radiated signals are then received by the responser and compared with a set of extremely stable reference levels and any departures from the specified performance are indicated by alarm signals.

The second system consists of a simulator unit which generates replies on any mode with any code. These replies can be made to appear at any range and bearing. By setting the range of the simulated replies slightly outside the maximum operational range, operators are able to keep a constant check on the decoder performance.

The SECAR control and indicator unit mounted above the normal p.p.i. display

Field Effect Transistors in D.C. Amplifiers

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

It is found experimentally and theoretically that by suitably biasing a field effect transistor its drain current can be made independent of temperature. The use of field effect transistors as single ended input stages for d.c. amplifiers is discussed in some detail. Drifts, which arise mainly in the power supplies or in the second stage, can readily be made less than 50µV/°C, or 5 × 10⁻¹¹A/°C at 20°C. Circuits of two typical amplifiers are given.

(Voir page 354 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 361)

WITHOUT resort to chopping techniques, it is possible to use bipolar transistors to make d.c. amplifiers which have very small voltage drifts ($10\mu V/°C$, say). It is difficult to keep the current drift as low as $1nA/°C$, and there are consequently severe restrictions on the resistance of the signal source.

The field effect transistor (f.e.t.) gives the possibility of attaining high input impedance and high current sensitivity provided temperature effects can be made small. While f.e.t. were being tested with a view to their use in balanced push-pull stages, it was found that at a certain value of gate bias V_{gT}, the drain current I_d was virtually independent of temperature over a wide temperature range. This suggests the possibility of using f.e.t. as a single-ended input stage.

The equivalent circuit assumed for a f.e.t. operated in the pinch-off region is given in Fig. 1. R_1 and R_2 are normally very much greater than $100M\Omega$, and R_3 much greater than $0{\cdot}1M\Omega$; in many applications they may be neglected. The capacitances are generally a few picofarads. The gate leakage current I_g is usually less than 1nA at 20°C. The equivalent circuit is similar to that of a pentode valve with an exceptionally large anode to grid capacitance.

Theory of Temperature Effects

In the pinch-off region, the drain current I_d is fairly independent of drain voltage, and depends almost wholly on V_g/V_p, where V_g is the gate to source voltage, and V_p is the pinch-off voltage. A simple relationship which fits most double diffused f.e.t. is:

$$I_d = I_{do}(1 - V_g/V_p)^2 \quad \ldots\ldots\ldots\ldots\ (1)$$

where I_{do} is the drain current when $V_g = 0$.

The mutual conductance g is then given by:

$$g = 2I_{do}(V_g - V_p)/V_p^2 \quad \ldots\ldots\ldots\ldots\ (2)$$

Now I_{do} is proportional to the conductivity of the channel region. If the impurities in this region are fully ionized the conductivity, and hence I_{do}, will increase with temperature as the mobility of the carriers increases at a rate of, say, k_1 parts per 1 000/°C where k_1 is generally between 3 and 8. However, as the temperature increases, the width of the depletion layer between gate and channel regions increases, causing I_d to decrease. This change in effective junction width with temperature occurs at all moderately doped p-n junctions and corresponds to a change in junction voltage, or in this case V_g, of, say, $V'/°C$, where V' is generally not much different from 2·3mV.

Hoerni and Weir[1] have pointed out that these two effects will cancel to leave I_d independent of temperature if g and I_d have suitable values g_T and I_{dT}. These values are given by:

$$k_1 I_{dt} = g_T V' \quad \ldots\ldots\ldots\ldots\ldots\ (3)$$

or:

$$I_{dT}/g_T = V'/k_1 = V_T \quad \ldots\ldots\ldots\ldots\ (4)$$

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where V_T is usually in the range 0·3 to 0·8V.

Since both temperature effects are reasonably linear, the cancellation holds good over a wide temperature range. Other sources of drift, such as variation of the internal source resistance with temperature may be postulated, but they are likely to be very much smaller than those discussed.

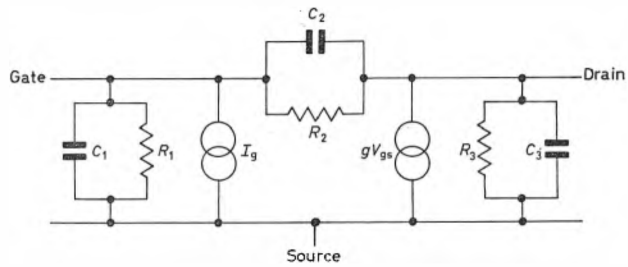

Fig. 1. Assumed equivalent circuit of field effect transistor in the pinch-off region

From equations (1) and (2) and (3) it follows that the gate bias V_{GT} for zero temperature drift is approximately $(V_P - 2V_T)$, and the mutual conductance g_T at this bias is $4I_{do}V_T/V_P^2$. By further manipulation it can be shown that if V_g differs by ΔV_g from V_{gT}, the input drift is about $1{\cdot}2\,\Delta V_g/V_T$ mV/°C. Alternatively, if I_d differs by a small amount ΔI_d from its zero drift value of I_{dT}, the drift is $1{\cdot}2\,\Delta I_d/I_{dT}$ mV/°C.

Choice of F.E.T.

From the foregoing theory it appears that almost any f.e.t. can be used to make an amplifier having very small temperature drifts ($< 10\mu V/°C$, say). However, changes in bias voltage V_{gT} are indistinguishable from changes in input signal, and since it is very difficult to establish any voltage with a drift of less than 0·001 per cent/°C, suitable f.e.t. must require a bias V_{gT} of not much more than 1V; that is, V_P must not be much more than 2 or 3V. The lower limit of V_{gT}, and hence V_P, is set by the requirement that the gate channel junction must not be appreciably forward biased.

If the second stage of a complete amplifier employs bipolar transistors, and has a current drift as small as 1nA/°C, say, then the transfer conductance G of the first stage must be at least $100\mu\Omega^{-1}$ if the component of drift arising from the second stage is to be less than $10\mu V/°C$. Generally G is rather less than g_T. In circuits where V_{gT} is provided by a source bias resistance R_s, given by V_{gT}/I_{dT}, the value G is less than or equal to $g_T/(1+g_T R_s)$, or $g_T/(1 + V_{gT}/V_T)$. Hence with auto bias, and with V_{gT} as large as 1V and V_T as small as 0·3V, g_T must be greater than $400\mu\Omega^{-1}$.

If the second stage is also a f.e.t., then a high mutual conductance is less necessary, but a small value of V_{gT} is still required.

The most suitable currently available f.e.t. appears to be the Ferranti n-channel type ZFT12, for which V_P is typically 1·5V and I_{do} typically 1mA. If V_T is assumed to be 0·5V, V_{gT} should be about 0·5V, and g_T about 0·9mA/V. Measurements made on eight samples are given in Table 1. The results lie within the range of values to be expected from the very simple theory given above. In order to confirm the validity of the postulated causes of temperature drift, k and V' were estimated independently. V' was found by measuring the change in gate to source voltage with temperature when a constant current (100μA) was passed through the junction in the forward direction with the drain electrode disconnected. To find k, the gate electrodes are disconnected and the resistance between drain and source measured at various temperatures. As expected V' had about the same value (2·3mV/°C) in all transistors, and the variations in V_T were mainly ascribable to variations in k.

The ZFT12 has a sharper knee in its characteristics than most f.e.t., and the slope of the I_d versus V_d curves is greater than 100kΩ at V_d = 5V. Even so, it is necessary to adopt a somewhat arbitrary criterion in determining the knee voltage V_P. The value of V_P given in Table 1 is the value of V_d at which I_d is 0·9 of its value at V_d = 5V.

All the f.e.t. quoted in Table 1 are suitable for use as low drift amplifiers. Two of them require the gate to be slightly positive with respect to the source (for these V_{gT} is given as negative in Table 1).

TABLE 1

ZFT12 No.	1	2	3	4	5	6	7	8
V_P (V)	1·5	0·55	1·2	1·8	1·7	1·9	2	2·1
I_{DO} (mA)	1·4	0·28	0·53	1·6	0·68	0·55	1·8	0·78
V_{gT} (V)	0·8	0	0·1	1·3	−0·1	−0·05	1·1	0·2
g_T (mA/V)	0·82	0·79	0·83	0·5	1	0·81	1	0·8
V_T (V)	0·35	0·37	0·5	0·33	0·78	0·74	0·43	0·75

Current Drifts

The gate current I_g of a f.e.t. is essentially that of a reverse (or very slightly forward) biased p-n junction. If its leakage component is small, I_g is a fairly slowly varying function of the various electrode potentials, but it increases by about 10 per cent/°C. All ZFT12 tested had gate currents of less than 10^{-9}A at 20°C and V_d less than 5V. Since both gate electrodes are available, this figure can be reduced at the expense of halving the mutual conductance, by applying the fixed bias to both gate electrodes, but the signal to one gate electrode only. Drifts of less than 3×10^{-11}A/°C at 20°C are then possible.

Input Stages

Single ended d.c. amplifier stages are very sensitive to variations in power supplies and associated components, and the stability of these elements is a major consideration in design.

At most three supplies are involved—the drain supply V_{dd}, the gate bias supply V_{gg} or source bias supply V_{ss}, and a supply V_{bb} to 'back-off' any standing potentials which would otherwise be passed to the second stage. It is usually preferable that the largest positive or negative supply line should not have to be excessively stable; it is relatively easy to derive a very stable supply of lower voltage and current from these lines by means of a Zener diode and resistance. Since I_d varies only slowly with V_d, it is fairly easy to avoid the need to make V_{dd} exceptionally stable.

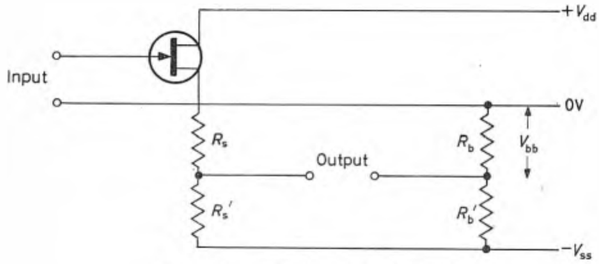

Fig. 2. Source follower input stage

If the f.e.t. requires negative gate bias, then the requirements on V_{ss} or V_{gg} can be eased by the use of an auto-bias circuit; that is, by the use of resistance R_s, given by V_{gT}/I_{dT}, in the source lead. The degeneration introduced by R_s is usually acceptable. The resistor must, of course, be a low temperature coefficient type. Unfortunately, it is difficult to devise auto bias circuits for f.e.t. which require a positive gate bias. However, any positive gate bias required is usually very small (< 0·1V, say), and a supply stable to 0·01 per cent/°C then gives an equivalent input drift of less than 10μV/°C.

Suitable input circuits based on these considerations are shown in Fig. 2 and Fig. 3. The circuit of Fig. 2 gives a voltage gain of rather less than unity; the circuit of Fig. 3 gives a voltage gain of about V_{bb}/V_T if R_s does not introduce appreciable degeneration. Both circuits give similar transfer conduc-

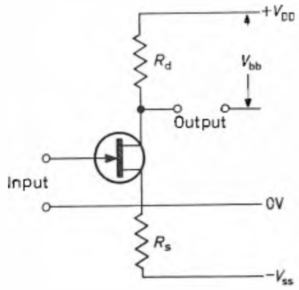

Fig. 3. Voltage amplifier input stage

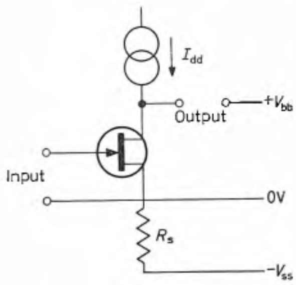

Fig. 4. High voltage-gain input stage

tances G of less than g_T. If bipolar transistors are used in the second stage, current gain is usually the important factor, and there is not a great deal to choose between the circuits on that account.

In the circuit of Fig. 2, if the required gate bias is negative, R_b is short-circuited to make V_{bb} zero, and R_s is given by V_{gT}/I_{dT}. If the gate bias is to be positive R_s is short-circuited and V_{bb} is made equal to V_{gT}. Assuming that V_{ss} is much greater than V_{gT} or V_T, a change of 1 per cent in V_{ss} gives an equivalent input change of 0·01 $(V_T + V_{gT})$.

In the circuit of Fig. 3, if the required gate bias is negative V_{ss} is made zero, and R_s provides auto bias. In that case a change of 1 per cent in V_{bb} corresponds to an input change of 0·01 $(V_T + V_{gT})$. If positive gate bias is needed, R_s can be omitted (unless required to make the source electrode available for negative feedback) and V_{ss} provides auto bias. In this case a change of 1 per cent in V_{bb} corresponds to an input change of 0·01 V_T, and a change of 1 per cent in V_{ss} to 0·01 V_{gT}.

It is clear that both circuits require supplies of the same order of stability, and that low values of V_T and V_{gT} are an advantage. However, the negative supplies only are critical in the circuit of Fig. 2, whatever the sign of bias required. In the circuit of Fig. 3, both positive and negative supplies are critical if f.e.t. requiring positive gate bias are to be accommodated. This may sometimes be a significant factor in the choice of circuit.

If a f.e.t. is used in the second stage, the voltage gain of the first stage is all important. The circuit of Fig. 3 gives a fairly small voltage gain unless V_{bb} and V_{dd} are very large. A more suitable circuit is shown in Fig. 4; here the voltage gain is determined by the amplification factor and internal feedback resistance of the f.e.t., and the internal resistance of the current generator, since this is not infinitely large in practice. These factors cannot usually be assessed very accurately from published data on devices, but voltage gains of several hundred are found experimentally to be easily obtainable. With so large a gain in the first stage, drifts in V_{BB} and the second stage are readily made negligible. A change of 1 per cent in I_{dd} corresponds to an input voltage change of about 0·01 $(V_T + V_{gT})$. The effect of any variation of V_{ss}, if V_{ss} is required to give a positive gate bias, is as before.

Overall Drifts

From the preceding discussion one can estimate that by using components of the highest stability (0·001 per cent/°C, say) and with selected transistors in the second stage, it is possible to design an amplifier with an overall drift of less than 20- to 50μV/°C, the exact value depending mainly on the value of V_{gT} and V_T for the particular f.e.t. employed. These figures assume all drifts are additive —some cancellation will occur in most amplifiers.

An alternative scheme is to use power supplies and resistors of moderate stability (0·01 per cent/°C, say) and a relatively simple second stage. The maximum drift may then be as large as 1mV/°C in some amplifiers—most will be very much better than this. All can be made to give a reasonable performance (less than 50μV/°C, say) by adjusting the bias in the first stage as the temperature of the complete amplifier, or its critical parts, is varied. The improvement that can be had in this way is limited by the difficulty of ensuring that the temperatures of components change at the same rate.

Setting up such an amplifier is less tedious than might be supposed, as an indication of the sense and amount of any bias offset required can be got from the theory given above. The components concerned are small and few, and can be heated or cooled relatively quickly. For rough work, and preliminary development, heating components with the fingers (about 10°C rise) or by bringing a soldering iron near them, often gives a useful indication of performance.

Source Follower Amplifiers

The input circuit of Fig. 2 has the advantage that, since its mean output is very nearly at earth potential, it can be used to precede most existing bipolar transistor amplifiers. The supply V_{ss} is usually best derived, via a Zener diode, from the main negative supply line; this gives a stable voltage and reduces the possibility of unwanted feedback via the supply lines. Overall feedback is taken either to the gate of the f.e.t., or, since there is some degeneration in the f.e.t. stage, to the spare base of the second stage transistor pair. The so-called ring of three amplifier[2] is particularly suitable in this connexion. If wide bandwidth is not required, the f.e.t. can be followed by a chopper d.c. amplifier; one of the features of chopper amplifiers is that they do not usually require or have very well stabilized power supplies. It is then sometimes convenient to use mercury cells to provide bias for the f.e.t.

Most of the f.e.t. amplifiers built by the author have used source follower input stages; the ease with which the circuit adapts itself to f.e.t. having widely different characteristics is its great advantage. No circuit diagrams of complete amplifiers are given, as the design of the succeeding bipolar transistor stages is now well understood and documented.

The two amplifiers to be described in detail demand f.e.t. for which the gate bias is negative. It is very easy to adapt them to use f.e.t. requiring positive bias if a well stabilized negative supply (< 0·1 per cent/°C) is available.

A Typical Amplifier

A typical amplifier using the input stage of Fig. 3 is shown in Fig. 5. The base resistors of the 2C111 are made approximately equal to minimize drifts. It is not necessary to select the 2C111 very stringently; but the combination of a ZFT12 having a low transconductance and 2C111 with a low current gain should be avoided. The

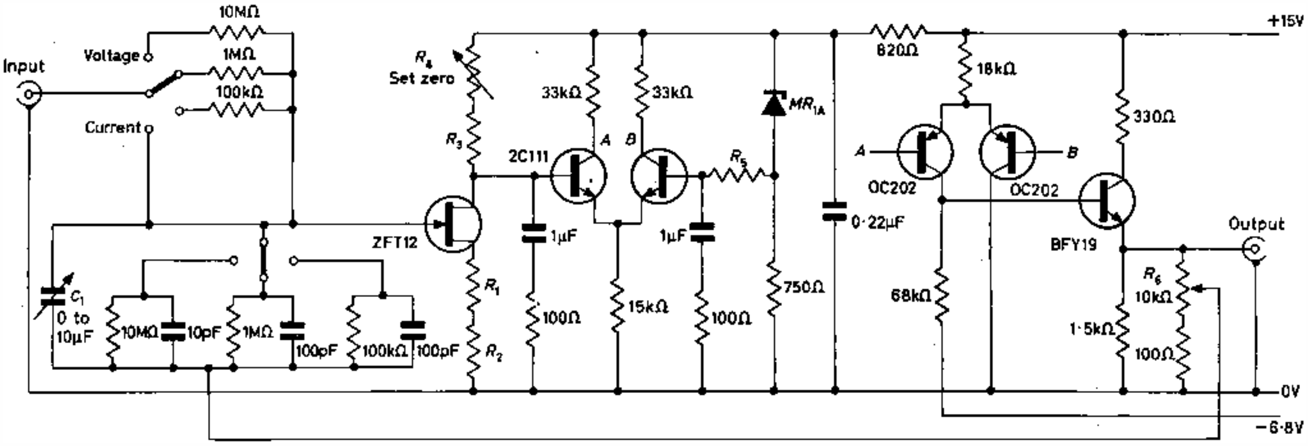

Fig. 5. Typical general purpose amplifier

One gate lead (pin 3) of the ZFT12 is earthed. R_1 to R_5 are chosen to suit the particular ZFT12 (see text). R_1 is rather less than V_{gT}/I_{dT}. R_2 is chosen on final test. R_5 is approximately equal to $(R_3 + \frac{1}{2}R_4)$. All resistors in first two stages are metal film or wire-wound types. R_6 is tapped to give required intermediate gain ranges. C_1 is switched to give required time constants

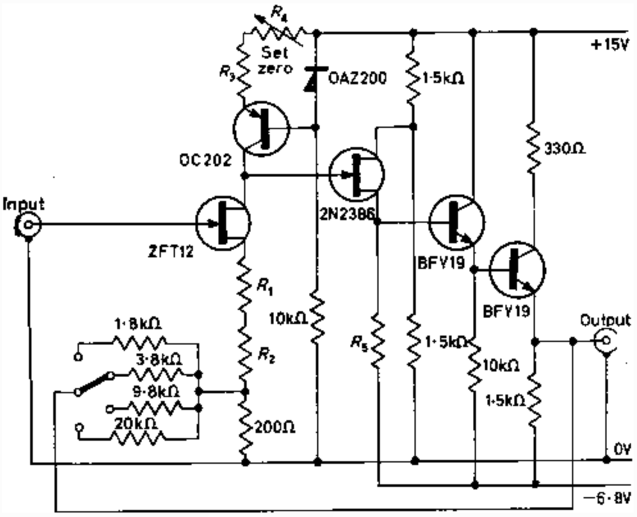

Fig. 6. Amplifier using two f.e.t.

The input is connected to both gate leads of the ZFT12. R_1 to R_5 are selected to suit the particular ZFT12 and 2N2386 (see text). R_1 is rather less than V_{gT}/I_{DT}. R_2 is selected in final test. All resistors in first two stages are metal film or wire wound types. Input voltage attenuator and switched current shunts have been omitted from diagram

gain without feedback is more than 3 000, and the input voltage drift is less than 300μV/°C before the bias on the f.e.t. is finally adjusted to give a drift of less than 50μV/°C. One of the prototype amplifiers using a good ZFT12 (V_{gT} about 0·1V) gives a drift of less than 30μV/°C without any bias adjustment; the current drift at 20°C is less than 4×10^{-11}A/°C.

An output of at least ±1V into 1kΩ is available. By varying the fraction of the output applied to the feedback resistance, and the feedback resistance, 1V output can be obtained for input currents of between 10^{-9} and 10^{-5}A. By adding resistance in series with the input voltage ranges of 100μV to 100V are made available. No attempt has been made to achieve a very fast response; the bandwidth varies from 1kc/s to 10kc/s as the gain is adjusted. In many applications a slow response is desirable to reduce the effect of noise and interference, and a wide range of time-constants can be obtained by changing the feedback capacitance.

Amplifier Using Two F.E.T.

A neat amplifier can be made if an n-channel f.e.t. is followed by a p-channel f.e.t. (or vice versa) as shown in Fig. 5. The gain of the first stage is so large (typically 200) that the bias on the second stage f.e.t. is relatively uncritical, so far as temperature effects are concerned. However, since the value of g/I_D, and hence the stage gain, increases with increasing bias, it is usually convenient to operate the f.e.t. near the zero temperature drift point at which g/I_d is about $1/V_T$.

The open loop gain of over 5 000 is reduced to 100, 50, 20 or 10 by means of feedback to give bandwidths of over 10kc/s at the highest gain, and over 100kc/s at lowest gain. In some particular amplifiers trimmer capacitances across the feedback resistances may be required to give critical damping. The input capacitance is less than 10pF. An output of at least ±1V into 1kΩ is available. A conventional switched voltage attenuator preceding the amplifier has been omitted from the circuit; in conjunction with the variable feedback it gives ranges of 10mV to 100V input for 1V output. Current ranges of 10^{-9} to 10^{-3}A are also provided in the usual manner.

The temperature drift before final adjustment of bias should be less than 500μV/°C. One of the prototype amplifiers uses a ZFT12 for which V_{gT} is as high as 1·3V; the overall drift is 300μV/°C with this value of bias. By increasing the bias to about 1·5V, the overall drift is reduced to less than 50μV/°C. The input current drift is less than 4×10^{-11}A/°C at 20°C.

This amplifier, and the preceding one, require the +15V supply to vary by less than 5mV/°C. The negative supply line is relatively uncritical—variations of ±1 per cent can be tolerated.

Conclusion

The bipolar transistor cannot compete with the thermionic valve as the input stage of a straight d.c. amplifier for measuring very small currents. However, voltage drifts arising from changes in emission and microphony in the valve, especially in directly heated types, are serious drawbacks and limit the useful voltage sensitivity. As a consequence the measurement of small currents requires very large resistances in the amplifier, and a very high resistance source. The field effect transistor can be readily made to give a voltage drift comparable with that of a bipolar transistor stage (or more than two orders of magnitude less than a miniature electrometer valve) combined with high input impedance. It has a sufficiently low input current to compete with the valve in measuring currents of more than 1nA, and it can be used with a much lower resistance source. It requires fairly stable power supplies, but these need give very much less current than the equally stable supplies required for a thermionic valve. The noise generated by a f.e.t. can be less than that of a bipolar transistor arranged to have similar input impedance, but is in general greater than with the noise other than that arising from microphony in a thermionic valve.

For long term recording under laboratory conditions f.e.t. amplifiers give acceptable drifts when f.s.d. is obtained for 10mV or 10^{-9}A. When used as an oscilloscope pre-amplifier, a sensitivity of 1mV/cm will not give more observable drift than that given by most commercially available oscilloscope amplifiers operated at 100mV/cm.

This article has discussed only single-ended non-chopper amplifiers. Differential amplifiers are often necessary for biophysical applications. Those described in the literature[3] rely on matching the temperature drifts of two f.e.t. There are possibilities of improved performance and easier selection of transistors if two f.e.t.'s each biased for zero drift are used in a differential stage. It is hoped that this may be discussed in a later article.

The main objections to the use of f.e.t. at present available are their wide spread of characteristics coupled with their high cost which makes it uneconomic to reject more than a small proportion of samples as unsuitable. In a laboratory making a wide range of instruments this objection is not particularly important as a f.e.t. unsuitable for one type of amplifier may be used for some other purpose. The large scale production of a single specific design may, however, present some difficulty until the cost of f.e.t. is reduced to a level at which either the transistor manufacturer or the user can afford to select f.e.t. more rigorously. In the present state of affairs the source follower amplifier is probably the most suitable type of input stage for large scale production.

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A Video Logarithmic Amplifier with Quick Recovery

By G. Kossoff*, C. N. Liu* and D. E. Robinson*

The design of a video, quick recovery, logarithmic amplifier operating on the 'successive detection and addition' principle is outlined. The dynamic range of the amplifier is 75dB and the recovery from overload is 0·15μsec. This recovery is achieved by employing successive differential stages with high frequency diodes across the anodes. The characteristics of amplifiers using variable successive stages are discussed. Circuits of 2Mc/s and 8Mc/s amplifiers are presented.

(Voir page 355 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 362)

ULTRASONIC echoscopes used for the mapping of internal structures of the human body operate on principles similar to radar, displaying echoes which originate at discontinuities of acoustic impedance of tissue. The amplitude of the received echoes has a large dynamic range depending on three factors; the reflection coefficient of the interface, the angle of incidence of the scanning beam, and the depth of penetration, since tissue attenuates ultrasound at 1dB/cm per Mc/s.

To compress this dynamic range, a sensitivity-time-control (s.t.c.) and a logarithmic amplifier have been employed. The s.t.c. amplifier, in which the gain is controlled by a voltage having a prescribed time function, is useful when the echo amplitude varies in a known manner, e.g. uniform attenuation. However, for most applications of ultrasonic echoscopy, the amplitude variation of the returned echoes cannot be fully compensated by s.t.c. and a logarithmic amplifier must also be employed.

The 'Successive Detection and Addition' Logarithmic Amplifier

The required dynamic range of the logarithmic amplifier

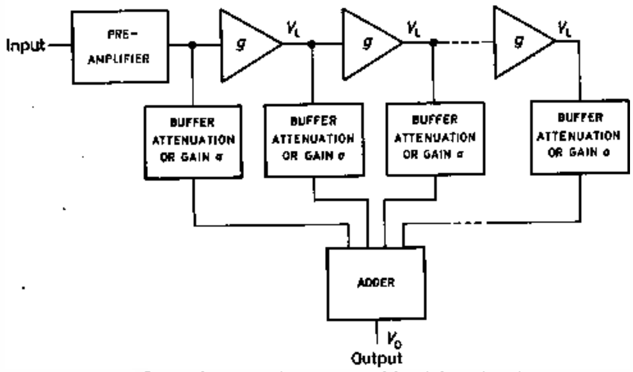
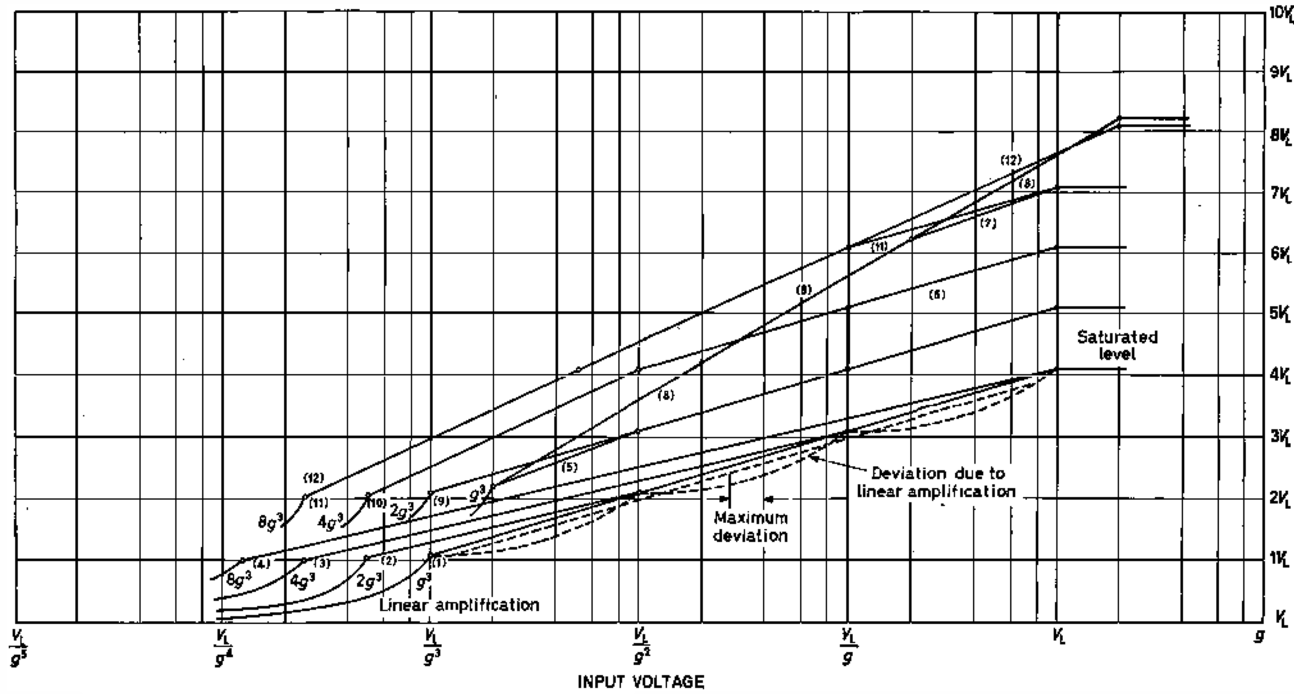


Fig. 1. The 'Successive detection and addition' logarithmic amplifier

Fig. 2. The characteristic of a three stage logarithmic amplifier whose individual stages do not have equal gain (g) or saturation level (V_L)

(1) All g, all V_L. (2) Last 2g, all V_L. (3) Last two 2g, all V_L. (4) All 2g, all V_L. (5) All g, last $2V_L$. (6) All g, last two $2V_L$. (7) All g, all $2V_L$. (8) All g, all $2V_L$, the pre-amplifier limited output $2V_L$ $(1 + (1/g))$. (9) First two g and V_L, last 2g and $2V_L$. (10) First g and V_L, last two 2g and $2V_L$. (11) All 2g and $2V_L$. (12) All 2g and $2V_L$, the pre-amplifier limited output $2V_L$ $(1 + (1/2g))$



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is determined by the performance of the scanning transducer and the dynamic range of the display unit, and was set at 75dB. The returned echoes consist of only a few cycles of a video frequency and the amplifier must have a quick recovery from overload so as not to block small echoes immediately following large ones.

The chosen logarithmic amplifier operates on the 'Successive Detection and Addition' principle as shown in Fig. 1. This amplifier was originally developed with successive detection from i.f. stages[1] and has been used extensively for anti-clutter purposes in radar receivers[2,3,4]. For input signals below the input level which saturates the last stage, the amplifier is linear and has a gain of g^n. For an input V_i, the adder output V_o is:

$$aV_i\,(g^n + g^{n-1} + \ldots . g + 1)$$

Suppose V_L is the limited or saturated output of each stage. Then, for an input of V_L/g^n, the last stage just saturates and the adder output is:

$$aV_L\,(1 + (1/g) + (1/g^2) + \ldots . + (1/g^n))$$

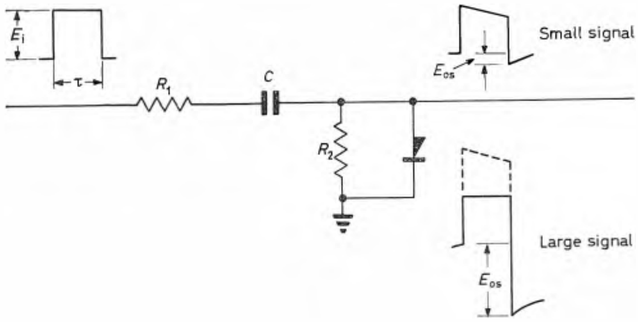

Fig. 3. Overshoot characteristic of a coupling network

The input and output levels, when the x^{th} stage just limits, are given by:

$V_i = (V_L/g^{n+1})\,g_x$, and $V_o = aV_L\,(x+(1/g))$, if $(1/g^2) \ll 1$.

These can be rewritten as $V_o = K_1 \log V_1 + K_2$, where K_1 and K_2 are the appropriate constants. Thus the amplifier has a logarithmic output for inputs at which any stage just becomes saturated. For other inputs, deviations from the logarithmic output depend on the amplification of each stage before saturation. A logarithmic amplification in each stage gives a logarithmic response for all inputs. With linear amplification, the deviation depends on the gain and its shape is illustrated in Fig. 2. For a stage gain of 15dB, the maximum deviation is 1·5dB.

So far it has been assumed that all the stages have equal gain g and saturate at V_L. Fig. 2 illustrates the characteristics of a three stage amplifier whose individual stages have a gain of either g or $2g$ ($g = 10$) and saturate at either V_L or $2V_L$. The pre-amplifier acts as an input limiter of output $V_L\,(1 + (1/g))$. The input is plotted in terms of V_L/g and the output in terms of V_L. Curve (1) shows the adder output when the three stages all have the same gain g and the same saturation level V_L. The circles represent the inputs when a stage saturates and logarithmic stage amplification has been assumed. These curves are useful to analyse the characteristics of a logarithmic amplifier which does not have identical stages.

Overshoot

In video amplifiers, overshoot is caused by coupling and by-pass time-constants. Fig. 3 shows a typical RC coupling network with simple limiting. For small signals the diode presents a high impedance and the overshoot is given by $A = E_{os}/E_i = \tau/t_r$ where τ is the pulse duration, and $t_r = (R_1 + R_2)\,C$ is the recovery time-constant.

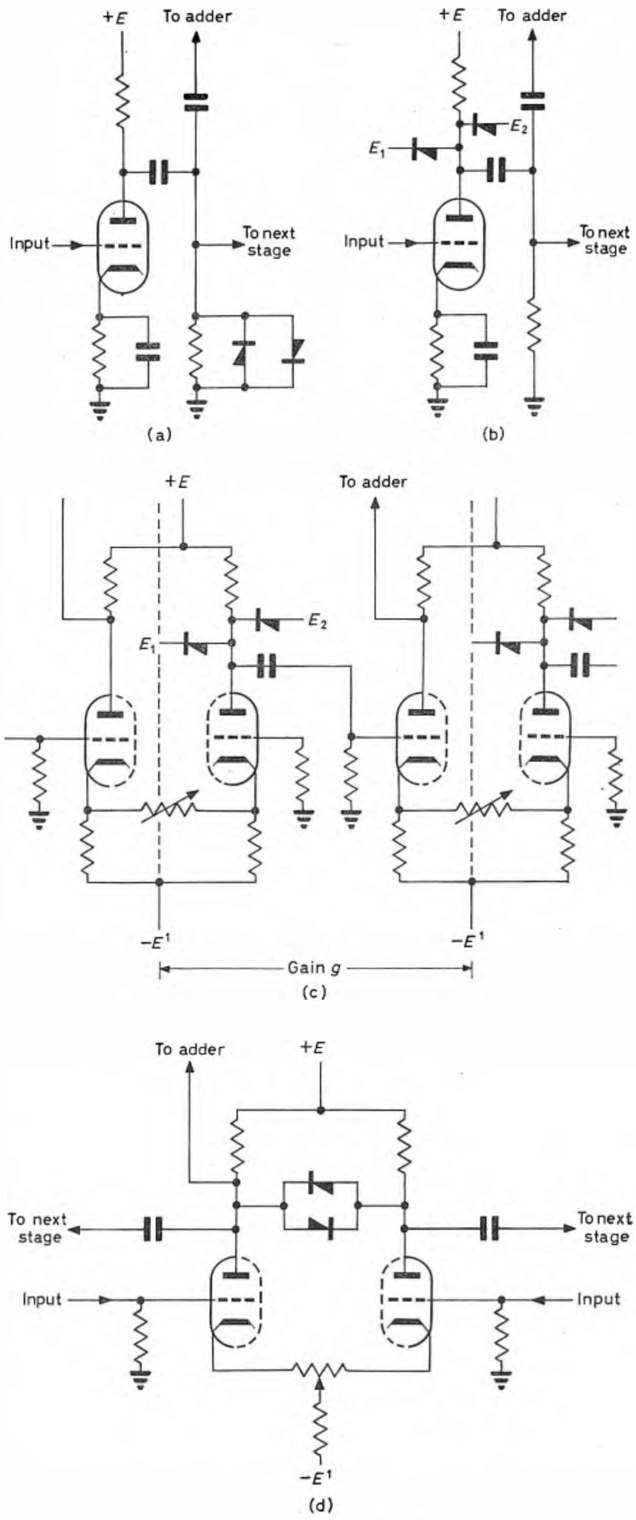

Fig. 4. Basic circuits of the individual amplifier stages

For large signals, the diode clamps the input signal and the capacitor C charges with a time-constant R_1C. After the pulse, C discharges with a recovery time-constant $(R_1 + R_2)\,C$, giving a large overshoot which blocks the following signals. Overshoot can be reduced by increasing the coupling time-constant, but R_1 is determined by the required high frequency response and the maximum value of C is limited by physical considerations. This problem can be overcome by not allowing saturation to take place

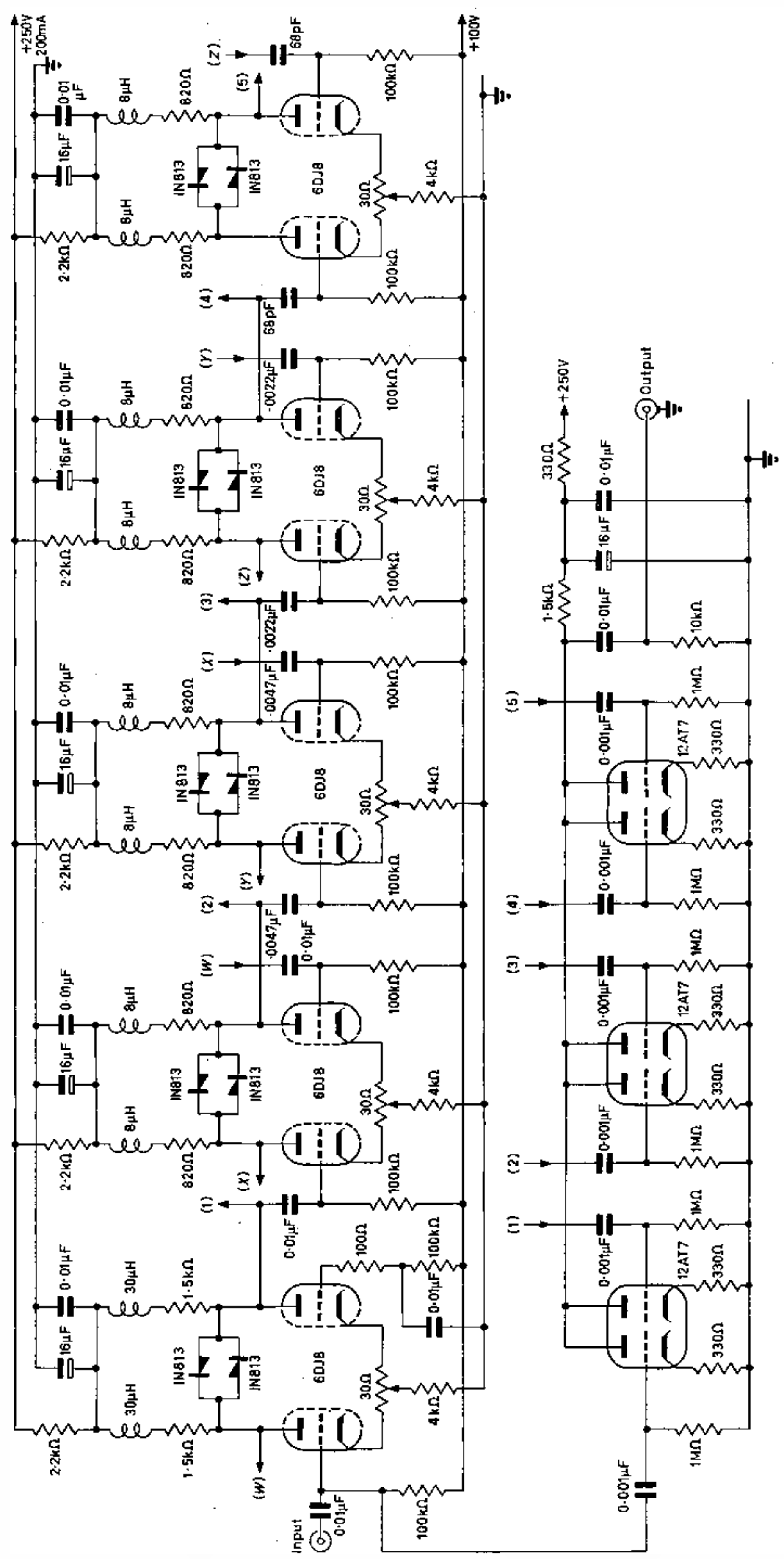

Fig. 5. The 2Mc/s logarithmic amplifier

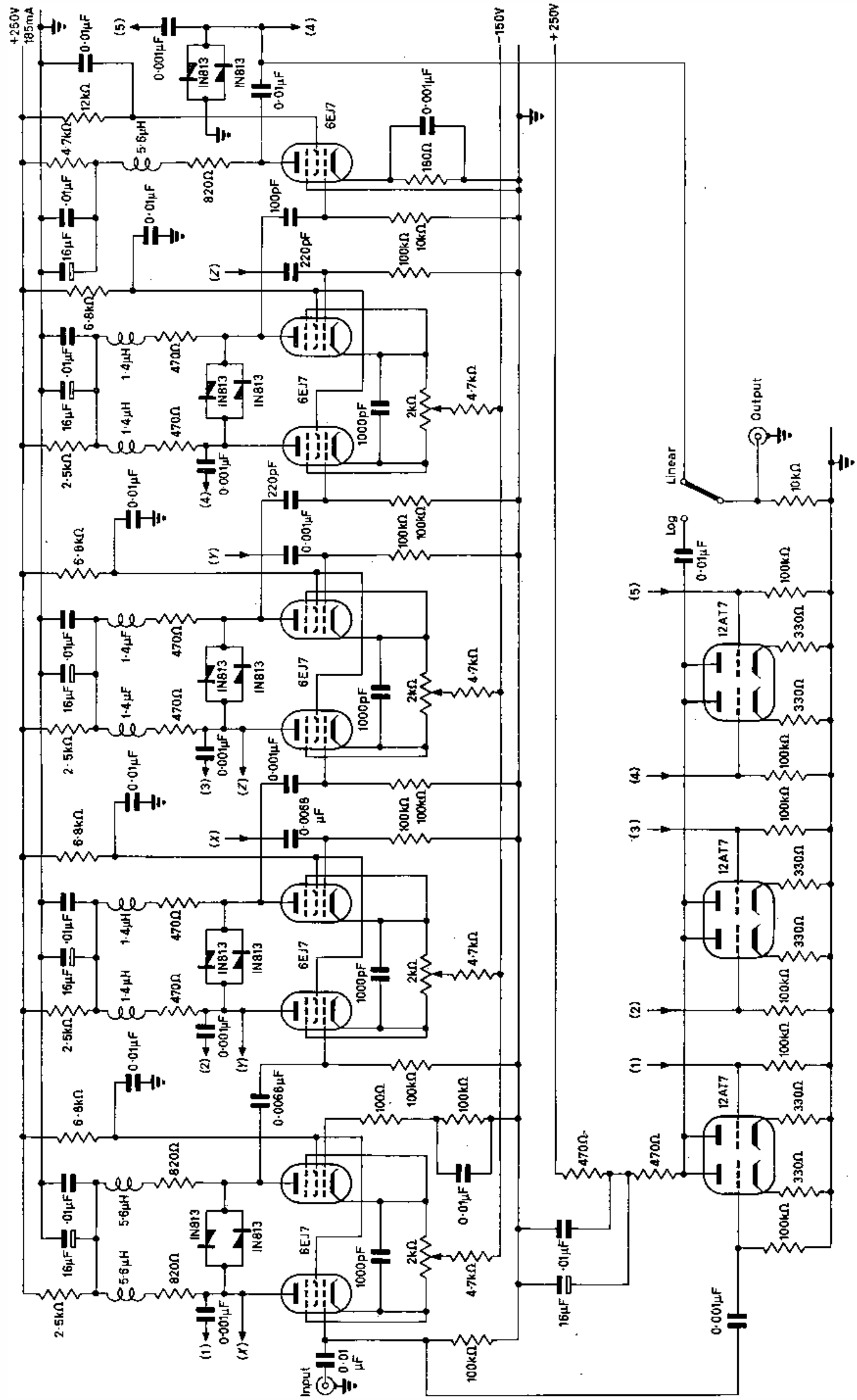

Fig. 7. The 8Mc/s logarithmic amplifier

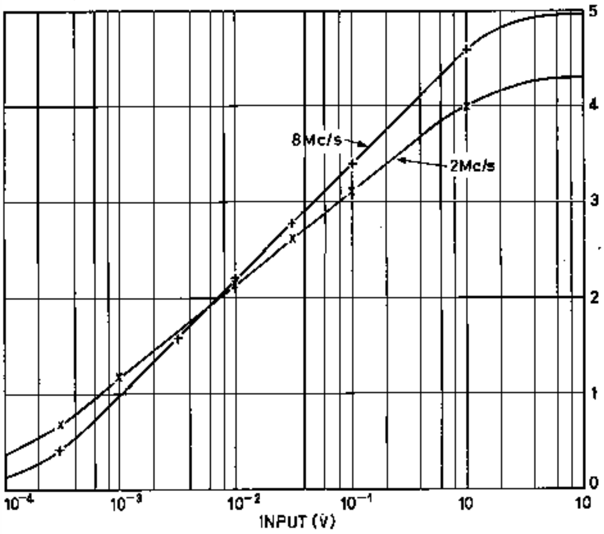

Fig. 6. The characteristic of the logarithmic amplifier

immediately after a coupling network. The overshoot of the first stage must then be lower than the overall dynamic range so that no weak signals are blocked by overshoot. The overshoot characteristics of an intermediate stage must be lower than the level of the amplified weakest signal reaching that stage.

Phase and Stability

In an i.f. system, all the video outputs have the same polarity, and a simple delay line can be used to compensate for the time delay of the video pulses. In the video system the possible phase reversal of the signals must also be considered. Since the signals at the adding circuit in the i.f. system are of video frequencies, the i.f. amplifier can be sharply tuned so that its video frequency gain is less than one, and a common load at the output of the delay line can provide good stability. In the video system such an arrangement will cause oscillation due to coupling at the adding circuit and buffer networks must be employed.

Basic Circuit

Fig. 4(a) shows a single-ended stage with a simple limiting arrangement. Fig. 4(b) shows an improved version where the limiting is done before the coupling capacitor. This stage requires a bias supply which must be stable to maintain a constant limiting level. Due to phase reversal, phase invertors are required for half the stages of the amplifier. Fig. 4(c) shows a cathode-coupled circuit, providing in-phase gain[6]. Since the same limiting arrangement is used, good bias stability is still required. The developed amplifier shown in Fig. 4(d) uses a differential stage, which provides in-phase gain, eliminates the bias supply and, due to common mode rejection, renders the limiting level independent of power supply variation. The d.c. levels of the anodes are balanced by the potentiometer at the cathodes.

