Electronic Engineering

VOL. 39. No. 473.

COMMENTARY

IN March last year the Ministry of Aviation formally accepted, on behalf of the Government, the recommendations of the second report* recently published of the Burghard Committee, set up in 1963 to codify common standards for electronic parts. (The first report, submitted in 1961, was not implemented for various reasons). In this connexion the phrase 'electronic parts' refers only to 'parts appearing in circuit diagrams' so that, for instance, mechanical structures associated with assembly are outside the Committee's terms of reference. Both active and passive parts are considered.

The Committee's broad objective was to devise a unified system of national standards. The parts considered are classified into three categories: (1) for general application, meeting 60% of Service requirements and 80% of the needs of the capital equipment industry, (2) for special application where some higher assurance of performance is given for the environmental stress levels of the first category and (3) for commercial application in which a lower assurance of performance is given at the environmental stress levels of the first category. The general application category is to be the common standard and it is to this that the Committee has primarily addressed itself and believes that if this category is adequately standardized economic benefits in production and testing will accrue. In short, the proposed common standards should satisfy the majority of Service requirements and the majority of industrial capital equipment requirements.

A consideration of reliability data required by various users has led the Committee to recommend a system of certified reliability data sheets which are to be prepared by manufacturers. This departure from the American MIL method of stratifying reliability into levels or ranges of failure rates has been made for three reasons. The first of these is that the smaller scale of production in the UK mitigates economically against large-scale endurance testing; secondly, stratification does not give complete life-test information and lastly, the Committee's recommendations would give actual test results and flexibility will be greater.

All electronic parts supplied by a manufacturer to common standards are considered to be subject to a qualification approval to show that they meet the relevant specifications and that they can be made in quantity under quality control. The British Standards Institution is to set up an 'Electronics Parts (Inspection) Committee' with overall responsibility for inspection. The actual inspecting authority, whether it be EID or some other body, is to be under Ministerial control as at present.

The British Standards Institution is recommended to manage the proposed system of common standards and to publish general specifications. The pattern of common standard documentation will be in what is termed by the Committee as 'blocks'. Block 1 is a collection of documents relating to subjects which are applicable to several types of part e.g. terminology, testing procedures etc. The material for Block 1 is already largely available in the form of DEF specifications or IEC specifications. Information on specific families of parts such as resistors, will be collected together in Block 2. The third block contains detailed specifications for particular parts ordinarily made by more than one manufacturer and Block 4 will specify the products of particular manufacturers. The final Block 5 contains 'Certified Reliability Data Sheets', prepared by a manufacturer and recording the results of the relevant tests of Blocks 3 and 4. The Committee recommend that these c.r.d.s be prepared every six months, mandatorily for parts in the general and special categories, and optionally for those in the commercial category.

To summarize: there will be a system of common standards within the general application category of parts, the technical content of which 'should conform with IEC recommendations or with Nato basic specifications'. All parts made under the scheme will be certified to conform to specifications and test results will be accumulated and issued by manufacturers as certified test records. The consequence of the new system will be the gradual replacement of a number of DEF and electronic valve and semiconduction diode (CV) specifications and existing British Standards.

With regard to the time-scale of the recommendations, the Committee indicated that within a year of acceptance of the report (which occurred in March 1966), the system should be in operation. In fact, the British Standards Institution has been involved in the work since March last year and specifications have already been prepared by technical committees of the Institution. Three other committees, with suitable representation, have been organized to co-ordinate and oversee the work.

The Ministry of Technology, the industry and the B.S.I. have undertaken to propagate abroad the ideas contained in the Burghard report so that other countries may participate in the scheme. The sting here is in the tail—if other countries are persuaded to take part then obviously compromises on detailed specifications will almost surely have to be made to get agreement; it then behoves the Ministry of Technology to ensure that any such compromises do not in their turn compromise UK component manufacturers.

* *Ministry of Technology: Second Report of the Committee on Common Standards for Electronic Parts. 1967, 36 pp. Price 6s. 6d. from H.M.S.O.*

All-Magnetic Circuits Incorporating New Coupling-Loop Configurations

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All-magnetic logic circuits, consisting of only multi-aperture devices and their connecting wires, have usually required a specially shaped device for each logic function or for a set of logic functions. This article describes how a single type of device can be used to perform the various logic functions, such as AND, OR and INHIBIT, by suitable arrangement of the coupling loops which connect the devices.

(Voir page 475 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 476)

ALL-MAGNETIC logic circuits are being used in systems where high reliability and long life are required. In this type of application, only the magnetic devices and their connecting wires are used within the logic circuits. One of the important applications of all-magnetic circuits is in fail-safe logic control such as nuclear-reactor safety control, chemical-plant safety control, etc.