Circuits

2Mc/s Amplifier

Fig. 5 shows the circuit for a logarithmic receiver for use with 2Mc/s transducers. It consists of five differential stages using twin triodes and high frequency silicon diodes. A potentiometer of 30Ω is sufficient to balance the d.c. level of the anodes. Each stage of the amplifier provides a gain of 15dB and a bandwidth of 11Mc/s. The overall gain is 75dB and recovery time less than 0·5μsec. The overall bandwidth through the video amplifier without limiting is 4Mc/s. The bandwidth at the adder output depends on the video and adder bandwidth and the input signal level. The adder bandwidth in this case is 5Mc/s. Due to the wide bandwidth of the individual stages, time delay compensation is not employed. The input is single-ended and the stage requires larger anode loads to maintain the 15dB gain. The input coupling time-constant is designed to keep the overshoot below 75dB for a 0·5μsec square pulse. The time-constants of the following stages are reduced to raise the lower half power frequency and thus lower the noise level. The adding circuit uses three twin triodes with a common anode load. This arrangement provides some gain and adequate isolation between output and input to ensure stability. Fig. 6 shows the logarithmic characteristics of this amplifier.

8Mc/s Amplifier

Fig. 7 shows the circuit of an 8Mc/s logarithmic amplifier which consists of four differential stages and a single-ended output stage. Pentodes are used to give the required high frequency response. Larger cathode resistances are required to stabilize the differential d.c. conditions and by-pass capacitors are employed. Each stage has a gain of 15dB and the overall dynamic range is 75dB with a bandwidth of 12Mc/s and recovery time of less than 0·15μsec. A switch is provided to change from a logarithmic to a limited linear characteristic. The logarithmic characteristic is shown in Fig. 6.

The single-ended stage with a simple limiting arrangement may be used in the later stages of the amplifier where overshoot requirements are not so stringent. This results in lower current requirements and less possibility of oscillation due to stray couplings. The limiting level, however, is double that of the differential stage if the same diode type is used. This can be overcome by increasing the gain of the stage by a factor of two and attenuating the limited signal by the same amount before feeding it to the adder. This attenuation, however, is not necessary to retain the input dynamic range as shown by curve (9) of Fig. 2. Alternatively diodes with half the forward voltage drop may be employed. If this is not done, the input dynamic range will be reduced by 6dB at the low end as illustrated in curve (5), and the curve will also change slope at inputs above those which saturate the previous stage. If more than one stage is employed the characteristic of the amplifier is shown by the appropriate curve of Fig. 2.

Conclusion

A video logarithmic amplifier consisting of differential stages has the best recovery characteristic. It is especially useful for low frequency applications when twin-triodes can be used to reduce the number of envelopes involved. For higher frequencies a combination of differential and single-ended stages is recommended. The single-ended stage reduces the current requirement and improves the stability of the amplifier.

Acknowledgment

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A Step-by-Step Design of a Two-Way Binary Counter

By D. Zissos*, B.Sc.

A step-by-step design is described for a d.c. coupled binary divider using NOR *modules when the complement of the input signal is not available. Static hazards are allowed to exist in the circuit if they are non-critical, although a method of eliminating them is described. A circuit technique simplifying* NOR-NOR *(and* NAND-NAND*) circuits is introduced. Because the modules are d.c. coupled the circuit is insensitive to rise and fall times of the input signal and the circuit will operate with either a square wave or a sine wave.*

(Voir page 355 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 362)

THROUGHOUT the article negative logic will be used, that is, the logical '1's and '0's correspond to the presence of negative and positive signals respectively. A NOR module for negative signals therefore generates the NAND function of positive signals.

Because it is easier to think in terms of the logical AND and OR functions rather than in terms of NAND and NOR functions, a NOR gate will be thought of as generating the AND function of the complements of the input signals[1] as shown in Fig. 1.

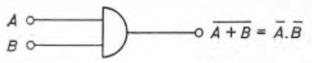


Fig. 1. A NOR module for negative signals

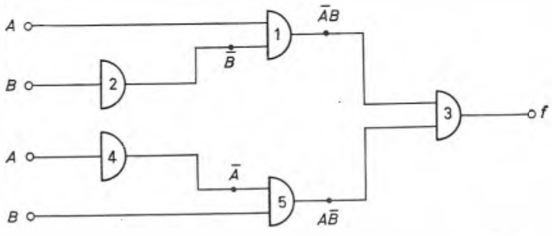


Fig. 2. A non-minimal circuit realizing $f = A\,B + \bar{A}\,\bar{B}$

To realize the NOR-NOR circuit of a given logical function f, its complement $\bar{f}$ is first expressed as a sum of products. If signals corresponding to each product term are next fed separately into a NOR module, its output is f. For example, consider the realization of the function $\bar{A}\,\bar{B} + AB$. $\bar{f}$ is first expressed as a sum of products, namely $\bar{f} = A\,\bar{B} + \bar{A}\,B$. The two signals $A\,\bar{B}$ and $\bar{A}\,B$ are next separately generated and fed into module 3 of Fig. 2. The output of module 3 is $\overline{\bar{A}\,B}\,.\,\overline{A\,\bar{B}} = (A + \bar{B})\,(\bar{A} + B) = \bar{A}\,\bar{B} + A\,B = f$.

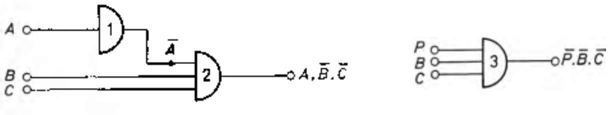


Fig. 3. A circuit simplification technique

Although the circuit of Fig. 2 realizes the function f, it is not necessarily in its minimal form. The next step is to apply the circuit technique shown in Fig. 3 for eliminating redundant modules.

If P has any one of the values $\bar{A}\,\bar{B}$ or $\bar{A}\,\bar{C}$ or $\bar{A}\,(\bar{B} + C)$ or $\bar{A}\,(B + \bar{C})$ or $\bar{A}\,(\bar{B} + C)$ or $\bar{A}\,\bar{B}\,\bar{C}$ the output of module 3 is the same as the output of module 2. Applying this circuit technique it is noted that the outputs of modules 1 and 5 in Fig. 2 will be the same if $\bar{A}\,\bar{B}$ is fed into the modules 1 and 5 instead of $\bar{B}$ and $\bar{A}$. Now $\bar{A}\,\bar{B}$ can be realized by simply feeding A and B into one module. The circuit therefore simplifies to that shown[2] in Fig. 4.

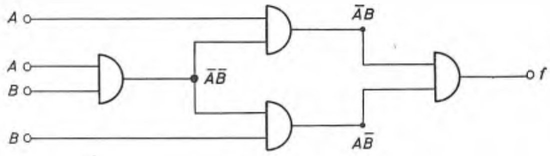


Fig. 4. A minimal NOR/NOR circuit realizing the function $f = A\,B + \bar{A}\,\bar{B}$

For NAND-NAND circuit realizations see Appendix (1).

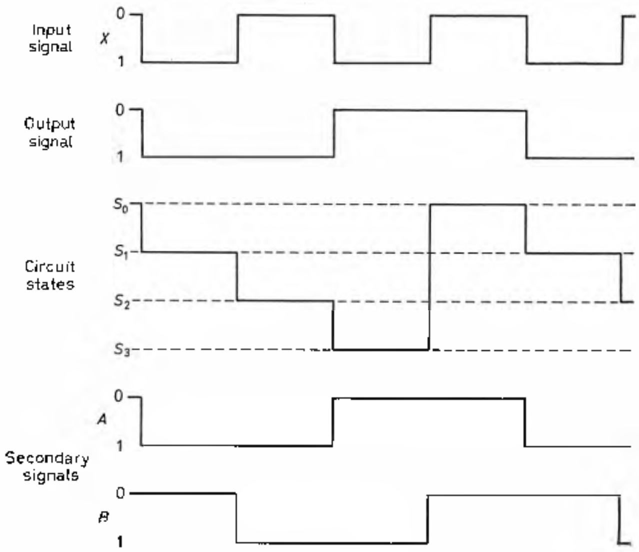
Designing the Binary Divider[3,4,5]

Step 1

In the first step the waveforms of the input, output and secondary signals are drawn.

Fig. 5 shows the input and output signals and the internal circuit states that the divider must have. Four states can be generated by two secondary signals A and B. Let the values of A and B corresponding to the four circuit

Fig. 5. State diagram of the two-way counter



* *City of Liverpool College of Technology.*

states S_0, S_1, S_2 and S_3 be 00, 10, 11, and 01 respectively. The values of A and B have been chosen so that when the circuit switches from one circuit state to another, it does so by changing one secondary signal only. This eliminates critical races[4].

Step 2

In this step a sequence (flow) table, Fig. 6, is used to describe the terminal action of the circuit. The table is

	X = 0	X = 1	AB
S_0	(S_0) 1	S_1 2	00
S_1	S_2 3	(S_1) 4	10
S_2	(S_2) 5	S_3 6	11
S_3	S_0 7	(S_3) 8	01

AB	X = 0	X = 1
00	00 1	10 2
10	11 3	10 4
11	11 5	01 6
01	00 7	01 8

Fig. 6 (left). The sequence (flow) diagram
Fig. 7 (right). The excitation table

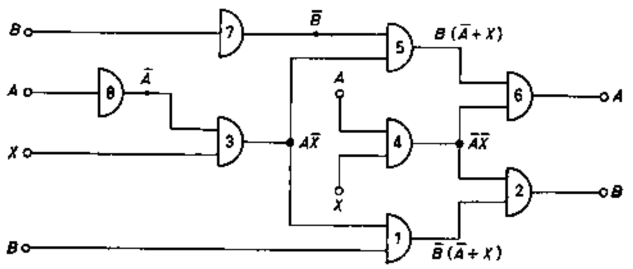


Fig. 8. A non-minimal realization of equations (1) and (2)

derived directly from Fig. 5. Each column heading is associated with each value of the input signal and each row with a circuit state.

Each entry in the table indicates the destination of the circuit, that is, the next circuit state, when the circuit is in a state corresponding to the row heading and the input signal is that specified by the column heading. For example, from Fig. 5 it is seen that if the circuit is in state S_0 and the input signal becomes '1', the circuit changes to S_1. S_1 is therefore entered in square 2. Similarly S_1 is entered in square 4 to indicate that the circuit remains in the S_1 state when $X = 1$.

If an entry is the same as the row heading, i.e. if the destination of the circuit is the same as its state at a given instant then the total circuit state is stable, and the entry may be circled to denote this fact.

Step 3

In the third step the excitation table and the logical equations of the circuit are derived. Consider entry 2. The S_1 entry indicates that the circuit state must change from S_0 to S_1 when $X = 1$. The circuit changes state from S_0 to S_1 by making $A = 1$ and $B = 0$. '10' is therefore entered in square 2. The same entry is naturally entered in square 4. In square 1 '00' is entered, since in state S_0, $A = B = 0$.

Similarly all the other entries are made. The resultant table is shown in Fig. 7. From it the logical equations for $\bar{A}$ and $\bar{B}$ are derived.

$$\bar{A} = \bar{A}\,\bar{X} + B\,X + \bar{A}\,B = \bar{A}\,\bar{X} + B(\bar{A} + X) \quad \ldots\ldots\ldots (1)$$

$$\bar{B} = \bar{A}\,\bar{X} + \bar{B}\,X + \bar{A}\,\bar{B} = \bar{A}\,\bar{X} + \bar{B}(\bar{A} + X) \quad \ldots\ldots\ldots (2)$$

The equations are then realized using the design procedure already explained. Namely to generate A, the two signals $\bar{A}\,\bar{X}$ and $\bar{B}(\bar{A} + \bar{X})$ are fed separately into a module etc. This results in the circuit shown in Fig. 8.

Before connecting the outputs of modules 6 and 2 to all input terminals marked A and B respectively, the circuit is further examined for redundant modules. Module 7 can be eliminated if into module 5 instead of $\bar{B}$, $\bar{B}\,.\,\overline{A\,\bar{X}} = \bar{B}\,(\bar{A} + X)$ is fed. But this is the output of module 1. Thus module 7 is redundant. Similarly module 8 can be eliminated if the output of module 4 is fed into module 3 instead of $\bar{A}$. The circuit is next re-drawn, as in Fig. 9.

Step 4

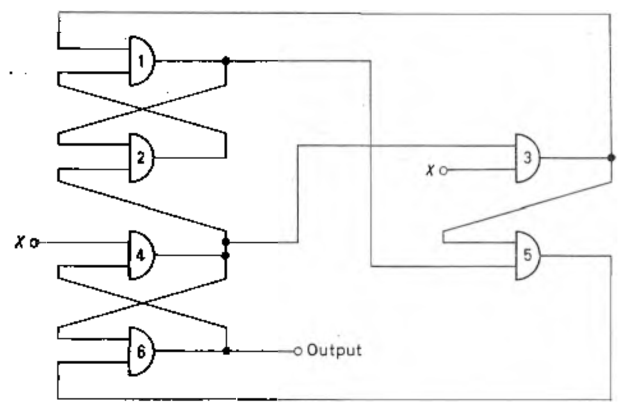
In this step the designer investigates the presence, if any, of static hazards.

The circuit waveforms are next re-drawn in Fig. 10 taking time delays into account. It will be assumed that all modules have identical propagation times, denoted by t. It will further be assumed that at the end of the first quarter period of the cycle (corresponding to circuit state S_0) $A = B = 0$. The waveform of the input signal is first drawn.

Consider the second quarter period of the cycle. During the first t seconds nothing has changed since it takes t seconds for the input signal to get through modules 3 and 4. Therefore at the end of t seconds modules 3 and 4 can only change. Because during the second quarter period the input signal = 1, the output of modules 3 and 4 becomes '0' if not already at '0', and remains at the value for at least a complete quarter period. Module 4 is the only module whose output has changed after t seconds in the second quarter of the cycle, since the output of module 3 was already at '0'. At the end of $2t$ seconds only modules 2, 3 and 6 can change state, since only these modules are connected to the output of module 4. Now the output of module 2 is held at '0' by the '1' signal from module 1. The output of module 3 is held at '0' by the presence of the input signal X. Module 6 changes state since the second signal fed into it (output of module 5) during the time under consideration is zero. Using this procedure the time diagram of Fig. 10 is completed.

From it, it can be seen that a static hazard is present at the output of module 3 during the first quarter of the cycle. That is the output of module 3 instead of remaining "static" at '0' during the first quarter of the cycle, changes temporarily to '1' for t seconds due to the relative propagation speeds of the various signals through the circuit. If the propagation speeds of the signals were infinite then the output of module 4 would change to

Fig. 9. A two-way binary counter based on equations (1) and (2)



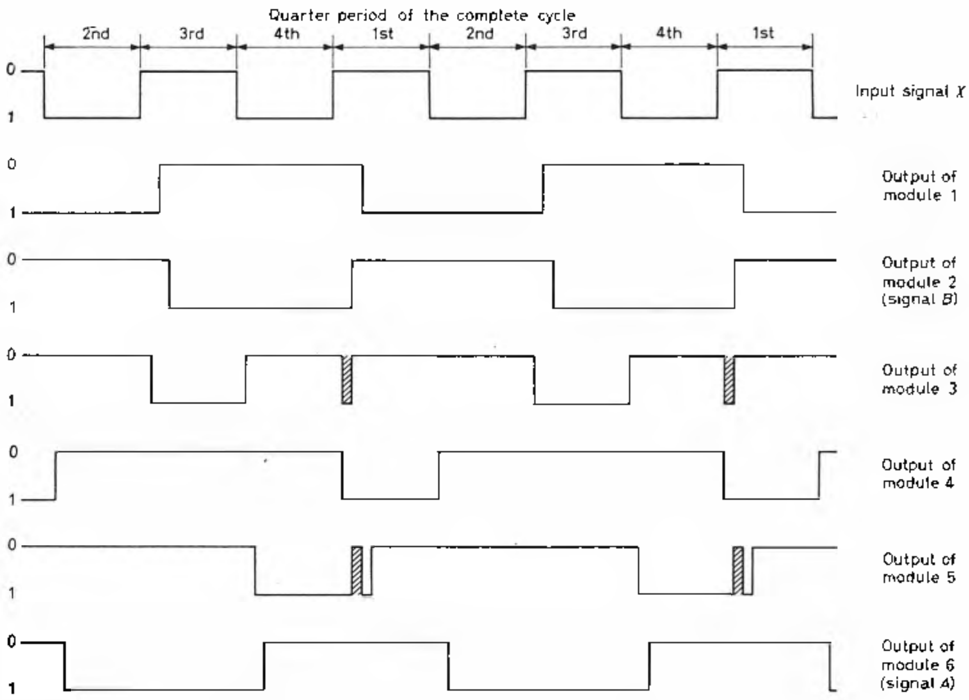


Fig. 10. The time diagram of the binary counter

'1' simultaneously with X changing to '0'. That is the input signals to module 3 would change from '10' to '01'. Since a '1' would always be present in the input of module 3 its output would remain at '0' during this change of the input signals. However, from the time diagram it can be seen that for the first t seconds of the first quarter cycle the output of module 4 is '0' instead of '1'. That is, the input signal of module 3 is 10, 00, 01, causing its output to go temporarily to '1'. The static hazard at the output of module 3 in turn causes a static hazard to appear at the output of module 5 t seconds later.

The static hazard at the output of module 5 is fed to the input of module 6. But it is prevented from affecting the output signal because it is held at '0' by the presence at the other input terminal of a second signal (output of module 4) which is '1' during its presence. It must be appreciated at this stage that if the output of module 4 is delayed by at least 2 module delays in reaching the input of module 6, the static hazard will be allowed to appear at the output of module 6. This was found to be true in practice. In such a case there are two alternatives that the designer can use.

Firstly, he can allow the static hazards to exist in the circuit and possibly take the output from modules 1 or 2. This is only true if the propagation time of two modules is always greater than the propagation time of one module.

Alternatively he can eliminate completely the static hazards from the circuit. A static hazard is present at the output of module 3 because the two signals feeding it are X and $\bar{A}\,\bar{X}$. During the first quarter period of the cycle because $A = 0\,(\bar{A} = 1)$ its output should remain at '0' during the change of X, since when $X = 1$, the output is maintained at '0' by X, and when $X = 0$ the output is maintained at '1' by $\bar{A}\,\bar{X}$. But because of the time delays $\bar{A}\,\bar{X}$ becomes '1' t seconds after X has changed to zero in the first quarter period.

Therefore the output changes momentarily to '1'. The static hazard can be eliminated by feeding a third signal into module 3 which will maintain the output at '0' during this time interval. Such a signal could be A or any other signal which is '1' during the first t seconds of the first period and '0' during the whole of the third quarter period. During the remaining part of the cycle it can be either '0' or '1'. This signal can be provided by the output of module 5, as can be seen from the time diagram of Fig. 10.

This follows from the fact that when the output of a NOR module is '0', any signals 0's or 1's can be fed into it, since such signals will have no effect on the output of the module. If, however, the output of the NOR module is '1', only '0' signals can be fed into it, since a '1' signal at the input will cause the output to become '0'. This circuit technique often allows the designer of logical circuits to eliminate the presence of static hazards using signals that already exist in the circuit rather than generating them using additional modules.

STEP 5

In the fifth step the circuit is built and tested before the design is finally accepted.

The circuit of Fig. 9 was designed on the assumption that the propagation times of all modules in the circuit are the same. Such an assumption is essential otherwise no systematic design could have been used. In practice however, no two modules are identical and their propagation times will invariably be different. The designed circuit must therefore be tested to ensure that it is independent of relative propagation times by introducing time delays in all the parts of the circuit.

Eliminating the static hazard in the divider by connecting the output of module 5 to module 3 resulted in the circuit becoming independent of relative propagation times although at low speeds, about 2kc/s, the circuit functions satisfactorily without the additional connexion.

Conclusions

In this article (a) a method of designing sequential circuits with NOR modules for negative signals and NAND modules for positive signals has been explained (b) a circuit technique of eliminating redundant modules has been introduced and (c) a method of eliminating static hazards in sequential circuits has been suggested. A method of designing combinational circuits using NOR and NAND modules has also been introduced.

The design technique used has been that of allowing static hazards to be present in the circuit when they are non-critical.

Acknowledgments

The author wishes to express his gratitude to his

colleague Mr. Brian Wilson, for his contribution in the preparation of this article.

Appendix

(1) NAND-NAND Circuit Realization for Negative Logic

To realize a given function using NAND modules, interchange the 'pluses' and 'points' in the algebraic expression of the function.

(a)

AB \ X	0	1
00	$A = 0$ (1)	$A = 1$ (2)
10	$A = 1$ (3)	$A = 1$ (4)
11	$A = 1$ (5)	$A = 0$ (6)
01	$A = 0$ (7)	$A = 0$ (8)

(b)

AB \ X	0	1
00	$B = 0$ (1)	$B = 0$ (2)
10	$B = 1$ (3)	$B = 0$ (4)
11	$B = 1$ (5)	$B = 1$ (6)
01	$B = 0$ (7)	$B = 1$ (8)

Fig. 11. Excitation tables for secondary signals A and B

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

Example

Assume the function $f = A \,.\, B + \bar{A} \,.\, \bar{B}$ is to be realized using NAND modules for negative signals. The function is first transformed to:

$f' = (A+B)\,.\,(\bar{A}+\bar{B}) = A\,\bar{A} + \bar{A}\,B + A\,\bar{B} + B\,\bar{B} = A\,\bar{B} + \bar{A}\,B$

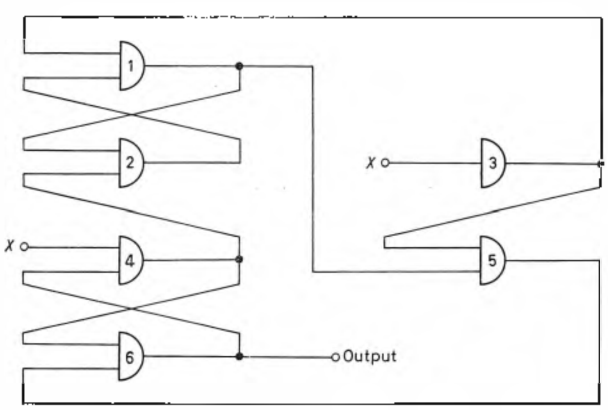
The realization of this function using NOR modules for negative signals is shown in Fig. 4. If now each module of Fig. 4 were a NAND gate for negative signals, the circuit would realize the function[2] $f = A\,B + \bar{A}\,\bar{B}$.

(2) Derivation of the Logical Equations from the Excitation Tables[2]

Figs. 11(a) and 11(b) show the excitation tables for the secondary signals A and B respectively.

To generate the signals A and B, $\bar{A}$ and $\bar{B}$ are expressed in their minimal AND-OR forms, as already explained at the beginning of the article under the heading of 'Designing with NOR Modules'. The minimal expressions for $\bar{A}$ and $\bar{B}$ were derived directly from the excitation tables of Fig. 11.

Fig. 12. A two-way binary counter based on equations (3) and (4)



From Fig. 11(a) it is seen that $A = 0$ in:

Square 1 i.e. when $A = 0$, $B = 0$, and $X = 0$;
therefore $\bar{A} = \bar{A}\,\bar{B}\,\bar{X}$

Square 6 i.e. when $A = 1$, $B = 1$, and $X = 1$;
therefore $\bar{A} = A\,B\,X$

Square 7 i.e. when $A = 0$, $B = 1$, and $X = 0$;
therefore $\bar{A} = \bar{A}\,B\,\bar{X}$

Square 8 i.e. when $A = 0$, $B = 1$, and $X = 1$;
therefore $\bar{A} = \bar{A}\,B\,X$

Hence:

$$\bar{A} = \bar{A}\,\bar{B}\,\bar{X} + A\,B\,X + \bar{A}\,B\,\bar{X} + \bar{A}\,B\,X$$

Simplifying:

$$\bar{A} = \bar{A}\,\bar{X}\,(\bar{B} + B) + B\,X\,(A + \bar{A})$$
$$= \bar{A}\,\bar{X} + B\,X \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

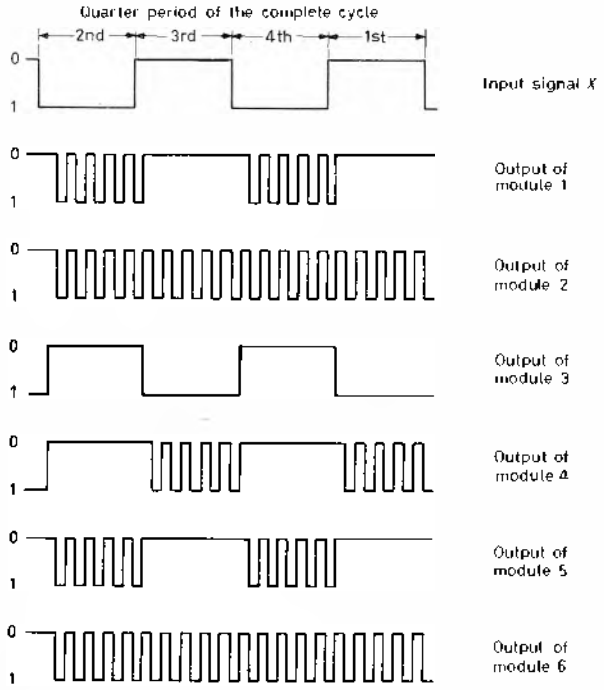


Fig. 13. The time diagram of circuit in Fig. 12

Similarly from Fig. 11(b) it is seen that $B = 0$ in:

Square 1 i.e. when $A = 0$, $B = 0$, and $X = 0$;
therefore $\bar{B} = \bar{A}\,\bar{B}\,\bar{X}$

Square 2 i.e. when $A = 0$, $B = 0$, and $X = 1$;
therefore $\bar{B} = \bar{A}\,\bar{B}\,X$

Square 4 i.e. when $A = 1$, $B = 0$, and $X = 1$;
therefore $\bar{B} = A\,\bar{B}\,X$

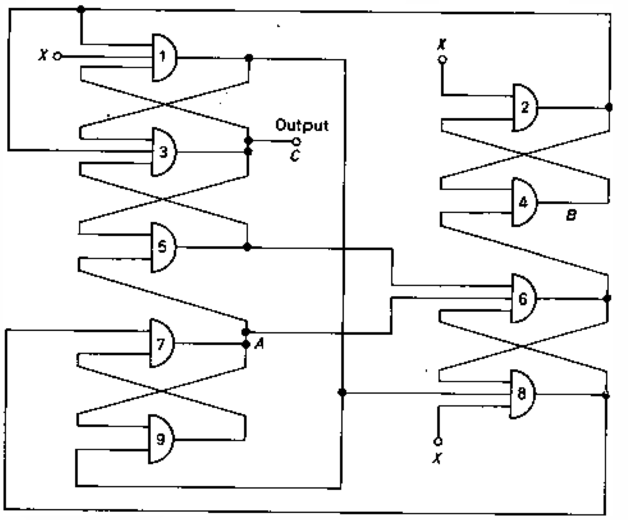
Square 7 i.e. when $A = 0$, $B = 1$, and $X = 0$;
therefore $\bar{B} = \bar{A}\,B\,\bar{X}$

Hence:

$$\bar{B} = \bar{A}\,\bar{B}\,\bar{X} + \bar{A}\,\bar{B}\,X + A\,\bar{B}\,X + \bar{A}\,B\,\bar{X}$$
$$= \bar{A}\,\bar{X}\,(\bar{B} + B) + \bar{B}\,X\,(\bar{A} + A)$$
$$= \bar{A}\,\bar{X} + \bar{B}\,X \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

Realizing equations (3) and (4) to generate the secondary signals A and B respectively, the minimal circuit of Fig. 12 was obtained.

The time diagram Fig. 13 of the circuit was next drawn assuming (a) that all modules have identical propagation times and (b) that at the end of the first quarter period of the cycle the secondary signals A and B are both equal to '0'.



DESIGN NOTES

Input signal	*Secondary signals*
X	*A B C*
0	0 0 0
1	0 1 0
0	1 1 0
1	1 1 1
0	0 1 1
1	0 0 1

Logical equations

$\overline{A} = \overline{A}(\overline{B}+C+X)+\overline{X}(\overline{B}+C)$
$\overline{B} = \overline{A}CX+\overline{B}\overline{X}$
$\overline{C} = AC+BX+BC\overline{X}$

Fig. 14. A three-way counter

The time diagram was found to be inconclusive. This is because modules 1 and 2 are connected 'back to back' forming a bistable element, their other inputs being $\overline{X}$ and $\overline{A}\,\overline{X}$ respectively. Now both these input signals are '1' during the first quarter period of the cycle when $A = X = 0$ and they both become simultaneously '0' at the end of t seconds in the second quarter period of the complete cycle (see Fig. 13). Depending on the relative propagation times of modules 3 and 4 B either remains at '0' or it becomes '1' during the beginning of the second quarter period of the cycle. If module 3 is faster than module 4, B remains at '0' but otherwise it changes to '1'. If B remains at '0' during this change or if A changes to '1' before B, the circuit moves from square 1 in Fig. 6 to square 4 and it functions as a binary pulse divider. If, however, B becomes 1 temporarily before A changes to 0 the circuit moves to square 8 which is a stable entry. From square 8 it then moves to square 1 when X changes to '0' and the circuit ceases to function as a binary pulse divider. All these observations have been verified experimentally. For example, the circuit when first built worked indicating that module 4 was slower than module 3. The circuit, however, ceased to function when modules 3 and 4 were interchanged.

A different circuit based on the same excitation tables of Fig. 11 had then to be designed. For this the Boolean expressions, equations (3) and (4) had to be modified. This was easily achieved by adding $\overline{A}\,B$ and $\overline{A}\,\overline{B}$ to $\overline{A}\,\overline{X} + B\,X$ and $\overline{A}\,\overline{X} + \overline{B}\,X$ respectively. This of course follows from the standard Boolean relationship:

$$X\,Y + \overline{X}\,Z = X\,Y + \overline{X}\,Z + Y\,Z$$

The equations thus derived were equations (1) and (2) which were used to realize the circuit of Fig. 9.

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Teaching Machines

Dawe Instruments Ltd are planning to produce a range of programmed teaching machines. The first of these is the type 2801 Primer, a simple battery-operated question-and-answer device. This is shortly to be followed by the Dawe type 2806 Universal Tutor, an altogether more elaborate machine with full linear and branching facilities.

The type 2801 Primer is a simple framing device for displaying a sheet of questions and answers. Up to thirteen questions are set out on the sheet, for which the corresponding answers must be chosen by pressing the appropriate button. Coloured lights indicate to the pupil whether he has selected the right or wrong answer. A scrambler switch prevents the pupil remembering which answer button corresponds to which question.

The simplest form of programme presents thirteen questions with thirteen possible similar solutions. A further stage is the presentation of a smaller number of questions with several alternative answers to each. The machine itself is equally suitable for teaching multiplication tables, historical dates or place names, missing words in sentences or the answers to arithmetic or physical problems. Its uses, in fact, are limited only by the ingenuity of the teacher, who normally prepares his own programmes, questions and answers being typed and then duplicated on ordinary quarto (10 by 8in) paper.

The type 2806 Universal Tutor is a true teaching machine designed to provide facilities for linear and branched programmes without the need for complicated equipment. A paper roll programme is presented frame by frame in a window on the top of the machine, a separate smaller roll being provided for written answers. The two rolls are synchronized to move forward together, the written answer on the smaller roll being moved below a transparent Perspex cover before the correct answer to the previous frame is revealed.

The exact mode of operation depends on whether the programme used is of the linear or branching type. Each frame of a linear programme presents only a small piece of information to which the student is required to respond. After responding, he moves forward to the next frame, proceeding linearly, frame by frame, through the roll. The next frame is selected by pressing one or two white buttons alternatively.

Four additional black buttons are used to handle branched programmes, a selected button staying down until the next button is pressed. With the two white buttons there are thus five possible alternatives which can be provided for each answer, giving four branches for following up mistakes. A reversing button is also provided. Information frames are always 8in wide, but may be 10, 5 or 2½in deep, depending on the type of material. A programme may comprise up to 120 frames of 10in, 240 frames of 5in, or 480 frames of 2½in.

A simple jig is provided for assembling and coding programmes, and although the experience of a specialist programmer is an obvious advantage, the machine is specifically designed to allow teachers to prepare their own material. The programme roll is assembled from quarto sheets fed into the jig, the frames being coded by fixing adhesive spots to the paper in positions indicated by slots in the base of the jig. These spots are searched by a pick-up when the programme is in the machine, the pick-up signals operating the switching circuits. A patent for this system has been applied for.

Facilities are also provided for synchronizing the universal tutor with a slide projector or tape recorder, and two variants, type 2806A and 2806B, will be offered for battery and mains operation respectively. They are of identical appearance, with a 16 by 20in sloping work panel and 6in maximum height, designed to rest on the student's desk top. Weight is approximately 12 lb.

Dawe Instruments are currently negotiating for a large supply of standard programmes (O-level maths, science, etc.) and are offering to adapt existing programmes to their machines. They do not at present plan to undertake programme writing, but are able to offer facilities in getting this done.

A Technique for Probability Density Function Measurement

By P. N. Nikiforuk* and G. Squires*

A method is described for measuring the probability density function of a random signal. Conventional operational amplifiers, an oscillator and a pulse counter are used for this purpose.

(Voir page 355 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 362)

IT is often necessary to measure the probability density function of a random signal of low frequency. This is particularly true in automatic control engineering where such signals are used to evaluate the performance of non-linear and adaptive systems. During the past decade a number of techniques have been developed for this purpose and they are now described in readily available textbooks[1,2,3]. Complexity is a common limitation of many of these techniques and it is the purpose of this article to describe a simple and accurate method of making such measurements. Conventional operational amplifiers, an oscillator and a pulse counter are used for this purpose.

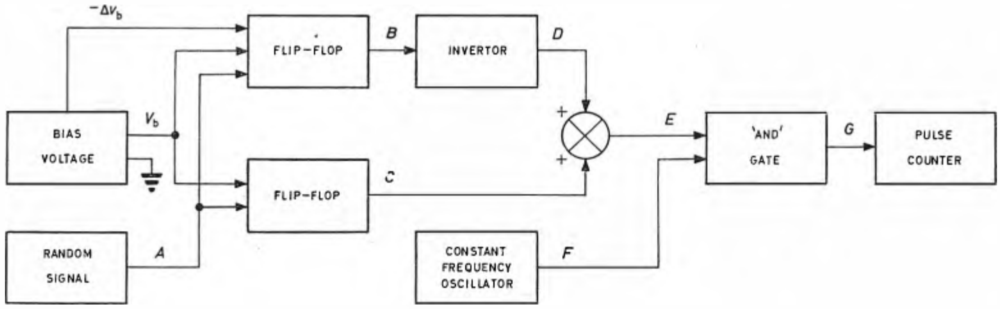


Fig. 1. Arrangement of probability density function analyser

Principle of Operation

Consider a random signal $v(t)$ for which a waveform of sample length T is available. In addition, let T_i be the total time during which this waveform has the amplitude:

$$(v_i - \Delta v) \leqslant v(t) \leqslant v_i \text{ volts}$$

The probability of finding $v(t)$ in this voltage range is T_i/T and the limit of this ratio when divided by Δv as $\Delta v \rightarrow 0$ is the probability density function $P(v_i)$ of $v(t)$ at v_i. Now if Δv is chosen small enough a valid approximation is that:

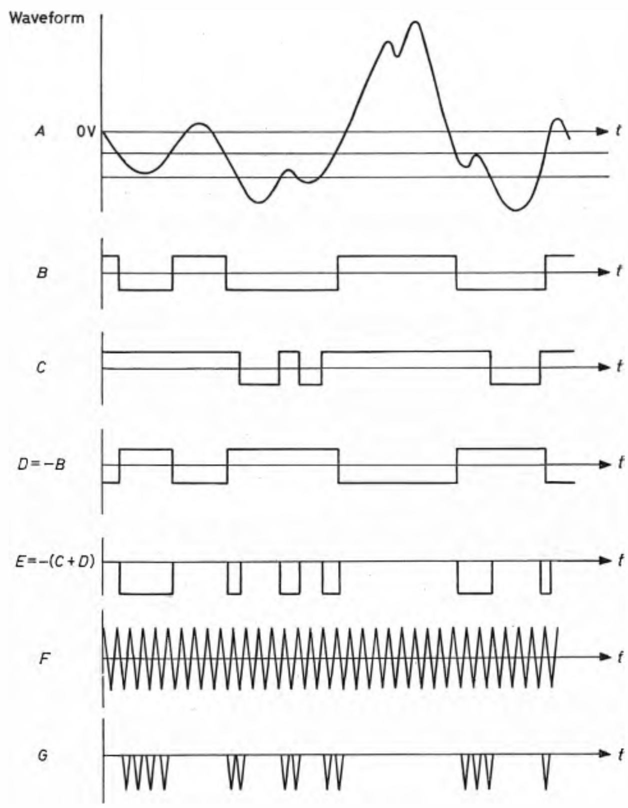
$$T_i/T\Delta v = P(v_i)$$

This expression shows that it is possible to determine the probability density function of a random signal by measuring the time that this signal spends in any given voltage range. A circuit for doing this is shown in Fig. 1.

The operation of this circuit is shown in Fig. 2. The random signal, waveform A, and an appropriate bias voltage V_b are applied to two flip-flops. An additional bias voltage $-\Delta v_b$ is also applied to one of the flip-flops. The effect of these bias voltages is to make the flip-flops change the state of their outputs at $v(t) = -V_b$ and $v(t) = -(V_b - \Delta v_b)$. This operation is shown by waveforms B and C. One of the latter waveforms, B in this particular case, is inverted and added to the other. This addition yields a waveform E which consists of a series of pulses whose widths are proportional to the time that is spent by the random signal in the voltage range determined by the difference in the bias levels.

To measure this time accurately the output from the summer is applied to an AND gate. A second input to this gate is the output, waveform F, of an oscillator whose frequency is kept constant. Since this gate is designed to operate only when both inputs are negative its output

Fig. 2. Typical waveforms associated with analyser



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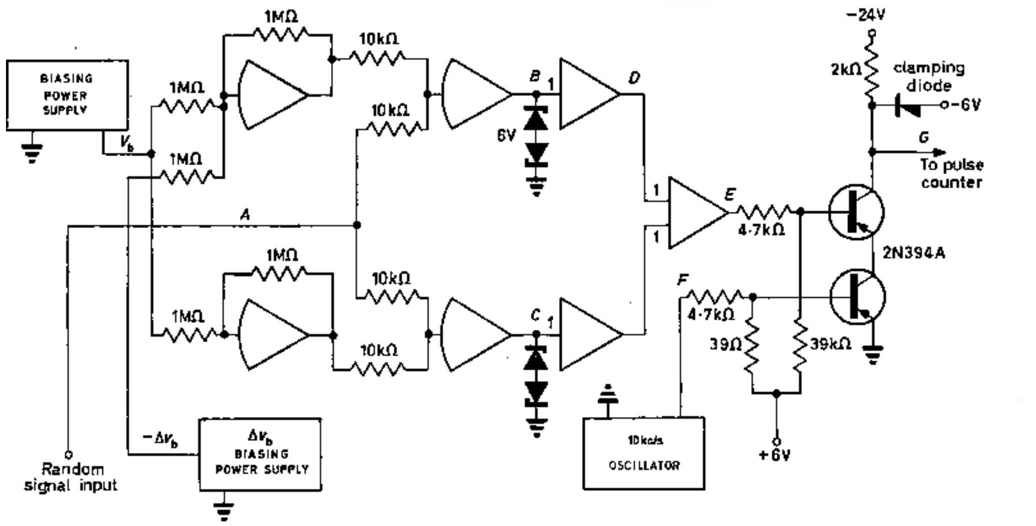


Fig. 3. Practical arrangement of analyser

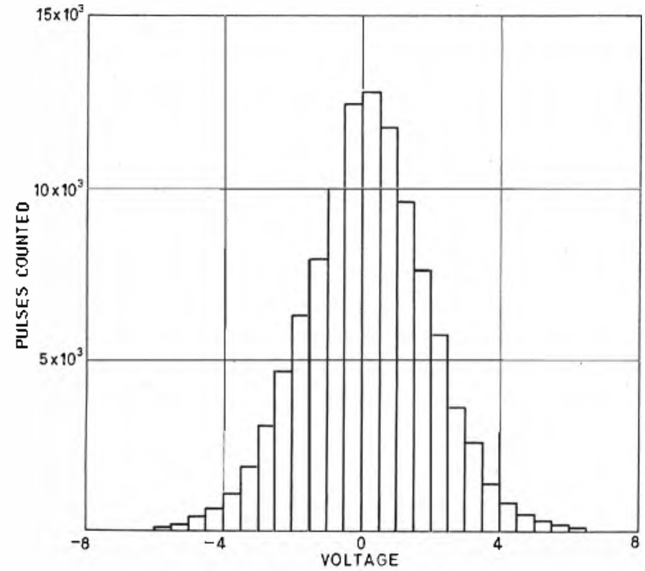
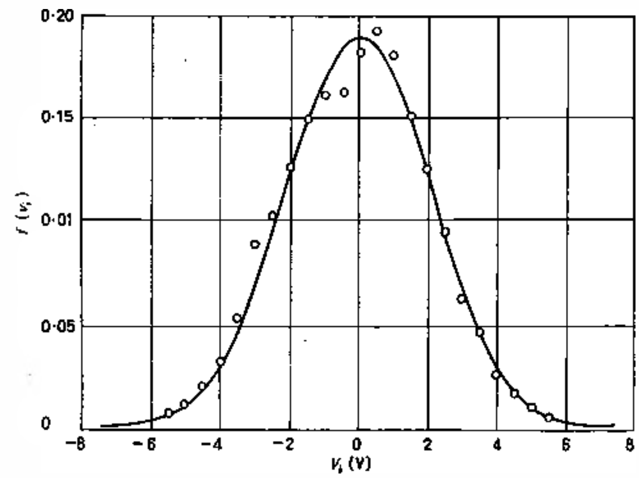


Fig. 4. Typical histogram of analyser

Fig. 5. Probability density function
Experimental points shown by circles
Theoretical curve shown by full line



consists of a series of pulses of the type shown by waveform G. This output is applied to a pulse counter which accurately determines the number of pulses in this waveform. Hence, if N is the maximum possible number of such pulses which can occur in a sample waveform, and n_i is the number actually present:

$$n_i = kT_i$$

$$\text{and } N = kT$$

where k is a proportionality factor. Since $\Delta v \equiv \Delta v_b$

$$n_i/N\Delta v_b = T_i/T\Delta v = P(v_i)$$

which is the desired result.

Experimental System

A circuit of a probability density function analyser which has been used for some time is shown in Fig. 3. Operational amplifiers having unity gain and 1MΩ input resistors are used for the input stage to prevent loading of the bias voltage supplies. Operational amplifiers connected in the open loop mode are also used for the flip-flops. A 10kc/s sinusoidal oscillator is used to supply one of the inputs to the AND gate. The output of this gate is applied to a commercially available pulse counter. A typical histogram that has been obtained in this manner is shown in Fig. 4. The corresponding probability density function is shown in Fig. 5 together with the theoretical curve. A value of $\Delta v_b = 0{\cdot}50$V was used to obtain the experimental curve. The value of Δv_b is somewhat large and a better approximation to the actual probability density function could have been obtained by using a smaller value of Δv_b.

A problem that is associated with this type of analyser is that the operational amplifiers occasionally display a small hysteresis effect when operated in the open loop mode. When this occurs the number of pulses indicated by the pulse counter must be adjusted accordingly.

Acknowledgments

The authors wish to acknowledge financial assistance of the National Research Council, Grant No. A-1819, and the Defence Research Board, Grant No. 4003-02.

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Operation of a Dekatron Valve Controlled by a Magnetic Field

By A. W. Gooder*, A.M.I.E.R.E. and M. H. N. Potok*, M.A., B.Sc., Ph.D., M.I.E.E., A.M.I.E.R.E.

It is shown, that a Dekatron can be successfully operated by voltage variations of arbitrary shape from d.c. to several kc/s in presence of a suitable magnetic field. The optimum circuit components and voltage have been arrived at experimentally.

(Voir page 355 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 362)

THE Dekatron[1] in its conventional mode of operation requires only electric fields to move the plasma beam from electrode to electrode. In a Trochotron[2] the electron beam in vacuum is made to progress by the application of a magnetic field. It has been observed that a Dekatron can also be made to step by the application of a magnetic field. In fact using the simple circuit of Fig. 1 and a magnet whose axis is concentric with that of the valve as in Fig. 2, the Dekatron is seen to work when driven by a sine wave, whose minimum amplitude depends on frequency as in Fig. 3.

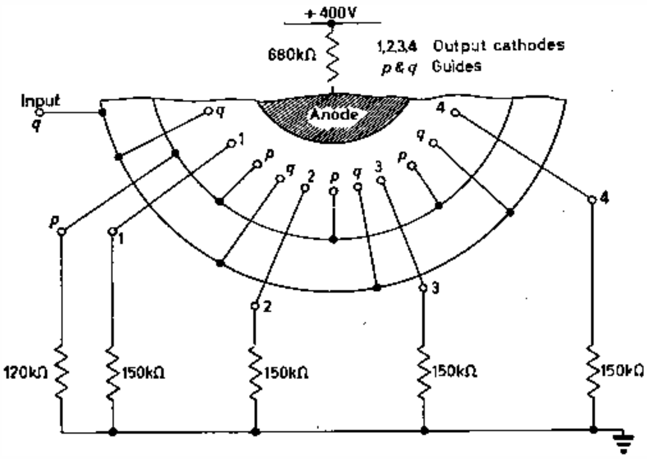


Fig. 1. Circuit used in the experiment (showing only 4 of the 10 cathodes)
The numbered electrodes act as cathodes, p and q refer to auxiliary cathodes internally connected as shown

These results should be compared with those quoted by the manufacturers for this particular Dekatron valve when operated in the conventional manner, i.e. driving pulse minimum amplitude 145V, maximum frequency 4kc/s.

It is useful to note that whereas there appears to be no deterioration at the upper frequency limit, the magnetic field permits operation right down to d.c. at very much reduced voltage changes.

This result led to further investigations.

Operation of Circuit when Driven by Variable Slope and Duration Pulses

VERY SLOW VOLTAGE CHANGES

A variable d.c. voltage was applied to the input of the circuit shown in Fig. 1. When the circuit was first switched on with guides q at earth potential discharge established itself between the anode and one of the guides q.

Raising the voltage to guides q slowly, it was found that when that voltage reached about 16V, the discharge transferred to the nearest cathode and guide p in the direction fixed by the magnetic field. The discharge continues to embrace both these electrodes until the voltage of q has fallen to below 4V, at which moment the discharge transfers to the next q electrode moving in the

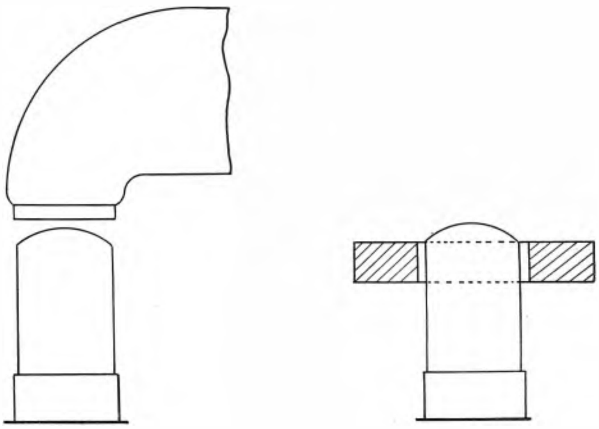
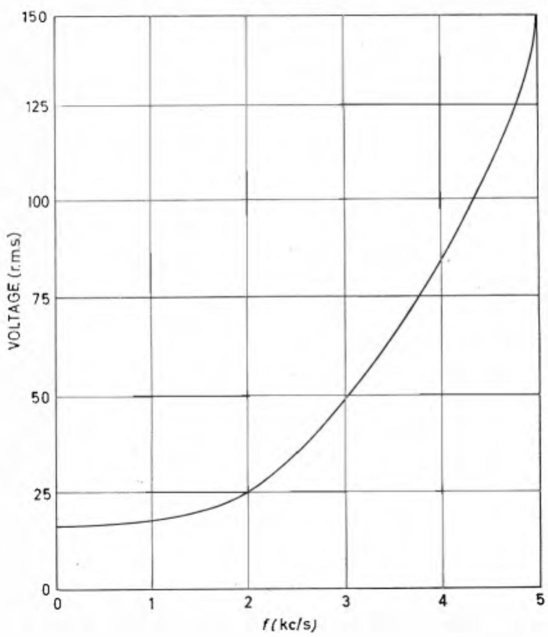


Fig. 2. Diagrammatic representation of the Dekatron and the magnet in position
On the left a 2 000 gauss magnetron magnet; on the right a 400 gauss Mullard FD112 focusing magnet used in most of the experiments

Fig. 3. Relation between input driving voltage and frequency with magnet in position



* Royal Military College of Science.

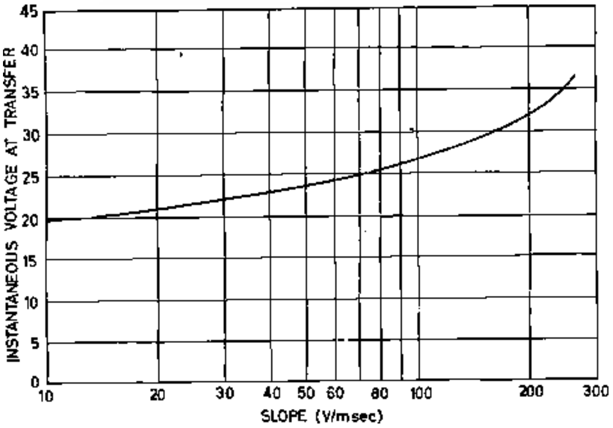


Fig. 4. Relation between voltage at which transfer occurs and the rate at which voltage increases at the input

same direction as before, fixed by the magnetic field. The voltages quoted are approximate and depend on the actual geometry within the valve, i.e. if the consecutive electrodes are slightly bent away from the nominal circle during manufacture, as seems very common on the valves examined, the voltages may vary, variations of up to ± 2V were observed.

Variable Slope Pulses

Pulses of variable rise time and height have been generated at a constant frequency of 100c/s. The slope of the pulse and the voltage at which transfer takes place has been measured. The results are given in Fig. 4. These were carried out with a 2000-gauss magnet.

Operation with Rectangular Pulses

A rectangular wave with variable mark-to-space ratio, but constant repetition frequency of 1 000c/s was applied. Both the zero level and pulse amplitude could be adjusted independently. While mark-to-space ratio was varied from 1:9 to 9:1 the amplitude and zero level were adjusted just to produce a transfer after each pulse edge. The results are shown in Fig. 5 giving minimum voltage for transfer from cathode to guide followed by the necessary change for transfer to next cathode for various lengths of pulses.

It will be seen that 100μsec is about the shortest pulse for reasonable switching voltage, with 100μsec minimum between pulses, i.e. a maximum p.r.f. of 5kc/s.

The pulse in the output is slightly delayed behind the input pulse, the delay depending on the leading edge slope and pulse height as seen in the c.r.t. traces reproduced in Fig. 6. This delay is constant for any one cathode but varies from cathode to cathode because of constructional imperfections. This results in a certain amount of pulse width noise of no consequence in counting.

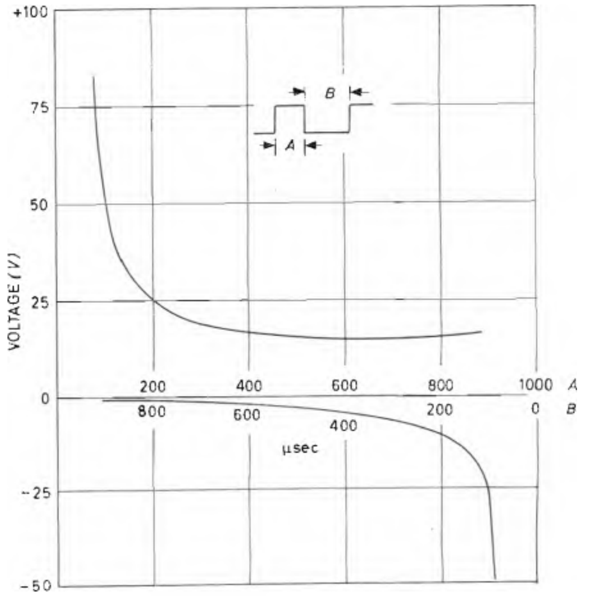


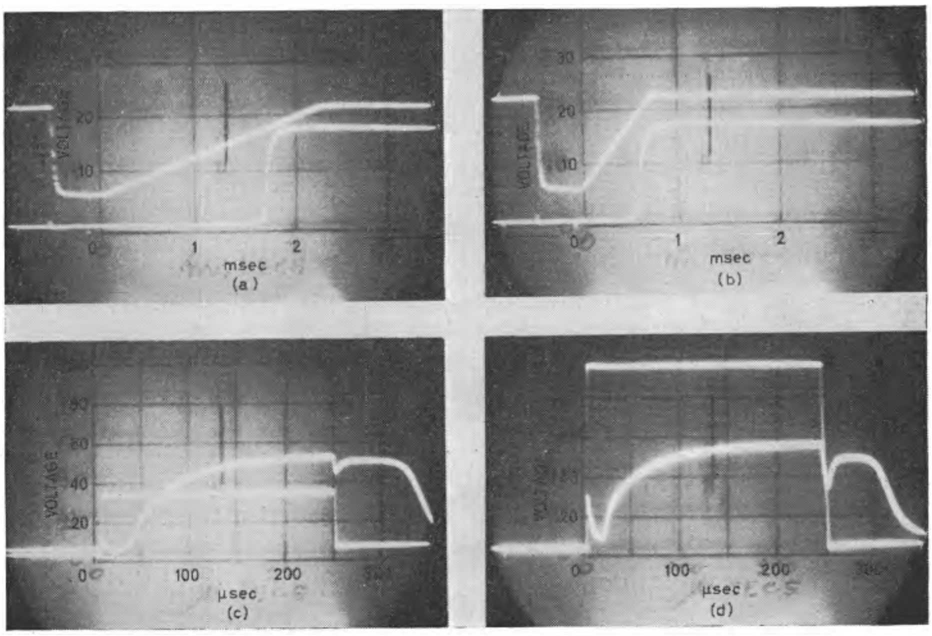
Fig. 5. Relation between length of pulse and pulse height required to achieve successive transfers from cathode to guide (top, scale A) and from guide to next cathode (bottom, scale B)

Choice of Circuit Parameters

Experiments were carried out to determine the optimum values for the resistors in the cathode and guide circuits of Fig. 1. This was done by choosing a value for the cathode resistor (R_k) and measuring the voltages required for transfer with different values of guide resistor (R_g). Some typical results are given in Fig. 7. The top (straight) lines correspond to the voltage on the guide q at which transfer takes place from a guide q to a cathode. To achieve transfer from a cathode, the voltage to q has to be reduced to a corresponding point on the curve below. The wide variations between the various cathodes is clearly shown here. The choice of optimum value is dictated by finding a range over which the switching can

Fig. 6. Traces showing the effect of increasing pulse forward slope (a and b) and pulse amplitude (c and d) on delay in transfer

The voltage scale shown refers to the driving pulse. The output pulse is magnified three times



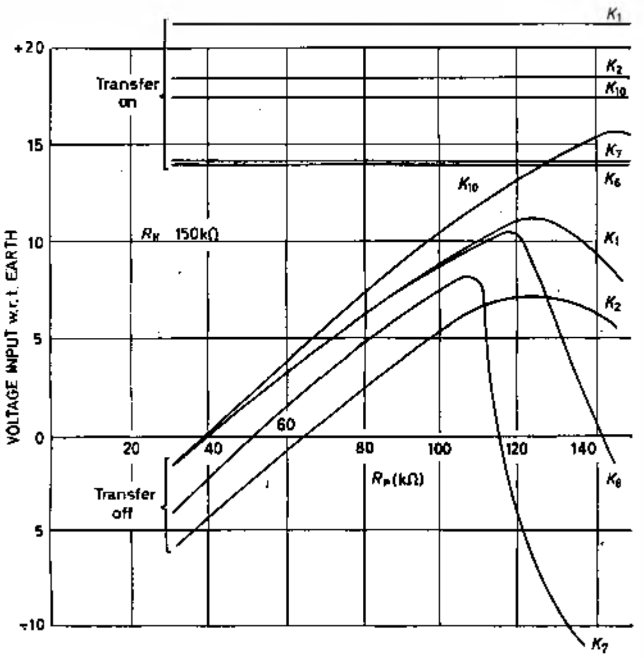


Fig. 7. Voltage input required to achieve transfer of plasma to successive electrodes for one value of R_k showing differences between cathodes

"Transfer off" refers to transfer from cathode to the following guide. "Transfer on" refers to transfer from guide to following cathode, completing the transfer

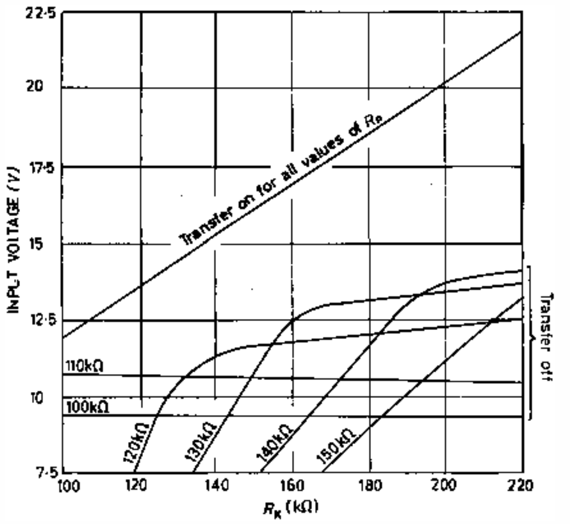


Fig. 8. Voltage input required to achieve transfer to guide (transfer off) and transfer from guide to cathode (transfer on) for various values of guide resistance R_p and cathode resistance R_k for one particular cathode—cathode sector of the Dekatron

be done reliably with least voltage change. It is seen to lie between 90 and 110kΩ in the guide circuit.

A similar experiment was carried out to find the best value for R_k. This time R_p was kept constant while R_k was changed and voltages were noted for transfer to and from the cathode. Results for a few values are shown in Fig. 8. These indicate that the best value for R_k is of the order of 150kΩ.

Conclusions

The results of these experiments show that although the application of a magnet makes it a little more difficult to reverse the direction of rotation and there is at best no improvement in the switching speed, many very useful characteristics can be obtained. The first of these is the simplicity of the driving mechanism, which allows one to use pulses or sine waves in the input without any need for processing or close amplitude control.

The second in order but not in importance is the fact that there is no lower frequency limit, i.e. the device can operate on a slowly varying or drifting input voltage. It can for example be made to count how often the voltage has risen above and fallen below certain preset limits without any need for intermediate switching circuits such as Schmitt triggers etc.

Thirdly, by using an electromagnet it should be possible to suppress or to reverse counting by variation of current through the magnet.

Fourthly, the output from the device is just large enough to drive the next Dekatron provided initial frequency is limited to below 1kc/s. It is possible that some changes in design could overcome even this limitation. At present the lack of precision in the cathode-guide circle means that when the signal is close to the minimum, the incidence of noise and structural irregularities may produce errors in counting. Possibly a magnetically controlled Dekatron needs closer tolerances or even some modifications in design.

Acknowledgments

The authors wish to thank the Dean Sir Donald Bailey, O.B.E., D.Eng., for his permission to publish this article.

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A New Television Camera

A new 4½in image orthicon television camera has recently been announced by the Marconi Co. Ltd. Fully transistorized, this camera, the mark V, weighs only 98lb and is the smallest and lightest 4½in image orthicon camera in production anywhere in the world. It is also the simplest to operate. The stability and reliability of the design have made it possible to relegate all controls for the camera electronics to the equipment racks, with the exception of the on/off functions. This leaves the operator entirely free to concentrate on controlling the position and field of view of the camera at the direction of the producer.

Other major features of the new camera include a single integrated zoom lens which frees the camera from the restriction of having a number of lenses of fixed focal length. A unique tilting viewfinder is provided, to ensure that the cameraman can always obtain the most comfortable viewing position at any camera angle. This viewfinder is four times brighter than its predecessors, and can display the picture from the camera, a picture from an outside source, or a mixture of the two.

Silicon transistors are used almost entirely, and the stability of the design is such that the output remains constant, irrespective of variations in the mains supply, changes in temperature, and external magnetic fields. This stability has made it possible to remove all the camera 'setting-up' controls to the equipment racks, where they can be treated as preset adjustments, normally left alone except for routine maintenance.

The use of transistors has also made it possible to combine the camera control unit, camera control panel and power supply unit, in a single rack mounted unit. The few controls which must of necessity be used during a production, controls which are affected by changes in scene illumination, are housed in a small operational control panel, designed to fit into the standard production console. Four of these units will fit into a single desk well.

Both units in this new camera channel have been professionally styled. The camera is finished in two colours with an almost indestructible enamel. The external lines are clean and functional, and the carrying handles down each side of the camera make the mark V extremely easy to carry, an important point on outside broadcasts.

An R.M.S. to Mean Convertor

By H. H. A. Bath*

A design procedure is given of a method for obtaining a d.c. output voltage from an r.m.s. voltage of any form factor having a maximum value of not more than 4. A number of diodes are used which convert the applied signal to follow a square function with respect to a charge current flowing in a capacitor.

(Voir page 355 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 362)

IN cases where it is required to record the r.m.s. value of a particular waveform and the form-factor of this waveform varies, solutions can be found by using conventional r.m.s. reading instruments. However, if this r.m.s. value is required to drive some auxiliary equipment and the poor overload and damping characteristics of a thermo-vacuo junction device are unsuitable, a solution can be found in a diode squaring circuit. Basically, this provides a non-linear resistive path to a waveform of any function and achieves this by approximating to a parabolic relationship between instantaneous input current and voltage. This method has been adopted by a number of manufacturers and it is currently being used for controlling the heater power in an evaporation coating plant.

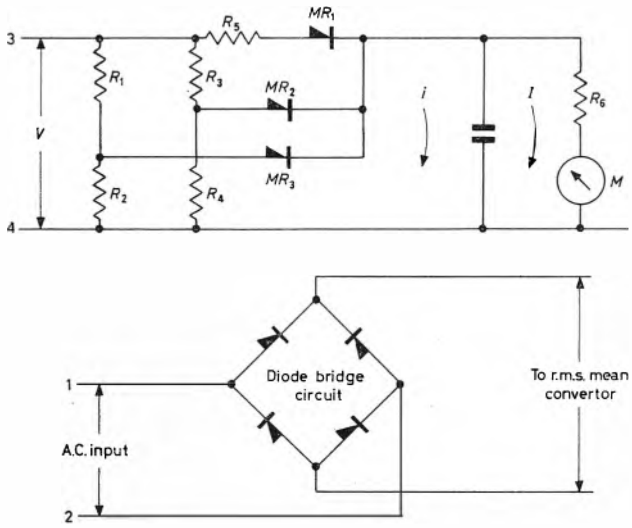


Fig. 1. The r.m.s. to mean convertor and a.c. bridge input circuits

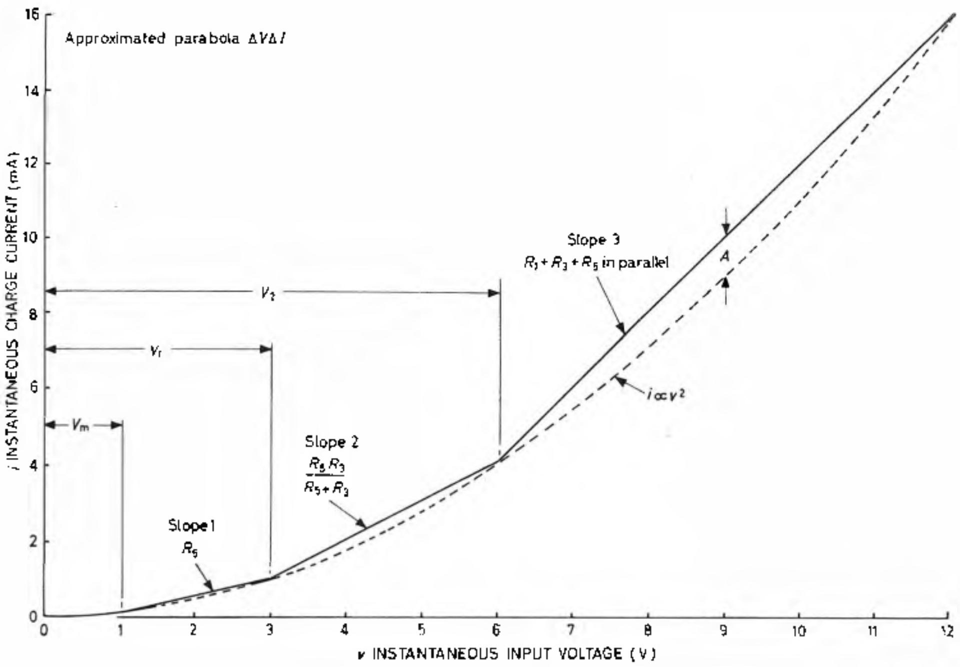
Circuit Design

The waveform to be measured is full wave rectified so that only the modulus of this waveform appears across the input to the convertor. In the circuit shown in Fig. 1, as the voltage V *rises* MR_1 conducts and charge current i flows through R_5. Each diode MR_1, MR_2 and MR_3 conducts successively at higher input voltages, thus making the effective resistive path to the capacitor lower. Charge current increases and follows the path of a predetermined parabola as indicated in Fig. 2.

The basic explanation of the circuit may be understood from the following. Fig. 2 represents an arbitrary curve $i \propto v^2$. The ratios of V_m to V_1 and V_2 also have conveniently been fixed i.e. 1 to 3, 1 to 6 etc. However, any combination of straight lines may be drawn on the parabola. V_m represents the d.c. voltage across C and V_M is its value at f.s.d. In Fig. 2 the scales shown apply when $V_m = V_M$, i.e. at full scale deflexion on M. When $V_m = KV_M$ both the voltage and current scales should be multiplied by K. Any input voltage waveform producing a mean charge I_{max} will give f.s.d. on the output meter and will have been produced from a waveform following a V^2 relationship. But I_{max} is equal to the instantaneous current when $v = V_1$ (d.c. conditions). Again taking the simplest case, that of d.c., at V_1 (r.m.s.) and referring to Fig. 2, $I_{mn} = 1$, and at $\frac{1}{2}V_m$, the r.m.s. voltage has dropped by $\frac{1}{2}$ to $V_1 = 1{\cdot}5$.

These results depend on the simple voltage divider network in Fig. 1 consisting of R_3 and R_4. The voltage break points at V_1 and V_2 are also in the same ratio as before. Hence a waveform producing f.s.d. working on the parabola (shown in Fig. 2) has an equation $i = av^2$. At $\frac{1}{4}$ f.s.d. of the output meter it can easily be shown that the value of this has halved. Similarly taking any arbitrary output

Fig. 2. Approximated parabola $\Delta V/\Delta I$



* Central Research Laboratories, Edwards High Vacuum International Ltd.

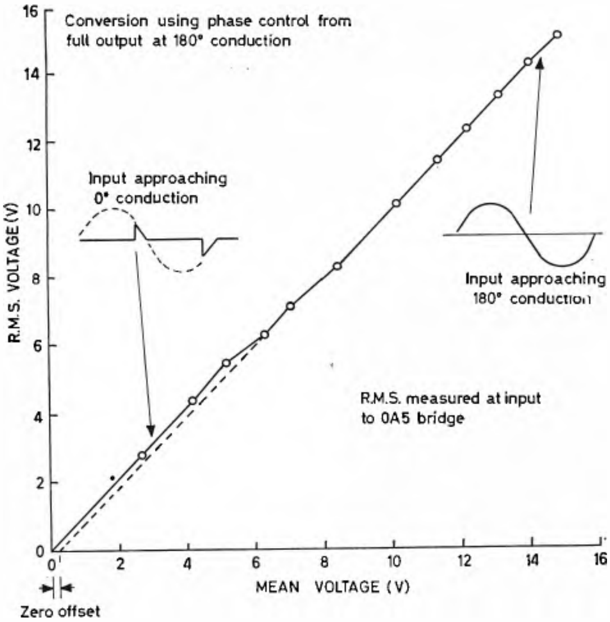
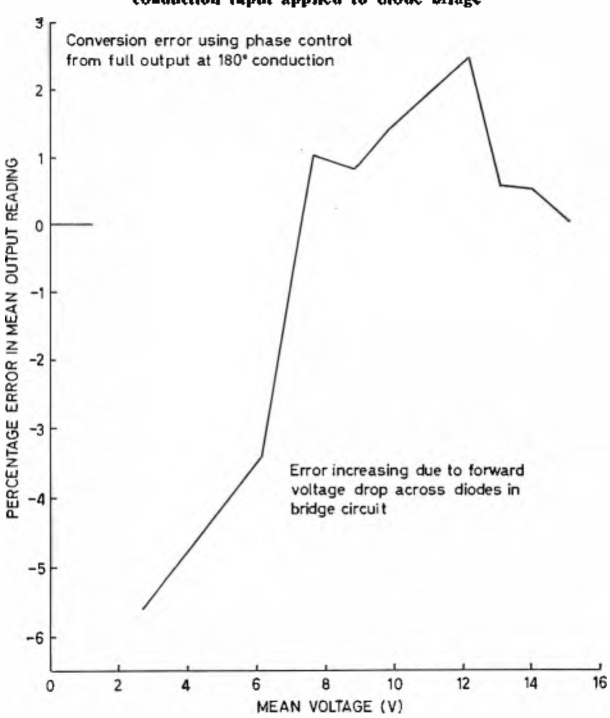


Fig. 3. Conversion using phase control from full output at 180° conduction. Input applied to diode bridge

reading, there will be a linear relationship between the mean output current and the ratios of the values of a. Thus, any r.m.s. input waveform which is varied over the range of the mean reading instrument produces a linear scale.

All this, however, is based on an ideal circuit and in practice errors can arise both from the values of circuit components and shape of input waveform. These can be minimized by careful selection of circuit components. From Fig. 1 the maximum possible error occurs at A which arises typically, for chopped input waveform (crest factor 3). This error can be reduced by incorporating an extra break-point, but after four such points practical limitations occur: the change in slope resistance will be determined solely by the variations of forward resistance of the diodes in circuit which also have a temperature dependent forward voltage drop. This can be made negligible by providing a high mean capacitor voltage. Furthermore the reverse leakage current of these diodes has to be of sufficiently low value not to affect greatly the capacitor charge current.

Fig. 4. Conversion error using phase control from full output at 180° conduction input applied to diode bridge



In the design of this circuit the source impedance of the supply also requires consideration. Thus in the above analysis it was assumed that the circuit would in no way load the supply to such an extent that regulation becomes a major factor.

Experimental Results

A requirement existed for a convertor which could accept an r.m.s. input voltage of 15V, be capable of working at this level with crest factors of up to 4, and that 5V d.c. be delivered across a 5kΩ resistive load. The forward voltage drop across each of the diodes MR_1 to MR_3 was included in the equation for calculating the circuit constants R_1 to R_5 shown in Fig. 1. A thermal milliameter together with a series resistance placed across

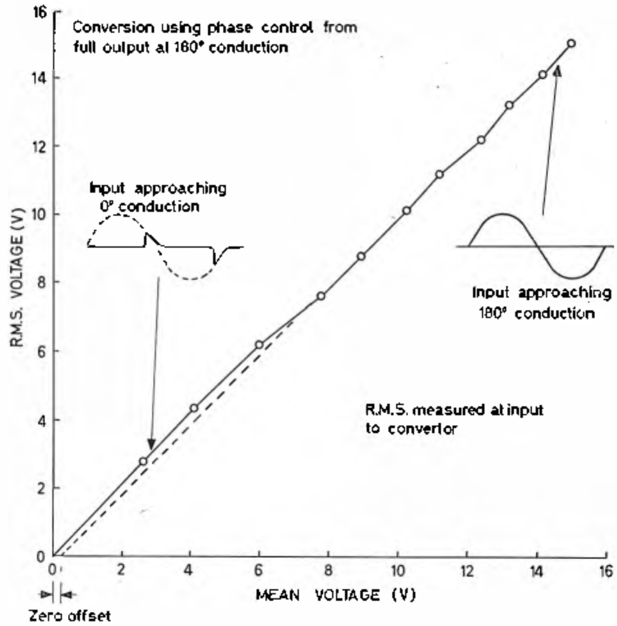


Fig. 5. Conversion using phase control from full output at 180° conduction input applied directly to convertor

the convertor input, monitored the r.m.s. value of applied voltage.

Since it was intended that the convertor should form part of a feedback loop with an a.c. phase controlled synchronous switch (typical waveforms are shown in Figs. 3 and 5) conversion characteristics were compiled with the r.m.s. voltage measured initially across the a.c. arms of the diode bridge and then at the input to the convertor. (See Fig. 1—points (1) and (2), (3) and (4). Figs. 3 and 5 show a slight zero offset indicating a constant error in the mean voltage. Fig. 4 shows the percentage error computed from the results from which Fig. 3 was obtained, and also the increasing error at the lower output voltage levels. Note that when the input is applied through a diode bridge the errors are larger than when the input is measured directly across the convertor as seen by comparison with Fig. 6. Fig. 5 shows the r.m.s. to mean conversion with the input measured across (3) and (4) (see Fig. 1).

Conclusions

With careful circuit design and close tolerance com-

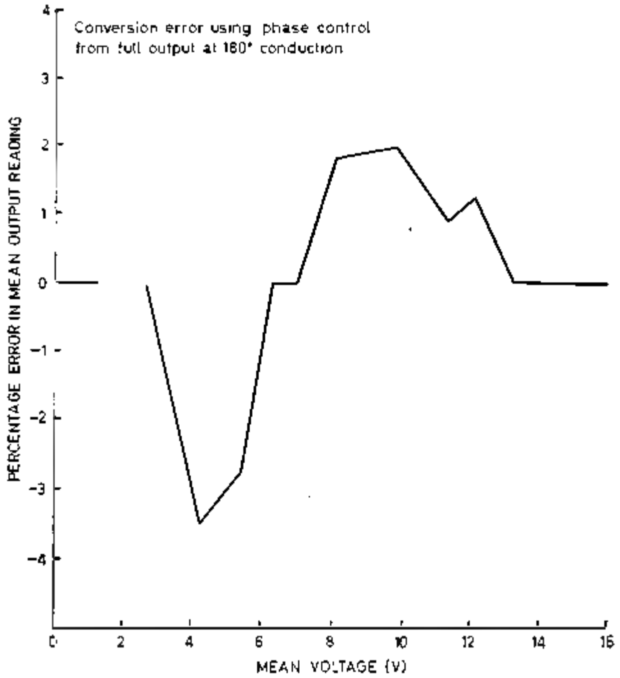


Fig. 6. Conversion error using phase control from full output at 180° conduction input applied directly to convertor

ponents, i.e. ±1 per cent there is no reason why the convertor should not give results approaching this value.

Since the conversion factor is theoretically constant at all values of input voltage the output meter can be linearly scaled. The chief source of error is due to the forward voltage drop across the diodes in circuit and thus at very low input levels the output reading is not predictable (see Figs. 3 and 5).

The total loading on the input supply increases virtually in steps as each section is cut in and will be considerably higher than the maximum value of instantaneous current, this step loading will have to be taken into account since adding resistance in series with the input produces a non-parabolic relationship between the two components of power.

If a.c. quantities are to be measured, it will be necessary to have a bridge rectifier preceding the convertor. The elements in the bridge circuit should not introduce a significant change in the voltage wave shape, i.e. to ensure that the r.m.s. or crest factor remains virtually constant. In order to minimize errors from this source the bridge should comprise low forward drop elements.

For current measurements this circuit will require a transformer capable of handling the high crest factors anticipated. Also distortion will be a limiting factor and the types of core material will play an important part. A low value current sensing non-inductive resistor placed in the load circuit could be used as the primary load of this transformer. Particular care would have to be paid to the total reactance and the resistance on the secondary side of the transformer. For reasons stated earlier these would have to be kept low.

Diode convertors have been used to transform waveforms from one function to another in a variety of applications, e.g. in flight simulator computers, low frequency oscillators and a number of manufacturers are using r.m.s to mean convertors for sensing r.m.s. load voltages and currents in commercial equipment.

As an r.m.s. reading instrument this circuit should prove a useful alternative to the conventional hot wire, thermocouple or moving-iron non-linear scaled meters and a d.c. output signal is certainly an attractive advantage.

Acknowledgment

The author wishes to express his gratitude to P. Granger and L. Higgins for their help with the apparatus and assistance in making measurements, to L. Holland and W. Steckelmacher for their continued encouragement and helpful criticism, and to Edwards High Vacuum International Ltd for kind permission to publish this article.

An Improved Raidotelephone System

Engineers at the Post Office Research Station, Dollis Hill, London, have developed a new system which will greatly improve long-distance overseas radiotelephone calls. Called LINCOMPEX, this application of electronic techniques enables satisfactory telephone service to be provided over long distance radio circuits even when radio reception conditions are so bad that no service would be possible using the conventional equipment.

In the new system, the syllable to syllable variations of loudness of speech are smoothed out, by means of a 'compressor' and the speech applied to the radio transmitter is at a constant high level or loudness. The original variations in loudness are, however, signalled to the receiver at the other end of the circuit over a separate 'control' channel. In the receiver the control signal, by means of an 'expander,' restores the loudness variations to the received speech. Hence the name LINCOMPEX which is short for 'LINked COMPressor and EXpander.'

With the conventional system, it is necessary to provide switching, operated automatically by the talkers' voices, to prevent the circuit from becoming unstable and breaking into oscillation. It is possible, however, when radio conditions are bad, for these switches to be operated by the noise on the radio circuit and to produce a state of 'lock-out.' when one talker cannot break in and is unable to make himself heard. LINCOMPEX requires no such voice-operated switching, so that there is no danger of 'lock-out.' In addition, it gives a much quieter background to the speech.

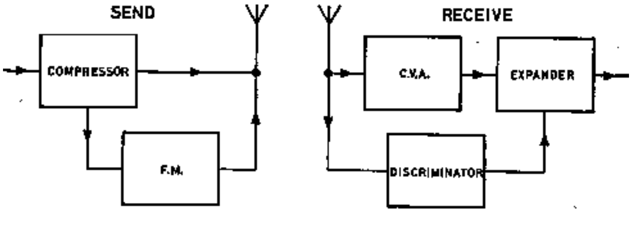
The principle of the new system is shown in Fig. 1. The control signal is transmitted over the radio link by frequency modulation of a sub-carrier, because this type signal is relatively immune to the effects of fading.

The control signal is derived at the sending end by rectification of the input speech. It is applied to the compressor and causes it to give a constant output level. The control signal is also applied to a frequency modulator the output of which is combined (by means of filters not shown) with the compressed speech, and passed to the radio transmitter.

At the receiver, the two signals are separated by filters (not shown in Fig. 1). The effect of variations of signal strength due to fading will have been considerably reduced by the normal automatic gain control of the radio receiver, but any residual variations of level of the speech are removed by a constant volume amplifier. The control signal is demodulated, in the conventional manner for frequency modulated signals, by a discriminator. The speech is then applied to an expander, which, under the control of the recovered control signal, restores the original speech level variations.

Since the speech and control channels are of widely differing bandwidth, their times of transmission from input at the sending end to output at the receiver necessarily differ appreciably. It has been found that an improvement in performance is obtained by equalizing these times. This is achieved by inserting delay networks (not shown in Fig. 1) in the speech channel.

Fig. 1. Principle of LINCOMPEX system



A Transistorized Demonstration Set for Quantized Pulse Amplitude Modulation

By J. Karfunkel, Dipl.-Ing.

All pulse modulation with a finite number of pulses per second shows quantization effects, depending on the pulse frequency in pulse amplitude and pulse time modulation systems, or on the number of stages in pulse code modulation. These effects are known as quantizing noise. To facilitate the study of quantizing noise and intelligibility under different conditions a demonstration set for quantized pulse amplitude modulated signals has been developed. One existing apparatus[1] is based on about 50 valves and controls, so transistorizing and simplicity were the main objects.

(Voir page 355 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 362)

THE demonstration set (Fig. 1) transforms audio frequency into quantized pulse amplitude modulation (q.p.a.m.) and back into audio frequency. For this purpose the audio signal is fed through a two-stage a.f. amplifier into a pulse amplitude modulator. A square wave from an external generator is also fed into the pulse amplitude modulator. The resulting pulse amplitude modulated wave (p.a.m.) triggers a group of ten blocking oscillators. Each of these blocking oscillators can be triggered at a different level and produces a controllable output pulse. The total of all pulses passes through a pulse amplifier and an emitter-follower to the output. Here it can be examined by an oscillograph or drive a pulse code modulator.

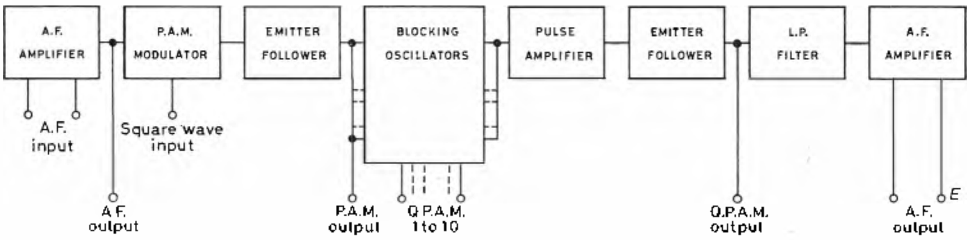


Fig. 1. Arrangement of demonstration set

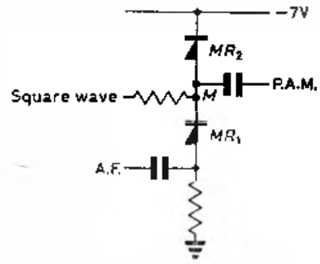


Fig. 2. Pulse amplitude modulator

In order to facilitate the aural and visual comparison between the incoming audio signal and the distorted signal which results after demodulation of the q.p.a.m. signal, a demodulator is attached. This consists of a low-pass filter with a volume controlled a.f. amplifier, which produces power for a 3Ω speaker.

The P.A.M. Modulator (Fig. 2)

The modulator contains passive elements only. Diode MR_1 restricts mixing point M to negative pulses only. Without any a.f. signal point M shows a negative going square wave with constant amplitude. The connected a.f. signal limits through diode MR_1 the square wave amplitudes, therefore producing a p.a.m. signal. As this p.a.m. modulator gives satisfactory results but is very sensitive to the load, it was coupled to the triggering stages by an emitter-follower.

The Blocking Oscillators (Fig. 3)

Several special requirements restricted the choice of the blocking oscillator.

(1) Besides producing a sharp pulse it should have a fairly constant trigger level, independent of the normal production spread of transistors.

(2) Most blocking oscillators show a transition zone—a zone in which the output pulse is not yet fully developed.

(3) They show a frequency dividing effect—triggering only every second, third or even tenth triggering input pulse.

(4) There should be no overshoot.

(5) The trigger level must be variable.

(6) The output pulse amplitude must be variable.

The blocking oscillator shown in Fig. 3 triggers at 0·5V with a transition zone of 2mV. The frequency dividing effect is eliminated by reducing the source voltage to 6·5V. Capacitors C_{12} and C_{13} take care of the overshoot, the transition zone and the sharpness of the pulse (Fig. 4).

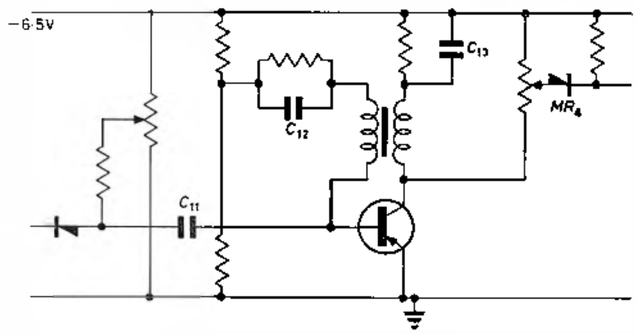


Fig. 3 (above). Trigger stage

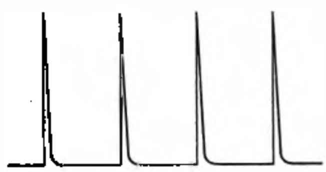


Fig. 4 (right). Output pulse

The coupling capacitor C_{11} was chosen very small in order to avoid triggering of a blocking oscillator from another through the base.

The pulses of all blocking oscillators are collected by a busbar. But only the pulse with the highest amplitude is wanted at the q.p.a.m. output, therefore all pulses are fed through diodes MR_4. With increasing loading there is an increasing interaction between the output pulses: the amplitude control of a lower pulse influences the amplitude of a higher pulse owing to the finite resistance of the diodes. Resistor R_{22} in series with the input impedance reduces this effect to a tolerable level.

The Complete Circuit (Fig. 5)

The a.f. and the square wave input are connected through potentiometers. The amplified a.f. signal can be observed before entering the p.a.m. modulator. The p.a.m. can be observed before triggering the blocking oscillators. There is a possibility of choosing stages 1 to 5, 6 to 10 or both together by using two stage switches; this facilitates the comparison of two different modulation characteristics.

Conclusion

The demonstration set is suitable for audio frequencies in the speech band from 100 to 4 000c/s, and for pulse repetition rates from 10kp/s to 50kp/s. Fig. 6(a) shows the p.a.m. of a 500c/s sine wave with a square wave of 15kp/s, Fig. 6(b) shows the q.p.a.m. by using a linear modulation characteristic and Fig. 6(c) shows the resulting a.f. wave after demodulation.

Acknowledgment

The apparatus described in this article was developed in the Electrical and Control Engineering Department of Battersea College of Technology, London. The author, who was then on leave from the Technion, Israel Institute of Technology, is grateful to the head of the department, Mr. W. F. Lovering, for granting financial support and laboratory equipment.

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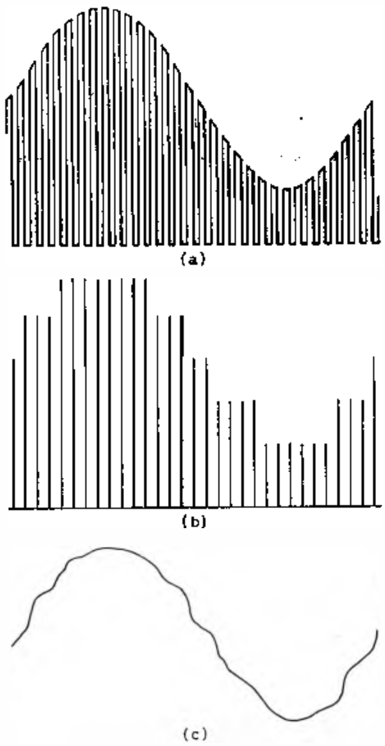
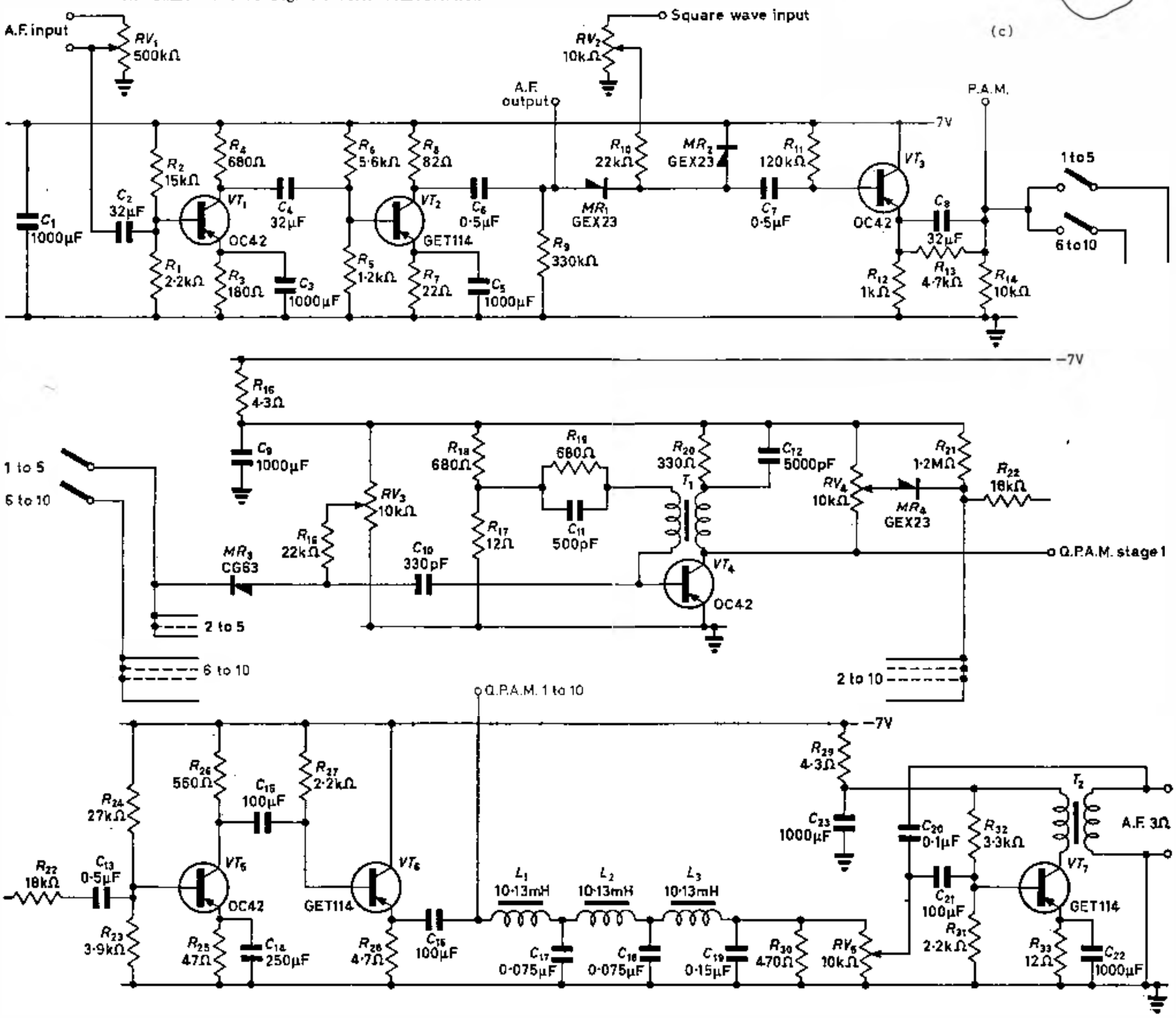
Fig. 5 (below). Complete circuit diagram

Fig. 6 (right). (a) Sine wave of 500c/s pulse amplitude modulated by a square wave of 15kp/s
(b) Same wave as Fig. (a) after linear quantization with 5 stages
(c) Same wave as Fig. (b) after demodulation





A Simple Transistor Servo-Simulator

By P. G. M. Dawe*, M.A., M.Sc.(Eng.)

A servo-simulator is described which consists of a number of single transistor shunt feedback amplifiers. These perform the functions of summing, integrating, phase-advancing and reversing. The amplifiers are capacitively coupled, and the integrators are compensated for quadratic error. The response to a step input at 50c/s can be examined for various damping arrangements, and a diode clamp is provided to damp the system on the negative half cycles of the input square wave.

(Voir page 355 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 362)

It may not be widely realized that an introduction to the features of an automatic control system, or servo-system, or to the ideas involved in setting up a transfer function on an analogue computer, can be simply given by the use of a few transistor amplifier circuits. Some circuits have already been given which illustrate the similarities between the error actuated control system and V_e, the error voltage is made proportional to θ_e. Normally the field current of the motor, I_f, is made proportional to this error voltage, and therefore the torque or rotating force produced by the motor on the output shaft is also proportional to V_e, since the torque depends only on the field current, if the armature current is maintained constant.

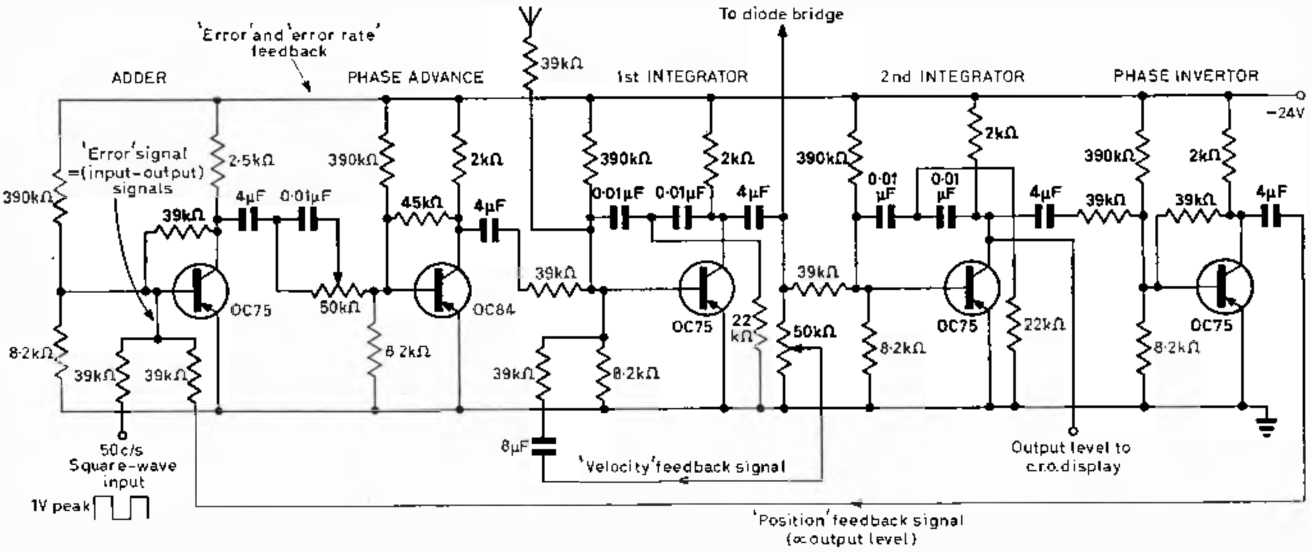


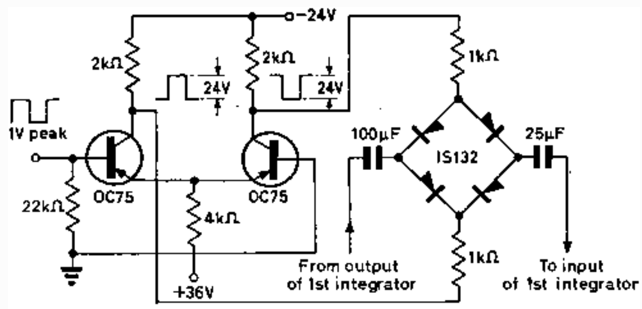
Fig. 1. The servo-simulator

the negative feedback amplifier[1,2]. It is suggested here, however, that the matter can be more fully considered in terms of representing the second order differential equation, normally characteristic of many servo or automatic control mechanisms, through simple integrating, summing and phase reversing operational amplifiers.

In a remote position control servo-system, such as may be employed to operate a gun turret, or steering mechanism of a ship, the position of an output shaft is controlled by the position of an input shaft. Power amplification is normally provided between the input and output shafts so that a considerable load at the output may be driven from a negligible operating force at the input. For this purpose, an electric motor, operating through a gear reduction, normally sets the position of the output shaft, and the motor is driven by a signal voltage, V_e, known as the error signal, which is proportional to the difference between the input and output shaft positions. The voltage V_e is obtained as the difference between the voltage outputs of two potentiometer transducers which are mounted on the input and output shafts respectively.

One has, then, that $\theta_i - \theta_o = \theta_e$ the error, where θ_i and θ_o are the angles of the input and output shafts, and

* *Institute of Experimental Psychology, University of Oxford.*



PARAPHASE AMPLIFIER DIODE GATE

The equation of motion of the output shaft relates its angular acceleration, $d^2\theta_o/dt^2$ to this driving torque, as follows:

$$\text{torque, } T \propto I_f \propto V_e = J\,(d^2\theta_o/dt^2) \quad \ldots\ldots\ldots \quad (1)$$

where J is the inertia referred to the output shaft.