The magnetic devices for these applications are of ferrite materials which have approximately rectangular hysteresis properties. These devices originally had only two apertures and were termed Transfluxors[1]. However, later, with the addition of several apertures arranged in various geometrical configurations, they were renamed multi-aperture devices. The logic capabilities of the devices come from the fact that it is possible to control the flux changes round specific apertures. A further feature is that isolation can be provided between the input and output circuits. Hence connexions between devices may be of wire only, making other components unnecessary. The system used for coupling the devices is known as MAD-R[3,4], because it uses only multi-aperture devices and connecting wires, where the value of the wire resistance is of great importance.

All-magnetic logic circuits have usually required a specially shaped device for each logic function[4]. Special devices have been designed for AND, OR negation etc. It has also been found that the simpler the shape of the device the better its operation, i.e. it provides a larger output signal/noise ratio and also allows a wider current range. More flux paths within the device increase the logic capabilities, but also reduce the operating tolerances. For this reason, devices with not more than two input or two output apertures are used, e.g. the AND device and the FAN-OUT device, respectively. Furthermore, simple devices are easier to manufacture and to wire. The design of magnetic circuits is best accomplished with a minimum number of differently shaped devices, and performing all logic functions with these devices. Most, but not all, logic functions can be achieved with two or three basic devices, and further techniques are required if no extra ones are to be introduced. These techniques use special wiring connexions between the devices. This article describes how the three basic logic functions can be achieved by using only one type of device configuration throughout the system.

The basic shaped multi-aperture device is that which performs the transfer operation. This can give destructive read-out pulses with a signal/noise ratio of over 20/1 and a 'prime' operating range of at least 4/1 (at room temperature). Using a number of devices of this shape, one can perform the logic functions of AND, OR and INHIBIT by wiring them together in a special manner. This method of coupling the devices reduces the operating tolerances, and hence greater care must be taken when designing them in circuits.

The transfer device was selected for its simplicity and its wide operating tolerances. However, one must add that the same coupling techniques described here may be applied to any shape of multi-aperture device.

The performance of the transfer device is described first and then its operation in the basic all-magnetic logic circuit. It is then shown how a number of these devices may be connected together.

The Transfer Device

The transfer device, which is normally used in shift registers, has three apertures: two small ones for input and output, and a large one used for controlling the operation. The device is shown in Fig. 1, which illustrates four distinct 'legs': legs 1 and 2 round the input aperture and legs 3 and 4 round the output aperture. Also illustrated are the windings required to operate the device, namely: Set, Block, Prime and Output.

The operation of this device will be explained with the aid of the flux patterns shown in Fig. 2. The flux lines will always remain in their given pattern provided no external field is applied.

In the initial state of the device, shown in Fig. 2(a), the flux round the large aperture is in a clockwise direction,

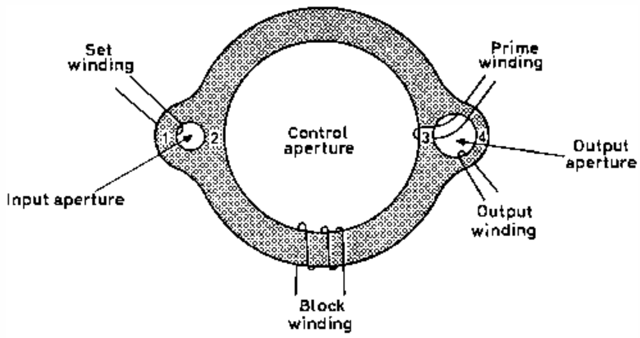

Fig. 1 The transfer device

i.e. in an upward direction in legs 1 and 2 and in a downward direction in legs 3 and 4. This initial state is established by the application of a current in the Block winding, of sufficient magnitude to switch magnetically all the material round the large aperture.

Consider now the effect of a current in the Prime winding round the output aperture, while the device is in its

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initial Block state. This current must be large enough to produce a field greater than the coercive force of the material round the small output aperture but not round the large aperture. This field is induced in a direction which tends to switch the flux round the output aperture in a counterclockwise direction, i.e. it tries to switch leg 3 in a downward direction and leg 4 in an upward direction. Since the flux in leg 3 is already saturated in a downward direction, as seen in Fig. 2(a), the Prime field will only drive it further into saturation. As it is impossible to reverse the flux in a path if any part of it is already saturated in the direction of the induced field, then the Prime field cannot switch any flux round the output aperture. It is for this reason the initial state is called 'blocked' since no output can be induced in the output winding.