The solution of this equation shows that for an input step angle of θ_i the output will take the form:

$$\theta_o = \theta_i\,(1 - \cos \omega_n t) \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

where $\omega_n = \sqrt{(k/J)} = 2\pi f_n$ and f_n is termed the natural undamped frequency of the servo-system.

Equation (2) shows that the system oscillates with an amplitude equal to the original displacement, and is inherently unstable since, although the angular acceleration becomes zero when the error term becomes zero, the motor will then have a maximum velocity which carries it beyond the required position. If a further time-lag is introduced into the system the torque can still be positive when the error is zero, and the oscillations will then build up until limited by the non-linearities of the system. Otherwise, damping may be introduced to reduce or prevent the oscillations in the system. Such damping is usually produced either by viscous friction, or through feedback signals which anticipate the behaviour of the system, such as a velocity feedback of an error rate feedback signal. In such a servo-system the motor, in effect, produces a double integration of the error signal to derive the output displacement, i.e.:

From equation (1)

$$\theta_o = \iint (d^2\theta_o/dt^2) = \iint (T/J) = k \iint \theta_e \quad \ldots\ldots (3)$$

and both the error and damping signals are supplied to the motor by feedback loops.

Electronic simulators, making use of electronic integrating circuits, have therefore been devised for the study of such systems, or use has been made of existing analogue computers for this purpose. Instead of the displacement of an input shaft forming the independent variable to the system, the level of potential across a resistance forms

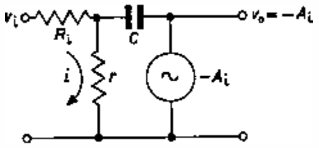


Fig. 2. Integrator circuit

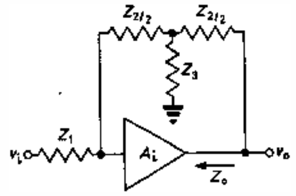


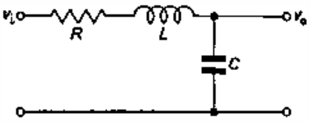
Fig. 3. Compensated feedback circuit

the input signal, and the output is the level of potential across another resistance. Such circuits, however, must be set up rather carefully if a real time analogue of the mechanical system is to be produced and will normally contain high gain d.c. amplifiers[3,4,5]. This is due to the fact that the operating time of the mechanical system may be relatively long, of the order of a few seconds. For instructional purposes it is not, however, necessary to simulate the mechanical system in real time, and in this case an a.c. or capacitively coupled circuit may be employed together with an oscillograph to show the response of the simulated analogue when operated with a repetitive step input of say, 50c/s.

In the circuit described, illustrated in Fig. 1 the transistors are employed as 'virtual earth'[6] operational amplifiers[7], and are capacitively coupled. Overall feedback loops provide the error signal, a combination of the phase-reversed output signal and the input signal, and a 'velocity' feedback signal is taken from the output of the first integrator. The forward loop also contains a phase advancing unit, or differentiator, which can be set to include various degrees of error rate voltage with the error voltage at the input to the integrators.

The theory of the shunt feedback single transistor amplifier has been given by M. G. Hartley[8] in terms of the open loop transfer impedance, A, and other circuit constants. It is found that for the integrator circuit of Fig. 2, subject to a step input of V_i volts, that the output signal V_o is given by the equation:

Fig. 4. 'Second-order' (LCR) network



$$V_o = -V_i(t/CR_i)[1 + (t/2CA') + \ldots.] \quad \ldots. (4)$$

where $A' = A/(1 + (r/R_i))$

Equation (4) shows that there is a quadratic error term, and also smaller high order errors. The second order, or quadratic, error, after t seconds is given as $(t/2CA') \times 100$ per cent. True integration is therefore not obtained. A paper by K. P. P. Nambiar and A. R. Boothroyd[9] has shown, however, that a compensating impedance can be found for this circuit so that the second order error is eliminated, i.e. for the circuit of Fig. 3:

$$Z_3 = -Z_2[Z_o(2 + A_i) + Z_2]/4(Z_2 + Z_o) \quad \ldots. (5)$$

For v_o/v_i to equal $-Z_2/Z_1$

In the case of the integrator this result can be further simplified, so that for true integration:

$$R_3 = -R_o/4(1 + A_i) \quad \ldots\ldots\ldots\ldots (6)$$

where R_o is the output resistance and A_i the current gain.

For an OC75 transistor the current gain, in grounded emitter, can be taken as -60, and R_o as 1·5kΩ in the circuit given. Hence the compensating impedance is 22kΩ.

A diode bridge, consisting of four diodes connected as shown in Fig. 1, operates as a switch when a square wave is applied to it, and applies full 'velocity' feedback to the system during negative half cycles of the input wave, so completely damping the response during this interval. The square waves for the bridge are derived from an OC75 paraphase amplifier, and are applied to opposite corners of the bridge. This damping ensures that the system is quiescent before each positive input step is applied.

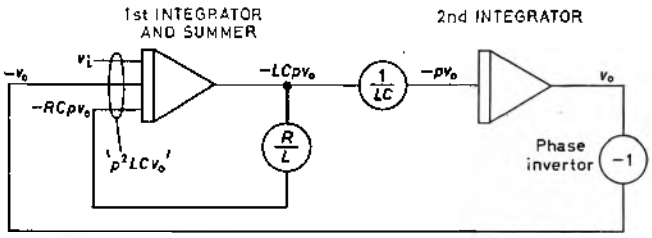
The ideas involved in the setting up of a transfer function on an analogue computer are also readily seen from the operation of this simulator. Each integrator performs a '$1/p$' operation, and there is included an adder and a phase invertor. The example most readily illustrated is therefore the transfer function of the simple L, C and R network shown in Fig. 4.

Expressing the Laplace transforms of the input and output voltages for an input voltage step one can say that:

$$v_o/v_i = \frac{1/pC}{R + Lp + (1/pC)} = \frac{1}{1 + pRC + p^2LC} \quad \ldots\ldots\ldots\ldots (7)$$

The transfer function expressed by equation (7) may be multiplied through and the terms rearranged so as to put the highest order differential on the left-hand side of the equation. This corresponds to its circuit position, since preferably only integration is performed on an analogue

Fig. 5. Flow diagram of simulator as computer



computer, differentiation tending to introduce unwanted noise signals.

i.e.:

$$p^2LCv_o = -RCpv_o - v_o + v_i \quad \dots\dots\dots \quad (8)$$

This equation may therefore be illustrated on the simulator, which expresses and solves the equation, as shown in the flow diagram of Fig. 5.

Thus the transfer function is represented by a summer, two integrators and some constant multipliers, and involves two feedback loops, one of which contains a phase invertor. The flow diagram illustrates the functioning of the circuit given in Fig. 1, and the output of the circuit for a step function input simulates the behaviour of the circuit shown in Fig. 4.

These matters are of some interest to the experimental psychologist since some of his investigations concern human and animal goal seeking behaviour and involve feedback circuits[10] similar to those in automatic control or cybernetic systems such as have been developed for the automatic piloting or landing of aircraft, etc. It is well known that the performance of such error actuated devices bears a close resemblance to certain types of animal and human behaviour.

Acknowledgment

Acknowledgment is made to Professor R. C. Oldfield for the laboratory facilities kindly made available for this investigation.

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A Marine Data Logger System

A marine data logger system has been designed and installed in a refrigerated cargo vessel by English-Electric-Leo-Marconi Computers Ltd in conjunction with the Shaw Savill Line, the ships' owners.

The ship, M.V. "Zealandic", a vessel of 9 500 tons with four refrigerated holds has been built for service between the United Kingdom and Australia and New Zealand via Suez, Panama or the Cape.

The data logger is mounted in the engine room and has its own temperature control system.

It continuously watches the operation of 350 points in the ship's main engines, auxiliary machinery, the vital refrigeration equipment and the refrigerated cargo spaces.

Immediately any malfunction occurs, the logger raises the alarm by sounding a klaxon, flashing a warning light and printing out in red type on the alarm record printer full details of location of fault and the time it occurred.

As long as the point continues in alarm, there is no further print-out or klaxon warning. When the point has returned to normal, further print-out in black will occur. Two separate pinboards are provided, one for main engine and auxiliaries and domestic refrigerated chambers, and the second for the refrigerated spaces and machinery. The pinboards allow for adjustment of the reference levels for any alarm point and independently for high or low.

Every four hours or on demand, a print-out of time, point identification and data takes place on a second strip printer. This total scan takes approximately 3 minutes.

The amount of information to be printed out during a logging cycle is determined by a selector switch which gives the following alternatives:

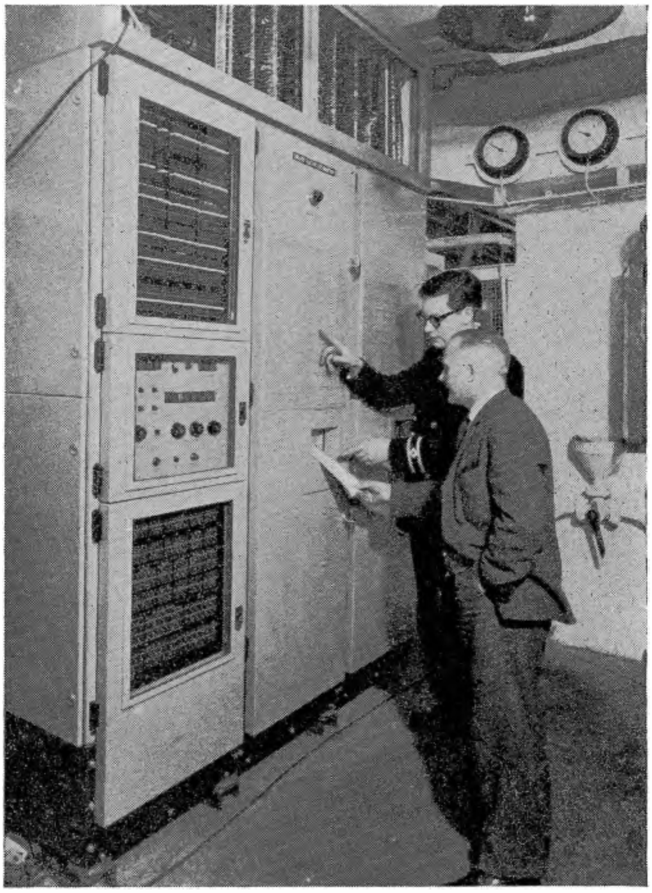
(a) Auxiliaries only.
(b) Auxiliaries and main engine.
(c) Auxiliaries and refrigeration.
(d) Complete system log.

Special features of the data logging system include the following:

(1) The measurement of peak fuel injection presures on the eight main engine cylinders. The fuel injection pressure points are each sampled in turn for a complete cycle and the peak value stored then being converted to engineering units for print-out or display.
(2) Conversion of suction and delivery pressures on the refrigeration machinery into the corresponding saturation temperatures. The points in question, although sensed as pressures, give print-out display as temperatures to the nearest $\frac{1}{2}$° F.
(3) Shaft horsepower is provided and available as a continuous signal. Calculation of s.h.p. is by means of continuous signals from a Siemens torsiometer and engine speed tachogenerator.

In addition to continuously monitoring the plant items, the equipment also self-monitors for equipment alarm conditions, for example, internal cubicle temperatures, test levels through amplifiers, analogue to digital convertor and the alarm comparator and power supplies. In the case of mains supply failure a klaxon will sound, the supply to the klaxon itself being derived from an independent source. In the case of the other equipment failures mentioned, alarm annunciation will be initiated.

The marine data logger unit installed in M.V. "Zealandic". The unit mounted in the centre of the left-hand rack gives a visual display of the number of the test point being measured, its actual value and the time at which the measurement is being made



A Staircase Function Generator
for Use in the Display of Engine Indicator Diagrams

By A. Watson*, B.Sc., A.M.I.Mech.E.

A staircase generator employing commercial transistor binary modules is described. The generator was used as a time-base for investigating the performance of internal combustion engines and several such applications are described.

(Voir page 355 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 362)

WHEN investigating the performance of internal combustion engines a pressure transducer fitted into the cylinder head can be used to give an output voltage proportional to the instantaneous value of the cylinder pressure.

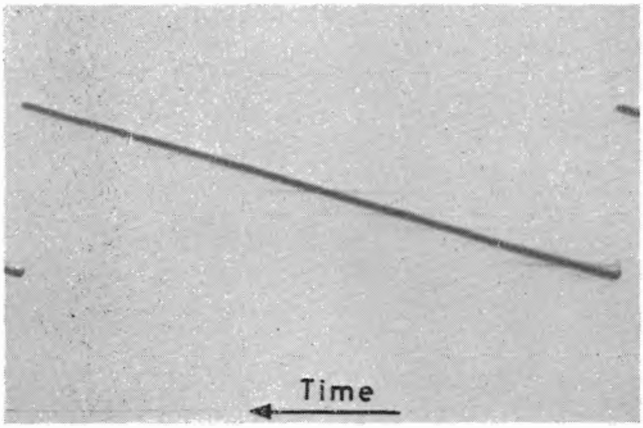


Fig. 1. Staircase sweep voltage proportional to degrees of crank angle

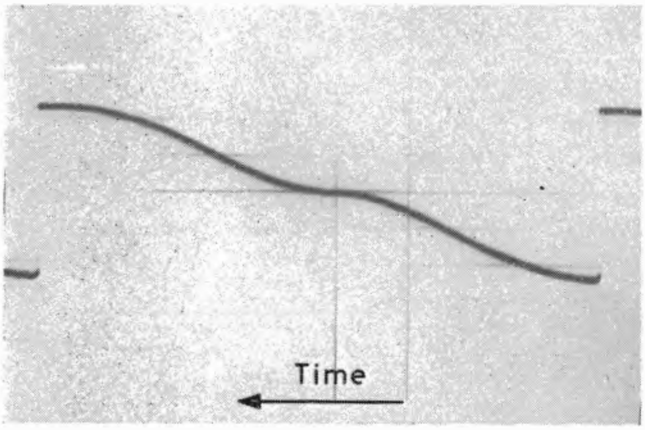


Fig. 2. Staircase sweep voltage proportional to engine piston displacement

This voltage is normally displayed against time on a cathode-ray oscilloscope by using the internal time-base of the c.r.o. To give a steady display the internal time-base is normally used in the triggered mode with the triggering pulse taken from a pick-up on the engine crankshaft in preference to triggering internally from the signal itself.

This method of display has the advantage of an extremely linear time-base, normally of an order of accuracy greater than that possessed by the pressure signal, but the sweep velocity must be altered with change in engine speed.

The sweep voltage generator described gives a voltage proportional to engine crank angle (Fig. 1) with good overall linearity, a constant sweep length regardless of engine speed, an extremely short flyback time and a switched shift facility. In addition it can be simply switched to provide a sweep voltage proportional to engine piston displacement (Figs. 2 and 3), which is convenient if the engine indicated mean effective pressure is to be measured from the diagram, as the area under the pressure curve is in this case already in units of work, and no graphical conversion is needed before measuring the area.

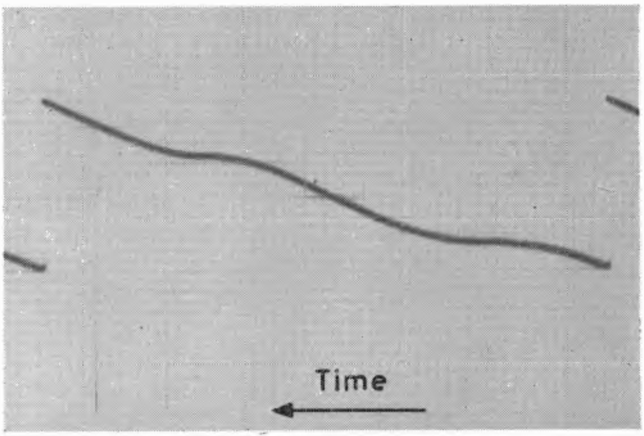


Fig. 3. Staircase sweep voltage proportional to engine piston displacement misphased

Against these advantages, there is the disadvantage that high frequency pressure variations such as might be present under abnormal engine running conditions cannot be observed.

Principle of Operation

Essentially the equipment consists of a binary staircase generator (Fig. 4) which is fed with pulses from a magnetic disk attached to the engine crankshaft. If the pulses are fed into the staircase generator at a constant rate the output voltage will be a linear staircase, whereas if the pulses are grouped into a predetermined pattern any desired shape of staircase can be produced (Fig. 5).

Pulse Generator

The pulse generator consisted of two Perspex disks mounted in ball bearings and driven from the engine crankshaft by an anti-backlash metallic disk coupling. The circumference of the disks was sprayed with a magnetic dispersion and had inductive pick-up heads fixed 0·001in clear of the periphery. The whole assembly was removed from the engine and mounted on a micro-dividing head with the disks clamped to the table and the pick-ups clamped in a fixed angular position. Three tracks of pulses were recorded on the disks, with an accuracy of seconds of the arc, by passing a current through the pick-up head which was sufficient to saturate the magnetic dispersion.

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Track 1. 1 pulse to correspond to engine top dead centre.

Track 2. 2^7 pulses at equal angular intervals.

Track 3. 2^7 pulses at angular increments to correspond to equal increments of piston displacement as calculated from the particular engine crankshaft-connecting rod geometry.

When the unit was coupled to the engine crankshaft these recorded pulses were produced at the pick-up heads with a minimum amplitude of 10mV peak-to-peak, over the range of engine running speeds. They were amplified using a two stage *RC* coupled transistor amplifier followed by an emitter-follower to give output pulses of 5V peak-to-peak.

Staircase Generator (Fig. 6)

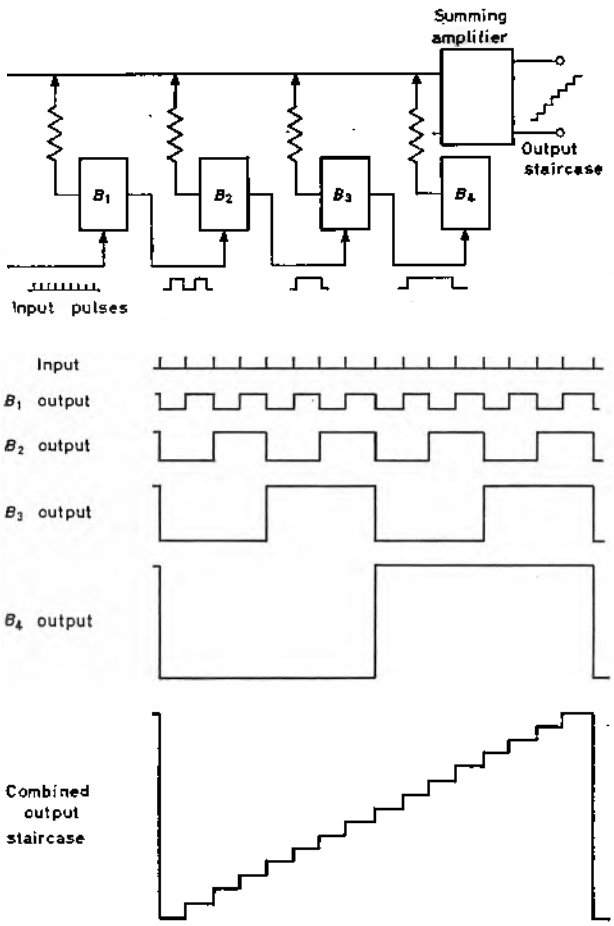
This consisted of a pulse shaper followed by seven binaries which fed a simple summing amplifier to give the required staircase output. The units were commercial transistor modules and the height of the output staircase was 0·3V.

An input switch directed pulses from either track 2 or track 3 into the shaper to provide a crank angle or displacement sweep as required. In addition to this the pulse from track 1 could be added to these pulses to produce a progressive shift of the staircase at the rate of one step per engine revolution.

Phasing the Pulse Generator

Before using the pulse generator the angular position of the pick-up was adjusted so that the top dead centre pulse on track 1 was produced at the exact moment that the engine piston reached top dead centre. This was carried out by marking a disk attached to the crankshaft with a pointer which coincided with a fixed pointer on the engine crankcase when the engine piston was at t.d.c. as determined statically. With the engine running a stroboscope was used to illuminate the disk while the strobo-

Fig. 4. Simple binary staircase generator



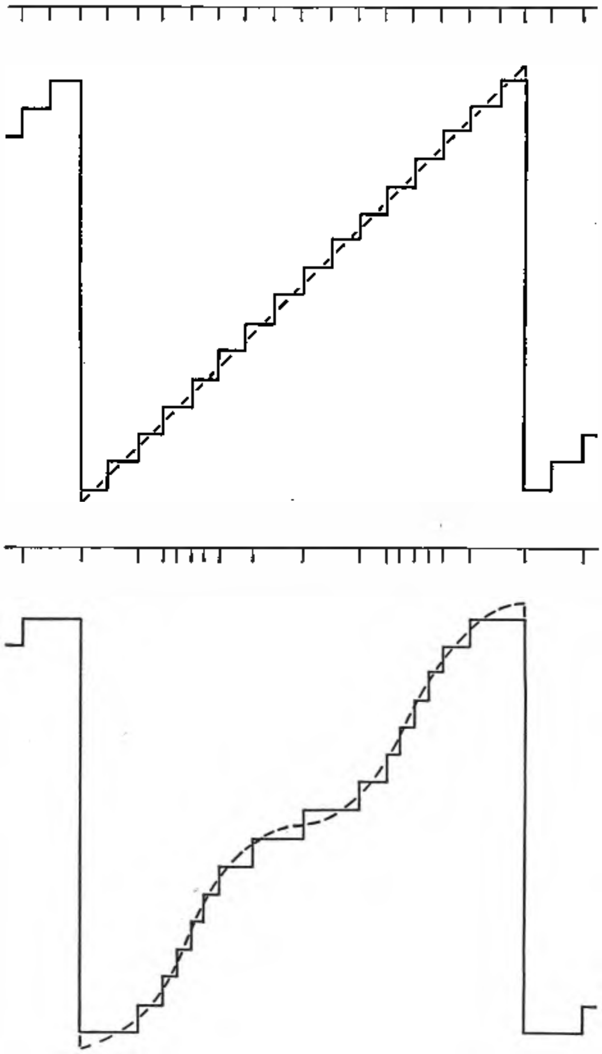
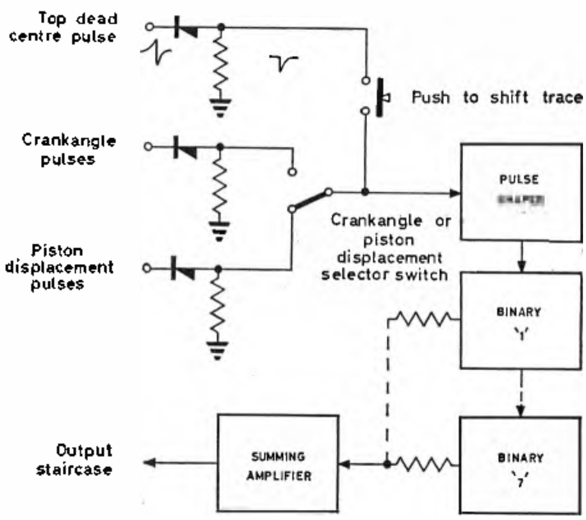


Fig. 5. Pulse grouping used to produce non-linear staircase

Fig. 6. Block diagram of equipment



scope discharge current pulse was monitored on a c.r.o. along with the t.d.c. pulse from track 1. The angular position of the pick-up head was adjusted so that the t.d.c. pulse coincided with the stroboscope current pulse when

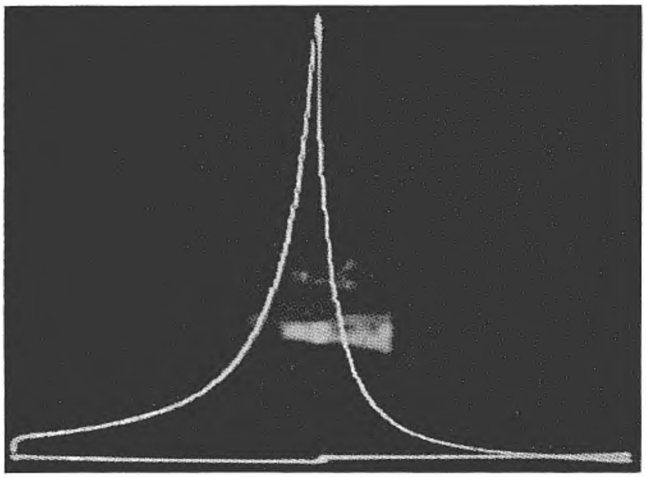


Fig. 7. Cylinder pressure against piston displacement (variable inductance pressure transducer)

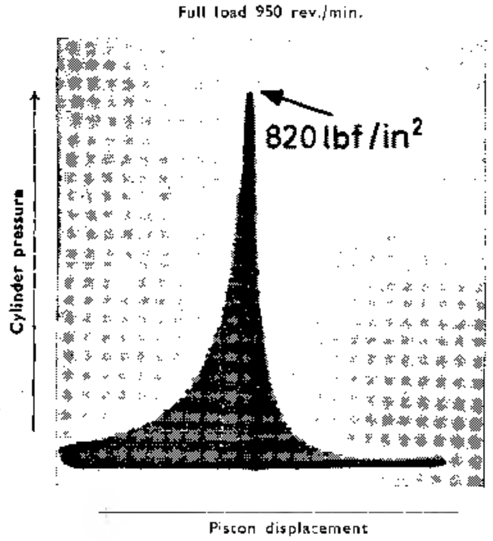
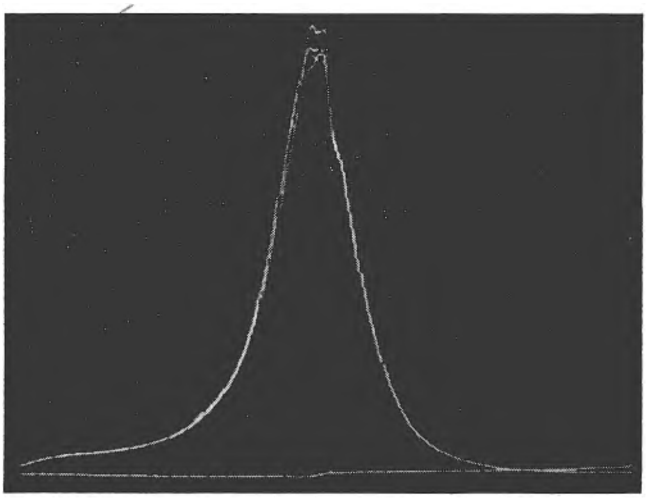


Fig. 8. Cylinder pressure against piston displacement (balanced pressure transducer) the engine loading being the same as in Fig. 7.

Fig. 9. Cylinder pressure against crank angle (variable inductance pressure transducer)



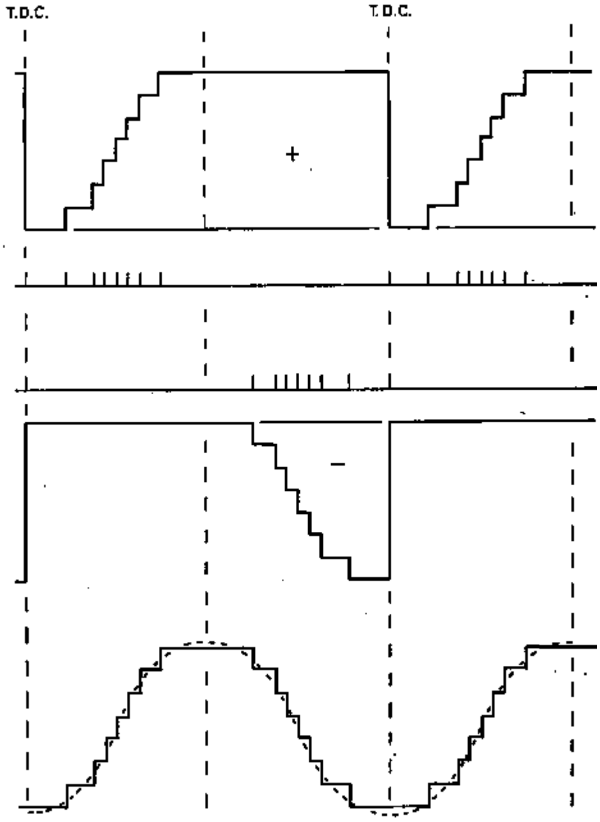


Fig. 10. Two non-linear staircases subtracted to give output voltage proportional to engine piston position

the stroboscope flash showed the disk pointer and the fixed pointer to be coincident. This determination of coincidence is made easy if a balanced c.r.o. input amplifier is used with equal height pulses in positive and negative directions on the one trace, at coincidence they disappear.

Conclusion

The piston displacement sweep voltage has been used in determining the indicated mean effective pressure of a single cylinder Diesel engine by using a variable inductance pressure transducer (Fig. 7), and also by using the balanced pressure technique with a diaphragm type pick-up (Fig. 8). In each case the position of top dead centre is automatically and accurately produced on the trace of cylinder pressure-piston displacement.

The crank angle sweep voltage has been used in teaching (Fig. 9), to show the effect of varying engine operating parameters, where despite the rapid engine speed changes encountered the pressure-crank angle diagram remained locked in position without elongating or slipping.

These sweep voltages can be used where the engine angular velocity is not constant over one revolution e.g. cranking, and by using two staircases (Fig. 10) it is possible to produce a reciprocating piston displacement sweep voltage to give a closed loop indicator diagram.

The disks must be re-recorded if the pulse generator is moved to an engine with different geometry but despite this fact along with the need of a direct crankshaft connexion the equipment has been found to be a useful accessory in determining engine i.m.e.p.s.

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Analysis of a Pulse Modulator

By J. Kontos*, Ph.D.

The analysis of a pulse modulator is presented that gives as output a pulse train whose repetition frequency and pulse duration both vary with the input voltage. A formula is derived giving the dependence of the pulse repetition frequency on the input voltage and the circuit parameters. This formula is tested experimentally by comparing calculated values of the frequency with ones obtained with a circuit built for this purpose. Agreement of experiment with theory is found.

(Voir page 355 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 362)

A MODULATOR giving as output a pulse train whose repetition frequency and pulse duration both vary with the input signal has already been described and its use for d.c. amplification demonstrated[1]. Further work has made possible the analysis and calculation of the frequency to input dependence and is presented here. The main physical fact that has emerged is the effect of the non-linearity of the transistor input resistance on the operation of the circuit. This effect complicates the mathematical analysis and in order to simplify it somewhat the circuit analysed here, which is shown in Fig. 1, has been made different from the original one by using a feedback filter consisting of a single RC section instead of two such sections.

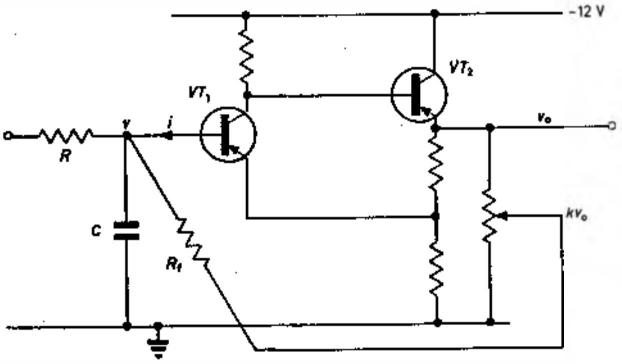

Fig. 1. The circuit for analysis

The effect of the non-linear transistor input resistance can be made apparent by disconnecting the feedback filter from the output terminal of the circuit. Under this condition the circuit can still be made to act as a pulse modulator provided the value of the input voltage is within the appropriate range. This behaviour strongly suggests a relaxation oscillator based on the non-linear input resistance of the active part of the circuit and very similar to the classical resistance-capacitance-neon relaxation oscillator[2].

The input resistance of the active part of the circuit can be approximated by a piecewise linear resistance particularly in view of the positive feedback from VT_2 to VT_1 through their common emitter load that causes switching behaviour. In fact, observation of current-voltage relationships at the base of VT_1 justifies the approximation of this resistance by one that switches between two constant values. The transitions between the two linear regimes take place whenever the voltage across the capacitor C reaches one of two specified values and depending on its previous history. This history dependence means that the current-voltage curve of this input resistance exhibits hysteresis taking a form like the one shown in Fig. 2. The switching values of the base voltage for the change in the input resistance of VT_1 are taken here to coincide with the ones at which the Schmitt trigger changes state.

In the next section a formula is derived giving the dependence of the pulse repetition frequency on the input voltage and the circuit parameters. This formula is subsequently tested experimentally.

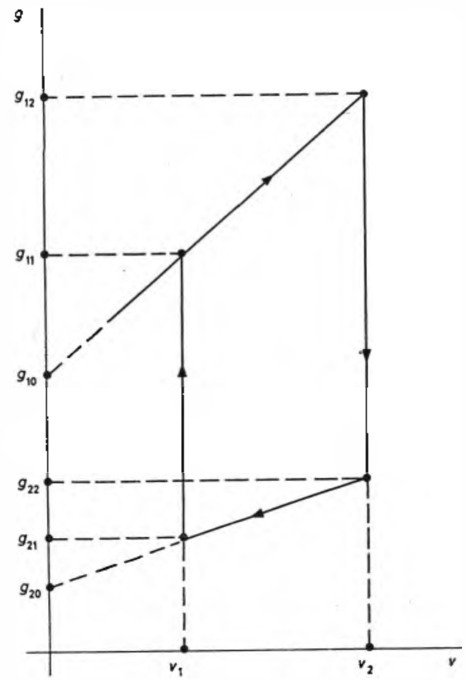

Fig. 2. Current-voltage curve showing hysteresis

Analysis

With reference to Fig. 1 and using capital letters to denote Laplace transforms of variables, the summation of currents at the base of VT_1 gives:

$$\frac{V_i - V}{R} + \frac{kV_o - V}{R_f} + I = sCV$$

or:

$$V = \frac{1-r}{1+sT(1-r)}\,(V_i + G + pV_o) \quad \ldots\ldots\ldots (1)$$

where:

$$r = \frac{R}{R+R_f},\ T = RC,\ p = k(R/R_f),\ g = iR$$

It is now assumed that g and v_o are functions of v of the form shown in Fig. 2 and Fig. 3. respectively. If one defines a new variable h given by:

$$h = g + pV_o$$

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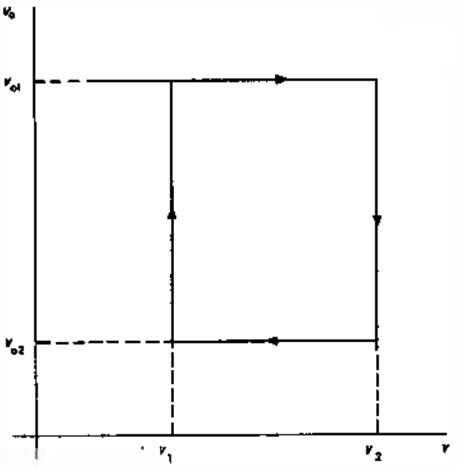

Fig. 3. v_o as a function of v

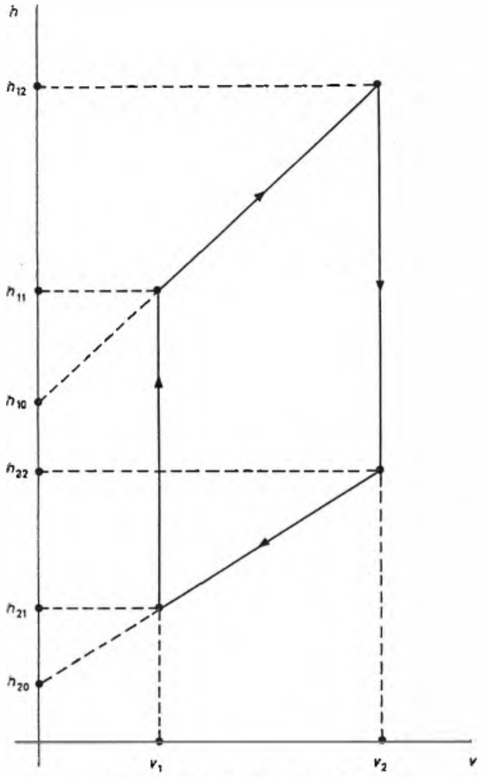

Fig. 4. h as a function of v

then h will be a function of v of the same form as g is of V.

The form of h is shown in Fig. 4 and

$$h_{11} = g_{11} + pv_{01},\ h_{12} = g_{12} + pv_{01},$$
$$h_{21} = g_{21} + pv_{02},\ h_{22} = g_{22} + pv_{02},$$
$$h_{10} = g_{10} + pv_{01},\quad h_{20} = g_{20} + pv_{02}$$

And now, one can write

$$h = h_0 + mv$$

so that

$$h_0 = h_{10} \text{ and } m = m_1,$$

for v going from v_1 to v_2, and

$$h_0 = h_{20} \text{ and } m = m_2,$$

for v going from v_2 to v_1, and

$$m_1 = \frac{g_{11} - g_{12}}{v_1 - v_2},\ m_2 = \frac{g_{21} - g_{22}}{v_1 - v_2}$$

Hence from equation (1) one can obtain the differential equation:

$$K\,(dv/dt) + v = U \qquad (2)$$

where:

$$K = \frac{AT}{1 - mA},\ U = \frac{A(h_0 + v_1)}{1 - mA},\ A = 1 - r,$$
$$K = K_1,\ U = U_1 \text{ for } m = m_1 \text{ and } h_0 = h_{10},$$
$$K = K_2,\ U = U_2 \text{ for } m = m_2 \text{ and } h_0 = h_{20}.$$

If it is now assumed that v oscillates between the two values v_1 and v_2 and if the time intervals for v to go from v_1 to v_2 and from v_2 to v_1 are denoted by t_1 and t_2 respectively, then from equation (2):

$$t_1 = K_1 . \ln \frac{1 - (v_1/U_1)}{1 - (v_2/U_1)} \qquad (3)$$

$$t_2 = K_2 . \ln \frac{1 - (v_2/U_2)}{1 - (v_1/U_2)} \qquad (4)$$

And finally, if f is the frequency of oscillation of the circuit, then:

$$f = \frac{1}{t_1 + t_2} \qquad (5)$$

In the next section, this formula is used for comparison with experimental results.

Experimental Test

The circuit shown in Fig. 1 was built and its frequency to input voltage dependence studied. The experimental results were compared with the ones calculated with the formulae derived in the previous section.

The transistors used were the ASZ21 type and the values of the main components were:

$$R = 50\text{k}\Omega,\ C = 0{\cdot}01\mu\text{F},\ R_t = 50\text{k}\Omega.$$

The base current of VT_1 was measured by inserting a 200Ω resistance between the capacitor C and the base of VT_1. The voltage-current relationship of the active part of the circuit was obtained by measuring the two small voltages across the capacitor C and across this resistance, not shown in Fig. 1, which was considered as included in the active part of the circuit.

The characteristics of the active part of the circuit were found to be of the form shown in Fig. 2 and Fig. 3 and the values at the break-points were, in volts:

$$v_1 = -1{\cdot}4,\ v_2 = -0{\cdot}7,\ v_{01} = -0{\cdot}4,\ v_{02} = -10{\cdot}4,$$
$$g_{11} = 138,\ g_{12} = 25,\ g_{21} = 0,\ g_{22} = 0.$$

From the above, the value of g_{10} was calculated as 88V. The calculated and the experimental frequencies are

TABLE 1

	k=0		k=0·4		k=0·8		k=1	
$-v_i$	f_E	f_C	f_E	f_C	f_E	f_C	f_E	f_C
5	8	8	19	18	29	27	33	32
6	11	10	21	21	31	30	35	34
7	14	13	23	22	33	31	37	35
8	17	15	25	25	35	33	39	36
9	19	18	28	26	36	35	40	38
10	21	20	29	29	37	36	41	39
11	23	22	31	32	39	38	42	41
12	26	24	33	33	40	40	44	43

v_i = *Input* (V)
f_E = *Experimental values of frequency (kc/s)*
f_C = *Calculated values of frequency (kc/s)*
k = *Amount of negative feedback*

listed in Table 1 for various values of the input voltage v_i and the amount of negative feedback k.

The experimental and calculated results listed in Table 1 are in good agreement in view of the experimental errors, the calculation round-off errors and the component tolerances involved. This agreement gives support to the theory of the pulse modulator presented in this article.

Acknowledgment

Thanks are due to G. Fragakis for building the circuit and taking the measurements.

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Elimination of Division Circuits by Variable Time Scale

By D. K. Taylor*

The analogue simulation of an equation involving the division of two variables has previously required a special division circuit, which is a large source of error. It is shown that in some cases it is possible to eliminate division circuits by introducing a time transformation into the equations. This adds an additional equation which gives the real time as a function of the computer time. For the case of a single first order linear differential equation the removal of a division circuit by a time transformation replaces the original equation by two coupled first order linear differential equations.

(Voir page 355 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 362)

THE division circuit is a notorious source of trouble in electronic analogue computers. It is often necessary to rescale the circuit when initial conditions are changed because a denominator suddenly gets too small and overloads the division amplifier. Also the division circuit can produce large errors. Consider the quotient q,

$$q = u/(v + \Delta v) \quad \ldots\ldots\ldots\ldots\ldots\ldots (1)$$

where u is a dividend and $v + \Delta v$ is a divisor with error Δv. If v is small compared to Δv, then the error in q, which is the reciprocal of a small number, will be large. In contrast, the error produced by an amplifier which multiplies by a constant is proportional to the error of the input and will not become as large.

It would, for these reasons, be desirable to do without division circuits whenever possible. In certain cases division circuits can be eliminated by introducing a variable time scale.

This technique will be developed here only for a specific case, rather than in more generality. The concept of a variable time scale is the important idea, and it is a small step to apply it to other cases.

An example of a case where it is possible to eliminate a division circuit by introducing a variable time scale is the analogue simulation of the following differential equation:

$$(dx/dt) + (f(x,t)/g(x,t)) = 0 \quad \ldots\ldots\ldots (2)$$

Its solution shall be represented by $x(t)$.

Introducing the transformation:

$$t = \int_0^T g\left(x(\xi), t(\xi)\right) d\xi \quad \ldots\ldots\ldots\ldots (3)$$

Then the derivative dx/dt is given by:

$$(dx/dt) = (dT/dt)\,(dx/dT) = \left(1/g(x,t)\right)\,(dx/dT) \quad \ldots\ldots (4)$$

The differential equation becomes:

$$(dx/dT) = f\left(x, t(T)\right) \quad \ldots\ldots\ldots\ldots\ldots (5)$$

and the real time t is obtained by solving equation (3).

As an illustration of the procedure consider the case where $f(x,t) = x - 1$ and $g(x,t) = x(t+1)$. This case can be solved analytically, but this does not imply that the method that is being described applies to only such cases. In this case the solution of equation (2) is:

$$(x - x_o) + \ln\left[(x-1)/(x_o-1)\right] + \ln(t+1) = 0 \quad \ldots\ldots (6)$$

where x_o is the value of x at $t = 0$.

The transformed equation is:

$$(dx/dT) + x = 1 \quad \ldots\ldots\ldots\ldots\ldots\ldots (7)$$

and its solution is:

$$x = (x_o - 1)\,e^{-T} + 1 \quad \ldots\ldots\ldots\ldots (8)$$

Equation (7) contains no divisions, and it is therefore more desirable to simulate it on an analogue computer than equation (2).

The real time is given by the integral equation:

$$t = \int_0^T x(t+1)\,dT \quad \ldots\ldots\ldots\ldots\ldots (9)$$

with the solution:

$$t = \exp\left[(x_o - 1)(1 - e^{-T}) + T\right] + 1 \quad \ldots\ldots (10)$$

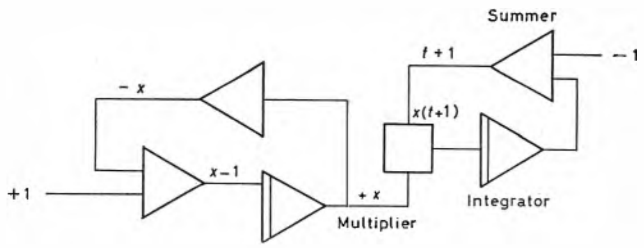


Fig. 1. Arrangement of computer

An unscaled computer wiring diagram for the simulation of the transformed equation (5) and for the computation of the real time t is shown in Fig. 1.

The circuit elements are all standard electronic analogue computer components, and the important thing is that no division circuits are used.

The quantities $x(T)$ and $t(T)$ can be displayed, for example, on a strip chart recorder. Then in order to locate the value of x at some particular time one needs only to follow the t channel until the t of interest is found and then to read off the corrresponding x.

Acknowledgment

Research was performed at the John Hopkins University Applied Physics Laboratory under Contract NOrd 7386 with the Bureau of Naval Weapons, Department of the Navy.

* *U.S. Naval Ordnance Laboratory, U.S.A.*

BOOK REVIEWS

Introduction to Semiconductor Physics

By R. B. Adler, A. C. Smith and R. L. Longini. 247 pp. Demy 8vo. John Wiley & Sons Inc. 1964. Price 34s.

THIS is a very good book for the student and also perhaps for graduates who are not physicists but who wish to work on semiconductors. While the treatment of the subject is deliberately simplified, the authors are careful to mention, in passing, situations that may arise when restrictions are lifted. On the other hand, the presentation could probably be improved by placing the whole of chapter 2, on the Energy-Band Model of a Semiconductor, somewhat earlier in the text.

The subject is presented in four chapters, each with a carefully selected set of problems, and an appendix containing detailed instruction on five laboratory experiments. In order to achieve a full understanding the student must attempt at least some of the problems on each chapter. Certain more mathematical parts of the text may be omitted at a first reading. The laboratory experiments described are those which might well be found in an undergraduate course, but the methods of carrying them out seem to assume a certain lack of equipment in the teaching laboratory (for instance, instructions on how to make your own spectrometer, and the widespread use of clothes-pegs).

The subject covered is the physics of semiconductors in bulk, up to the point at which the student is ready to be introduced to p-n junctions. Indeed, the last laboratory experiment described is the preparation and measurement of an alloyed germanium diode. This is the first of seven volumes in a series sponsored by the Semiconductor Electronics Education Committee, a U.S. organization, the other six of which are intended to deal with transistors and transistor circuits. At 34s., it is good value.

P. C. NEWMAN

Communication Nets Stochastic Message Flow and Delay

By Leonard Kleinrock. 209 pp. Med. 8vo. McGraw-Hill Book Co. 1964. Price £5.

THE preface explains that this monograph is based upon a doctoral thesis. Unfortunately insufficient attention has been paid to changing the original into a readable technical presentation. The main text occupies about two-thirds of the book, but it is almost impossible to deal with this without constant reference to the remaining one-third which consists of appendices. This results in loss of continuity and consequently detracts from the value of the information presented.

The theoretical approach is based upon store-and-forward communication nets, a cited example being the Western Union Plan 55A. This is a military system which has to deal with multi-delivery messages, a prime reason for choosing store-and-forward operation instead of circuit switching. However, the theory of queueing dealt with by the author ignores this requirement and, for simplicity, considers single-delivery messages of an exponential type message-length distribution. Unfortunately a designer of networks could not apply such theories to a practical requirement involving multi-delivery messages with other message length distributions, e.g. constant length, or the typical telegraph message distribution, a constant followed by an exponential form.

The theoretical chapters are of interest, contain original ideas and, within the limits of the chosen parameters, draw a number of conclusions which may be of value for determining the approximate behaviour of a given network. A conclusion is reached early in the monograph that probability theorems are of limited use when considering complex practical networks and that resort must be made to simulation. In consequence, Appendix E, which considers a simulation program, and Chapter 7, which describes results derived by this program, are of more practical value than the remainder of the monograph. The author does suggest further investigations; if these are pursued and the results published, a greater service would be rendered to the student of networks than does the existing monograph.

J. RICE

Transistor Circuits in Electronics

By S. S. Hakim and R. Barrett. 341 pp. Demy 8vo. Iliffe Books Ltd. 1964. Price 63s.