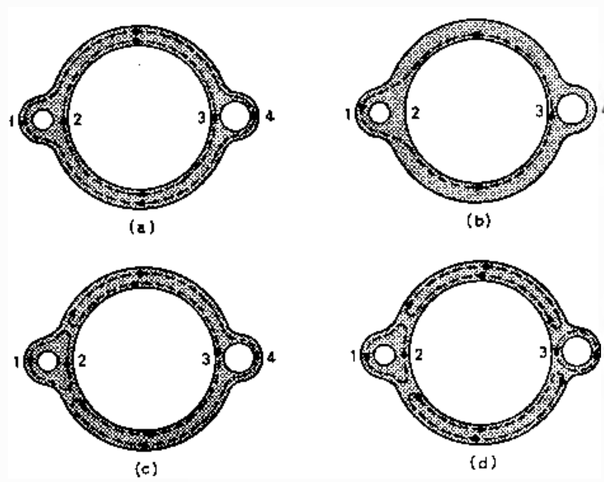


Fig. 2 (a) and (b) Flux-switching path during the input-set operation
(c) and (d) Flux patterns of the transfer device

A counterclockwise-direction input field, produced by a current in the Set winding, cannot switch flux round the input aperture. As in the case of the Prime field, the flux in leg 2 is already saturated in the direction of the induced field. If, however, the input field is large enough then it can switch the flux round the larger aperture along a path shown in Fig. 2(b). Flux switches along its shortest available path. For a field applied to leg 1, the shortest path is leg 2, but since this is not available for switching it will seek another available path. If the input field is large enough the path presented by leg 3 is then the shortest available path. Leg 4 also presents an available return path but it is not the shortest. After the input field is removed there will be some further domain rotation round the boundaries of the path, so as to preserve flux continuity through the device. The flux pattern after the removal of this input field is shown in Fig. 2(c); this is termed the Set state.

Once the device is in the Set state, the Prime field applied in the output small aperture is able to switch the flux round it. The flux round the output aperture, while the device is in the Set state, is in a clockwise direction. Thus, when the Prime field is produced it will switch the flux round the output aperture into a counterclockwise direction, as shown in Fig. 2(d). As the flux switches in leg 4 of the device, a voltage pulse is induced in the output winding.

When the next Block pulse is applied, it switches the device from its Primed state back into its Blocked state, inducing a second voltage pulse in the output winding, but of opposite polarity to that produced by the Prime field.

This Prime field must be limited to affect the output aperture only, and its magnitude must not be large enough to 'set' the device. To increase the allowable operational range of the Prime field, the Prime winding is shared between legs 3 and 4. This is possible because a Prime field applied to leg 3 alone cannot set the device, when it is in its Block state, as it only drives the flux in that leg further into saturation. Since the Prime operation cannot take place until after the device is set, then a d.c. Prime field may be used, thus reducing the requirement of an extra clock pulse.

The best practical results are obtained from a device in which the cross-sectional area of legs 1 and 2 is larger than that of 3 and 4. This makes the diameter of the input aperture smaller than the diameter of the output aperture. The effects of this have been described in a previous publication[3].

The All-Magnetic Transfer Operation

Within the Transfer device there are two main switching paths, one along legs 1 and 3 and the other along legs 3 and 4. With these two paths, it is possible to achieve the isolation of the input circuit from the output circuit. When the device is being set, the input switching path is round the large aperture along legs 1 and 3, whereas the output winding on leg 4 may be short-circuited without affecting the input operation. From this it is clear that the output leg can be heavily loaded by a very low-impedance coupling loop, without this impedance being reflected into the input circuit. The basic all-magnetic circuit, using only transfer devices and their coupling loops, is shown in Fig. 3.

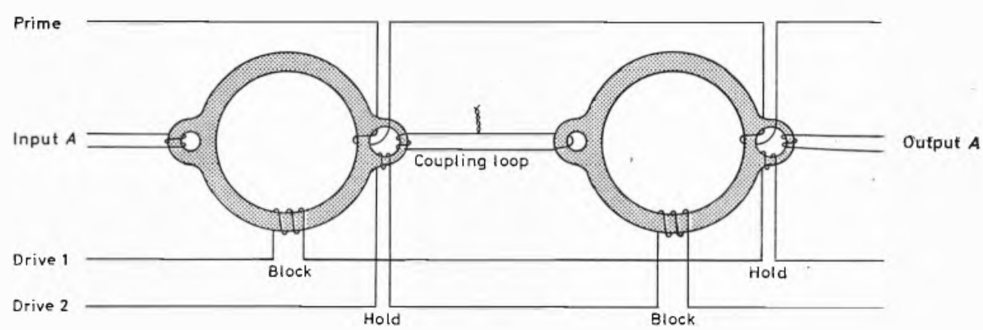


Fig. 3 Transfer operation

Although the low-impedance coupling loop, assembled on leg 4, does not affect the input operation, it does affect the Prime operation. During this operation, there will be flux reversal in leg 4 (as it switches from a downward to an upward direction), which produces a voltage output pulse and consequently causes a current to flow in the coupling loop. The coupling loop connecting the two devices is wired in such a way that the current caused by the Prime operation, will flow in a direction which tends to drive leg 1 of the following device further into saturation. That is, this current will not be able to cause any flux switching in the following device, and therefore, during the Prime operation, the following relationship holds in the coupling loop:

$$N_t\ (d\Phi_4/dt) + IR = 0 \quad \ldots\ldots\ldots\ldots \quad (1)$$

where the flux change in leg 4 is represented by $d\Phi_4$, N_t is the number of turns on the output winding, I is the current in the loop and R is the resistance of the coupling-loop wire.