THIS book has been written as an introduction to the use of the transistor in performing a wide range of electrical system functions. It begins with a brief description of transistor action, which is given in terms which are simple but adequate for the purposes of the book. The basic common base and common emitter connexions of the transistor are explained, and the fundamental limitations of the device in terms of permissible power dissipation are dealt with.

Next the transistor is considered as an amplifier, and its characteristics in circuits under various conditions of static and dynamic loading are dealt with. The small signal performance of the transistor is analysed and the *h* parameter representation and T equivalent circuit of the transistor are described.

A comprehensive survey of various types of transistor amplifier is given including d.c. amplifiers, wide band amplifiers, tuned amplifiers and power amplifiers. This section is a particularly useful one for the linear circuit engineer.

The final part of the book considers the transistor as a switch. The drawbacks of the transistor in terms of collector series resistance and charge storage in switching applications are satisfactorily discussed, but in the reviewer's opinion the book would have benefited if the authors had dealt more fully with the differences between various types of transistor in more detail.

The book includes a good chapter on the fundamentals of Boolean algebra and the use of logical switching circuits to perform Boolean algebraic functions.

On the whole the book can be recommended as a good introduction to transistor circuits for the systems engineer, although it does not indicate to him in any detail which types of transistor to use for different system requirements.

A. A. SHEPHERD

An Introduction to Numerical Control of Machine Tools

By O. S. Puckle and J. R. Arrowsmith. 272 pp. Med. 8vo. Chapman & Hall Ltd. 1964. Price 45s.

THE authors have done a difficult task competently in preparing this book—the first in the new and important field of numerical control.

The opening chapter discusses the basic history of control and emphasizes flexibility, thereby opening the door to the concept of numerical control. The chapters which follow bring out the essential details of point-to-point and contouring systems, the main items covered being information presentation, to numerical control machines, position measuring devices, amplifiers, mechanical power units and servo systems. The book is well illustrated with line drawings and in a table appended there is a family tree of numerical control possibilities.

A considerable degree of standardization, particularly of input tape format, has been attempted in recent months by organizations such as B.E.A.M.A., and therefore it is disappointing that emphasis has not been placed on this aspect of numerical control, which is desirable both from the point of view of the manufacturer and the operator. However, the chapter on tape codes and information gives considerable background information and the keen student should find the description of the wide variety of different tape codes which have been used of interest.

The book has a most excellent bibliography which is well presented and, in addition, the authors refer to the most important patents concerning measuring systems. The chapter on measuring systems is very well put together and covers the range of transducers which are currently employed by numerical control manufacturers.

Considerable importance is attached to the programming for numerical control and this particular aspect offers a very detailed account of the EMI analogue parabolic interpolator continuous path system. One would have preferred a more general analysis of the programming for point-to-point and continuous path and tied in with the general programme developments such as APT or Profiledata.

The economics of numerical control are not well understood and have not, up to now, been well presented. This book is no exception, and while reference is made to other published work, there is a general lack of firmness in handling this matter.

H. OGDEN

Radio Wave Propagation V.H.F. and Above

By P. R. Matthews. 155 pp. Crown 8vo. Chapman & Hall Ltd. 1965. Price 28s.

THIS book is one of a new series under the general title of 'Modern Electrical Studies' aimed at the post-graduate or final year student. As such it is in principle a welcome addition to the literature and superficially it achieves its object in the coverage which it gives to the subject.

There are nine chapters in all which include line-of-sight and trans-horizon propagation with a discussion of signal strength and system loss; the effects of the atmosphere and of diffraction by obstacles and in ducted modes; the influence of the ionosphere and the problem of noise, especially in relation to satellite communication. An initial chapter on the radio frequency spectrum and a final one on the use of frequencies above 30Mc/s for radio communication set the subject in its proper perspective.

Having said so much in commendation it is, however, necessary to warn the reader that the treatment is uneven, and it is obvious that the author is not equally at home with all his material. He believes, for instance, that the vertical incidence critical frequency of the ionosphere never exceeds 12Mc/s whereas considerably higher values were attained at the last sunspot maximum. In some places the mathematics is quite sophisticated, as in the use of phase integrals in duct theory and perturbation methods in the analysis of scattering. On the other hand the treatment of curved earth geometry and Fresnel diffraction theory, so important in the prediction of television service areas and of the performance of relay links, is scanty in the extreme.

There are many loose statements, some of which are inaccurate for want of qualifying conditions. A student who might be expected to know what is meant by 'free-space' may be bemused by being told that it is a region whose properties are isotropic, homogeneous and loss-free and which is remote from regions of differing properties, a description that might be applied to an extended ionized medium for some purposes. A figure is given purporting to show that at long wavelengths appreciable fields are formed in the shadow region, whereas at v.h.f. and above these diffracted fields become small. The curves are, however, for frequencies between 30Mc/s and 10 000Mc/s, i.e. v.h.f. and above and refer to the case when both transmitter and receiver are on the ground and there is no horizon distance.

We are correctly informed in the initial chapter that v.l.f. is the range 100km-10km, but in the chapter on diffraction it says that at v.l.f. the wavelength may be several thousand kilometres and the earth's diameter is only a few wavelengths. If the author meant e.l.f. then this is merely one of the many misprints too numerous to mention in detail. The majority of these can be laid at the door of the author as the text gives plenty of evidence of careless writing and proof-reading. The treatment of the ray theory in tropospheric refraction is full of errors and the description of modified refractive-index in relation to the k factor is very confused into the bargain.

In the final chapter the field strengths required for television reception in Bands I, III and V are given as 36·5, 47·5 and 60μV/m insead of decibels above 1μV/m. No less serious is the wrong set of values 50·5, 54·8 and 58·3km for the horizon distances of a transmitter at a height of 100m for k values of 1, 1·2 and 1·4. Had the author given the simple formula relating the horizon distance of a transmitter to its height above ground, the reader could at least have discovered that the values are all too large by a factor of about $\sqrt{2}$. One cannot help wondering how many other errors exist in quoted values that are intended to be taken on trust.

It is to be hoped that the publishers will request the author to carry out a careful revision of the text, and that they will bring out an amended edition without delay, for which the intending purchaser will be well advised to wait unless he is prepared to accept the attendant difficulties. If he successfully surmounts them he will find much in the book of value to the communications engineer concerned with v.h.f. and above, more especially in the meterological aspects where the author has some first-hand experience.

G. MILLINGTON

Practical Oscilloscope Handbook

By R. P. Turner. 235 pp. Demy 8vo. Iliffe Books Ltd. 1965. Price 25s.

The two volumes of the original American edition are here incorporated in a single binding. Volume 1 introduces the oscilloscope and explains applications primarily of interest to technicians, radio and television servicing engineers and hobbyists. Volume 2 is devoted to applications of interest to technologists and technicians in science and industry. Its aim is to give clear guidance in the use of the oscilloscope for specific tests and measurements.

Solar Plasma Geomagnetism and Aurora

By S. Chapman. 141 pp. Med. 8vo. Blackie & Sons Ltd. 1964. Price 32s. 6d.

This book is based on lectures given in July 1962 at the twelfth annual session of the Les Houches Summer School of Theoretical Physics, sponsored by the University of Grenoble and directed by Mme. C. DeWitt.

Lectures on Magnetoionic Theory

By K. G. Budden. 82 pp. Med. 8vo. Blackie & Sons Ltd. 1964. Price 30s.

The author's object in his first chapter is to provide a thorough understanding of the basic ideas of magnetoionic theory. The homogenous medium and the very slowly varying medium are discussed in this and the following two chapters. The fourth deals with applications of magnetoionic theory to the ionosphere and exosphere and the theory is extended to deal with media which vary in space. An outline of the basis and the main uses of the phase integral method in radio propagation problems is given in the last chapter.

Quantitative Linguistics

By G. Herdan. 284 pp. Royal 8vo. Butterworth & Co. 1964. Price 75s.

This book aims to provide a complete picture of the area of linguistic science in which statistical methods are used to establish laws and trends of linguistic structure and development. Even the most elementary statistical concepts, such as the average and the standard deviation, are explained and care is taken to justify the use of quantitative methods in an essentially non-mathematical field by establishing their theoretical foundations.

Probability and Information Theory with Applications to Radar

By P. M. Woodward. 136 pp. Demy 8vo. 2nd Edition. Pergamon Press. 1964. Price 35s.

Although this book was first published in 1953 the original text remains substantially unchanged in the new edition. One chapter is added describing briefly the way in which direct probabilities are usually applied to the decision problem in radar design.

Dictionary of Data Processing

By L. Trollhann & A. Wittmann. 300 pp. Med. 8vo. Elsevier Publishing Co. 1964. Price 95s.

This dictionary contains approximately 5 000 entries in English/American, German and French relating to: digital and analogue computers and their most important components; mathematical terms in constant use; office automation and organization relative to commercial data processing; technical and scientific applications of computers; and civil and military data transmission terms. Listed alphabetically in English/American, each term is followed by its German and French equivalent and a reference to usage.

Analogue Computing for Beginners

By K. A. Key. 155 pp. Crown 8vo. Chapman & Hall. 1965. Price 25s.

This book is for the student or the man in industry with a problem and no experience of computing. It describes in simple language the common components of the analogue computer and how they can be connected together and controlled.

While the emphasis is principally on the simpler type of computer, the more expensive items such as multipliers, function generators and recorders are also described.

An Introduction to the Mathematical Formulation of Self-Organizing Systems

By J. Formby. 200 pp. Demy 8vo. E. & F. N. Spon. 1965. Price 35s.

This book has been written particularly for technologists and engineers concerned with large scale automative systems.

The place of the concept of a mathematical model is emphasized, and the evolution of mathematical ideas on models is traced, from their origins in the seventeenth and eighteenth centuries.

The treatment of each topic begins with simple, clearly stated elements and ends at a stage well beyond the scope of most academic courses.

The chapters on digital and analogue computers and on error-actuated devices will be a useful introduction to modern practice here. The later chapters, concerned with self-regulating, self-adjusting and self-optimizing systems, include the latest advances in these fields.

The selected bibliography, and the biographical notes, gives access to the chief contributors in the field of complex automation.

Dictionary of Electronics, Communications and Electrical Engineering, Volume II

By H. Wernicke. 576 pp. Demy 8vo. Rohde & Schwarz. 1964. Price £2 16s.

Volume II is the German-English edition of the dictionary and contains some 6 000 additional words drawn from the fields of data processing, aeronautical electronics, physics, radar and space research.

It also contains some 2 000 letter symbols and abreviations which are grouped together at the beginning of each letter.

Volume I (the English-German edition) contains some 6 000 words and was published in 1962.

Ionized Gases

By A. von Engel. 325 pp. Med. 8vo. 2nd Edition. Oxford University Press. 1965. Price 55s.

The second edition of this book first published in 1955 has maintained its original pattern as a textbook for undergraduates and research workers in the fields of discharge and plasma physics.

An enlarged chapter is included dealing with collision processes leading to excitation and dissociation.

Proceedings of the Conference on Signal Recording on Moving Magnetic Media

470 pp. Crown 4to. Kultura. 1964. Price L4,6,0

This contains the papers presented at the conference held in Budapest, Hungary, in October 1962.

The papers are published the original languages, English, German, French and Russian; and deal with the whole field of magnetic recording including the application in cybernetics and automation.

Nonlinear and Parametric Phenomena in Radio Engineering

By A. A. Kharkevich. 190 pp. Demy 8vo. Iliffe Books Ltd. 1965. Price 35s.

The present book serves as an introduction to the theory of non-linear systems and its practical application to typical electronic problems. It opens with a chapter on non-linear processes. The second chapter deals with the generation of oscillations, stability criteria, and various types of oscillator, including those for microwaves. The third chapter is concerned with the response of non-linear systems to external signals: rectifier operation, locking, the regenerative receiver, pulsed synchronization and synchronization of a relaxation oscillator. The final chapter discusses circuits with variable, time-dependent, parameters.

Originally published in Russia, where the author is currently Professor of Theoretical Radio Engineering at the Moscow Electrotechnical Institute of Communications, the book was subsequently translated, edited and up-dated in America.

Tensor Analysis of Networks

By G. Kron. 635 pp. Med. 8vo. Macdonald & Co. Ltd. 1965. Price 70s.

This is an unrevised re-publication of the book which first appeared in 1939, and which has been out of print for some years.

The author has written a new introduction for this issue.

Conference Proceedings Volumes 24 and 25

By A.F.I.P.S. Vol. 24, 643 pp. Vol. 25, 629 pp. Demy 4to. Macmillan Co. 1965. Price £6 10s.

Volume 24 contains the full text of the 55 papers presented at the 1963 Fall Computer Conference held in Las Vegas in November 1963. Volume 25 contains the papers presented at the 1964 Spring Joint Computer Conference held at Washington D.C. in April 1964. Both conferences were sponsored by the American Federation of Information Processing Societies (A.F.I.P.S.).

Analysis Instrumentation 1964

By L. Fowler, R. J. Harmon and D. K. Roe. 272 pp. Demy 4to. Plenum Press. 1964. Price $14.50

This book contains the proceedings of the Tenth National I.S.A. Analysis Instrumentation Symposium held at San Francisco in June 1964 and sponsored by the Analysis Instrumentation Division of the Instrument Society of America.

Space-Age Acronyms

By R. C. Moser. 240 pp. Crown 4to. Plenum Press. 1964. Price $17.50

This book contains some 10 000 acronyms on abbreviations, with more than 17 000 definitions, relating to industrial, military and space projects.

Included is a list of missile, aircraft, ship and communications electronic equipment designating systems.

Proceedings of the American Documentation Institute Volume I

By A. W. Elias. 521 pp. Demy 4to. Macmillan & Co. 1965. Price £6 6s.

This contains the proceedings of the American Documentation Institute held at Philadelphia, Pennsylvania, in October 1964.

LETTERS TO THE EDITOR

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

A D.C. Amplifier with Unity Voltage Gain

DEAR SIR,—I have read the article published by G. G. Bloodworth about a D.C. amplifier with unity voltage gain in the February issue (*Electronic Engng.* *37*, 112, 1965).

It may perhaps be interesting to note that the following simple device (Fig. 1) provides a very good performance.

It has a voltage gain very close to unity, a very high input impedance and a low output impedance.

As Mr. Bloodworth has pointed out, the combination of VT_1 and VT_2 is equivalent to an npn transistor connected as VT_1 in Fig. 1 and equivalent to a pnp transistor connected as VT_1 in Fig. 2 with enhanced current gain roughly equal to the product of the individual current gains.

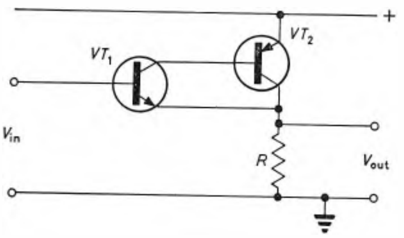


Fig. 1. Equivalent to npn transistor as VT_1

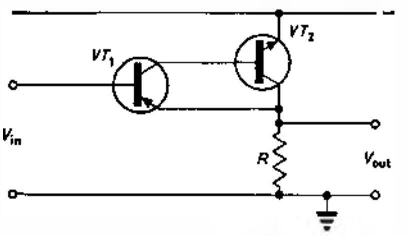


Fig. 2. Equivalent to pnp transistor as VT_1

The hybrid parameters of this combination are given by the following formulae.

$$[H] = \begin{bmatrix} \dfrac{h_{11}{}^1+\Delta h^1\, h_{11}{}^2}{1+h_{11}{}^2\, h_{22}{}^1} & 1-\dfrac{h_{12}{}^1\, h_{12}{}^2}{1+h_{11}{}^2\, h_{22}{}^1} \\ \dfrac{h_{21}{}^1\, h_{21}{}^2}{1+h_{11}{}^2\, h_{22}{}^1}-1 & \dfrac{h_{22}{}^2+\Delta h^2\, h_{22}{}^1}{1+h_{11}{}^2\, h_{22}{}^1} \end{bmatrix} \quad \ldots\ldots\ldots\ldots\ldots (1)$$

where the h^1 are the common emitter parameters of VT_1 and the h^2 are the common collector parameters of VT_2 and $\Delta = h_{11} \times h_{22} - h_{12} \times h_{21} = h_{11}\, h_{22} - h_{12}\, h_{21}$.

For a combination of typical transistors

$h_{11}{}^1 = 2\,000\Omega$, $h_{12}{}^1 = 16 . 10^{-4}$,

$h_{21}{}^1 = 49$, $h_{22}{}^1 = 50 . 10^{-6}\,\Omega^{-1}$

$h_{11}{}^2 = 2\,000\Omega$, $h_{12}{}^2 = 1$,

$h_{21}{}^2 = 50$, $h_{22}{}^2 = 50 . 10^{-6}\,\Omega^{-1}$

and although the transistors do not operate in the same conditions, one may compute the H parameters in order to get an idea:

$H_{11} = 1\,860\Omega$, $H_{12} = 0{\cdot}9986$,

$H_{21} = -2\,220$, $H_{22} = 2320 . 10^{-6}\,\Omega^{-1}$

The properties of the device are then easily computed, if $Zg = 500\Omega$ and $Z_L = R = 2\,000\Omega$.

$$Z_{in} = \frac{\Delta + H_{11} Y_L}{H_{22} + Y_L} = 800\text{k}\Omega$$

$$Z_{out} = \frac{H_{11} + Z_g}{\Delta + H_{22} Z_g} = 2\Omega$$

$$A_v = -\frac{H_{21}\, Z_L}{H_{11} + \Delta Z_L} > 0{\cdot}999$$

These properties can be favourably compared with those of Bloodworth's buffer amplifier.

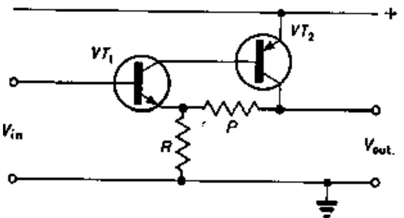


Fig. 3. Addition of resistance P

In order to get a voltage gain closer to unity it is always possible to add a resistance P of a few ohms as shown in Fig. 3.

It is important to note that this device is much better than the device described by Bloodworth in Fig. 1(c).

It is a Darlington's pair in the common collector connexion, and its H parameters are given[1] by equation (2) where now h^1 and h^2 are the common collectors parameters of VT_1 and VT_2.

$$[H] = \begin{bmatrix} \dfrac{\Delta h^1\, h_{11}{}^2+h_{11}{}^1}{1+h_{11}{}^2\, h_{22}{}^1} & \dfrac{h_{12}{}^1\, h_{12}{}^2}{1+h_{11}{}^2\, h_{22}{}^1} \\ -\dfrac{h_{21}{}^1\, h_{21}{}^2}{1+h_{11}{}^2\, h_{22}{}^1} & \dfrac{h_{22}{}^2+h_{22}{}^1\, \Delta h^2}{1+h_{11}{}^2\, h_{22}{}^1} \end{bmatrix} \quad \ldots\ldots\ldots\ldots\ldots (2)$$

for typical values we find

$H_{11} = 93\text{k}\Omega$, $H_{12} = 0{\cdot}9095$

$H_{21} = -2\,271$, $H_{22} = 2{\cdot}34 . 10^{-5}$

If $Z_g = 500\Omega$ and $Z_L = 2\,000\Omega$ we find

$Z_{in} = 820\text{k}\Omega$

$Z_{out} = 40{\cdot}9\Omega$

$A_v = 0{\cdot}974$

I naturally agree that the "D.C. Buffer Amplifier" provides a better performance regarding drift with temperature.

Yours faithfully,

J. AJDLER,

University of Brussels,
Belgium.

REFERENCE

1. AJDLER, J. Introduction d'elements actifs dans les ponts de mesure. (Universiti de Bruxelles).

The author replies:

DEAR SIR,—Mr. Ajdler's basic circuit comprises VT_2 and VT_3 of my circuit (Fig. 4) and I am grateful to him for pointing out that this combination alone provides a high input resistance R_{in}, a low output resistance R_{out} and a voltage gain A_V near to unity. The advantages of adding VT_1 and VT_4 are as follows:

(1) The voltage between base and emitter of VT_1 cancels the offset across VT_2, and the temperature drifts also cancel.

(2) With similar transistors, R_{out} is reduced because the effective source resistance is further reduced by the current gain of VT_1.

(3) R_{in} is virtually independent of the load ($\simeq \beta_1 R_1$).

(4) The voltage gain is nearer to unity because the incremental voltage between base and emitter of VT_1 cancels that of VT_2 to a first order unless R_L is very low.

If due to mismatch between the transistors the gain is more than unity it can be reduced by inserting a resistance in series with the emitter of VT_1, and if it is less than unity it can be increased by a resistance in series with the emitter of VT_2. (The latter increases the output resistance and in this respect is equivalent to the resistance P in Mr. Ajdler's circuit.) Unfortunately, this reintroduces the offset voltage which was removed by adjustment of R_1 and R_5. The latter resistances can be adjusted to drive VT_2, relative to VT_1, so that either $V_{be_1} = V_{eb_2}$ (no offset) or $\Delta V_{be_1} = \Delta V_{eb_2}$ (unity incremental gain). With different transistors it is not possible to satisfy both conditions together. By adding a linear resistance in series with the emitter of either VT_1 or VT_2 these conditions can be brought close together but cannot be made to coincide exactly because the emitter resistances are non-linear. If these non-linear resistances are swamped by large resistances in series with both emitters, the output resistance of the amplifier is increased. Similarly, the resistance P in Mr. Ajdler's circuit can be a small one to balance the incremental resistance of the transistor and give unity gain, or a large one to balance the chord resistance and give zero offset; however, these conditions are relatively far apart due to the lack of the compensating transistor and in the latter case the output resistance would be considerably increased.

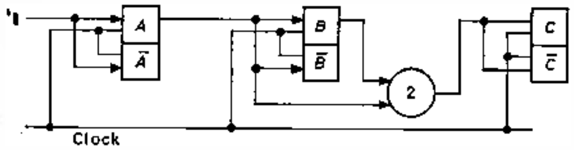
Yours faithfully,

G. G. Bloodworth,

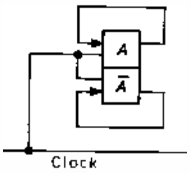
University of Southampton.

The Synthesis of Cyclic Code Generators

Dear Sir,—With regard to the article 'The Synthesis of Cyclic Code Generators' which appeared in the April 1964 issue. I wish to raise a few questions. The change equation for the first stage shown in Fig. 17 is $X_A = A + A'$. Does this imply that both the P and Q inputs to stage A are always at the voltage level corresponding to logic '1' as indicated below?



On the other hand, since $X_A = A + A'$ can be written as $X_A = A(A) + A'(A')$, the first stage then assumes the following configuration (a divide by-two circuit)



It seems to me that the first stage derived from the interpretation $X_A = A(1) + A'(1) = A + A'$ is not likely to operate as the flip-flop will be instructed to set and reset at the same time. However, the interpretation seems legitimate enough.

In the light of the above difficulty, then may one conclude that the P and Q expressions cannot contain a common term as in the following example.

	A	B	C	D
10	1	0	1	0
4	0	1	0	0
9	1	0	0	1
3	0	0	1	1
7	0	1	1	1
14	1	1	1	0
12	1	1	0	0
8	1	0	0	0
1	0	0	0	1
2	0	0	1	0
5	0	1	0	1
10	1	0	1	0
	etc.			

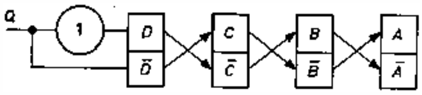
	$\bar{A}\bar{B}$	$\bar{A}B$	AB	$A\bar{B}$
$\bar{C}\bar{D}$	1_{0}	1_{4}	0_{12}	1_{8}
$\bar{C}D$	1_{1}	1_{5}	0_{13}	0_{9}
CD	0_{3}	1_{7}	1_{15}	0_{11}
$C\bar{D}$	1_{2}	0_{6}	0_{14}	0_{10}

The accompanying Karnaugh map has been derived according to conditions (a), (b) and (c) on page 256, namely $ABCD = 1$, $ABCD = 1$, and $(ABC\bar{D})(\bar{A}BCD) = 0{\cdot}1 = 0$.

$X_D = D(\bar{A}\bar{C} + BC) + \bar{D}(\bar{B}\bar{C} + \bar{A}\bar{C} + \bar{A}\bar{B})$.

Is this expression for X_D correct? Both P and Q are seen to contain $\bar{A}\bar{C}$.

If I assign a '1' to state 6, then $X_D = D(\bar{A}\bar{C} + BC) + \bar{D}(\bar{B}\bar{C} + \bar{A}\bar{C} + \bar{A}C) = D(\bar{A}\bar{C} + BC) + \bar{D}(\bar{B}\bar{C} + \bar{A})$. Can you tell me what has gone wrong in my application of your methods? Attacking the problem the conventional way, I obtain the following configuration:



$P = \dot{Q}$

$Q = A\bar{B}D + A\bar{B}\bar{C} + \bar{A}BC + \bar{A}\bar{C}\bar{D}$.

Yours faithfully,

J. S. Loui,

Framingham,

Mass., U.S.A.

The authors reply:

Dear Sir,—Regarding Mr. Loui's question about Fig. 17, the code is such that stage A changes state for every shift-pulse (e.g. it is a 'divide-by-two' circuit, as we pointed out in our article and as Mr. Loui confirms in his letter). P and Q could either be connected permanently to '1' (e.g. zero volts) or resistors RB_1 and RB_2 could be omitted altogether. In either case the stage would operate correctly. The suggestion that the stage would be instructed to set and reset simultaneously is not true. The change function X implies 'If $X = 1$, the stage will change state at the next shift-pulse; if $X = 0$, the stage will not change state at the next shift-pulse'. Thus, there can be no ambiguity; the X function merely determines whether or not the shift-pulse reaches the stage in question. Resistors RA_1 and RA_2 ensure that no pulse can ever reach both transistor bases simultaneously. This point is explained more fully in the section of our article subtitled 'Practical Realization of Feedback Shift Register' (p. 255). Consequently Mr. Loui's conclusion that P and Q must not contain a common factor is false, and both expressions for X_D in his example are correct, and generate the desired cyclic sequence. The only difference in performance is the possibility of a slightly different 'starting transient' when switching on; state 6 leads to state 12 if the first expression for X_D is used, and to state 13 if the second expression is used. In both cases the final cyclic behaviour is identical. State 6 is thus a 'don't care' state and may be assigned either '1' or '0' on the Karnaugh map, the condition giving the simplest logic normally being chosen.

Yours faithfully,

A. C. Davies,

Northampton College of Advanced Technology, London;

P. E. K. Chow,

University of Birmingham.

Photomultiplier Fatigue

Dear Sir,—The evidence produced by Mr. Gordy[1] to support his statement that photomultiplier fatigue does not manifest itself at anode currents of 10^{-6}A is not entirely convincing.

Fatigue measurements have been made on the 1P21 by Marshall, Coltman and Hunter[2], the results of three separate 2 hour tests showing that for initial currents of 10μA, 100μA and 1 000μA, the responses decreased to 50 per cent, 30 per cent and 12 per cent respectively. Considering the trend in these results, and the severity of the fatigue at the lowest initial current (10μA), one would expect the effect still to be present at 10^{-6}A.

As a general point, photomultiplier fatigue has been observed[3] at 10^{-7}A, a decrease in response to 92 per cent in 5min being recorded.

Yours faithfully,

R. J. Thresher,

The General Electric Co. Ltd,

Stanmore, Middlesex.

REFERENCES

1. Gordy, E. Photomultiplier Fatigue (Letter). *Electronic Engng.* 37, 120 (1965).
2. Marshall, —, Coltman, —, Hunter, —. *Rev. Sci. Instrum.* 18, 504 (1947).
3. Lenouvel, F., Daguillon, J. *J. Phys. Radium.* 15, 287 (1954).

The author replies:

Dear Sir,—Perhaps I can clarify matters by describing how we make our measurements, with respect to our previously published densitometer circuit[1]. After a 20min warm up period, a recording was made of the output of the densitometer for the next 16 hours. During this period, a drift of ·005 optical density units was recorded. This corresponds to ½ per cent of full scale on the 0 to 1 (most sensitive) range of optical density, or to 0·125 per cent of full scale on the 0 to 4 range. This recording would include drift in every component, photomultiplier tube, transistors, resistors, light source etc. I cannot say whence arises our overall drift, since opposing drifts may compensate each other. Since the overall performance of this instrument is what we intended to describe, this is what we reported. I might mention that a drift of this magnitude may perhaps arise from a diminished light output from the source lamp due to condensation of evaporated tungsten on to the bulb envelope, since the drift is in the direction of increasing optical density.

Returning to the question of photomultiplier fatigue, it may well be that it is present and is not observed because we do not begin to use the instrument immediately it is turned on, but rather wait for 20min for various stabilizations to occur.

Yours faithfully,

Edwin Gordy.

Roswell Park Memorial Institute

Buffalo, New York.

REFERENCE

1. Gordy, E., Hasenpusch, P., Sieber, G. F. A Four Decade Transistor Linear Densitometer *Electronic Engng.* 36, 808 (1964).

ELECTRONIC EQUIPMENT

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

(Voir page 349 pour la traduction en français; Deutsche Übersetzung Seite 356)

OXYGEN ANALYSER

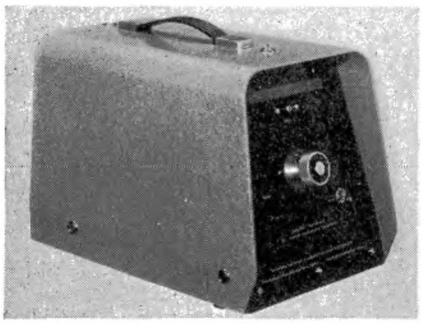
Servomex Controls Ltd, Crowborough, Sussex

(Illustrated below)

Servomex Controls Ltd is introducing a new version of its portable oxygen analyser, to be known as the type DCL.101 Mk. II. This analyser is of a design similar to the earlier version of the DCL.101 in that it uses the paramagnetic susceptibility measuring cell designed by the Distillers Company, having a platinum suspension and using electromagnetic feedback. The reading is obtained by manual setting of a dial calibrated 0 to 100 per cent oxygen and reading to 0·1 per cent oxygen. Power is normally obtained from dry batteries.

The Mk II version has an intrinsic safety certificate and is suitable for use in atmospheres containing gases of the ethylene, pentane and hydrogen classes as defined in B.S.1259. This means that the analyser may be used safely in many locations where hazardous conditions exist and it is particularly useful in refineries and chemical plants.

The improvements in the new analyser include a redesigned magnet which is completely free from external field



effects. There is now no restriction on the position of use, and this, together with its relative freedom from tilt effect (not exceeding 0·005 per cent oxygen per 1° of tilt) makes the DCL.101 Mk. II a suitable instrument for portable use.

Sampling equipment has now been included as standard. Each analyser is provided with a hand aspirator, a silica-gel drying tube and a sintered glass disk filter of novel design which incorporates very simple glandless pinch valves which can be easily renewed by the user.

EE 81 751 *for further details*

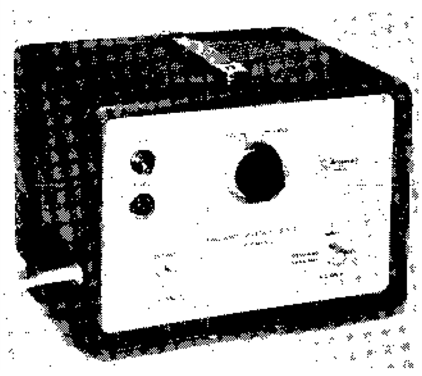
100Mc/s DIVIDER UNIT

Advance Electronics Ltd, Roebuck Road, Hainault, Ilford, Essex

(Illustrated above right)

To meet a demand for greater flexibility in the higher ranges of frequency measurement, Advance Electronics Ltd has added a 100Mc/s divider unit—the TCD100—to its growing list of products in this field.

The TCD100 is designed to extend the range of any electronic counter now commercially available from 1Mc/s upwards, to 100Mc/s. Measurement is



achieved by a division factor of 100 or 20, controlled through a front panel switch. Input and output parameters are sufficiently flexible to accommodate a wide range of input sources. The instrument is transistorized and utilizes tunnel diodes.

It has an input sensitivity of 50mV into 50Ω, and output pulses at 3V peak-to-peak are available from 10kc/s to 1Mc/s(÷100) and 50kc/s to 5Mc/s(÷20). The accuracy of the basic counter is unaffected by the divider unit.

EE 81 752 *for further details*

ARTIFICIAL TRANSMISSION LINE

Telefon Fabrik Automatic A/S, 7 Amaliegade, Copenhagen, Denmark

(Illustrated right)

The artificial transmission line type E4 has general characteristics that very closely simulate telephone subscribers lines.

In its principal application the artificial line is intended to be employed as a laboratory device to provide a simulation of the effects of linelength on subscriber installations, i.e. telephone sets, data transmission systems etc.

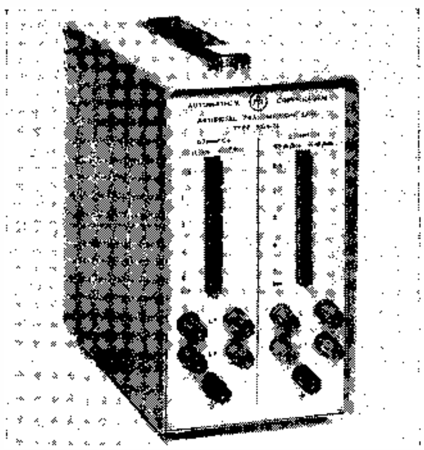
The artificial line consists of *RC*-sections each corresponding to a unit length of line of 0·5km (other unit length on request). By connecting a number of sections in tandem any line-length from 0·5km can be simulated in steps of 0·5km.

This type of artificial line comprises in fact two lines of a maximum of 27 sections resulting in a maximum of 13·5km of line. The two lines can of course be connected in series thereby obtaining a total linelength of 27km.

Tandem connexion of sections is established by a series of push-buttons. Five buttons are used for inserting linelengths of 0·5, 1, 2, 4 and 6km respectively. The resulting length of inserted sections is hence the sum of values corresponding to the activated push-buttons.

The artificial line is mounted in a metal cabinet. The front panel is divided in two symmetrical halves, one for each line. In addition to the push-buttons the terminals for the line are mounted on the front panel.

The artificial line is equipped with binding posts for input, output, and ground connexions. The components are mounted on printed circuit cards, each comprising 9 sections. The printed cards are mounted in the cabinet by means of multi-connectors and are easily removed for repair or inspection. The cards are all identical and hence interchangeable.



EE 81 753 *for further details*

ELECTRONIC ENGINEERING will occupy stand number 58 in the Grand Hall, Olympia, at the R.E.C.M.F. Exhibition from 18 to 21 May, and visitors will be welcome.

OSCILLOSCOPE UNITS

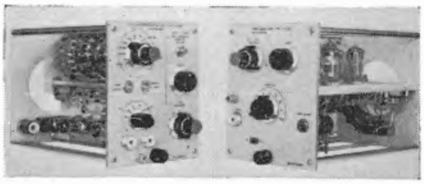
Solartron Electronic Group Ltd, Farnborough, Hampshire

(Illustrated on page 341)

Two new plug-in units are available for the Solartron types CD1220 and CD1212 (CT484) oscilloscopes.

The high gain differential plug-in unit CX1258 features high d.c. sensitivity (100μV/cm), high in-phase rejection (80dB—10 000 : 1) and very low drift. In addition a maximum bandwidth of 200kc/s enables the amplifier to be used for a wide range of general purpose applications.

The high gain input stage is designed to maintain an accurate balance for maximum in-phase rejection. For a.c. coupled conditions, d.c. feedback reduces a basically low d.c. drift to negligible proportions. Special Quality, Reliable



Valves are used in critical positions.
CX1258 Specification:
Sensitivity:
$100\mu V$/cm to 2V/cm in 14 calibrated ranges.
Bandwidth:
D.C. to 50kc/s at $100\mu V$/cm.
D.C. to 100kc/s at $200\mu V$/cm.
D.C. to 200kc/s at $500\mu V$/cm.
In-phase rejection:
Not less than 80dB from d.c. to 50kc/s.
Max. In-phase voltage:
$\pm$25V peak from $100\mu V$/cm to 20mV/cm.
$\pm$250V peak from 20mV/cm to 2V/cm.
Calibration accuracy:
3 per cent.

The CX1259 has a similar performance to the CX1256, but includes an additional $\times 10$ d.c. coupled transistor amplifier providing a maximum sensitivity of 5mV/cm at a bandwidth of d.c. to 24Mc/s. The basic performance of d.c. to 40Mc/s at 50mV/cm is still retained. Composite feedback enables excellent pulse response to be maintained on all ranges.

'Y' shift applied as a 'backing off' potential, ensures undistorted displays for signals with a d.c. content.

CX1259 Specification:
Sensitivity and bandwidth:
$\times 10$: 5mV/cm—d.c. to 24Mc/s.
$\times 1$: 50mV/cm—d.c. to 40Mc/s.
Ranges:
Calibrated in 1-2-5 steps with a fine control.
Calibration accuracy:
3 per cent.

In the illustration the high gain differential unit, CX1258 is on the left and the wide band unit CX1259 is on the right.

EE 81 754 *for further details*

'TEACHING' INSTRUMENTS

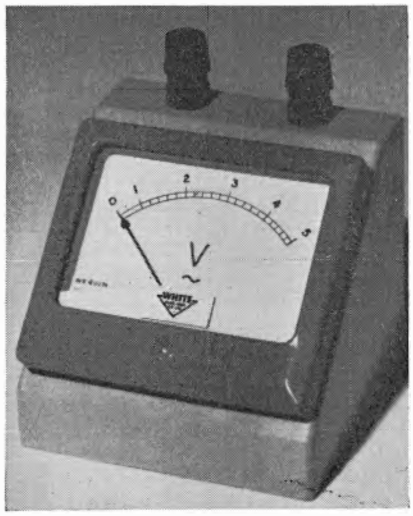
White Electrical Instrument Co. Ltd, 10 Amwell Street, Rosebery Avenue, London, E.C.1

(Illustrated above right)

Although suitable for general laboratory use, these instruments have been designed specifically for modern physics teaching. Known as type RM/AS they are of robust construction, designed especially to withstand the rigours of school use and yet at the same time incorporate a movement of the high quality and accuracy that is necessary for this new class of work.

They are 2½in dial instruments with clear easily-read scales, mounted in two-tone easy-clean pastel grey plastic inclined bench stand cases with 4mm socketed screw terminals on top.

The movements are the latest spring controlled pattern having polished hardened steel pivots and synthetic sapphire jewelled bearings. The moving-iron type RMI is pneumatically damped, suitable for a.c. or d.c., but is only available with a side zero. The moving-coil type RMC is only suitable for d.c. but is obtainable with either a side or centre zero. It is dead beat and electrodynamically damped. Both types of meter are available in a wide range of voltage and current calibrations.



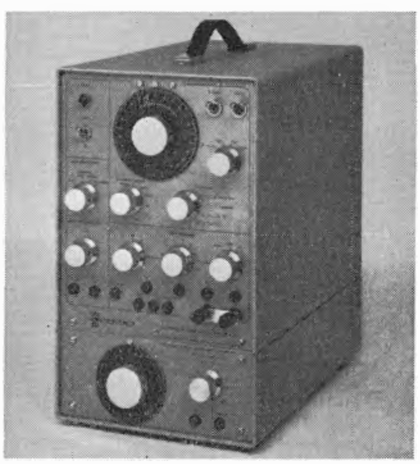
EE 81 755 *for further details*

VARIABLE PHASE ATTACHMENT

Servomex Controls Ltd, Crowborough, Sussex

(Illustrated below)

Servomex Controls Ltd has produced a variable phase attachment for use with the recently developed low frequency waveform generator type LF 141. The main purpose of the new unit, the VP. 142, is for rapid and accurate phase measurement over the range of frequencies from 2 000c/s down to 1 cycle in 500sec.



This is done by the well-known Lissajous figure method whereby an ellipse degenerates into a straight line when the phase is correct. By techniques which are explained in the data sheet, it is possible to make measurements even at very low frequencies in less than 1 cycle, using either a cathode-ray oscilloscope or some form of pen recorder. An amplitude control is provided so that when the Lissajous figure is inclined at 45° the amplitude can be read off the amplitude control directly, as well as the phase. On the rare occasions when there is too much noise or distortion in the signal under test it is possible to fit a suitable band-pass filter.

The variable phase voltage is derived from a potentiometer with a circular winding fed at 0°, 90°, 180°, and 270°. To enable the amplitude measuring method to work, a special circuit has been provided to keep the amplitude constant as the phase control is rotated.

This unit is screwed to the bottom of the LF 141 and receives its inputs from a multiway plug and socket. The combined equipment, in addition to being used as a source of low frequency signals of many sorts for phase measurement in power systems can be employed for a variety of other purposes such as, for example, the demonstration of the 'starting surge' in a.c. systems. By removing a paralysing voltage from the gate terminal a train of sine waves can be started at any desired angle.

Among the many types of waves that can be produced are square waves with variable rise time in which both the repetition rate and the rise time are separately controllable and also sine-squared pulses which generate harmonics inside an accurately known bandwidth.

For medical and other applications a train of pulses of known number can be produced. The counter can be counting square pulses of fast time from the auxiliary square-wave output terminal while the main output terminal is emitting whatever waveform is required.

The Servomex VP 142 variable phase attachment has been designed with particular attention to ease of maintenance. Its electronic components are all on printed circuit boards hinged for easy servicing. Most of the components are chosen from the Interservice approved list and are run at the proper Service ratings.

EE 81 756 *for further details*

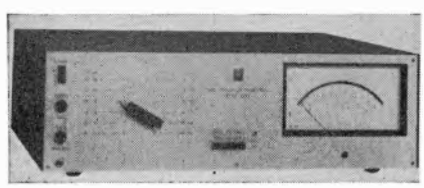
A.F. MILLIVOLTMETER

Elektromesstechnik Wilhelm Franz KG, 80 Kaiserstrasse, 763 Lahr/Schwarzwald, Germany

(Illustrated below)

The fully transistorized voltmeter EMT 125 has been specially designed for exact measurements on audio amplifiers and equipment as well as electro-acoustic installations. Its extremely low consumption permits permanent mains connexion without noticeable heat development.

The modern flat case allows stacking of several units; all terminals, push-buttons and switches are arranged on the front panel together with the meter.



The scale is calibrated in volts and decibels.

The unit has 12 measuring ranges between 1mV f.s.d. and 300V f.s.d. The smallest readable value is 100μV. The frequency range extends to 200kc/s; a low-pass filter is incorporated which can be inserted by push-button operation and limits the highest frequency at which measurements may be carried out to 20kc/s. Waveforms differing a great deal from sine waves can also be measured. Measurements in r.m.s. or peak values can be selected by push-button.

Measuring r.m.s. values in accordance with the German standard DIN 45 402 and at a pulse frequency of 200c/s, the instrument achieves a duty cycle of 1:10 at full scale and 1dB deviation. Peak value measurements are suitable for the determination of extraneous voltages, e.g. noise voltages, when frequencies above 20kc/s may be filtered out.

Used as amplifier, it gives the highest amplification of 60dB in the 1mV range; it drops 10dB per range. 1V into 4kΩ is available at the output.

A stabilized reference voltage source is incorporated to facilitate re-calibration.

EE 81 757 *for further details*

DIGITAL CLOCKS

Distributed by: Systems Engineering Services Ltd. 35-39 South Ealing Road, Ealing, London, W.5

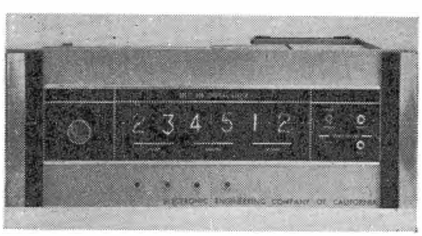
(Illustrated below)

Two new precision digital clocks to provide accurate time of day as a decimal display, and in binary-coded-decimal form are being manufactured by the Electronic Engineering Company of California.

The new instruments, the EECO 815 digital clock and the EECO 816 digital clock, both provide parallel logic levels; in addition the 816 provides serial read-out. They provide direct time codes to data logging or data recording systems, computers, strip chart recorders, printers, etc. Both models feature all solid-state design with no mechanical devices such as motors, relays, or stepping switches.

Both the EECO 815 and EECO 816 are synchronized to power line frequency, or may be supplied with an internal frequency source. In addition to the time codes, six square wave pulse rate outputs are simultaneously available. Each model includes a power failure indicator which remains on until a power failure reset is pressed.

The EECO 815 (parallel output code



only) provides an accurate time code indicating hours, minutes, and seconds in binary coded decimal form for direct entry into data recording or logging systems using paper tape, magnetic tape, or direct printing systems; or may be used for direct input to computers.

The EECO 816 (both parallel and serial read-outs) can be used for the above, and additionally for recording time serially on strip chart recorders or on magnetic tape recorders. A modulated carrier (100c/s or 1 000c/s carrier) output can be provided, as an option, for recording the serial time code as a width-coded, amplitude-modulated carrier on analogue magnetic tape.

EE 81 758 *for further details*

TRANSFORMER RATIO ANALOGUE COMPUTER

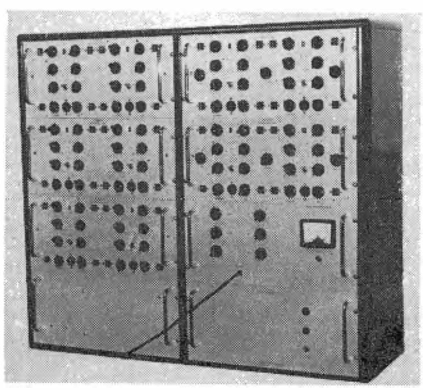
F. C. Robinson & Partners Ltd, Davies House, 181 Arthur Road, Wimbledon, London, S.W.19

(Illustrated below)

Specially designed to meet the needs of colleges and universities, this analogue computer, based on the well-known principles of using the turns ratio of pairs of computing transformers to represent complex numbers, has wide applications in circuit and network analysis.

The computer consists of:

Three twin universal impedance units



(for direct representation of impedance or admittance at any phase angle).

Two twin special universal impedance units (impedance or admittance, voltage or current source, transformer or mutual impedance).

One 400c/s transistorized oscillator.

One null metering unit using a single computing transformer, logarithmic amplifier and meter.

One power supply unit.

The accuracy of circuit constant settings and of measurement is 0·1 per cent.

EE 81 759 *for further details*

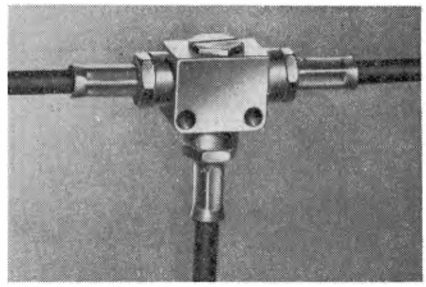
SUBMINIATURE COAXIAL CABLE CONNECTOR

Sealectro Ltd, Hersham Trading Estate, Walton-on-Thames, Surrey

(Illustrated above right)

A 75Ω, crimp-type 'T' junction block for subminiature coaxial cables is the latest addition to the connectors available from Sealectro Ltd. The 5116 connector utilizes crimp type cable connexions which are particularly suitable for high production rate assembly and provide a cable termination that is stronger than the cable itself.

This connector incorporates clearance



holes for panel or bulkhead mounting and has provision for stacking units when a cable junction centre is required. Additionally, the 5116 'T' junction block features ·0001in gold plating which, when combined with ConheX design, assures maximum electrical efficiency.

EE 81 760 *for further details*

ULTRASONIC FLAW DETECTOR

Ultrasonoscope Co. (London) Ltd, Sudbourne Road, London, S.W.2

(Illustrated below)

The mark 4 Ultrasonoscope is designed to give the performance and versatility needed to solve the most difficult inspection problems. It is suitable for routine testing in the works or advanced research in the laboratory. With a monitor unit it can operate a fully automatic scanning and recording system.

Applications in different industries often demand conflicting instrument characteristics and so the mark 4 amplifier is interchangeable. A special amplifier can be designed and easily fitted for a special application. The standard amplifier has a wide bandwidth and dynamic range at each of the frequencies of 10, 5, 2½, 1¼ and ½Mc/s. It can resolve short pulses and is linear.

Scatter from the structure of a material can be suppressed with a continuously variable reject control. The calibrated attenuator has a range of 0 to 80dB with a reading accuracy of 1dB. The powerful transmitter and high gain amplifier give greatly increased penetration on single or double probe operation.

Either rectified, unrectified or B-scope presentation may be selected. On B-scope the defect echoes brighten the time-base

and the display is similar to a radiograph.

The cathode-ray tube has a 5in flat face with photographic phosphor and optical filter for improved contrast. The time-base is linear and the fastest speed is 2μsec so that ¼in of steel may be displayed across the full screen width for accurate thickness and velocity measurement. The maximum range is 20ft.

The time-base delay is continuously variable from 0 to 300μsec by a calibrated control allowing magnification of any section of the trace. The pulse repetition frequency is from 50 to 1 000c/s and the mark 4 is fully stabilized against mains voltage fluctuations.

EE 81 761 *for further details*

PRINTED CIRCUIT DRILL

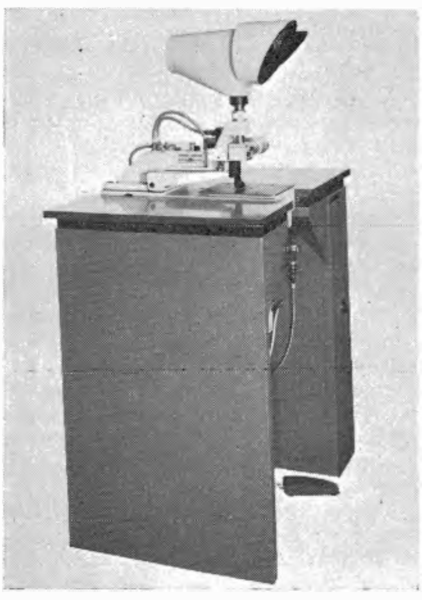
Distributed by: Alfred Jacoby (Machine Tools) Ltd, 54 Manchester Street, London, W.1

(Illustrated below)

The Posalux 'Copyfor' drilling machine, manufactured by Frautschi & Monney, Switzerland, was developed especially for the manufacture of printed circuits although it is equally suitable for the drilling of aluminium foil name plates and similar components.

Drilling takes place by means of a vertically mounted 'Unifor' 10 hydro-pneumatic drill unit, operating in an upward direction. The drilling cycle can be either hand-operated or semi-automatic. In either event, the stack of components, carrying the drilling template on top, is moved until the spring-loaded location pin which is positioned in the same axis as the drill spindle, locates one of the holes in the template. When working automatically, this will cause the components to be pneumatically clamped and the drill to feed forward, drilling being almost instantaneous. On hand operation, the clamping and drilling cycle is initiated manually.

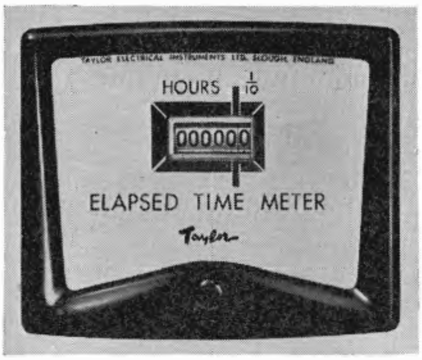
The projection screen or microscope, which can be supplied with the machine, enable drilling templates to be manufactured on the machine itself by spotting off the original printed circuit. This arrangement is also beneficial for small quantity batches where the manufacture of individual templates would not be warranted. The projection screen has a corrected image with high light intensity and magnification.



The machine is capable of high speed production. This is partly due to the design of the index pin, the pressure on which is adjustable to suit working conditions. Its movement is extremely sensitive, thereby eliminating scrap due to accidental starting up of the drilling spindle. The head tilts backwards, giving easy access to the unit for purposes of drill changing, and ample provision is made for swarf clearance. A suction unit can be supplied as an optional extra and fan ring, forming part of the drill chuck, ensures that all swarf is blown into the disposal chute.

High spindle speeds are employed to reduce drilling time and prevent formation of burrs. Maximum drill size is 3mm (⅛in) diameter and the machine will accommodate boards up to 420mm (16½in) wide.

EE 81 762 *for further details*



ELAPSED TIME METER

Taylor Electrical Instruments, Montrose Avenue, Slough, Buckinghamshire

(Illustrated above)

Taylor Electrical Instruments (M.I. Group) has introduced a new elapsed time meter, which matches the Vista range of meters and has the same frontal appearance as the Vista model 40.

Models are available for use on either 110 to 125 or 200 to 250V, 50 or 60c/s mains controlled supply and they are capable of recording hours and tenths of an hour up to 99999·9 hours after which the counter automatically sets itself to zero.

A self-starting, synchronous motor enables the indication to remain correct even though the periods of use may be of short duration.

EE 81 763 *for further details*

PULSE GENERATOR

Hewlett Packard, Dallas Road, Bedford

(Illustrated above right)

Hewlett-Packard GmbH—the Hewlett-Packard research and manufacturing facility in Germany, has introduced a compact, low cost pulse generator. The Hewlett-Packard model 8000A pulse generator is designed for making

response measurements on broad band instruments, circuits and semiconductors.

The generator's 10V output (into 50Ω) is essentially a flat topped pulse with a repetition rate of 100kc/s and a rise time of less than 20nsec. All pulse top variations, such as overshoot and ringing, are held to less than 2 per cent; the flat top is maintained for at least 100sec. Fall time is less than 20nsec. Step recovery diodes developed by Hewlett-Packard are utilized as the pulse shaping elements in the output circuit.

Among the features of the instrument are step attenuated output (0·1 to 10V in six steps), positive or negative output polarity, and an advanced trigger output. The trigger pulse eliminates any need for a delay line when working with sampling or other high speed oscilloscopes. It has a 200nsec advance, has a rise time of less than 6nsec, a width of 20nsec and an amplitude of 0·5V into 50Ω.

EE 81 764 *for further details*

SOLID STATE RELAYS AND TIMERS

Keyswitch Relays Ltd, 120-132 Cricklewood Lane, London, N.W.2

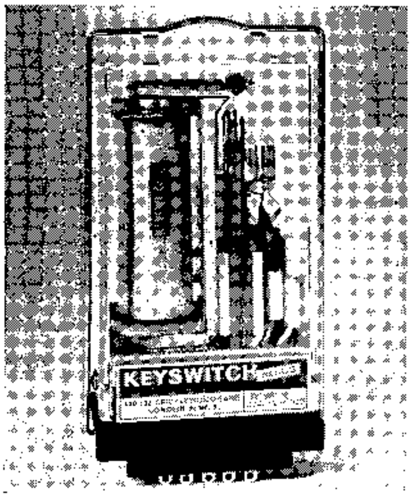
(Illustrated on page 344)

Keyswitch Relays Ltd has introduced the P33VT range of plug-in timers and transistorized relays, designed for low output control and available in a standard module. Developed from the P33 relay series, the new range can be used in conjunction with that series and with the recently-announced type 304 component board unit.

Applications of the P33VT include atmosphere, temperature and thermal control, detecting devices, machine tool automation, voltage and current sensing, container filling control, alarm circuits, and a wide variety of data processing operations.

There are four standard types in the range, each with a large number of possible design variations according to specific requirements. P33VT-1 puts a low signal of the order of 10μA a.c./d.c. into an impedance of 10kΩ, controlling a circuit to switch the relay. Voltage supply is normally 24V a.c./d.c., contacts being either heavy duty changeover (5A 230V a.c. resistive), or light duty changeover up to 1A.

The P33VT-3 relay series is operated



either by the closing of external contacts passing a low input current of 500μA or less, or by varying external resistance. For indicating meter applications, meter contact bounce and fluctuations are compensated.

P33VT-7 provides for direct switching of the relay with high-level photocells, for 15μA low level light sources using a two-stage solid-state amplifier, or for control by external light source into an in-built photocell.

The P33VT-8 series comprises stabilized timing relays with snap action, and delay ranges of 0 to 10sec, 5 to 30sec, 10 to 60sec, 15 to 120sec, and up to 5min.

Working voltage of the P33VT range plug and socket assembly is 750V a.c. or d.c., and insulation resistance better than 1 000MΩ at 500V d.c.

EE 81 765 *for further details*

PRESSURE TRANSDUCERS

Intersonde Ltd, The Forum, High Street, Edgware, Middlesex

(Illustrated below)

New into the Intersonde range of bonded strain gauge pressure transducers is the B.M. series, made specifically to cover the low pressures: 0 to 100 lb/in² to 0 to 750 lb/in².

The transducers are based on a pressure responsive element in the form of a closed-end beryllium copper tube to which four bonded strain gauges are attached and connected in a four-arm bridge configuration. Gaseous or liquid pressure media admitted to the interior of the tube produce minute dimensional changes which are instantly converted into changes in resistance by the two active arms of the bridge. The bridge unbalance signal is proportional to applied pressure.

The transducers have an output resistance of 350Ω and produce 30mV output at full rated pressure with 20V excitation. The combined non-linearity and hysteresis error is within ±1·0 per cent of full range and the operating temperature limits extend from −40°C to +120°C.

EE 81 766 *for further details*

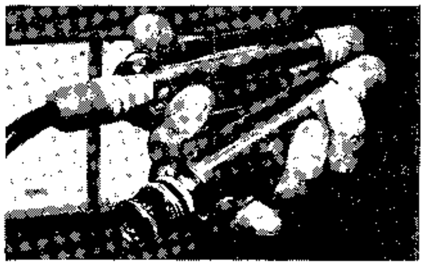
MINIATURE TELEVISION CAMERA

EMI Electronics Ltd, Hayes, Middlesex

(Illustrated below)

A new all-transistor closed-circuit television camera, only 4½in long and weighing 1¾lb, has been developed by EMI Electronics Ltd.

Despite its small size, the camera is extremely rugged and produces very high resolution pictures, so it is particularly useful in inaccessible locations where rough handling and poor illumination are expected. It can be operated on 405, 525 and 625 line standards, and can be changed from one standard to another merely by pressing a button.

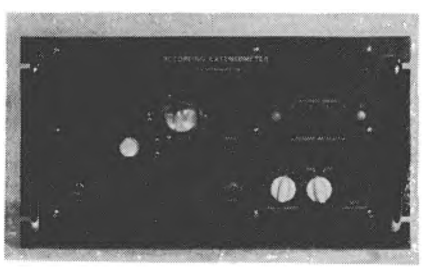


Camera head-equipment is contained in two sealed stainless steel cylinders, each 4½in long and of 1·7in diameter. The lens head unit, which weighs 1¼ lb, contains an EMI ½in vidicon camera tube. It can easily be held in one hand, leaving the other free to focus by rotating the lens, or it can be mounted in restricted locations.

Contained in the other cylinder is the amplifier head unit, weighing ¾ lb, which can be up to 100ft away from the lens head and joined to it by cable.

Camera control unit and other units comprising the camera channel can be up to 1 000ft away.

EE 81 767 *for further details*



RECORDING EXTENSOMETER

C.N.S. Instruments Ltd, 61 Holmes Road, London, N.W.5

(Illustrated above)

C.N.S. Instruments Ltd announce that their recording extensometer is now available with automatic range switching. When fitted with this facility the instrument is supplied in 19in rack mounting form with output of 0 to 10mV (0 to 50mV to special order) and is intended for use with a potentiometric recorder such as a Honeywell-Brown Electronik.

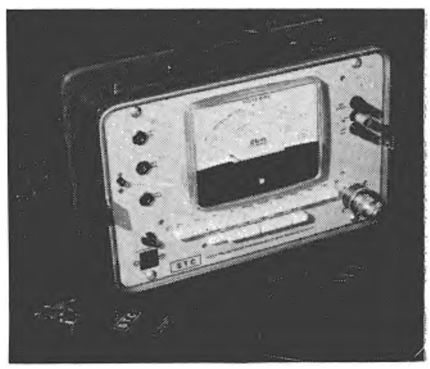
Six ranges are supplied as standard. 250 and 500μin (6 and 12 microns), 0·001, 0·01, 0·025 and 0·1in (0·025, 0·25, 0·625 and 2·5mm). Thus on a 10in (250mm) chart, increments of 0·000002in (0·05 microns) may readily be measured.

Other ranges to customers' requirements are available to special order.

The range switching facility is particularly useful in creep and other mechanical tests where the rate of extension is not a linear function of time. Permanent records overnight and over weekend may therefore now be obtained without skilled personnel having to be present. Thus a high sensitivity setting may be used at the start of the test and as the test progresses, and full scale deflexion in each range is reached the instrument is automatically switched to the next lower range.

Particular care to minimize temperature effects in the transducers, which are available singly or as summating pairs, has been taken and the windings of the static and moving coils are carried on quartz formers suitable for operation at temperatures up to 45°C as standard or higher to special order subject to an absolute maximum of 100°C.

EE 81 768 *for further details*



MILLIVOLTMETER

Standard Telephones & Cables Ltd, Transmission Testing Apparatus Division, Corporation Road, Newport, Monmouthshire

(Illustrated above)

A versatile new high impedance millivoltmeter has been introduced by Standard Telephones and Cables Ltd, for general purpose and telecommunications use.

Known as the 74251 the unit operates throughout the frequency range 20c/s to 20Mc/s and permits measurements in 10 ranges covering 0 to 1mV to 0 to 30V. An active probe gives an input impedance greater than 100kΩ in parallel with 40pF.

Two versions are available, the 'A' version being for operation on self-contained batteries and the 'B' version for use with self-contained batteries or from a.c. mains.

The 74251 is a portable transistorized instrument designed for general application and for use in the baseband frequency range of microwave radio link systems. It may also be used as a bridge detector, and its calibrating oscillator will supply a 100kc/s (approx) signal to external circuits. The internal amplifier has an outlet to permit its use in external circuits.

The meter has two voltage scales and also a decibel scale for use on 75Ω circuits. The condition of the batteries can also be checked on the meter.

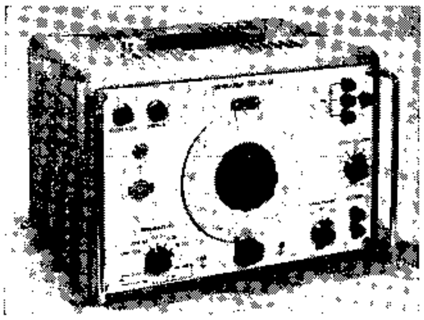
EE 81 769 *for further details*

V.L.F. GENERATOR

Distributed by: Claude Lyons Ltd, 76 Old Hall Street, Liverpool, 3

(Illustrated below)

The C.R.C. type GB64 v.l.f. generator provides five 10:1 ranges, calibrated in frequency and period, covering the range 0·005c/s to 500c/s, and alternative sine, square or triangular outputs are available with a balanced output variable up to 40V peak-to-peak into 10kΩ from an internal impedance of 100Ω. The sine wave output has 2 per cent maximum distortion, square wave output a rise time of 25μsec, and triangular output a maximum slope error of 2 per cent. Random noise level is 20mV maximum. Frequency stability of 0·5 per cent and level stability of 0·2dB for ±10 per cent mains variations is ensured by stabilized supplies.



A synchronization pulse output (25V peak, 50μsec) comprising alternative positive and negative pulses, is provided for such purposes as oscilloscope triggering.

EE 81 770 *for further details*

HIGH RESOLUTION VIDICONS

The English Electric Valve Co. Ltd, Chelmsford, Essex

(Illustrated above right)

The range of English Electric Valve Company high resolution, separate mesh 1in vidicons, has been further extended with the introduction of two types having high peak response in the 'blue' region of the spectrum.

These tubes, the 8625 (P846) and 8626 (P847), incorporate entirely new photosurfaces providing: very high and uniform sensitivity; extremely low lag and reduced long term sticking characteristics; correct panchromatic response with tungsten illumination; high resolution at high signal currents.

Excellent signal uniformity is achieved with improved manufacturing techniques whereby photo surfaces are 'prefabricated' to ensure an overall even deposition. Studio lighting levels may be

considerably reduced due to the very short lag times promoted by the new photo surfaces.

A further important advantage is extremely good reproduction of flesh tones, due to the well balanced spectral sensitivity of these vidicons.

The introduction of the new photo surfaces, in combination with separate field mesh construction, makes the 8625 (P846) and 8626 (P847) vidicons eminently suited for live and telecine pickup in both colour and monochrome television. The extra blue sensitivity gives a useful improvement in signal-to-noise ratio in colour television cameras, and also promotes high signal output in applications where light with high blue content is used e.g. fluorescent lamps, natural daylight etc.

The 8625 (P846) has a standard 6·3V/0·6A heater, and the 8626 (P847) has the new low consumption 6·3V/·095A heater.

EE 81 771 *for further details*

DRY REED INSERT

Hivac Ltd, Stonefield Way, South Ruislip, Middlesex

A new dry reed insert—XS10—has been added to the Hivac range of standard types. Designed to meet the same electrical specification as the XS2 series, it has a much shorter glass length and a reduced diameter.

The dimensions are: glass length 1·4in, overall length 3·13in, and maximum diameter 0·185in.

The contact surfaces are plated with gold which is partially diffused into the base material, and will give long life operation at currents up to 250mA. The standard version of the XS10 operates at between 30 and 70At and has a minimum breakdown voltage of 450V d.c.

Operate time is less than 1·3msec, bounce duration less than 0·5msec and release time 0·3msec, when measured at twice just operate ampere turns.

The XS10 provides a British made equivalent to some imported types of reed insert, and augments the already wide range of Hivac high performance types.

EE 81 772 *for further details*

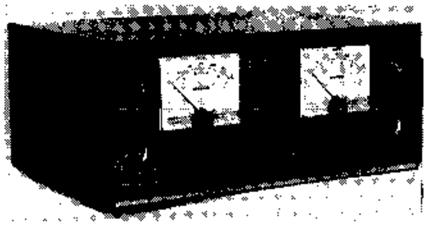
POWER SUPPLIES

D.E.B. Electronics Ltd, New Barnes Mill, Cottonmill Lane, St. Albans, Hertfordshire

(Illustrated above right)

A new range of laboratory power supplies from D.E.B. Electronics Ltd feature low ripple and close regulation, and use silicon semiconductors throughout. The basic bench unit is a double supply, each half of which delivers 0 to 1A at 0 to 30V, with a ripple of around 300μV, and a regulation of better than 10mV. Stability is better than 1 000:1.

Constant current protection is incorporated, the maximum current available being selected on a seven-position rotary switch. The output characteristic is so nearly rectangular that the unit can be used as a current source with a current regulation of 5mA. Four-terminal sensing is provided.



One per cent, clear front measuring instruments, with scale lengths of 4in are fitted, one per half of the double supply, to display voltage or current.

The voltage adjustment is by means of a single vernier knob giving a stepless sweep over the whole range of voltage.

Each half of the unit is floating, and may be connected in series or in parallel with the other half, without loss of performance or protection. Protection is also provided to prevent damage in the event of connexions of opposite polarity being made. Supplies are available to give a total of 8A in the parallel mode or 120V in the series mode.

All units are completely compatible and are housed in identical instrument cases, which are rapidly detachable for rack mounting directly on the 19in by 7in front panels. They have been designed for reliability and ruggedness, and will run continuously at full current into a short-circuit.

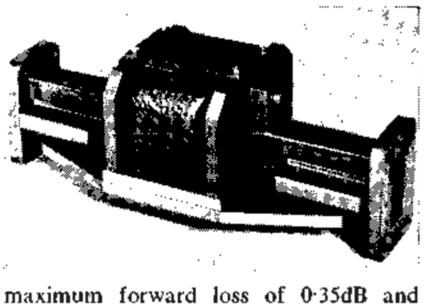
EE 81 773 *for further details*

FERRITE ISOLATORS

The M-O Valve Co. Ltd, Brook Green Works, London, W.6

(Illustrated below)

The M-O Valve Co. Ltd has introduced a new range of ferrite isolators covering the frequency bands 5 925 to 6 425Mc/s, 5 925 to 6 175Mc/s and 6 175 to 6 425Mc/s. The CIC4 (illustrated) is a field displacement isolator with a



maximum forward loss of 0·35dB and a minimum reverse loss of 35dB. The v.s.w.r. is 1·02:1. The CIC5 and CIC6 are resonance isolators with forward and reverse loss of 0·5dB min. and 25dB max. respectively. Both have a v.s.w.r. of 1·06:1. The r.f. connexions on all three devices are waveguide 14.

These devices are particularly suitable for use in broadband radio communication systems.

EE 81 774 *for further details*

Short News Items

The Council of the Institution of Electrical Engineers has decided to modify the learned-society structure and set up a Control and Automation Division in order to take account of the growing significance of control engineering and automation and of the corresponding importance of their place in the activities of the Institution.

At the same time they will set up a Science and Education Joint Board to provide a guiding and co-ordinating body for the professional groups dealing with basic science and education, which are subjects of common interest to all members and Divisions of the Institution.

The amended learned-society structure will come into effect on 1 October 1965 and will comprise:

Control and Automation Division

Electronics Division

Power Division

Science and Education Joint Board.

Each Division will continue to work through specialized professional groups. The professional groups (at present in the Science and General Division) that deal with basic science and education will be constituted as joint professional groups, each being represented on those Divisional Boards that share an interest in its work, their activities being co-ordinated by the Science and Education Joint Board.

The Radio and Electronic Component Manufacturers' Federation is organizing a second 'Exports Day' at their headquarters at 6 Hanover Street, London, W.1, on Tuesday, 11 May 1965.

This year the theme of the conference will be 'The Challenge of Improving the Exports of the Electronics Industry' and the programme will include speakers from the industry, both at home and overseas, as well as discussion periods.

Apart from the addresses, the Exports Day also provides an opportunity for over 200 member firms to meet for discussion on many aspects of exporting which are of common interest.

In 1964, U.K. exports of electronic products totalled £100M to which the components sectors contributed nearly £60M in 'direct' exports.

An International Symposium on Microminiaturization in Automatic Control Equipment and in Digital Computers is to be held in Munich, Germany, on 21 to 23 October 1965 under the auspices of both the International Federation of Automatic Control (IFAC) and the International Federation of Information Processing (IFIP).

The main subjects of the symposium will be: new pneumatic and hydraulic and optical components; storage components; integrated circuits in planar, thin film and hybrid techniques, construction and interconnexion techniques of miniaturized components; microminiature measuring devices and transducers; trends and limits of microminiaturization.

Attendance at the symposium will be by registration only. Registration forms may be obtained from: Verein Deutscher Ingenieure—Abt. 0, P.O. Box 10 250, Dusseldorf, West Germany.

An international conference on 'U.H.F. Television' will be held in London on 1 to 2 September 1965.

Aspects to be covered include receiver and transmitter design, propagation, receiving and transmitting aerials, parametric amplifiers, and test equipment.

The conference has been timed to take place during the Radio Show, which this year will be held from 24 August to 4 September and will have an international content, so that visitors to the Show can participate.

The sponsoring bodies, the Institution of Electrical Engineers Electronics Division, the Institute of Electrical and Electronics Engineers, the Institution of Electronic and Radio Engineers, and the Television Society, are inviting papers and contributions. In the first instance synopses of about 250 words should be submitted to the U.H.F. Television Conference Joint Secretariat, 8-9 Bedford Square, London, W.C.1, by 20 April 1965.

Complete papers, and contributions of not more than 2 000 words or the equivalent including diagrams, should arrive by 1 June.

A Sectional Meeting of the World Power Conference is to be held in Tokyo on 16 to 20 October 1966 and the theme is 'Problems of Future Years in Energy Utilization'.

A committee under the chairmanship of Sir Josiah Eccles has been set up by the British National Committee to arrange for the preparation of British papers to be presented at the Tokyo meeting.

Offers of papers within the scope of the Technical Programme should be sent without delay to: The Secretary, British National Committee, World Power Conference, 201-202 Grand Buildings, Trafalgar Square, London, W.C.2.

The Institute of Mathematics and its Applications is to hold a conference on 'The State of the Art in Numerical Analysis' at the University of Birmingham on 18 to 21 July 1965.

The conference aims to appraise present-day methods in numerical analysis in a way which will be of interest to mathematicians in industry, Government establishments, universities, colleges and schools. The emphasis will be on the survey of the practical techniques of numerical analysis rather than specialist treatment in depth.

Applications to attend the conference should be made before 14 June to The Registrar, Institute of Mathematics and its Applications, c/o Imperial College of Science and Technology, London, S.W.7.

A Machine Tool and Production Equipment Group has been formed by the Scientific Instrument Manufacturers' Association, bringing together those member firms whose interests include instrumentation and control in this field.

Under its terms of reference the Machine Tool and Production Equipment Group will keep itself informed of the requirements of the users of instrumentation and control equipment in this field. They will make known the activities of the group members to those concerned and they will establish liaison with users particularly through appropriate trade associations and research establishments.

A Steering Committee has been set up to give detailed consideration to action to be taken by the group and to make initial contacts with relevant organizations.

Newly elected Fellowships of the Royal Society include Professor R. V. Jones, Department of Natural Philosophy in the University of Aberdeen, elected for his work on fundamental physical measurements by novel instruments of exceptional precision and for application of scientific methods to defence; and Professor T. Kilburn, Department of Computer Engineering in the University of Manchester, elected for his contributions to the development, design and engineering of electronic digital computers and high speed computer components.

The Council of the Institution of Electrical Engineers has decided to disband its Committee on Radio Interference.

Constituted originally as the Committee on Electrical Interference with Broadcasting in 1933, the Committee was reconstituted with its present title in April 1945, and until the early 1950's was concerned both in the formulation of basic policy and in the detailed assess-ment of measures and limits for the suppression of interference. But these two roles are now adequately covered by the Advisory Committees on interference suppression appointed by the Postmaster-General, as a result of the Wireless Telegraphy Act, 1949 (for which the I.E.E. provides the secretariat), and by the International Special Committee on Radio Interference (C.I.S.P.R.), which deals with international aspects of interference suppression.

The BBC is to bring into service five more major transmitting stations for operation on BBC-2 television by the end of this year.

Details are as follows:

	Opening date (1965)	*Channel*
Wenvoe	12 September	51
Winter Hill	17 October	62
Emley Moor	17 October	51
Rowridge	14 November	24
Black Hill	12 December	46

In addition, a permanent station at Sutton Coldfield, replacing the present temporary station, will begin operating on 4 October using Channel 40.

The British Standards Institution has recently published B.S.3860 covering monophonic and stereophonic audio-frequency amplifiers—suitable for domestic, public address and similar systems. Methods are given for measuring and expressing the performance of these amplifiers, operating from d.c. or a.c. mains supply, from a primary or secondary battery or from a combination of these.

The Standard covers equipment employing valves, transistors or a combination of both for sound amplification, but excludes the signal generating source for instance, a gramophone pick-up, microphone or tape deck—and loudspeakers. Separate methods are included where appropriate for measurements on pre-amplifiers, power amplifiers, and integrated amplifiers in which the pre-amplifier and power amplifier are constructed as one unit.

Copies of B.S.3860 may be obtained from the BSI Sales Branch, 2 Park Street, London, W.1, price 6s. each. (Postage will be charged extra to non-subscribers.)

Standard Telephones and Cables Ltd has received orders worth about £300 000 from the General Post Office for a new high-capacity microwave telephone and television link between Bristol, Plymouth and the communication satellite station at Goonhilly Down in Cornwall.

Installation along the 182-mile route will be speeded up to meet the expected completion date of Goonhilly's second satellite aerial in the spring of 1966. The system will be the first in the U.K. to operate in the 'upper 6 000Mc/s' frequency band and will employ STC's latest transistorized equipment.

The new system will link up with the £200 000 microwave system that STC is now installing between Bristol and the 600ft Post Office tower in the centre of London. STC is also installing new permanent microwave telephone/television links between London and France.

Burndept Electronics Ltd of Erith, Kent, has received an order valued at £10 000 from Bristow Helicopters Ltd for the supply of 117 miniature radio beacons and eight floating beacons.

Bristow Helicopters, an air charter associate of British United Airways, has contracted to fly oil drill crews to the drilling platforms of ten oil companies in the North Sea.

The personal beacons ordered are the Sarbe 301 worn by R.A.F. aircrew to identify their position in a crash at sea or forced landing. The transmitter measures 5½in by 2½in by 1in and weighs 36oz with the battery. It has a steel strip aerial which springs erect when a small toggle is pulled; range of the S.O.S. signal sent out is about 80 miles to an aircraft at 10 000ft.

The floating beacons—the Sarbe 311—are larger and operate attached to a liferaft by means of a nylon cord. The aerial springs erect and starts operating automatically as the cord unwinds from the floating beacon. Designed for civil airlines, to identify the position of crash survivors at sea, the 311 signal is picked up 100 miles away by an aircraft at 10 000ft, 150 miles away at 20 000ft, and 200 miles away at 30 000ft.

Westinghouse Brake and Signal Co. Ltd Automation Division is to supply a data logging installation for a power station operated by the Steel Company of Wales at Port Talbot.

The equipment, which is transistorized, will be used to record the temperature of 47 bearings, the flow rate of nine condensate and steam flows, and three vacuums, associated with five turbo-alternators.

The hourly log of information will be recorded on preprinted sheets by an I.B.M. 'golf ball' typewriter, and out-of-limit alarms will be logged by an Addo strip printer.

The point values will be constantly monitored for deviation from limits which will be set on a pin board, individual settings being provided for each point.

Provision is to be made for future extension to an additional four machines.

An oceanographic buoy has been developed by EMI-Cossor Electronics Ltd, a Canadian company in the EMI Group. Oceanographic surveys require a large number of underwater observations to be made. Up to the present, these have been done by lowering suitable instruments into the water from ships, but this is an expensive and slow method.

This new device is a vehicle for gathering underwater data for recording or relaying to a monitor ship, aircraft or shore station. It consists of a sub-surface float containing the measuring equipment, which is moored to the bottom by a taut wire. A slack rope carrying the electric cables links this float with a surface buoy containing the radio and recording equipment.

Buoys can remain on station for many months and one ship could monitor a large number of buoys, so effecting a considerable saving in expense and time.

A symposium on automatic control in electricity supply, arranged by the North Western Centre of the Institution of Electrical Engineers under the aegis of the United Kingdom Automation Council, will be held at the Renold Building, Manchester College of Science and Technology, Manchester, from 29 to 31 March 1966.

The aim of the symposium is to present and discuss the most recent experience in operation and practical design aspects of automatic control in the supply of electric power and to express views and expectations on possible future trends. It is intended to hold an exhibition of suitable equipments and to arrange organized visits to power station plant and to manufacturers' works.

Over 30 offers of contributions have so far been received, but material can still be considered for inclusion in the programme. Those wishing to offer material are asked to contact the Honorary Secretary of the Symposium Organizing Committee without delay.

Further details and registration forms will be available in due course from the Honorary Secretary of the Symposium Organizing Committee, Dr. R. Feinberg, Transformer Division, Ferranti Ltd, Hollinwood Avenue, Chadderton, Lancashire.

Electronic Control Systems for Industry is the title of a symposium to be held at the University of Durham from 8 to 10 September 1965. The object of the meeting is to bring to the attention of industry modern electronic control techniques and ways of increasing productivity and reducing costs, while at the same time improving quality and reliability. The Northern Regional Office of the Ministry of Technology (formerly D.S.I.R.) is co-operating with the North Eastern Section of the Institution of Electronic and Radio Engineers in organizing the three-day symposium. Three fields will be covered: computer analysis of information; advanced telecommunication techniques; use of automotive systems in production.

The papers and discussions will concentrate on the practical application of these techniques.

Further details are available from Ministry of Technology, Northern

Regional Office, Wellbar House, Gallowgate, Newcastle upon Tyne 1.

Cossor Electronics Ltd has received an order from the Ministry of Aviation for the supply of an RBDE-5 16in radar bright display. The equipment is already installed in Aerodrome Control at R.A.E. Farnborough where successful operational trials over the past five months have taken place. Scan conversion techniques are employed, incorporating a 945-line television raster and radar information from the existing airfield radar is displayed in p.p.i. form. The picture thus presented of traffic in the circuit, approaching or departing, can easily be viewed without the use of a viewing hood in all lighting conditions encountered. The equipment is readily available from the range of radar bright display equipment designed by the Raytheon Company for the Federal Aviation Agency. Both 16in vertical and 22in horizontal displays are available and both types can be incorporated into complex systems with full alphanumeric marking capability.

University College, London, following a promise by Mullard Ltd to contribute £65 000 towards space science research, is now to establish an outstation of the Physics Department specifically devoted to the subject. They have purchased Holmbury House, near Dorking, Surrey, which will be known as the Mullard Space Science Laboratory, and will be headed by Professor Boyd. The other group in the college engaged in space research, led by Dr. G. V. Groves, will remain in London.

Scientists at the Mullard Space Science Laboratory will be engaged both on the planning and design of experiments using instruments to be flown in satellites or rockets, and on the subsequent analysis of information received from these flights, together with related laboratory studies. The launchings will be under U.K., American or European auspices.

The main investigations with which they will be concerned are in the fields of ultra-violet and X-ray astronomy both of the sun and of other celestial objects and the study of the ionosphere and the interplanetary medium. They have already made considerable contributions to knowledge in these areas, particularly in the development and use of new and specialized techniques. While the primary aim of the experiments is to increase the fundamental knowledge of the world around us, the findings are expected to have a bearing on such widely different spheres as satellite communications and space travel.

Initially there will be a staff of about 50 at Holmbury House, a large Victorian mansion. It is hoped that the laboratory will be fully operative by the beginning of the next academic year (starting October 1965), and it is anticipated that the group will grow to about 100 people in five to seven years' time. A number of post-graduate research students, scientific and technical staff will be resident.

British Aircraft Corporation announces that an Industrial Products Group has been established within the Stevenage Works of its Guided Weapons Division.

For some time the Stevenage Works has been applying experience and knowledge, gained in missile technology to the needs of industry in general, and the formation of the Industrial Products Group is a further step in the expansion of these activities.

The Industrial Products Group is already engaged in the design and manufacture of automatic test equipment; computer and data handling equipment; servo control equipment for industry; industrial instrumentation. Many other projects are in hand, including measuring equipment using infra-red techniques, laser applications, industrial servo-systems and data handling equipment.

A new type of telephone system came into public service recently which will enable more telephone calls to be carried by existing Post Office cables.

Using pulse code modulation (p.c.m.) this system is now in operation between Reading and Marlow for user trials, following extensive engineering tests by the GPO.

P.C.M. is a development of major commercial importance since it enables the message carrying capacity of existing telephone systems to be substantially increased without the cost, disruption and delay of laying new cables. The system enables 24 conversations to take place simultaneously over wires originally designed to carry only a single circuit. The technique is expected to find extensive application between busy city exchanges.

P.C.M. works on a time sharing basis. In the present installation, which was supplied by Automatic Telephone & Electric Co. Ltd—a member of the Plessey group of companies—the speech content of each of 24 separate channels is sampled, in turn, 8 000 times every second. As each sample is taken, it is converted into an electronic code consisting of a short burst of pulses.

The transmission path to the distant exchange is allocated to each of the 24 channels in turn. At the receiving terminal the channels are separated and the information contained in each code is used to re-constitute the original speech, which is that connected to subscribers' telephones in the normal manner.

This system has been developed at the Taplow laboratories of British Telecommunications Research Ltd, a subsidiary of Automatic Telephone & Electric Co. Ltd

The Companhia Portuguesa Radio Marconi (C.P.R.M.) has ordered over £75 000 worth of Marconi self-tuning (MST) transmitting equipment. This equipment, which will be delivered early this summer, is part of the new h.f. point-to-point communications system introduced by The Marconi Company last year.

This order, which brings the total sales of MST transmitting equipment to over £1 700 000 since its introduction last summer, covers the supply of four 7½kW MST transmitters, and a number of transistorized synthesizer drive assemblies and wideband aerial matching transformers.

These transmitting equipments, which will compatibly fit into any existing h.f. system, will supplement existing Marconi communications equipment at C.P.R.M.'s Alfragide station, just north of Lisbon.

General Precision Systems Ltd is currently building a colour visual simulation system for Australia's Qantas Empire Airways.

As with other similar systems recently ordered from GPS by major airlines in the United Kingdom and the U.S.A., the equipment is based upon the use of a Marconi colour camera viewing the surface of a 2 000 : 1 three-dimensional model constructed on an endless band, the resulting picture being displayed on to a screen in front of the simulator pilot's windshield by the latest Marconi three-tube projector.

A GPS designed optical probe on the camera gives a picture of high definition and allows complete freedom to aircraft manoeuvres in the yawing and rolling planes.

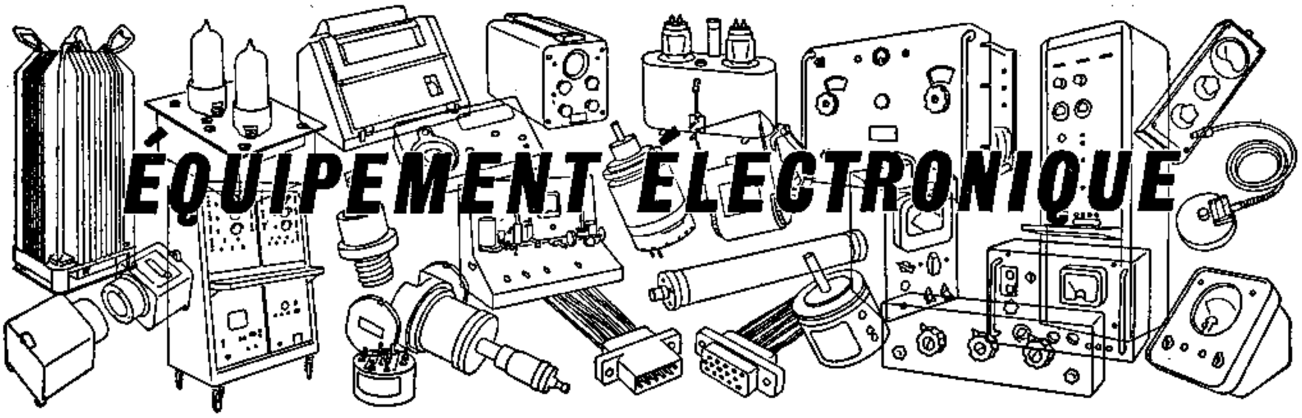
The system will be installed early next year at the Qantas Training Unit, Sydney, and will be switchable for operation with their Link Boeing 707-138B analogue and Boeing 707-338C digital simulators.

Components and Materials used in Electronic Engineering is the subject of the conference which has been arranged jointly by the Institution of Electrical Engineers Electronics Division, the Institution of Electronic and Radio Engineers and the United Kingdom and Eire Section of the Institute of Electrical and Electronics Engineers, and will be concurrent with the 1965 Radio and Electronic Components Show at Olympia.

Among the 44 papers to be read at the conference, to be held at the I.E.E. 17 to 20 May 1965, are contributions from Canada, Poland, Switzerland and the United States. Over 400 people are expected to attend.

Subjects to be discussed are recent developments in active and passive components, including integrated circuits, and in the materials of which they are made.

ÉQUIPEMENT ÉLECTRONIQUE



Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

Traduction des pages 340 à 345

ANALYSEUR D'OXYGÈNE

Servomex Controls Ltd, Crowborough, Sussex

(*Illustration à la page 340*)

La société Servomex Controls Ltd a introduit une nouvelle version de son analyseur d'oxygène portatif: le type DCL.101 Mk 11. Cet analyseur est semblage au modèle précédent car il utilise la cellule de mesure à sensibilité paramagnétique conçue par la Distillers Company, ayant une suspension au platine et utilisant la réaction électromagnétique. La lecture s'obtient par réglage manuel d'un cadran étalonné de 0 à 100% d'oxygène et donnant une lecture allant jusqu'à 0,1% d'oxygène. Le courant nécessaire s'obtient normalement de batteries sèches.

Le modèle Mk 11 est doté d'un certificat de sécurité intrinsèque et il peut être utilisé dans des atmosphères contenant des gaz d'éthylène, de pentane et d'hydrogène, tels que définis dans la norme britannique B.S.1259. L'analyseur peut donc être utilisé sans danger dans de nombreux emplacements présentant des conditions dangereuses et il est particulièrement utile dans les raffineries et les installations chimiques.

Les perfectionnements apportés au nouvel analyseur comportent un nouvel aimant entièrement insensible aux effets de champ extérieurs. Il n'y a donc plus aucune restriction à la position d'emploi et cet avantage, allié à l'absence relative de tout effet de basculement (ne dépassant 0,005% d'oxygène par 1° d'inclinaison), rend l'appareil DCL.101 Mk 11 particulièrement indiqué pour l'emploi portatif.

Les accessoires d'échantillonnage ont été ajoutés comme pièce standard. Chaque analyseur est muni d'un aspirateur à main, d'un tube de séchage et d'un filtre à disque de verre d'un dessin inédit, comprenant des valves très simples pouvant être facilement renouvelées par l'utilisateur.

EE 81 751 *pour plus amples renseignement*

DIVISEUR DE 100 MHz

Advance Electronics Ltd, Roebuck Road, Hainault, Ilford, Essex

(*Illustration à la page 340*)

Pour répondre à la demande d'une plus grande souplesse dans les gammes supérieures de mesure de fréquence, la société Advance Electronics Ltd a ajouté un diviseur de 100 MHz, le type TCD100, à sa liste croissante de produits de ce genre.

Le TCD100 a été étudié pour étendre la gamme de tout compteur électronique vendu dans le commerce de 1 MHz à 100 MHz. La mesure s'effectue par un facteur de division de 100 ou 20, contrôlé au moyen d'un commutateur du panneau avant. Les paramètres d'entrée et de sortie sont suffisamment souples pour recevoir une gamme étendue de sources d'entrée. L'instrument est transistorié et utilise des diodes tunnel.

Sa sensibilité d'entrée est de 50 mV dans 50 Ω et des impulsions de sortie à 3 V de crête sont fournies de 10 kHz à 1 MHz ($\div$ 100) et de 50 kHz à 5 MHz ($\div$ 20). La précision du compteur de base n'est pas affectée par le diviseur.

EE 81 752 *pour plus amples renseignements*

LIGNE DE TRANSMISSION ARTIFICIELLE

Telefon Fabrik Automatic A/S, 7 Amaliegade, Copenhague, Danemark

(*Illustration à la page 340*)

La ligne de transmission artificielle type EG4 a des caractéristiques générales qui simulent très fidèlement les lignes téléphoniques ordinaires.

Cette ligne artificielle a été principalement conçue pour servir de dispositif de laboratoire simulant les effets de longueur de ligne sur les installations téléphoniques tels que les postes téléphoniques, les systèmes de transmission de données, etc.

La ligne artificielle se compose de sections RC qui correspondent chacune à une longueur unitaire de ligne de 0,5 km. D'autres longueurs unitaires peuvent être prévues sur demande. En reliant un certain nombre de sections en tandem, n'importe quelle longueur de ligne de 0,5 km peut être simulée par plots de 0,5 km.

Ce type de ligne artificielle comprend en fait deux lignes d'un maximum de 27 sections donnant un maximum de 13,5 km de ligne. Les deux lignes peuvent évidemment être reliées en série donnant ainsi une longueur de ligne totale de 27 km.

La connexion en tandem des sections s'effectue par une série de boutons-poussoirs. Cinq boutons sont utilisés pour insérer des longueurs de ligne de 0,5, 1, 2, 4 et 6 km respectivement. La longueur des sections insérées qui en résulte est donc la somme des valeurs correspondant aux boutons-poussoirs activés.

La ligne artificielle est montée dans un cabinet métallique. Le panneau frontal est divisé en deux moitiés symétriques, une pour chaque ligne. En plus des boutons-poussoirs, les bornes de lignes sont montées sur le panneau frontal.

La ligne artificielle est équipée de postes de liaison pour les connexions d'entrée, de sortie et de masse. Les composants sont montés sur cartes de circuit imprimé, chaque carte comprenant 9 sections. Les cartes imprimées sont montées dans le coffret au moyen de multiconnecteurs et sont facilement enlevées pour pouvoir être éventuellement réparées ou contrôlées. Les cartes sont toutes identiques et par conséquent interchangeables.

EE 81 753 *pour plus amples renseignements*

ELECTRONIC ENGINEERING occupera le stand 58 dans le Grand Hall à l'Olympia de Londres durant le Salon de R.E.C.M.F. du 18 à 21 mai et les visiteurs y seront les bienvenus.

ELÉMENTS D'OSCILLOSCOPE

Solartron Electronic Group Ltd, Farnborough, Hampshire

(*Illustration à la page 341*)

Deux nouveaux éléments à fiche ont été réalisés pour les oscilloscopes Solartron types CD1220 et CD1212 (CT484).

L'élément à fiches différentiel à gain élevé CX1258 se caractérise par une sensibilité de courant continu élevée (100 µV/cm), un rejet en phase élevé (80 dB — 10 000:1) et une très faible dérive. En outre, une largeur de bande maxima de 200 kHz permet d'utiliser l'amplificateur pour une gamme étendue d'applications universelles.

L'étage d'entrée à gain élevé a été étudié pour maintenir un équilibre précis pour le rejet en phase maximum. Pour les conditions de couplage au courant alternatif, la réaction à courant continu réduit une dérive de courant continu fondamentalement faible à des proportions négligeables. Des valves sûres et de qualité spéciale sont utilisées dans les positions critiques.

Spécification de l'élément CX1258:

Sensibilité:
: 100 µV/cm à 2 V/cm dans 14 gammes étalonnés.

Largeur de bande
: Du courant continu à 50 kHz à 100 µV/cm
: Du courant continu à 100 kHz à 200 µV/cm
: Du courant continu à 200 kHz à 500 µV/cm.

Rejet en phase
: Non inférieur à 80 dB du courant continu à 50 kHz.

Tension en phase maxima
: ±25 V de crête de 100 µV/cm à 20 mV/cm
: ±250 V de crête de 20 mV/cm à 2 V/cm.

Précision d'étalonnage
: 3 pour cent.

L'élément CX1259 a une performance semblable à celle du CX1256, mais comprend un amplificateur à transistors accouplé au courant continu x 10 fournissant une sensibilité maxima de 5 mV/cm à une largeur de bande allant du courant continu à 24 MHz. Il conserve la performance de base du courant continu à 40 MHz à 50 mV/cm. La réaction composée permet d'obtenir une excellente réponse d'impulsion sur toutes les gammes.

Le décalage "Y", appliqué comme potentiel d'équilibrage, assure un affichage non déformé des signaux à courant continu.

Spécification de l'élément CX1259:

Sensibilite et largeur de bande
: x 10 : 5 mV/cm; du courant continu à 24 MHz
: x 1 : 50 mV/cm; du courant continu à 40 MHz

Gammes
: Etalonnées en 1-2-5 plots avec réglage précis

Précision d'étalonnage
: 3 pour cent.

L'élément différentiel à gain élevé CX1258 se trouve à gauche sur notre gravure et l'élément à large bande CX1259 se trouve à droite.

EE 81 754 *pour plus amples renseignements*

INSTRUMENTS D'ENSEIGNEMENT

White Electrical Instrument Co. Ltd.
10 Amwell Street, Rosebery Avenue, London, E.C.1

(*Illustration à la page 341*)

Bien que convenant à l'usage général de laboratoire, ces instruments ont été spécifiquement conçus pour l'enseignement moderne de la physique. En effet, ces instruments, type RM/AS, sont de construction robuste et spécialement prévus pour résister aux rigueurs de l'utilisation dans les écoles. Ils comprennent, cependant, un mouvement d'une précision et de la haute qualité nécessaires à ce nouveau genre de travail.

Les cadrans de 6,35 cm de ces instrument comportent des échelles offrant une lecture claire et facilement lisible, montées dans des coffrets de banc d'essai inclinés en matière plastique gris pastel à deux tons. Ces coffrets sont facilement nettoyables et ils sont munis sur le haut de bornes à vis à douille de 4 mm.

Les mouvements sont du types à ressort le plus récent et comprennent des pivots en acier poli ainsi que des paliers à saphir synthétique. Le type RMI à fer mobile est à atténuation pneumatique convenant tout aussi bien pour le courant alternatif que pour le courant continu mais il n'est livrable qu'avec un zéro latéral. Le type RMC à cadre mobile ne convient que pour le courant continu mais peut être obtenu soit avec zéro latéral soit avec zéro central. Il est apériodique et à atténuation électrodynamique. Ces deux types d'instruments de mesure sont prévus dans une gamme étendue d'étalonnages de tension et de courant.

EE 81 755 *pour plus amples renseignements*

ACCESSOIRE À PHASE VARIABLE

Servomex Controls Ltd, Crowborough, Sussex

(*Illustration à la page 341*)

La société Servomex Controls Ltd a réalisé un accessoire à phase variable pouvant être utilisé avec le générateur de formes d'ondes à basse fréquence qu'elle vient de mettre au point, à savoir le type LF141. Le but principal du nouvel appareil, type VP.142, est de permettre la mesure de phase précise et rapide dans la gamme de fréquences de 2000 Hz à 1 Hz en 500 sec.

La mesure s'effectue suivant la méthode bien connue dite du chiffre de Lissajous par laquelle une éllipse dégénère en une ligne droite lorsque la phase est correcte. En suivant la procédure expliquée dans la feuille de données, on peut effectuer des mesures même à des fréquences très basses en moins d'un Hz, en utilisant soit un oscilloscope cathodique soit une forme quelconque d'enregistreur à plume. Une commande d'amplitude est prévue de sorte que lorsque le chiffre de Lissajous est incliné à 45°, l'amplitude peut être lue directement de la commande d'amplitude, de même que la phase. Aux rares occasions où il se produit un bruit ou une distorsion excessive dans le signal soumis au contrôle, l'appareil peut être muni d'un filtre passe bande approprié.

La tension de phase variable s'obtient d'un potentiomètre à enroulement circulaire alimenté à 0°, 90°, 180°, et 270°. Pour permettre l'utilisation de la méthode de mesure d'amplitude, un circuit spécial a été prévu qui maintient l'amplitude constante à mesure que tourne la commande de phase.

Cet élément est vissé au fond du LF141 et il reçoit son entrée d'une fiche et d'une douille multivoies. L'équipement combiné, tout en étant utilisable comme source de diverses sortes de signaux de basse fréquence pour la mesure de phase de systèmes de puissance, peut être employé également pour une variété d'autres usages tels que, par exemple, la démonstration de la surtension d'amorçage dans les systèmes à courant alternatif. En supprimant une tension paralysante de la borne de porte, on peut déclencher un train d'ondes sinusoïdales à n'importe quel angle voulu.

Parmi les nombreux types d'ondes pouvant être produits, on compte les ondes carrées à temps de montée variable dans lesquelles la fréquence de répétition et le temps de montée peuvent être contrôlés séparément ainsi que les impulsions carrées sinusoïdales qui produisent des harmoniques à l'intérieur d'une largeur de bande connue avec précision.

Pour les applications médicales ainsi que pour les autres applications un train d'impulsions d'un nombre connu peut être produit. Le compteur peu compter des impulsions carrées à temps de montée rapide à partir de la borne de sortie auxiliaire à ondes carrées, tandis que la borne de sortie principale émet n'importe quelle forme d'onde voulue.

L'accessoire à phase variable Servomex VP142 a été conçu en vue de lui assurer une facilité d'entretien particulière. Ces composants électroniques sont tous sur plaquette de circuit imprimé à charnières, facilitant ainsi leur entretien. La plupart des composants sont choisis de la liste approuvée des services des forces armées et ils sont utilisés suivant les spécifications militaires appropriées.

EE 81 756 *pour plus amples renseignements*

MILLIVOLTMÈTRE A.F.

Elektromesstechnik Wilhelm Franz KG,
80 Kaiserstrasse, 763 Lahr/Schwarzwald, Allemagne

(*Illustration à la page 341*)

Le voltmètre entièrement transistorisé EMT 125 a été spécialement conçu pour la mesure exacte sur les amplificateurs et le matériel basse fréquence ainsi que les installations électro-acoustiques. Sa consommation extrêmement faible permet de le brancher au secteur de manière permanente sans qu'il s'en suive un échauffement appréciable.

Le coffret plat et de lignes modernes permet l'empilage de plusieurs éléments. Les bornes, boutons-poussoirs et commutateurs sont tous disposés sur le panneau frontal, de même que l'instrument de mesure. L'échelle est étalonnée en volts et en décibels.

Le millivoltmètre comporte douze gammes de mesure allant de 1 mV de déviation sur la totalité de l'échelle à 300 V de déviation sur la totalité de l'échelle. La plus petite valeur pouvant être indiquée est de 100 μV. La gamme de fréquence s'étend jusqu'à 200 kHz; un filtre passe-bas est incorporé et peut être inséré par bouton-poussoir. Il limite la plus haute fréquence à laquelle une mesure peut être effectuée à 20 kHz. Des formes d'onde très différentes des ondes sinusoïdales peuvent être mesurées également. Des mesures en valeur de pointe ou en valeur efficace peuvent être faites par bouton-poussoir.

Pour la mesure des valeurs efficaces suivant la norme allemande DIN 45 402 à une fréquence d'impulsion de 200 Hz, l'instrument réalise un cycle de régime de 1:10 sur la totalité de l'échelle et avec une déviation de 1 dB. Les mesures de valeur de pointe sont indiquées pour pouvoir déterminer les tensions extérieures telles que les tensions de bruit permettant le filtrage de fréquence dépassant 20 kHz.

Utilisé comme amplificateur, il donne une amplification maxima de 60 dB dans la gamme de 1 mV; cette amplification tombe de 10 db par gamme. Une tension de 1 V dans 4 kΩ est fournie à la sortie.

Une source de tension de référence stabilisée est incorporée pour pouvoir faciliter un nouvel étalonnage.

EE 81 757 *pour plus amples renseignements*

MONTRES NUMÉRIQUES

Distributeurs: Systems Engineering Services Ltd, 35-39 South Ealing Road, Ealing, London, W.5

(Illustration à la page 342)

Deux nouvelles montres numériques de précision indiquant l'heure sous forme d'affichage décimal, ainsi que sous forme décimale codée binaire, ont été réalisées par la société Electronic Engineering Co. de Californie.

Les nouveaux instruments, à savoir la montre numérique EECO 815 et la montre numérique EECO 816, fournissent tous deux des niveaux logiques parallèles; de plus le 816 fournit la lecture en série. Ils fournissent des codes de temps directs au système de concentration ou d'enregistrement des données, aux calculatrices, aux enregistreurs à diagramme, aux imprimeuses, etc. Les deux modèles sont entièrement constitués de corps solides et ne comportent aucun dispositif mécanique tels que moteurs, relais ou commutateurs à gradins.

Tant le EECO815 que le EECO816 sont synchronisés en fonction d'une fréquence de ligne de puissance ou peuvent être fournis avec une source de fréquence intérieure. En plus des codes de temps, six sorties d'impulsions à onde carrée peuvent être obtenues simultanément. Chaque modèle comprend un indicateur de panne de puissance jusqu'à ce que le bouton de panne de puissance soit réenclenché.

Le EECO815 (code de sortie parallèle seulement) fournit un code de temps précis indiquant les heures, les minutes et les secondes sous forme décimale codée binaire pour l'enregistrement direct dans les systèmes de concentration ou d'enregistrement des données à bande de papier, à bande magnétique ou à impression directe; en variante, il peut être utilisé pour la transmission directe aux calculatrices.

Le modèle EECO816 (à lecture parallèle et à lecture en série) peut être utilisé pour l'usage décrit ci-dessus ainsi que pour l'enregistrement du temps en série sur enregistreurs à diagramme ou sur enregistreurs à bande magnétique. Une sortie porteuse modulée (100 Hz ou 1000 Hz) peut être prévue au choix pour l'enregistrement du code de temps en série sous forme codée en largeur ou sous forme de porteuse modulée en amplitude sur bande magnétique analogique.

EE 81 758 *pour plus amples renseignements*

CALCULATRICE ANALOGIQUE POUR RAPPORT DE TRANSFORMATEUR

F. C. Robinson & Partners Ltd, Davies House, 181 Arthur Road, Wimbledon, London, S.W.19

(Illustration à la page 342)

Spécialement conçue pour répondre aux besoins des colleges et universités, cette calculatrice analogique, basée sur le principe bien connu d'utilisation du rapport des tours de paires de transformateurs de calcul pour représenter des nombres complexes, trouve des applications étendues dans l'analyse de circuits et de réseaux.

La calculatrice comprend:

Trois éléments d'impédance universels jumelés (pour la présentation directe de l'impédance ou de l'admittance à n'importe quel angle de phase);

Deux éléments d'impédance universels spéciaux jumelés (impédance ou admittance, source de tension ou de courant, impédance mutuelle ou de transformateur);

Un oscillateur transistorisé de 400 Hz;

Un élément de mesure de zéro utilisant un transformateur de calcul unique, un amplificateur logarithmique et un enregistreur;

Un bloc d'alimentation.

La précision des réglages constants de circuit et de la mesure est de 0,1%.

EE 81 759 *pour plus amples renseignements*

CONNECTEUR DE CÂBLES COAXIAUX SUBMINIATURE

Sealectro Ltd, Hersham Trading Estate, Walton-on-Thames, Surrey

(Illustration à la page 342)

La société Sealectro Ltd vient d'ajouter à sa gamme de connecteurs un bloc de jonction "T" du type à sertissage de 75 Ω pour câbles coaxiaux subminiature. Le connecteur 5116 utilise des connections de câble du type à sertissage qui sont particulièrement indiquées pour l'assemblage à taux de production élevée et fournissent une terminaison de câble plus résistante que le câble lui-même.

Le connecteur comprend des trous d'espacement pour montage sur panneau ou sur cloison et permet l'entassement d'éléments lorsqu'un centre de jonction de câble est nécessaire. De plus, le bloc de jonction 5116 "T" est à revêtement doré, ce qui lui assure une efficacité électrique maxima lorsqu'il est allié au connecteur Conhex.

EE 81 760 *pour plus amples renseignements*

DÉTECTEUR DE PAILLES À ULTRASONS

Ultrasonoscope Co. (London) Ltd, Sudbourne Road, London, S.W.2

(Illustration à la page 342)

L'Ultrasonoscope Mark 4 a été conçu pour assurer la performance et la souplesse d'emploi nécessaires à la solution des problèmes de contrôle les plus difficiles. Il est prévu pour le contrôle normal en usine ainsi que pour les travaux de recherche en laboratoire. Avec son contrôleur, il peu actionner un système d'enregistrement et de balayage entièrement automatique.

Les applications dans les diverses industries exigent souvent des instruments à caractéristiques différentes. L'amplificateur Mark 4 est donc interchangeable. Un amplificateur spécial peut être réalisé et monté facilement sur le détecteur pour des applications spéciales. L'amplificateur standard a une largeur de bande étendue et une gamme dynamique à chacune des fréquences de 10, 5, 2½, 1¼ et ½ MHz. Il peut résoudre de courtes impulsions et il est linéaire.

La dispersion de la structure d'un matériau peut être supprimée à l'aide d'une commande de rejet à variation continue. L'atténuateur étalonné a une gamme de 0 à 80 dB et une précision de lecture de 1 dB. L'émetteur puissant et l'amplificateur à gain élevé donnent une pénétration grandement accrue dans le fonctionnement à double sonde ou à sonde unique.

On peut obtenir la présentation redressée, non-redressée ou à oscilloscope B. Sur l'oscilloscope B, les échos de défauts éclairent la base de temps et l'image est semblable à celle d'un radiographe.

Le tube cathodique a une face plate de 12,5 cm avec surface luminescente photographique et filtre optique pour assurer un meilleur contraste. La base de temps est linéaire et la vitesse la plus rapide est de 2 μsec de sorte que 0,63 cm d'acier peuvent être affichés sur toute la largeur de l'écran pour la mesure précise de l'épaisseur et de la vitesse. La gamme maxima est de 6 m.

Le délai de la base de temps est à variation continue de 0 à 300 μsec au moyen d'une commande étalonnée permettant l'agrandissement de n'importe quelle section de la trace. La fréquence de répétition des impulsions s'étend de 50 à 1000 Hz et le détecteur Mark 4 est pleinement stabilisé contre les fluctuations de tension secteur.

EE 81 761 *pour plus amples renseignements*

FORET À CIRCUIT IMPRIMÉ

Distributeurs: Alfred Jacoby (Machine Tools) Ltd. 54 Manchester Street, London, W.1

(Illustration à la page 343)

La machine à percer Posalux "Copyfor" que construit la société Frautschi et Monney, Suisse, a été spécialement réalisée pour la fabrication de circuits imprimés, bien quelle puisse également être utilisée pour le perçage de plaques indicatrices en feuilles d'aluminium et de composants analogues.

Le perçage s'effectue au moyen d'un foret hydropneumatique "Unifor" 10, monté verticalement et fonctionnant dans le sens vertical. Le cycle de perçage peut être soit manuel soit semi-automatique. Dans les deux cas, la pile de composants, portant le gabarit de perçage au sommet, est déplacée jusqu'à ce que la broche de localisation à ressort qui se trouve dans le même axe que la tige de perçage trouve l'un des trous du gabarit. Durant le travail automatique, ce procédé permettra de fixer les composants de manière pneumatique et de poursuivre l'avance du foret, le perçage étant presque instantané. Pour les opérations manuelles, le cycle de serrage et de perçage est déclenché à la main.

L'écran de projection ou microscope qui peut être fourni avec la machine, permet de fabriquer les gabarits de perçage sur la machine même en suivant le circuit imprimé original. Cette méthode est également pratique pour les petits lots qui ne justifieraient guère la fabrication de gabarits individuels. L'écran de projection a une image corrigée à agrandissement et à intensité lumineuse élevés.

La machine est propre à la production à grande vitesse. Cet avantage est dû en partie au modèle de la broche indicatrice qui est soumise à une pression réglable en fonction des conditions d'utilisation. Son mouvement est extrêmement sensible, éliminant ainsi les riblons dus à la mise en marche accidentelle du pivot de perçage. La tête peut être inclinée vers l'arrière, permettant ainsi de changer facilement le foret et de nettoyer l'intérieur de la machine. Un élément d'aspiration peut être fourni comme supplément facultatif et un anneau à ventilateur, faisant partie du mandrin du foret, assure le dégagement des copeaux dans la gouttière appropriée.

La machine est utilisée à des vitesses de pivotage élevées afin de réduire la durée du perçage et d'empêcher la formation d'ébarbures. Le diamètre maximum du foret est de 3 mm et la machine peut recevoir des plaquettes d'un largeur maxima de 420 mm.

EE 81 762 *pour plus amples renseignements*

INDICATEUR DE TEMPS ÉCOULÉ

Taylor Electrical Instruments, Montrose Avenue, Slough, Buckinghamshire

(Illustration à la page 343)

La société Taylor Electrical Instruments (M.I. Group) vient de réaliser un nouvel indicateur de temps écoulé qui s'harmonise avec la gamme des indicateurs Vista et a la même apparence frontale que le modèle Vista 40.

Il existe des modèles pour alimentation secteur contrôlé de 110-125 ou 200-250 V, 50 ou 60 Hz, pouvant enregistrer les heures et les dixièmes d'heure jusqu'à 99999,9 heures, après quoi le compteur se met automatiquement à **zéro.**

Un moteur synchrone auto-démarreur assure l'exactitude de l'indication même pendant les périodes de courte durée.

EE 81 763 *pour plus amples renseignements*

GÉNÉRATEUR D'IMPULSIONS

Hewlett Packard, Dallas Road, Bedford

(Illustration à la page 343)

La société Hewlett-Packard GmbH, c'est à dire la filiale allemande de recherche et de fabrication de la société Hewlett-Packard a réalisé un générateur d'impulsions compact et à bas prix. Le générateur d'impulsions Hewlett-Packard, modèle 8000 A, a été conçu pour effectuer des mesures de réponse sur les instruments à large bande, les circuits et les semi-conducteurs.

La sortie de 10 V (dans 50 Ω) du générateur consiste essentiellement en une sortie à sommet plat avec un taux de répétition de 100 kHz et un temps de montée inférieur à 20 sec. Toutes les variations de pointe d'impulsion, telles que la surmodulation ou l'appel sont maintenues à un niveau inférieur à 2 %; le sommet plat est maintenu pendant au moins 100 sec. Le temps de chute est inférieur à 20 sec. Les diodes de rétablissement de plots réalisées par Hewlett-Packard sont utilisées comme éléments de formation d'impulsions dans le circuit de sortie.

L'instrument se caractérise par sa sortie atténuée par plots (de 0,1 à 10 V en six plots), par sa polarité de sortie négative ou positive ainsi que par sa sortie de déclenchement avancée. L'impulsion de déclenchement rend superflue toute ligne de retard lorsque l'appareil est utilisé avec des oscilloscopes d'échantillonnage ou d'autres oscilloscopes à action rapide. Sa progression est de 200 sec, son temps de montée est inférieur à 6 sec, sa largeur est de 20 sec et son amplitude de 0,5 V dans 50 Ω.

EE 81 764 *pour plus amples renseignements*

RELAIS ET MINUTERIES CONSTITUÉS DE CORPS SOLIDES

Keyswitch Relays Ltd, 120-132 Cricklewood Lane, London, N.W.2

(Illustration à la page 344)

La société Keyswitch Relays Ltd a réalisé la gamme P33VT de minuteries et de relais transistorisés à fiches conçus pour la commande de faibles sorties et livrables sous forme de modules normaux. La nouvelle gamme qui dérive de la série des relais P33, peut être utilisée en liaison avec cette dernière ainsi qu'avec l'élément à plaquette de composants type 304, dont on vient d'annoncer l'apparition.

Les applications de la gamme P33VT comprennent le contrôle thermique d'atmosphère et de température, les dispositifs de détection, l'automatisation des machines-outils, la perception de tension et de courant, le contrôle de remplissage de conteneurs, les circuits d'alarme et une variété étendue d'opérations de traitement des données.

La gamme comprend quatre types standard, dont chacun permet un grand nombre de variations de dessin selon les besoins spécifiques. Le modèle P33VT1 introduit un faible signal de l'ordre de 10 μA c.a./c.c. dans une impédance de 10 kΩ, contrôlant un circuit qui branche le relais. L'alimentation en tension est normalement de 24 V c.a./c.c., les contacts étant soit du type à permutation de puissance (5 A, 230 V c.a. résistifs) ou du type à permutation légère allant jusqu'à 1 A.

Le modèle de relais P33VT 3 est actionné soit en fermant les contacts extérieurs qui font passer un courant à faible entrée de 500 μA ou un courant inférieur, soit en variant la résistance extérieure. Pour les applications d'indicateurs, les fluctuations de contact sont compensées.

Le modèle P33VT 7 permet la commutation directe du relais à l'aide de cellules photoélectriques à niveau élevé, ainsi que la commutation de source lumineuse à niveau réduit de 15 μA, à l'aide d'un amplificateur constitué de corps solides à deux étages, ou le contrôle par une source lumineuse extérieure à l'aide d'une cellule photoélectrique incorporée.

Le modèle P33VT 8 comprend des relais de minutage stabilisé à action rapide, et le retard s'étend de 0 à 10 sec, de 5 à 30 sec, de 10 à 60 sec, de 15 à 120 sec et jusqu'à 5 min.

La tension de régime de l'assemblage à fiche et douille de la gamme P33VT est de 750 V c.a. ou c.c. et la résistance à l'isolement est supérieure à 100 MΩ à 500 V c.c.

EE 81 765 *pour plus amples renseignements*

TRANSDUCTEURS DE PRESSION

Intersonde Ltd, The Forum, High Street, Edgware, Middlesex

(Illustration à la page 344)

La société Intersonde Ltd, vient d'introduire la série B.M. à sa gamme de transducteurs de pression extensométriques. Cette nouvelle série est spécifiquement prévue pour les faibles pressions: 7 kg/cm² à 53 kg/cm².

Les transducteurs sont basés sur un élément réagissant à la pression sous forme d'un tube de cuivre au béryllium à extrémités fermées auxquelles sont fixés quatre extensomètres reliés en un montage en pont à quatre bras. Les corps de pression liquides ou gazeux admis à l'intérieur du tube produisent des changements dimensionnels très réduits qui sont instantanément convertis en changements de résistance par les deux bras actifs du pont. Le signal de déséquilibre du pont est proportionnel à la pression appliquée.

Les transducteurs ont une résistance de sortie de 350 Ω et produisent une sortie de 30 mV à la pression nominale totale avec une excitation de 20 V. L'erreur combinée de non-linéarité et d'hystérésis est de ± 1,0% de la gamme totale et les limites de température de fonctionnement s'étendent de —40° C à +120° C.

EE 81 766 *pour plus amples renseignements*

CAMÉRA MINIATURE DE TÉLÉVISION

EMI Electronics Ltd, Hayes, Middlesex

(Illustration à la page 344)

Une nouvelle caméra de télévision entièrement transistorisée à circuit fermé ne mesurant que 11,25 cm de long et pesant environ 1,5 kg a été réalisée par la société EMI Electronics Ltd.

En dépit de ses faibles dimensions, cette caméra est extrêmement robuste et produit des images à très haute résolution. Elle est donc particulièrement utile dans les lieux inaccessibles où elle risque d'être un peu malmenée et où l'éclairage est généralement faible. Elle peut être utilisée sur 405, 525 et 625 lignes et on peut la faire passer de l'une à l'autre de ces normes sur simple pression d'un bouton.

L'équipement de tête de la caméra est logé dans deux cylindres en acier inoxydable scellés, mesurant tous deux 11,25 cm de long et 3,9 cm de diamètre. L'élément de tête à lentille qui pèse environ 0,80 kg contient un tube de caméra vidicon EMI de 1,25 cm. Il peut être aisément tenu dans une main de manière à laisser l'autre libre pour la focalisation par rotation de la lentille. En outre il peut être monté en des lieux où l'espace est fort restreint.

Dans l'autre cylindre se trouve la tête d'amplificateur qui peut être placée à une distance de 30,48 mètres de la tête de lentille et reliée à cette dernière par un câble.

Le contrôleur de la caméra et les autres éléments comprenant le canal de la caméra peuvent être placés jusqu'à une distance de 305 mètres.

EE 81 767 *pour plus amples renseignements*

EXTENSOMÈTRE ENREGISTREUR

C.N.S. Instruments Ltd, 61 Holmes Road, London, N.W.5

(Illustration à la page 344)

La société C.N.S. Instruments Ltd, annonce que son extensomètre enregistreur peut être livré maintenant avec commutation de gamme automatique. Lorsqu'il est muni de ce dispositif de commutation, l'instrument est fourni sous forme de montage sur bâti de 47,5 cm avec sortie de 0-10 mV (0-50 mV sur commande spéciale) et il peut alors être utilisé avec un enregistreur potentiométrique tel que le Honeywell-Brown Electronik.

Six gammes sont normalement prévues: six et douze microns 0,025, 025, 0,625 et 2,5 m. On peut ainsi mesurer facilement des augmentations de 0,05 microns sur une bande de 250 m.

D'autres gammes peuvent être prévues sur commande spéciale.

Le dispositif de commutation de gamme est particulièrement utile pour les essais de déformation et autres contrôles mécaniques où le taux d'extension n'est pas en fonction linéaire du temps. On peut donc obtenir des enregistrements permanents de nuit et pendant les week-ends en l'absence de personnel qualifié. On peut ainsi utiliser un réglage à haute sensibilité dès le début du contrôle. A mesure que celui-ci s'effectue et qu'on atteint la déviation sur la totalité de l'échelle dans chaque gamme, l'instrument est automatiquement commuté à la gamme inférieure suivante.

Un soin particulier a été apporté à la réduction des effets de température dans les transducteurs qui sont fournis individuellement ou en paires de sommation. Les bobinages des cadres mobiles ou statiques sont sur supports de quartz, propres à l'utilisation à des températures atteignant 45° C ou à des températures supérieures sur commande spéciale, avec un maximum absolu de 100° C.

EE 81 768 *pour plus amples renseignements*

MILLIVOLTMÈTRE

Standard Telephones & Cables Ltd, Transmission Testing Apparatus Division, Corporation Road, Newport, Monmouthshire

(Illustration à la page 344)

Un nouveau millivoltmètre à impédance élevée et d'une grande souplesse d'emploi vient d'être mis au point par la Standard Telephones & Cables Ltd pour l'usage général ainsi que pour les télécommunications.

Le nouveau millivoltmètre qui porte le numéro de référence 74251, fonctionne sur toute la gamme de fréquences s'étendant de 20 Hz à 20 MHz et permet d'effectuer des mesures en dix gammes allant de 0 à 1 mV à 0 à 30 V. La sonde active dont il est muni donne une impédance d'entrée supérieure à 100 kΩ en parallèle avec 40 pF.

Deux modèles sont prévus, à savoir le modèle A pour fonctionnement sur batteries autonomes et le modèle B pour fonctionnement soit sur batterie autonome soit sur courant de secteur alternatif.

Le millivoltmètre 74251 est un instrument transistorisé portatif pour les applications générales ainsi que pour la gamme de fréquence de bandes de base des systèmes de liaison par faisceaux hertziens microondes. Il peut être utilisé également comme détecteur de pont, et son oscillateur d'étalonnage peut fournir un signal d'environ 100 kHz à des circuits extérieurs. L'amplificateur intérieur a une sortie permettant de l'utiliser dans les circuits extérieurs.

Le millivoltmètre 74251 comporte deux échelles de tensions ainsi qu'une échelle de décibels pour les circuits de 75 Ω. On peut également vérifier l'état des batteries sur le millivoltmètre.

EE 81 769 *pour plus amples renseignements*

GÉNÉRATEUR TRÈS BASSE FRÉQUENCE

Distributeurs: Claude Lyons Ltd, 76 Old Hall Street, Liverpool, 3

(Illustration à la page 345)

Le générateur très basse fréquence, type GB64, comprend cinq gammes de 10:1, étalonnées en fréquences et en périodes, couvrant la gamme de 0,005 Hz à 500 Hz. Un choix de sorties triangulaires carrées ou sinusoïdales est prévu avec sortie équilibrée variable jusqu'à 40 V de crête à crête dans 10 kΩ, à partir d'une impédance intérieure de 100 Ω. La sortie d'ondes sinusoïdales a une distorsion maxima de 2%, la sortie d'ondes carrées a un temps de montée de 25 µsec et la sortie triangulaire a une erreur de pente maxima de 2%. Le niveau de bruit aléatoire est de 20 mV au maximum. Une stabilité de fréquence de 0,5% et une stabilité de niveau de 0,2 dB pour une variation de secteur de ±10 % sont assurées par des alimentations stabilisées.

Une sortie d'impulsions à synchronisation (25 V de pointe, 50 µsec) comprenant des impulsions négatives et positives est prévue pour le déclenchement oscilloscopique.

EE 81 770 *pour plus amples renseignements*

VIDICONS À HAUTE RÉSOLUTION

The English Electric Valve Co. Ltd, Chelmsford, Essex

(Illustration à la page 345)

La gamme des vidicons de 2,5 cm à haute résolution et à maille séparée de la société English Electric Valve Co. Ltd a été étendue par l'introduction de deux nouveaux types à réponse de pointe élevée dans la zone "bleue" du spectre.

Ces tubes, modèles 8625 (P846) et 8626 (P847), comprennent des surfaces photoélectriques entièrement nouvelles d'une sensibilité uniforme et très élevée, a très faible retard et à caractéristiques de blocage à long terme réduites, à réponse panchromatique exacte avec éclairage au tungstène et à haute résolution aux courants à signaux élevés.

L'excellente uniformité du signal a été réalisée grâce à l'emploi de techniques de fabrication perfectionnées permettant de "préfabriquer" les surfaces photoélectriques afin d'assurer un dépôt égal sur l'ensemble de ces surfaces. Les niveaux d'éclairage des studios peuvent être considérablement réduits grâce aux temps de retard très courts permis par les nouvelles surfaces photoélectriques.

Un avantage supplémentaire d'importance est offert par la reproduction extrêmement bonne des couleurs de la chair, cet avantage étant dû à la sensibilité spectrale parfaitement équilibrée de ces vidicons.

L'introduction des nouvelles surfaces photoélectriques, alliées à la construction séparée des mailles, rendent les vidicons 8625 (P846) et 8626 (P847) particulièrement indiqués pour les prises de vues en direct ou de télécinéma en couleurs et monochrome. La sensibilité bleue supplémentaire constitue un perfectionnement utile dans le rapport signal/bruit des caméras de télévision en couleurs et fournit en outre une sortie de signaux élevée dans les applications où la lumière à teneur bleue élevée est utilisée comme, par exemple, dans les lampes fluorescentes, l'éclairage du jour naturel, etc.

Le modèle 8625 (P846) comporte un élément de chauffage standard de 6,3 V/0.6 A tandis que le modèle 8626 (P847) comprend le nouvel élément de chauffage à faible consommation de 6,3 V/0,095 A.

EE 81 771 *pour plus amples renseignements*

AUXILIAIRE À LAME VIBRANTE

Hivac Ltd, Stonefield Way, South Ruislip, Middlesex

Un nouvel auxiliaire à lame vibrante, XS10, a été ajouté à la gamme des types normaux Hivac. Etudié pour répondre aux mêmes spécifications électriques que la série XS2, il a une longueur de verre beaucoup plus courte et un moindre diamètre.

Ses dimensions sont: Longueur de verre 35 mm; longueur hors-tout 79,5 mm; diamètre maximum 4,7 mm.

Les surfaces de contact sont dorées et assurent ainsi un fonctionnement de longue durée à des courants allant jusqu'à 250 mA. La version standard du XS10 fonctionne entre 30 et 70 atmosphères et sa tension de rupture minima est de 450 V c.c.

Le temps de fonctionnement est inférieur à 1,3 msec, la durée de rebondissement est inférieure à 0,5 msec et le temps de déclenchement est de 0,3 msec lorsqu'il est mesuré à deux tours d'ampères de fonctionnement.

Le XS10 constitue l'équivalent britannique de certains types de composants à lame vibrante importés et il augmente la gamme déjà étendue des composants Hivac à performance élevée.

EE 81 772 *pour plus amples renseignements*

BLOCS D'ALIMENTATION

D.E.B. Electronics Ltd, New Barnes Mill, Cottonmill Lane, St. Albans, Hertfordshire

(*Illustration à la page 345*)

La nouvelle gamme des blocs d'alimentation de laboratoire réalisés par la D.E.B. Electronics Ltd, se caractérise par son degré réduit d'ondulation et son réglage précis, ainsi que par le fait qu'elle est entièrement équipée de semi-conducteurs au silicum. L'élément de base de bande d'essai est un double bloc dont chaque moitié fournit de 0 à 1 A 0 à 30 V avec une ondulation d'environ 300 μV et une régulation supérieure à 10 mV. La stabilité est supérieure à 1000 : 1.

Le bloc comprend un dispositif de protection constante du courant, le courant maximum fourni étant choisi au moyen d'un commutateur rotatif à 7 directions. La caractéristique de sortie est quasi rectangulaire de sorte que le bloc peut être utilisé comme source de courant avec une régulation de 5 mA. La perception à quatre bornes est prévue. Le bloc comprend des instruments de mesure avec largeur d'échelle de 10 cm pour l'affichage du courant ou de la tension.

Le réglage de tension s'effectue au moyen d'un seul bouton à démultiplicateur assurant un balayage sans plots sur l'ensemble de la gamme de tensions.

Chacune des deux moitiés du bloc est flottante et peut être reliée en série ou en parallèle à l'autre moitié sans perte de performance ou de protection. La protection est également assurée pour empêcher tout endommagement lorsque des connextions de polarité opposées sont effectuées. Des alimentations sont prévues pour donner un total de 8A dans le mode parallèle ou de 120 V dans le mode série.

Tous les blocs sont entièrement compatibles et logés dans des coffrets à instruments identiques, rapidement amovibles pour montage direct sur panneau frontaux de 47,5 cm × 17,5 cm. Ce sont des blocs robustes et d'une grande sécurité de fonctionnement continu à plein courant dans un court-circuit.

EE 81 773 *pour plus amples renseignements*

ISOLATEURS EN FERRITE

The M-O Valve Co. Ltd, Brook Green Works, London, W.6

(*Illustration à la page 345*)

La M-O Valve Co. Ltd, a réalisé une nouvelle série d'isolateurs en ferrite couvrant les bandes de fréquence de 5925 à 6425 MHz, de 5925 à 6175 MHz et de 6175 à 6425 MHz. Le modèle CIC4 que l'on voit dans notre gravure est un isolateur à déplacement de champ avec perte avant maxima de 0,35 dB et perte arrière minima de 35 dB. Le rapport d'amplitude de tension d'ondes stationnaires est de 1,02:1. Les modèles CIC5 et CIC6 sont des isolateurs à résonance avec pertes avant et arrière de 0.5 dB minimum et 25 dB maximum respectivement. Tous deux ont un rapport d'amplitude de tension d'ondes stationnaires de 1,06 : 1. Les connexions HF sur les trois modèles sont du type de guide d'ondes 14.

Ces dispositifs sont particulièrement indiqués pour l'emploi dans les systèmes de radio-communication à large bande.

EE 81 774 *pour plus amples renseignements*

Résumés des Principaux Articles

Le Laddic dans les circuits logiques entièrement magnétiques par D. J. Morris

Résumé de l'article aux pages 290 à 296

Cet article décrit l'appliquation et le comportement du Laddic dans les circuits logiques entièrement magnétiques dont la connexion entre les dispositifs magnétiques est en fil métallique. Le Laddic est utilisé dans ces circuits comme dispositif d'entrée pouvant assurer l'opération "ET", ainsi que comme convertisseur de courant continu en impulsions (les signaux d'entrée étant sous forme de couran continu et la sortie sous forme d'impulsions à répétition).

L'auteur présente une analyse des systèmes de commutation de flux du Laddic et suggère également des systèmes de commutation de flux pouvant constituer des variantes.

Un tachymètre à réseaux de diffraction circulaires par B. V. Jayawant et D. P. Rea

Résumé de l'article aux pages 297 à 301

L'appareil décrit utilise deux réseaux de diffraction circulaires ayant leurs axes exactement en coïncidence pour fournir un signal dont la fréquence est proportionelle à la vitesse de la machine à laquelle les réseaux sont associés. Un échantionnage à faible période produit un signal, dont l'amplitude est proportionelle à la fréquence et, donc, à la vitesse, qui peut être observé sur un oscilloscope à rayons cathodiques. Puisque la mésure de la fréquence se fait rapidement, l'appareil peut suivre et enregistrer les phénomènes transitoires dans la machine et permet l'étude immédiate de régimes transitoires.

Transistors à effet de champ dans les amplificateurs à courant continu par J. C. S. Richards

Résumé de l'article aux pages 302 à 305

L'auteur démontre, tant expérimentalement que par la théorie, que le fonctionnement d'un transistor à effet de champ, avec une polarisation appropriée, est tel que le courant peut etre rendu indépendent de la température. Il discute l'emploi du transistor à effet de champ comme étage d'entrée unipolaire dans les amplificateurs à courant continu. La dérive, qui prend naissance surtout dans l'alimentation ou dans le deuxième étage, peut être rendu, sans grande difficulté, inférieure à 50 V/°C ou 5×10^{-13} A/°C à 20°C. On donne les schémas de deux amplificateurs typiques.

Un amplificateur vidéo logarithmique à faible temps de remise par G. Kossoff, C. N. Liu et D. E. Robinson

Résumé de l'article aux pages 306 à 310

On esquisse l'étude d'un amplificateur vidéo logarithmique à remise rapide dont le principe de fonctionnement est celui de détection et addition successives. La gamme dynamique de l'amplificateur est 75 dB et le temps de remise après surcharge est 0,15 µsec. On obtient ce faible temps de remise au moyen d'étages successifs de différentiation ayant des diodes hautes fréquences entre les anodes. On donne un aperçu des caractéristiques d'amplificateurs à étages successifs variables. Les schémas d'amplificateurs de 2 MHz et de 8 MHz sont donnés.

Une méthode d'étude par approximations successives d'un compteur binaire double par D. Zissos

Résumé de l'article aux pages 311 à 315

L'auteur decrit une méthode d'étude par approximations successives d'un diviseur binaire à couplage direct qui utilise des modules NI en l'absence du complément du signal d'entrée. On introduit un moyen de simplifier les circuits NI—NI (et NON-ET—NONET). Puisque les modules sont à couplage direct, le circuit est insensible aux temps de montée et de descente du signal d'entrée et fonctionne indifféremment avec une onde rectangulaire ou une onde sinusoidale.

Une méthode de mésure de la probabilité de densité d'une fonction par P. N. Nikiforuk et G. Squires

Résumé de l'article aux pages 316 à 317

On décrit une méthode de mésure de la probabilité de densité d'une fonction aléatoire. Des amplificateurs opérationnels courants, un oscillateur et un compteur d'impulsions sont utilisés.

Le fonctionnement d'un Dekatron commandé par un champ magnétique par A. W. Gooder et M. H. N. Potok

Résumé de l'article aux pages 318 à 320

On montre qu'un Dekatron peut fonctionner d'une manière satisfaisante à partir de tensions de forme quelconque, depuis le continu jusqu'à plusieurs kHz., en présence d'un champ magnétique approprié. Les valeurs optimum des éléments du circuit et de la tension ont été trouvées expérimentalement.

Un appareil pour convertir une valeur efficace en valeur moyenne par H. H. A. Bath

Résumé de l'article aux pages 321 à 323

L'auteur présente une méthode pour obtenir une tension continue à partir de la valeur efficace d'une tension de forme quelconque à condition que le facteur de crête ne dépasse pas la valeur 4. On se sert de diodes, pour convertir le signal en une fonction rectangulaire, et du courant de charge d'un condensateur.

Un appareil de demonstration de la modulation d'impulsions par amplitude avec des effects de quanta par J. Karfunkel

Résumé de l'article aux pages 324 à 325

Tout système de modulation par impulsions dont le nombre d'impulsions par seconde est limité fait apparaître des effets de quanta qui dépendent, dans le cas de modulation d'impulsions par amplitude ou pûr durée, de la fréquence d'impulsion et, pour la modulation d'impulsions codée, du nombre d'étages. Ces effets se présentent sous la forme de quanta de bruit. Afin de faciliter l'étude de ce bruit et de l'intelligibilité dans diverses conditions, on a mis au point un appareil de demonstration. Puisqu'un appareil actuel est basé sur l'emploi d'environ 50 tubes et commandes, on a voulu créer un appareil simple et transistorisé.

Un simulateur de servomécanismes élémentaire transistorisé par P. G. M. Dawe

Résumé de l'article aux pages 326 à 328

L'auteur décrit un simulateur de servomécanismes qui se compose d'un certain nombre d'amplificateurs à un seul transistor avec contre-réaction en parallèle. Ceux-ci effectuent les opérations d'addition, d'intégration, d'avancement de phase et d'inversion. Les amplificateurs sont à couplage par capacité et les erreurs de déphasage des intégrateurs sont compensées. La réponse à un signal d'entrée rectangulaire d'une fréquence de 50 Hz. peut etre étudiée pour divers amortissements. Une diode sert à supprimer la partie négative du signal d'entrée.

Générateur de fonctions du type "Escalier" pour l'affichage de diagrammes de fonctionnement de moteurs par A. Watson

Résumé de l'article aux pages 329 à 331

Un générateur à escalier utilisant des modules binaires à transistors commerciaux est décrit dans cet article. Ce générateur a été utilisé comme base de temps pour étudier le comportement de moteurs à combustion interne ainsi que pour plusieurs autres applications de ce genre, qui sont également décrites.

L'analyse d'un modulateur d'impulsions par J. Kontos

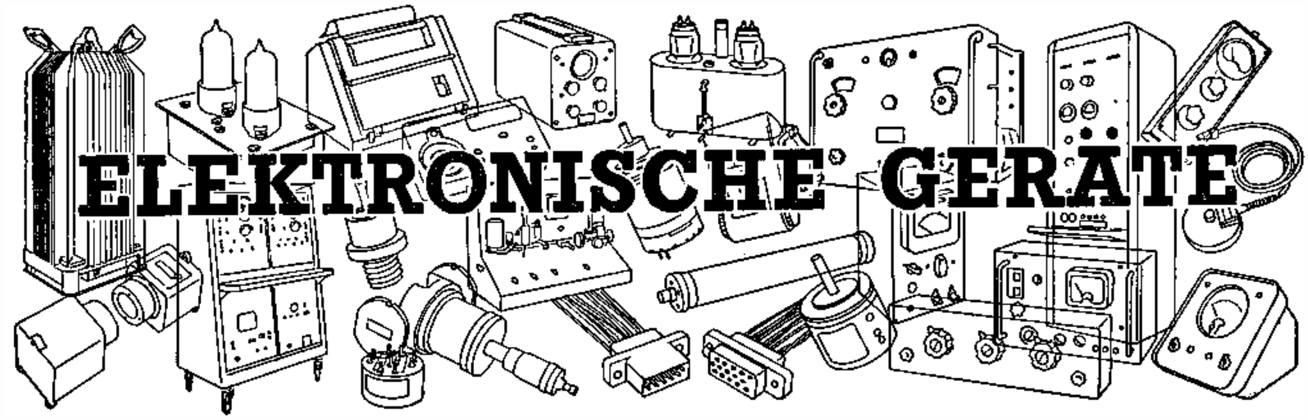
Résumé de l'article aux pages 332 à 333

L'auteur présente l'analyse d'un modulateur d'impulsions dont la fréquence de répétition et la durée de l'impulsion varient avec la tension d'entrée. Il developpe une expression qui relie la fréquence de répétition, la tension d'entrée et les paramètres du circuit. Cette expression est vérifiée expérimentalement en faisant la comparaison entre des valeurs calculées de la fréquence et celles obtenues au moyen d'un circuit spécialement construit à cette fin. On constate qu'il-y-a accord entre l'expérience et la théorie.

L'élimination des circuits de division par l'échelle de temps variable par D. K. Taylor

Résumé de l'article aux pages 334

La simulation analogique d'une équation comportant la division de deux variables exigeait jusqu'ici un circuit de division spécial, ce dernier étant souvent une cause d'erreurs importantes. L'auteur de cet article montre qu'on peut en certains cas se dispenser des circuits de division en utilisant la transformation du temps dans les équations. Cette méthode ajoute une équation supplémentaire qui indique le temps réel en fonction du temps calculé. Dans le cas d'une équation différentielle linéaire unique du premier degré, l'emploi de la transformation du temps au lieu du circuit de division remplace l'équation originale par deux équations différentielles linéaires accouplées du premier degré.



ELEKTRONISCHE GERÄTE

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 340 bis 345

Sauerstoffanalysator

Servomex Controls Ltd, Crowborough, Sussex

(Abbildung Seite 340)

Servomex Controls Ltd stellt eine neue Ausführung ihres tragbaren Sauerstoffanalysators unter der Bezeichnung DCL.101 Baumuster 11 vor. Dieser Analysator entspricht insofern den wachsenden Forderung nach grösserer Anpassungsfähigkeit älteren Ausführungen des DCL.101, als die von der Distillers Company entwickelten Messzellen für paramagnetische Suszeptibilität mit Platinaufschlämmung und elektromagnetischer Rückführung Verwendung finden. Die Anzeige wird durch Handeinstellung einer von 0...100% Sauerstoff geeichten Skala mit Unterteilung in 0,1% Sauerstoff erreicht. Die Energie wird normalerweise Trockenbatterien entnommen.

Das Baumuster 11 hat ein Eigensicherheitszertifikat und ist für Einsatz in Gasen der Gruppen Äthen, Pentan und Wasserstoff enthaltenden Atmosphären nach Norm BS.1259 geeignet. Das bedeutet, dass der Analysator in vielen Örtlichkeiten, in denen gefährliche Bedingungen bestehen, verwendbar und vor allem in Raffinerien und chemischen Werken nützlich ist.

Unter den Verbesserungen des neuen Analysators sind zu nennen: ein neuconstruierter, von Einflüssen externer Störfelder vollkommen freier Magnet; keine Beschränkung der Position beim Einsatz und verhältnismässige Freiheit vom Neigungseffekt (nicht mehr als 0,005 Prozent Sauerstoff pro 1° Neigung). Der D.C.L.101 Baumuster 11 eignet sich daher für tragbaren Einsatz.

Einrichtungen zum Probenehmen gehören jetzt zur Standardausrüstung, und mit jedem Analysator wird folgender Zubehör geliefert: ein Handsaugapparat, ein Kieselgel-Trockenrohr und ein Sinterglasscheibenfilter neuartiger Konstruktion mit sehr einfachen, eingebauten stopfbuchslosen Quetschhähnen, das der Benutzer ohne Schwierigkeiten auswecheln kann.

EE 81 751 *für weitere Einzelheiten*

100-MHz-Teiler

Advance Electronics Ltd, Roebuck Road, Hainault, Ilford, Essex

(Abbildung Seite 340)

Advance Electronics Ltd hat die wachsende Liste ihrer Erzeugnisse für die Frequenzmesstechnik durch den 100-MHz-Teiler TCD100 erweitert, um der im Bereich der höheren Frequenzen nachzukommen.

Der TCD100 ist für die Bereicherweiterung jedes zur Zeit kommerziell greifbaren elektronischen Zählgerätes—von 1 MHz aufwärts—auf 100 MHz bestimmt. Zum Messen wird auf der Frontplatte mittels eines Schalters ein Divisor von 100 oder 20 eingestellt. Die Ein- und Ausgangsparameter sind anpassungsfähig genug, um eine breite Auswahl von Eingangsquellen zu akkommodieren. Das Gerät ist transistorisiert und mit Tunneldioden bestückt.

Es hat eine Eingangsemfindlichkeit von 50 mV an 50 Ω; Ausgangsimpulse von 3 V_{ss} stehen von 10 kHz...1 MHz (:100) und 50 kHz...5 MHz (:20) zur Verfügung. Die Messgenauigkeit des Grundzählgerätes wird durch den Teiler nicht beeinflusst.

EE 81 752 *für weitere Einzelheiten*

Künstliche Übertragungsleitung

Telefon Fabrik Automatic A/S, 7 Amaliegade, Kopenhagen, Denmark

(Abbildung Seite 340)

Die künstliche Übertragungsleitung E.G.4 hat Kenndaten, die diejenigen von Fernsprechteilnehmerleitungen sehr genau simulieren.

Das Hauptverwendungsgebiet der künstlichen Leitung ist im Labor, wo sie die Einwirkung von Leitungslängen auf Teilnehmeranlagen, d.h. Fernsprecher, Datenübertragungssysteme usw., simuliert.

Die künstliche Leitung besteht aus RC-Abschnitten, die jeweils einer Einheitsleitungslänge von 500 m (auf Wunsch auch anderen Längen) entsprechen. Durch Kaskadenschalten einer Anzahl von 500-m-Abschnitten kann man jede beliebige Leitungslänge als Mehrfaches des 500-m-Abschnittes simulieren.

Dieser Typ der künstlichen Leitung umfasst in Wirklichkeit zwei Leitungen von höchstens 27 Abschnitten, die zusammen eine Leitungslänge von 13,5 km ergeben. Beide Leitungen können natürlich in Serie geschaltet werden, wodurch man eine Gesamtlänge von 27 km erhält.

Kaskadenschalten von Abschnitten wird durch Betätigung einer Reihe von Drucktasten erreicht. Zum Einfügen der Leitungslängen 0,5, 1, 2, 4 bzw. 6 km werden fünf Drucktasten benutzt. Die sich ergebende Länge für die eingefügten Abschnitte entspricht daher der Summe der mit den betätigten Drucktasten korrespondierenden Werte.

Die künstliche Leitung ist in einem Metallgehäuse untergebracht, dessen Frontplatte in zwei symmetrische Hälften unterteilt ist: eine für jede Leitung. Ausser den Drucktasten sind die Anschlüsse für die Leitung auf die Frontplatte montiert.

Die künstliche Leitung hat Klemmen für Eingang, Ausgang und Masseverbindungen. Die Bauelemente sind auf gedruckte Schaltungen montiert, die jeweils 9 Abschnitte umfassen. Die gedruckten Schaltungen werden mittels Mehrfachsteckverbindungen in das Gehäuse eingeschoben und können leicht für Prüfung und Reparatur entfernt werden.

Die Schaltungsplatten sind alle identisch und daher gegeneinander austauschbar.

EE 81 753 *für weitere Einzelheiten*

Einschübe für Oszillografen

Solartron Electronic Group Ltd, Farnborough, Hampshire

(*Abbildung Seite 341*)

Für die Solartron-Oszillografen CD1220 und CD1212 (CT484) gibt es zwei neue Einschübe.

Der Differential-Einschub mit hoher Verstärkung Typ CX1258 hat mit einem Ablenfaktor von 100 μV/cm hohe Gleichspannungsempfindlichkeit, hohe Gleichtaktunterdrückung (80 dB-10 000:1) und sehr niedrige Drift. Ausserdem gestattet die Höchstbandbreite von 200 kHz Einsatz als Universal-Verstärker mit umfangreichen Anwendungsmöglichkeiten.

Die Eingangsstufe gibt hohe Verstärkung und ist für Aufrechterhaltung genauer Symmetrie für höchste Gleichtaktunterdrückung ausgelegt. Bei gleichspannungsfreier Kopplung wird die niedrige Gleichspannungsdrift durch Gleichspannungsrückführung vernachlässigbar klein. In kritischen Positionen werden zuverlässige Sonderqualitätsröhren eingesetzt.

Technische Daten des CX1258:

Ablenkfaktor
: 100 μV/cm...2 V/cm in 14 geeichten Bereichen

Bandbreite
: 0...50 kHz bei 100 μV/cm
: 0...100 kHz bei 200 μV/cm
: 0...200 kHz bei 500 μV/cm

Gleichtaktunterdrückung
: nicht weniger als 80 dB zwischen 0 und 50 kHz

Gleichtakthöchstspannung
: $\pm$25 V_s von 100 μV/cm...20 mV/cm
: $\pm$250 V_s von 20 mV/cm...2 V/cm

Eichunsicherheit
: 3%

Der CX1259 hat ähnliche Kenndaten wie der CX1256, jedoch einen zusätzlichen direkt gekoppelten x10-Transistorverstärker, der bei einer Bandbreite von 0...24 MHz einen maximalen Ablenkfaktor von 5 mV/cm gibt. Die Grundkenndaten von 0...40 MHz bei 50 mV/cm sind beibehalten. Durch zusammengesetzte Gegenkopplung ergibt sich in allen Bereichen ausgezeichnete Impulswiedergabe.

Benutzung der Y-Verschiebung als Kompensationsspannung gewährleistet unverzerrte Darstellung von Signalen mit Gleichspannungsgehalt.

Ablenkfaktor u. Bandbreite
: x10 : 5 mV/cm 0...24 MHz
: xl : 50 mV/cm 0...40 MHz

Bereiche
: Geeicht in 1-2-5-Stufen mit Feineinstellung.

Abgebildet ist links der Differential-Einschub mit hoher Verstärkung CX1258 und rechts der Breitband-Einschub CX1259.

EE 81 754 *für weitere Einzelheiten*

Unterrichts-Messgeräte

White Electrical Instrument Co. Ltd, 10 Amwell Street, Rosebery Avenue, London, E.C.1

(*Abbildung Seite 341*)

Diese Instrumente wurden speziell für den modernen Physikunterricht entwickelt, sind aber auch für allgemeine Laborarbeiten geeignet. Sie werden als Typ RM/AS angeboten, sind robust gebaut und so konstruiert, dass sie einerseits den rauhen Ansprüchen im Schuldienst genügen und andererseits mit Messwerken ausgerüstet sind, deren hohe Qualität und Genauigkeit den Ansprüchen dieser neuen Aufgaben genügen.

Die 63,5-mm-Skalen sind deutlich lesbar beschriftet; die geneigten Tischgehäuse der Instrumente sind in zweitonigem, pastellgrauen Kunststoff ausgeführt, oben mit 4-mm-Klemmen ausgerüstet und mühelos sauber zu halten.

Die Messwerke des neusten Federtypes haben gehärtete und polierte Stahlzapfen, die in künstlichen Saphirlagern laufen. Die Dreheisentypen RM1 sind pneumatisch gedämpft, für Gleich- und Wechselstrom geeignet, aber nur mit Nullpunkt an der Seite zu haben. Der Drehspultyp RMC eignet sich nur für Gleichstrom, ist aber mit dem Nullpunkt an der Seite oder in der Mitte lieferbar. Er ist aperiodisch und elektrodynamisch gedämpft. Beide Typen sind in breiter Auswahl für Spannung und Strom geeicht lieferbar.

EE 81 755 *für weitere Einzelheiten*

Phasenmesszusatz

Servomex Controls Ltd, Crowborough, Sussex

(*Abbildung Seite 341*)

Servomex Controls Ltd hat die Fertigung eines regelbaren Phasenzusatzes für den kürzlich entwickelten NF-Wellenformgeber LF 141 aufgenommen. Der Hauptzweck des neuen Gerätes VP.142 ist schnelles und genaues Messen von Phasen im Frequenzbereich von 2 kHz bis zu 1 Periode in 500 Sekunden herunter.

Das geschieht nach dem bekannten Verfahren der Lissajousschen Figuren, nach dem eine Ellipse in eine gerade Linie entartet, wenn die Phase korrekt ist. Durch eine im Datenblatt erläuterte Technik ist es möglich, diese Messungen bei Tiefstfrequenzen in unter einer Periode auszuführen, wobei entweder ein Oszillograf oder ein Schreiber irgendwelcher Art erforderlich ist. Amplitudenregelung ist vorgesehen, so dass bei einer Schrägstellung der Lissajousschen Figur um 45° ausser der Phase auch die Amplitude direkt am Amplitudenregler abgelesen werden kann. In den seltenen Fällen, in denen das zu messende Signal zu viel Rauschen oder Verzerrung enthält, kann man ein geeignetes Bandfilter benutzen.

Die regelbare Phasenspannung wird von einem Potentiometer mit kreisförmiger Wicklung abgeleitet, die bei 0°, 90°, 180° und 270° gespeist wird. Eine Spezialschaltung hält während des Durchdrehens des Phasenreglers die Amplitude konstant, um das angewandte Amplitudenmessverfahren zu ermöglichen.

Der Zusatz wird an den Boden des LF 141 geschraubt; Eingänge werden über eine Mehrfach-Steckverbindung zugeführt. Das kombinierte Gerät kann ausser als Quelle für niederfrequente Signale vieler Art auch zum Messen von Phasen in Energiesystemen für die verschiedensten Zwecke eingesetzt werden, z.B. für Demonstration des Einschwingvorganges in Wechselstromsystemen. Durch Abschalten einer an der Torschaltung liegenden Sperrspannung lässt sich ein Sinuswellenzug in jedem gewünschten Winkel auslösen.

Zu den vielen erzeugten Wellenformen gehören u.a. Rechteckwellen mit veränderlicher Anstiegszeit, in denen sowohl Impulsfrequenz wie Anstiegszeit getrennt regelbar sind, sowie glockenförmige Impulse, die innerhalb einer genau bekannten Bandbreite Harmonische erzeugen.

Für medizinische und andere Zwecke können Impulsreihen mit bekannter Impulzahl erzeugt werden. Der Zähler kann Rechteckimpulse mit kurzer Anstiegszeit von Rechteck-Hilfsausgangsanschlüssen zählen, während am Hauptausgang die gewünschte Wellenform abgegeben wird.

Der regelbare Phasenzusatz Servomex VP 142 wurde speziell für erleichterte Wartung ausgelegt. Die elektronischen Bauelemente sind aus diesem Grund auf schwenkbaren Leiterplatten angeordnet. Die meisten Bauelemente wurden von der Liste der für die Streitkräfte zugelassenen Typen gewählt und entsprechend der Zulassung belastet.

EE 81 756 *für weitere Einzelheiten*

NF-Millivoltmeter

Elektromesstechnik Wilhelm Franz KG, 763 Lahr/Schwarzwald Kaiserstrasse 80, Deutschland

(*Abbildung Seite 341*)

Das volltransistorisierte Voltmeter EMT 125 wurde speziell für exakte Messungen an NF-Verstärkern und -Geräten sowie elektrostatischen Anlagen entwickelt. Seine Leistungsaufnahme ist äusserst gering, und es kann ohne nennenswerte Erwärmung ständig angeschlossen bleiben.

Das moderne flache Gehäuse gestattet Übereinanderstellen mehrerer Geräte; alle Anschlüsse, Tasten und Schalter sind zusammen mit dem Anzeigeinstrument auf der Frontplatte angeordnet. Die Skala ist in Volt und Dezibel geeicht.

Das Gerät hat 12 Messbereiche mit Skalenendwerten von 1 mV bis zu 300 V. Der kleinste ablesbare Skalenwert beträgt 100 μV. Der Frequenzbereich des Gerätes geht bis zu 200 kHz; ein eingebauter Tiefpass kann durch Tastendruck eingeschaltet werden und beschränkt die oberste messbare Frequenz auf 20 kHz. Auch Wellenformen, die stark von der Sinusform abweichen, können gemessen werden. Mit einer Drucktaste kann man wahlweise auf Effektivwert- oder Spitzenwert-Messungen schalten.

Bei Effektivwert-Messungen nach DIN 45 402 und einer Impulsfolgefrequenz von 200 Hz erreicht das Gerät bei einer Abweichung von 1 dB ein Tastverhältnis 1:10 bei Vollausschlag. Die Spitzenwert-Messung kann man zur Fremdspannungsmessung von z.B. Rauschspannungen verwenden, wobei Frequenzen über 20 kHz dann ausgefiltert werden können.

Bei Einsatz als Messverstärker ist die grösstmöglichste Verstärkung im Messbereich 1 mV 60 dB; sie fällt pro Messbereich um 10 dB ab. Am Ausgang steht 1 V bei Vollausschlag und Abschluss mit 4 kΩ zur Verfügung.

Zur inneren Eichung ist eine stabilisierte Referenzspannungsquelle eingebaut.

EE 81 757 *für weitere Einzelheiten*

Digitaluhr

Vertrieb: Systems Engineering Services Ltd, 35-39 South Ealing Road, Ealing, London, W.5

(*Abbildung Seite 342*)

Zwei neue Präzisions-Digitaluhren, die die genaue Tageszeit in Dezimalform darstellen und in binär-kodierter Dezimalform abgeben, werden von der Electronic Engineering Company in Kalifornien hergestellt.

Die neuen Digitaluhren EECO 815 und EECO 816 geben beide parallele logische Pegel, und Typ 816 gibt ausserdem Serienentnahme. Weiterhin geben sie direkte Zeitcodes für Datenerfassungs- und Datenaufzeichnungssysteme, Elektronenrechner, Streifenschreiber, Druckwerke usw. Beide Modelle sind in Festkörpertechnik ohne Motoren, Relais oder Schrittschalter ausgeführt.

EECO 815 und EECO 816 sind beide mit der Netzfrequenz synchronisiert, aber auch mit interner Frequenzquelle lieferbar. Ausser den Zeitcodes können gleichzeitig sechs Rechteck-Impulsraten entnommen werden. Ausfall der Netzspannung wird in jedem Modell angezeigt, bis eine Rückstelltaste gedrückt wird.

Modell EECO 815 hat nur Parallel-Ausgangscode und gibt einen genauen Zeitcode für Stunden, Minuten und Sekunden in binär-kodierter Dezimalform ab, der direkt in Datenaufzeichnungs- und Datenerfassungssysteme mit Lochstreifen, Magnetband oder direkten Druckwerken gespeist werden kann oder für Direkteingabe in Datenverarbeiter zur Verfügung steht.

Modell EECO 816 hat Parallel- und Serienausgänge und ist ausser für die obigen Zwecke auch für serienförmige Zeitregistrierung auf Streifenschreibern oder Magnetbandgeräten einzusetzen. Ein modulierter Trägerausgang (100 oder 1000 Hz Trägerfrequenz) kann im Sonderauftrag vorgesehen werden, um den Serien-Zeitcode als breitenkodierten, amplitudenmodulierten Träger auf Analogmagnetbändern aufzuzeichnen.

EE 81 758 *für weitere Einzelheiten*

Übersetzungsverhältnis-Analogrechner

F. C. Robinson & Partners Ltd, Davies House, 181 Arthur Road, Wimbledon, London, S.W.19

(*Abbildung Seite 342*)

Dieser besonders für technische Lehranstalten und Hochschulen entwickelte Analogrechner beruht auf dem bekannten Konzept, nach dem das Windungsverhältnis von Rechentransformatoren komplexe Zahlen darstellt. Dieses Prinzip hatte breite Anwendungsmöglichkeiten in der Schaltungs- und Netzwerkanalyse.

Der Rechner besteht aus:

drei Zwillingsimpedanzen (zur Darstellung von Impedanz oder Admittanz bei jedem Phasenwinkel),

zwei Zwillingsspezialimpedanzen (Impedanz oder Admittanz), Spannungs- oder Stromquelle, Übertrager oder Kopplungswiderstand),

einem transistorisierten 400-Hz-Oszillator,

einem Nullabgleichgerät mit Einzelrechentransformator, logarithmischem Verstärker und Anzeigeinstrument, sowie

einer Stromversorgung.

Die Unsicherheit für das Einstellen der Schaltungskonstanten und das Messen ist 0,1 Prozent.

EE 81 759 *für weitere Einzelheiten*

Anschlussstück für Subminiaturkoaxialkabel

Sealectro Ltd, Hersham Trading Estate, Walton-on-Thames, Surrey

(*Abbildung Seite 342*)

Ein 75-Ω-T-Anschlussstück mit lötfreiem Anschlag für Subminiaturkoaxialkabel ist das neuste der von Sealectro Ltd lieferbaren Verbindungselemente. Die Verbindung 5116 hat Quetschanschlüsse, die besonders für hohe Produktionsraten geeignet sind und einen Kabelabschluss ergeben, der dauerhafter als das Kabel selbst ist.

Das Anschlussstück hat Befestigungslöcher für Montage auf Frontplatte oder Scheidewand und ist für Stapeln in Verbindungszentren für Kabel geeignet. Ausserdem hat der T-Anschluss 5116 eine 0,0025 mm starke Vergoldung, die der Einrichtung zusammen mit der ConheX-Konstruktion höchste elektrische Leistungsfähigkeit gibt.

EE 81 760 *für weitere Einzelheiten*

Überschall-Prüfgerät

Ultrasonoscope Co. (London) Ltd, Sudbourne Road, London, S.W.2

(*Abbildung Seite 342*)

Das Ultrasonoscope Baumuster 4 wurde für die zur Lösung der schwierigsten Kontrollaufgaben erforderliche Leistungsfähigkeit und Vielseitigkeit entwickelt. Es ist für Routineprüfungen im Werk ebenso wie höhere wissenschaftliche Arbeiten im Labor verwendbar. Mit einem Monitor kann es als vollautomatisches Abtast- und Registriersystem arbeiten.

Da Einsatz in verschiedenen Industrien oft zu miteinander in Widerspruch stehenden Geräteeigenschaften führen kann, ist der Verstärker des Baumusters 4 auswechselbar. Für Sonderzwecke kann ohne Schwierigkeiten ein Spezialverstärker gebaut werden. Die Standardverstärker haben bei jeder der Frequenzen 10, 5, 2½, 1¼ und 0,5 MHz grosse Bandbreite und Dynamik; sie können kurze Impulse unterscheiden und sind linear.

Streuung von der Struktur des Materials kann mittels einer kontinuierlich regelbaren Sperre ausgefiltert werden. Der geeichte Abschwächer hat einen Bereich von 0 ... 80 dB mit einer Einstellunsicherheit von 1 dB. Der leistungsfähige Geber und die hohe Verstärkung des Verstärkers geben wesentlich höhere Durchdringung im Betrieb mit einem oder zwei Wandlern.

Es kann entweder gleichgerichtete, nicht gleichgerichtete oder B-Schirm-Darstellung eingestellt werden. Bei Darstellung auf dem B-Schirm hellen die von Fehlern herrührenden Echos die Zeitbasis auf, und die Darstellung ist einem Röntgenbild ähnlich.

Die Oszillografenröhre hat einen 127-mm-Planschirm mit fotografischem Leuchtstoff und optischem Filter zur Erhöhung des Kontrastes. Die Zeitbasis ist linear und die höchste Geschwindigkeit 2 μs, so dass 6,35 mm Stahl für genaue Dicken- und Geschwindigkeitsmessungen auf dem vollen Schirm untersucht werden können. Der Höchstbereich ist 6 m.

Die Zeitverzögerung ist zwischen 0 und 300 μs kontinuierlich einstellbar, und jeder beliebige Abschnitt der Spur lässt sich mittels eines geeichten Reglers vergrössern. Die Impulsfolgefrequenz ist 50 ... 1000 Hz, und das Baumuster 4 ist gegen Netzschwankungen voll stabilisiert.

EE 81 761 *für weitere Einzelheiten*

Bohrmaschine für Leiterplatten

Vertrieb: Alfred Jacoby (Machine Tools) Ltd, 54 Manchester Street, London, W.1

(*Abbildung Seite 343*)

Die von Frautschi & Monney in der Schweiz erzeugte Posalux-Bohrmaschine "Copyfor" wurde hauptsächlich für die Herstellung von gedruckten Schaltungen entwickelt, ist aber ebenso zum Bohren von Firmenschildern aus Aluminiumfolie oder ähnliche Zwecke geeignet.

Gebohrt wird von unten nach oben mittels einer vertikal angeordneten Öldruck - Luftdruck - Bohreinrichtung "Unifor 10". Der Arbeitsgang wird entweder manuell oder semiautomatisch gesteuert. In jedem Fall wird der Plattenstapel mit der Bohrschablone oben solange verschoben, bis der in derselben Achse wie die Bohrspindel liegende gefederte Positionierbolzen in eins der Schablonenlöcher einrastet. Bei selbsttätigem Betrieb werden die Leiterplatten dann automatisch mit Druckluft festgeklemmt und der Bohrer vorgeschoben;

das Bohren findet fast momentan statt. Bei Handbetrieb werden Festspannen und Bohren manuell eingeleitet.

Mit dieser Maschine kann auch ein Projektionsschirm oder Mikroskop geliefert werden, die Erstellung von Bohrschablonen auf der Maschine und von Original-Leiterplatten gestatten. Auf diese Weise kann man dann auch Kleinserien herstellen, für die sich Anfertigung von Schablonen nicht lohnt. Der Projektionsschirm gibt ein korrigiertes Bild hoher Lichtstärke und Vergrösserung.

Mit der Maschine lässt sich ein hoher Ausstoss erreichen, was z.T. auf die Konstruktion des Rastbolzens zurückzuführen ist, dessen Druck den Arbeitsbedingungen entsprechend einstellbar ist. Da die Bewegung des Bolzens sehr fein ist, wird vorzeitiges Anlaufen und damit Ausschuss vermieden. Der Kopf kann nach hinten gekippt werden, was leichtes Auswechseln der Bohrer und müheloses Entfernen der Späne erlaubt. Gegen Aufschlag sind eine Absaugeinrichtung und ein in das Bohrfutter eingebauter Gebläsering lieferbar, die gewährleisten, dass alle Späne in den Abfallschacht gehen.

Zur Verkürzung der Bohrzeit und Verhinderung von Gratbildung wird mit hohen Spindelgeschwindigkeiten gearbeitet. Die Maschine ist für Bohrer bis zu 3 mm sowie bis zu 420 mm breite Leiterplatten ausgelegt.

EE 81 762 *für weitere Einzelheiten*

Zeitmesser

Taylor Electrical Instruments, Montrose Avenue, Slough, Buckinghamshire

(Abbildung Seite 343)

Taylor Electrical Instruments, ein Mitglied der M-I-Gruppe, hat einen neuen Zeitmesser herausgebracht, der in das Vista-Programm passt und von vorn gesehen dieselbe Ausführung wie das Vista Modell 40 hat.

Das Instrument ist für 110 ... 125 V oder 200 ... 250 V, 50 oder 60 Hz Netzanschluss lieferbar und kann Stunden sowie zehntel Stunden bis zu 99999,9 Stunden registrieren, wonach sich der Zähler automatisch wieder auf Null zurückstellt.

Ein selbstanlaufender Synchronmotor gestattet selbst bei kurzen Perioden korrektes Registrieren.

EE 81 763 *für weitere Einzelheiten*

Impulsgeber

Hewlett Packard, Dallas Road, Bedford

(Abbildung Seite 343)

Hewlett-Packard GmbH—die Forschungs- und Fertigungsstätten der Hewlett Packard in Deutschland—hat einen kompakten, preisgünstigen Impulsgeber herausgebracht. Der Hewlett-Packard-Impulsgeber 8000A ist für die Bestimmung des Frequenzganges von Breitbandinstrumenten, Schaltungen und Halbleitern ausgelegt.

Die Ausgangsspannung des Generators ist 10 V an 50 Ω und im wesentlichen ein abgeflachter Impuls mit 100 kHz Folgefrequenz und unter 20 ns Anstiegszeit. Alle Impulskopfabweichungen, wie z.B. Überschwingen und gedämpfte Schwingungen, werden unter 2 Prozent gehalten, der abgeflachte Impuls wird für mindestens 100 Sekunden bewahrt. Die Abfallzeit ist kürzer als 20 ns. In der Ausgangsschaltung werden als impulsformende Elemente von Hewlett-Packard entwickelte stufenweise Entionisierungsdioden verwendet.

Unter den Eigenschaften des Instrumentes sind zu nennen: Ausgangsabschwächung von 0,1 bis 10 V in sechs Stufen, positive und negative Ausgangspolarität und ein vorlaufender Triggerausgang. Bei Arbeiten mit Sampling- oder anderen sehr schnellen Oszillografen behebt der Triggerimpuls die Notwendigkeit für Verzögerungsleitungen. Er läuft 200 ns vor, hat eine Anstiegszeit von unter 6 ns, 20 ns Dauer und eine Amplitude von 0,5 V an 50 Ω.

EE 81 764 *für weitere Einzelheiten*

Festkörperrelais und -zeitgeber

Keyswitch Relays Ltd. 120-132 Cricklewood Lane, London, N.W.2

(Abbildung Seite 344)

Einsteckzeitgeber und transistorisierte Relais der Serie P33VT wurden von Keyswitch Relays Ltd zur Steuerung kleiner Ausgangsleistungen eingeführt und sind als standardisierter Baustein lieferbar. Diese neue Serie entstand aus dem Relaistyp P33 und kann mit diesem sowie mit der jüngstens bekanntgegebenen Bauelemente-Schaltplatte 304 verwendet werden.

Anwendungsmöglichkeiten für die Serie P33VT bestehen u.a. in der Regelung von Atmosphäre, Temperatur und Wärme, in Suchgeräten, Werkzeugmaschinensteuerung, Spannungs- und Stromfühlern, Füllstandwächtern, Alarmschaltungen und den verschiedensten Abläufen der Datenverarbeitung.

In dieser Serie gibt es vier Standardtypen, die jeweils bestimmten Anforderungen entsprechend in vielen Abwandlungen lieferbar sind. P33TV1 speist ein Kleinsignal von 10 μA≃ in eine 10-kΩ-Impedanz, die das Schalten des Relais steuert. Die Speisespannung ist üblicherweise 24 V≃, die Kontaktbestückung ein Umschalter für hohe Belastung (5 A, 230 V~ ohmisch) oder für leichte Belastung bis zu 1 A.

Das Relais P33VT3 wird entweder durch Schliessen externer Kontakte und den dadurch hervorgerufenen niedrigen Eingangsstrom von 500 μA oder weniger, oder durch Änderung eines externen Widerstandes betätigt. Wo Verwendung mit Anzeigegeräten infrage kommt, werden Kontaktprellen und Schwankungen des Messwerkes kompensiert.

P33VT7 gestattet direktes Schalten des Relais mittels Hochleistungsfotozellen, einer mit einem zweistufigen Festkörperverstärker 15 μA erzeugenden schwachen Lichtquelle, oder Steuerung mittels externer Lichtquelle über eine eingebaute Fotozelle.

P33VT8 ist eine Serie stabilisierter Zeitrelais mit Schnappkontakt und Verzögerungsbereichen von 0...10 s, 5...30 s, 10...60 s, 15...120 s und bis zu 5 Minuten.

Die Betriebsspannung für die Typen der Serie P33VT ist 750≃ einschliesslich Sockel und Fassung, der Isolationswiderstand bei 500 V— besser als 1 GΩ.

EE 81 765 *für weitere Einzelheiten*

Druckgeber

Intersonde Ltd. The Forum, High Street, Edgware, Middlesex

(Abbildung Seite 344)

Intersonde Ltd hat in ihr Programm von Druckgebern mit gekitteten Dehnstreifen die Serie B.M. aufgenommen, die besonders für niedrigere Drucke von 7 kg/cm²... 53 kg/cm² entwickelt wurde.

Die Druckgeber beruhen auf einem druckempfindlichen Element in Form eines einseitig geschlossenen Berylliumkupferrohres, auf das vier in Brückenschaltung verbundene Dehnstreifen gekittet sind. Gasförmige oder flüssige Druckträger, die Zutritt zum Inneren des Rohres haben, rufen sehr kleine Abmessungsänderungen hervor, die momentan durch die beiden aktiven Brückenzweige in Widerstandsänderungen umgesetzt werden. Die Diagonalarmspannung ist dem angewandten Druck proportional.

Die Messwertwandler haben einen Ausgangswiderstand von 350 Ω und erzeugen bei 20 V Erregung und voller Nennlast eine Ausgangsspannung von 30 mV. Der kombinierte Nichtlinearitäts- und Hysteresefehler ist innerhalb ±0,1 Prozent des Skalenendwertes und der Betriebstemperaturbereich −40°C ... +120°C.

EE 81 766 *für weitere Einzelheiten*

Miniatur-Fernsehkamera

EMI Electronics Ltd, Hayes, Middlesex

(Abbildung Seite 344)

EMI Electronics Ltd hat eine neue volltransistorisierte Industrie-Fernsehkamera entwickelt, die nur 114,3 mm lang ist und nur etwa 800 g wiegt.

Da die Kamera trotz ihrer kleinen Abmessungen äusserst robust ist und Bilder mit sehr hoher Auflösung erzeugt, ist sie besonders in unzugänglichen Positionen nützlich, wo rauhe Behandlung und schlechte Beleuchtung zu erwarten sind. Sie kann mit 405, 525 und 625 Zeilen arbeiten; Umschaltung erfolgt über Drucktasten.

Die Kamera ist in zwei dichten Zylindern aus rostfreiem Stahl untergebracht, die jeweils etwa 114 mm lang sind und 43 mm Durchmesser haben.

Die nur etwa 800 g wiegende Objektiveinheit enthält ein ½"-EMI-Vidikon und kann bequem in einer Hand gehalten werden; mit der anderen Hand kann man durch Drehen des Objektivs fokussieren oder in einen beschränkten Raum einbauen.

Der andere Zylinder enthält den Vorverstärker und wiegt nur 340 g; er kann bis zu ca. 30 m von der Objektiveinheit entfernt stehen und durch Kabel angeschlossen werden.

Das Kontrollgerät und andere zum Kamerazug gehörige Geräte kann man bis zu 300 m entfernt aufstellen.

EE 81 767 *für weitere Einzelheiten*

Registrierender Dehnungsmesser

C.N.S. Instruments Ltd, 61 Holmes Road, London, N.W.5

(Abbildung Seite 344)

C.N.S. Instruments Ltd gibt bekannt, dass ihr registrierender Dehnungsmesser nunmehr mit automatischer Bereichumschaltung lieferbar ist. Das mit dieser Einrichtung ausgerüstete Instrument wird als Einschub in ein 19"-Gestell mit 0 ... 10 mV Ausgangsspannung (0 ... 50 mV gegen Sonderauftrag) geliefert und ist für Einsatz mit einem Kompensationsschreiber, z.B. einem Honeywell-Brown Electronik, gedacht.

Die Standardausführung kommt in sechs Bereichen mit Endwerten von 6 und 12 Mikron, 0,025, 0,25, 0,625 und 2,5 mm. Auf einem 250 mm breiten Registrierstreifen können daher Sprünge von 0,05 Mikron gemessen werden.

Andere Bereiche sind Kundenwünschen entsprechend gegen Sonderauftrag lieferbar.

Die automatische Bereichumschaltung ist besonders bei Dauerstandversuchen und anderen mechanischen Tests nützlich, in denen die Dehnungsrate keine lineare Funktion der Zeit ist. Dauerprotokolle sind jetzt während der Nacht und übers Wochenende ohne Überwachung durch geschultes Personal möglich. Die empfindlichste Einstellung kann dann für die Anfangsphase des Tests eingestellt werden, und während des Versuchsablaufes wird das Gerät sich jeweils bei Erreichung des Skalenendwertes auf den nächsten Bereich umschalten.

Besondere Aufmerksamkeit wurde der Kleinsthaltung der Temperatureinflüsse auf die einzeln oder als summierendes Paar lieferbaren Messwertgeber gewidmet, und die statischen sowie die Drehspulen sind auf Quarzspulenkörper gewickelt, die für Temperaturen bis zu 45°C., im Sonderfall bis zu höchstens 100°C aushalten können.

EE 81 768 *für weitere Einzelheiten*

Millivoltmeter

Standard Telephones & Cables Ltd, Transmission Testing Apparatus Division, Corporation Road, Newport, Monmouthshire

(Abbildung Seite 344)

Ein vielseitiges neues hochohmiges Millivoltmeter wurde von Standard Telephones & Cables Ltd für allgemeine Verwendung und die Fernmeldetechnik herausgebracht.

Modell 74 251 ermöglicht es, Messungen im Frequenzbereich 20 Hz ... 20 MHz in zehn Teilbereichen von 0 ... 1 mV bis zu 0 ... 30 V vorzunehmen. Ein aktiver Messkopf hat eine Eingangsimpedanz von über 100 kΩ parallel zu 40 pF.

Das Gerät ist in zwei Ausführungen erhältlich: Ausführung 'A' wird aus eingebauten Batterien gespeist, Ausführung 'B' aus eingebauten Batterien oder dem Wechselstromnetz.

Modell 74 251 ist ein tragbares, transistorisiertes Instrument für Mehrzweckeinsatz und Verwendung im Basisband-Frequenzbereich von Mikrowellen-Richtfunkverbindungen. Es kann auch als Brückendetektor Verwendung finden, und der eingebaute Eichoszillator gibt an externe Schaltungen bei ca. 100 kHz ein Signal ab. Der interne Verstärker hat Ausgangsanschlüsse, um ihn in externen Schaltungen verwenden zu können.

Das Messinstrument hat zwei Spannungsskalen und auch eine Dezibelskala für 75-Ω-Schaltungen. Die Batteriespannung kann auch mit dem Messinstrument kontrolliert werden.

EE 81 769 *für weitere Einzelheiten*

Tiefstfrequenzgenerator

Vertrieb: Claude Lyons Ltd, 76 Old Hall Street, Liverpool, 3

(Abbildung Seite 345)

Der C.R.C.-Tiefstfrequenzgenerator GB64 hat fünf in Frequenz und Periode geeichte 10:1-Teilbereiche, die 0,005 ... 500 Hz überstreichen, und gibt von einer internen 100-Ω-Impedanz wahlweise Sinus-, Rechteck- oder Dreieckwellenformen mit symmetrischer Ausgangsspannung von bis zu 40 V_{ss} an 10 kΩ ab. Die maximale Verzerrung der Ausgangssinuswellenform ist 2 Prozent, die Rechteckwelle hat eine Anstiegszeit von 25 μs, und der Steigungsfehler der Dreieckwelle ist höchstens 2 Prozent. Der Höchstpegel des statistischen Rauschens ist 20 mV, die Frequenzkonstanz ist 0,5 Prozent und die Pegelstabilität bei ±10% Netzschwankungen dank der Konstantstromversorgungen 0,2 dB.

Man kann einen Synchronisierungsimpulsausgang (25 V_S, 50 μs), der aus abwechselnden positiven und negativen Impulsen besteht, entnehmen, z.B. zum Triggern von Oszillografen.

EE 81 770 *für weitere Einzelheiten*

Vidikon mit hohem Auflösungsvermögen

The English Electric Valve Co. Ltd, Chelmsford, Essex

(Abbildung Seite 345)

Das Programm der English Electric Valve Company für 1"-Vidikons mit hohem Auflösungsvermögen und getrennter Maschenelektrode wurde durch Einführung von zwei neuen Typen ergänzt, deren Höchstempfindlichkeit im blauen Gebiet des Spektrums liegt.

Diese Röhren—die 8625 (P846) und 8626 (P847)—haben eine völlig neue Fotoschicht, die sehr hohe und gleichförmige Empfindlichkeit, sehr niedriges Nachziehen und verringertes Langzeitkleben, sowie bei Wolframbeleuchtung die korrekte panchromatische Empfindlichkeit und selbst bei hohen Signalströmen ein hohes Auflösungsvermögen hat.

Durch verbesserte Fertigungsverfahren, nach denen die Fotoschicht zur Erzielung überall gleichförmiger Abscheidung "vorfabriziert" wird, kann man ausgezeichnete Gleichförmigkeit des Signals erzielen. Die Studiobeleuchtung kann wegen der durch die neue Fotoschicht erreichten sehr kurzen Nachziehzeiten wesentlich herabgesetzt werden.

Ein weiterer Vorteil ist die äusserst gute Wiedergabe von Fleischfarbtönen auf Grund der sehr gut ausgeglichenen Spektralempfindlichkeit dieser Vidikons.

Einführung der neuen Fotoschicht macht die 8625 (P846) und die 8626 (P847) im Zusammenhang mit der getrennten Feldmaschenbauart für Live- und Zwischenfilmabtastung sowohl im Farb- wie im Schwarzweissfernsehen hervorragend geeignet. In Farbfernsehkameras gibt die blaue Extraempfindlichkeit einen höheren Rauschabstand und auch höhere Ausgangssignale, wo Lichtquellen mit hohem Blaugehalt vorkommen, z.B. bei Leuchtstoffröhren, bei natürlichem Tageslicht usw.

Die 8625 (P846) hat einen Standardheizfaden (6,3 V; 0,6 A), die 8626 (P847) dagegen den neuen Heizfaden mit geringer Stromaufnahme (6,3 V; 0,095 A).

EE 81 771 *für weitere Einzelheiten*

Schutzgaskontakt

Hivac Ltd, Stonefield Way, South Ruislip, Middlesex

Das Programm der Hivac-Standardtypen wurde durch den neuen Schutzgaskontakt XS10 erweitert. Die elektrischen Daten entsprechen denen der Serie XS2, doch ist das Glasrohr kürzer und der Durchmesser kleiner.

Das Glasrohr ist ca. 35 mm lang, die Gesamtlänge ist 79,5 mm und der grösste Durchmesser 4,7 mm.

Die Kontaktflächen sind mit Gold plattiert, das teilweise in das Grundmaterial diffundiert ist, und geben bei Strömen bis zu 250 mA eine lange Lebensdauer. In der Standardausführung wird der XS10 mit 30 ... 70 Amperewindungen betätigt und hat eine Mindestdurchbruchsspannung von 450 V—.

Die Ansprechzeit liegt unter 1,3 ms, das Prellen dauert weniger als 0,5 ms, und die Abfallzeit beträgt 0,3 ms, wenn sie bei der doppelten Amperewindungszahl, die die Kontakte gerade betätigen kann, gemessen werden.

Der XS10 ist das britische Äquivalent

bisher importierter Schutzgaskontakttypen und ergänzt das bereits umfangreiche Sortiment der Hivac-Hochleistungstypen.

EE 81 772 *für weitere Einzelheiten*

Stromversorgungen

D.E.B. Electronics Ltd, New Barnes Mill, Cottonmill Lane, St. Albans, Hertfordshire

(*Abbildung Seite 345*)

Ein neues Programm für Labor-Stromversorgungen mit geringer Restwelligkeit und enger Regelung, die durchweg mit Silizium-Halbleitern bestückt sind, wurde von D.E.B. Electronics Ltd bekanntgegeben. Die Tisch-Grundausführung ist eine Doppelstromversorgung; jede Hälfte gibt 0 ... 1 A bei 0 ... 30 V mit etwa 300 μV Restwelligkeit und einer Regelung von besser als 10 mV ab. Die Konstanz ist besser als 1 000 : 1.

Schutzschaltungen für konstanten Strom sind vorhanden, und der zur Verfügung stehende Höchststrom wird durch einen 7stufigen Drehschalter eingestellt. Die Ausgangskennlinie ist so nahezu rechteckig, dass das Gerät als Stromquelle mit 5 mA Stromregelung eingesetzt werden kann. Vierpolige Fühler sind vorhanden.

Ein Messinstrument zur Anzeige von Spannung und Strom mit 1 Prozent Genauigkeit auf einer 101,6 mm langen Skala ist für jede Hälfte eingebaut und durch eine klare Frontscheibe ablesbar.

Die Spannung wird über den ganzen Bereich kontinuierlich mittels eines Einzeleinreglers eingestellt.

Beide Gerätehälften sind erdfrei, so dass sie sich ohne Leistungs- oder Schutzverluste in Serie oder parallel schalten lassen. Die Schutzschaltung ist auch effektiv, wenn entgegengesetzte Polaritäten zusammengeschaltet werden. Stromversorgungen, die insgesamt 8 A parallel geschaltet und 120 V in Serie abgeben, sind lieferbar.

Alle Geräte sind völlig kompatibel und in identische Instrumentgehäuse eingebaut, die auf den 19" (483 mm) × 178 mm Frontplatten leicht ausgebaut und in Gestelle montiert werden können. Sie sind auf Zuverlässigkeit und Haltbarkeit konstruiert und können kontinuierlich mit vollem Sollstrom in einen Kurzschluss arbeiten.

EE 81 773 *für weitere Einzelheiten*

Ferrit-Richtungsleiter

The M-O Valve Co. Ltd. Brook Green Works, London, W.6

(*Abbildung Seite 345*)

Die M-O Valve Company hat eine Reihe neuer Ferrit-Richtungsleiter für die Frequenzbänder 5925 ... 6425 MHz, 5925...6175 MHz und 6175...6425 MHz herausgebracht. Der abgebildete Typ CIC4 ist ein Feldverschiebungs-Isolator mit einem maximalen Verlust von 0,35 dB in Durchlassrichtung und einem Mindestverlust von 35 dB in Sperrichtung. Das Stehwellenverhältnis ist 1,02:1. Typen CIC5 und CIC6 sind Resonanzisolatoren mit Durchlass- und Sperrverlusten von mindestens 0,5 dB bzw. maximal 25 dB. Beide haben ein Stehwellenverhältnis von 1,06 : 1. Die HF-Anschlüsse für alle Typen sind Hohlleiter Nr.14.

Diese Elemente eignen sich vor allem für Verwendung in Breitband-Funkverkehrsystemen.

EE 81 774 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

"Laddic" in vollmagnetischen logischen Schaltungen von D. J. Morris

Zusammenfassung des Beitrages auf Seite 290-296

Dieser Beitrag beschreibt Anwendung und Arbeitsweise des "Laddic" in vollmagnetischen logischen Schaltungen, in denen die Verbindung zwischen magnetischen Elementen nur aus Draht besteht. Laddic wird in diesen Schaltungen als Eingangselement benutzt, das die UND-Operation ausführen kann; man kann Laddic aber auch als Gleichstrom-Impulsumsetzer (mit Eingangssignalen in Gleichstromform und Ausgangssignalen in Form sich wiederholender Impulse) einsetzen.

Eine Analyse des Kraftlinien-Schaltmusters im Laddic zusammen mit einem Vorschlag für ein alternatives Kraftlinienmuster wird gegeben.

Ein Tachometer mit kreisförmigen Beugungsgittern von B. V. Jayawant und D. P. Rea

Zusammenfassung des Beitrages auf Seite 297-301

Das beschriebene Instrument arbeitet mit zwei kreisförmigen Beugungsgittern mit genau zentrierten Achsen zur Erzeugung eines Signals, dessen Frequenz der Geschwindigkeit der mit den Gittern gekuppelten Maschine proportional ist. Durch kurzzeitige Abtast-Frequenzmessung kann auf einem Oszillografen eine kontinuierliche Spur dargestellt werden, deren Pegel der Frequenz und damit der Geschwindigkeit proportional ist. Da die Geschwindigkeit sehr schnell gemessen wird, kann das Instrument dem Anlaufverhalten der Maschine folgen und es registrieren, wodurch direktes Untersuchen der Einschwingvorgänge ermöglicht wird.

Feldeffekt-Transistoren in Gleichstromverstärkern von J. C. S. Richards

Zusammenfassung des Beitrages auf Seite 302-305

Sowohl experimentell wie auch theoretisch wurde gefunden, dass der Drainstrom sich durch geeignete Vorspannung temperaturunabhängig machen lässt. Der Einsatz von Feldeffekt-Transistoren als Eintakt-Eingangsstufen für Gleichstromverstärker wird eingehend besprochen. Drifts, die hauptsächlich in Stromversorgungen oder in der zweiten Stufe auftreten, können ohne Schwierigkeiten auf weniger als 50 μV/°C oder 5×10^{-11} A/°C bei 20°C gebracht werden. Schaltungen für zwei typische Verstärker werden gegeben.

Ein logarithmischer Videoverstärker mit kurzer Erholungszeit von G. Kossoff, C. N. Liu und D. E. Robinson

Zusammenfassung des Beitrages auf Seite 306-310

Die Konstruktion eines logarithmischen Videoverstärkers mit kurzer Erholungszeit, der nach dem Prinzip der "Kettengleichrichtung und Addition" arbeitet, wird umrissen. Der Dynamikbereich des Verstärkers ist 75 dB, die Erholungszeit von Übersteuerung 0,15 µs. Die Erholung wird durch Verwendung aufeinanderfolgender Differentialstufen mit Hochfrequenzdioden parallel zu den Anoden erzielt. Die Eigenschaften des Verstärkers mit regelbaren aufeinanderfolgenden Stufen werden besprochen. Schaltungen von 2-MHz- und 8-MHz-Verstärkern werden gegeben.

Die schrittweise Konstruktion eines binären Vor-Rückwärts-Zählers von D. Zissos

Zusammenfassung des Beitrages auf Seite 311-315

Die schrittweise Konstruktion eines galvanisch gekoppelten binären Teilers mit NOR-Moduln wird beschrieben für den Fall, dass das Kompliment des Eingangssignals nicht zur Verfügung steht. Soweit sie nicht kritisch sind, kann man statistische Gefahren in der Schaltung gestatten, jedoch wird ein Verfahren zu ihrer Beseitigung gegeben. Eine Technik zur Vereinfachung von NOR-NOR- (und NAND-NAND-)Schaltungen wird eingeführt. Die Schaltung ist durch die galvanische Kopplung der Moduln von den Anstiegs- und Abfallzeiten des Eingangssignals unabhängig und kann sowohl mit Rechteck- wie auch Sinuswellen arbeiten.

Eine Messtechnik für die Wahrscheinlichkeitsdichtefunktion von P. N. Nikiforuk und G. Squires

Zusammenfassung des Beitrages auf Seite 316-317

Ein Verfahren zum Messen der Wahrscheinlichkeitsdichtefunktion eines beliebigen Signals wird beschrieben. Herkömmliche Funktionsverstärker, ein Oszillator und ein Impulszähler werden für diesen Zweck benutzt.

Arbeitsweise einer magnetfeldgesteuerten Dekatronröhre von A. W. Gooder und M. H. N. Potok

Zusammenfassung des Beitrages auf Seite 318-320

Es wird gezeigt, dass ein Dekatron in Gegenwart eines geeigneten Magnetfeldes erfolgreich durch Spannungsänderungen willkürlicher Form von Gleichstrom bis zu mehreren Kilohertz betrieben werden kann. Die optimalen Schaltungselemente und Spannung wurden versuchsmässig bestimmt.

Ein Effektiv- zu Mittelwertumsetzer von H. H. A. Bath

Zusammenfassung des Beitrages auf Seite 321-323

Das für eine Methode zur Umsetzung einer Effektivspannung mit jedem beliebigen Spitzenwert, der 4 nicht übersteigt, in eine Ausgangsgleichspannung erforderliche Entwurfsverfahren wird besprochen. Eine Anzahl von Dioden wandelt das angelegte Signal so um, dass es in bezug auf einen in einen Kondensator fliessenden Ladestrom einer quadratischen Kennlinie folgt.

Ein transistorisiertes Vorführgerät für quantisierte Impuls-Amplitudenmodulation von J. Karfunkel

Zusammenfassung des Beitrages auf Seite 324-325

Impulsmodulation aller Art mit einer endlichen Impulszahl per Sekunde zeigt Quantisierungseffekte, die in Impulsamplituden- und Impulszeitmodulationssystemen von der Impulsfrequenz und in der Impulskodemodulation von der Anzahl der Stufen abhängig sind. Diese Effekte werden als Quantisierungsgeräusch bezeichnet. Zur Erleichterung des Studiums von Quantisierungsgeräuschen und der Verständlichkeit unter verschiedenen Bedingungen wurde ein Vorführgerät für quantisierte IAM-Signale entwickelt. Ein bereits bestehendes Gerät (1) beruht auf etwa 50 Röhren und Reglern, so dass die Hauptaufgabe in Transistorisieren und Vereinfachen bestand.

Ein einfacher Transistor-Servosimulator von P. G. M. Dawe

Zusammenfassung des Beitrages auf Seite 326-328

Ein hier beschriebener Servosimulator besteht aus einer Anzahl von Einzeltransistorverstärkern mit Parallelgegenkopplung. Diese führen die Funktionen Summieren, Integrieren, Phasenschieben und Phasenumkehren aus. Die Verstärker sind kapazitiv gekoppelt und die Integratoren für quadratische Fehler kompensiert. Das Ansprechen auf einen Sprungeingang bei 50 Hz kann für verschiedene Dämpfungsanordnungen untersucht werden, und eine Klemmdiode zur Dämpfung des Systems während der negativen Halteperiode einer Eingangsrechteckwelle ist vorhanden.

Ein Treppenfunktionsgeber zur Darstellung von Kraftmaschinen-Indikatordiagrammen von A. Watson

Zusammenfassung des Beitrages auf Seite 329-331

Ein Treppengenerator mit kommerziellen Transistor-Binärmoduln wird beschrieben. Der Generator wurde als Zeitbasis für die Untersuchung der Leistung von Verbrennungskraftmaschinen eingesetzt, und es werden verschiedene Anwendungsmöglichkeiten dieser Art erläutert.

Analyse eines Pulsmodulators von J. Kontos

Zusammenfassung des Beitrages auf Seite 332-333

In diesem Beitrag wird ein Pulsmodulator analysiert, der eine Impulsreihe abgibt, deren Folgefrequenz sowie Impulsbreite mit der Eingangsspannung schwanken. Es wird eine Formel abgeleitet, die die Abhängigkeit der Impulsfolgefrequenz von der Eingangsspannung und den Schaltungsparametern gibt. Die Formel wird experimentell durch Vergleich mit berechneten Frequenzwerten getestet, die man von einer für diesen Zweck gebauten Schaltung erhält. Zwischen Versuch und Theorie wurde Übereinstimmung gefunden.

Eliminieren von Teilungsschaltungen durch einen veränderlichen Zeitmassstab von D. K. Taylor

Zusammenfassung des Beitrages auf Seite 334

Die analoge Nachbildung einer Gleichung, die die Division von zwei Variablen umfasst, erforderte bisher Teilungsschaltungen, die eine grosse Fehlerquelle bilden. Es wird gezeigt, dass die Teilungsschaltungen in einigen Fällen durch Einführung einer Zeittransformation in die Gleichungen eliminiert werden können. Dadurch wird eine zusätzliche Gleichung erforderlich, die die Echtzeit als Funktion der Rechnerzeit gibt. Für den Fall einer einfachen linearen Differentialgleichung erster Ordnung wird durch Entfernung der Teilungsschaltung und Einführung der Zeittransformation die Originalgleichung durch zwei gekoppelte lineare Differentialgleichungen erster Ordnung ersetzt.