During the Prime operation, the coupling loop acts as a partial short circuit on leg 4 of the first device. From equation (1) it can be seen that the resistance must be greater than zero to permit any flux switching to occur in leg 4. Unfortunately, as will be shown later, the wire resistance must be low enough to allow sufficient transfer of energy from the first device to the second device. In fact the wire resistance is of the order of milliohms. As explained in the previous section, the magnitude of the Prime field is limited, and thus the current in the loop is small. From this it can be seen that the switching time of the Prime operation, in order to produce a flux reversal in leg 4, will be of a relatively long duration, in the range of 100μs.

From equation (4) it can be seen that a high flux gain will be achieved if the wire resistance is as small as possible, with the ideal gain obtained for $R = 0$.

The choice of a resistance value for the connecting wire is a compromise between the conflicting requirements of the Prime and the transfer operations; if R is too large no transfer is possible and if it is too small the Prime duration is too long. The optimum value of resistance is only a fraction of an ohm, obtained in practice by a suitable choice of length and gauge of the coupling loop wire.

As the first device is switched back to its blocked state, the second device switches to its set state, i.e., the information stored in the first device is transferred to the second device. Since a device cannot receive and transmit information at the same time, i.e. switch both to the Set and to the Block states; thus two clock drive pulse systems must be used. Drive 1 blocks the first device and sets the second device and drive 2 blocks the second device. The full operation of the transfer circuit can be followed from Fig. 4, which shows the clock pulses and the directions of the flux in the four legs of both devices at each operation stage.

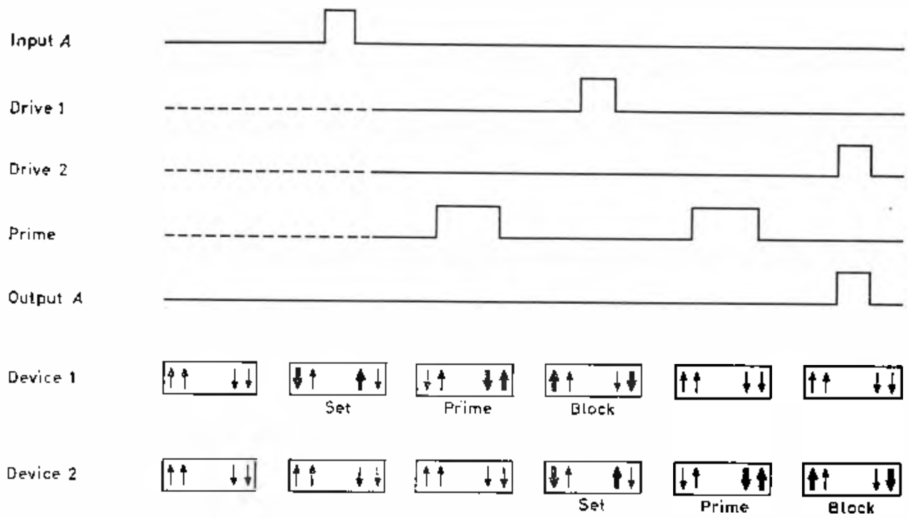


Fig. 4 Transfer timing diagram

Once the device is in its Prime state, the application of a current pulse in the drive winding round the large aperture will reset the device back to its Block state. This operation will reverse the flux in leg 4 and thus induce an output voltage pulse in the coupling loop of an opposite polarity to that induced by the Prime operation. The pulse will produce a current in the loop which will now be in a direction to switch the following device into its set state. Since leg 1 of the following device switches, the coupling loop does not appear as a short circuit during this operation. The driving operation, when both devices switch, is characterised by the following:

$$N_t\,(d\Phi_4/dt) + IR + N_r\,(d\Phi_1/dt) = 0 \quad \ldots\ldots \quad (2)$$

where N_r is the number of turns on the input winding of the second device, and $d\Phi_1$ is the flux change in leg 1. This equation can be written as:

$$N_t\,\Delta\,\Phi_4 + N_r\,\Delta\,\Phi_1 + R\,Idt = 0 \quad \ldots\ldots\ldots \quad (3)$$

In order that the first device will set the second device, one must achieve what is known as 'flux gain', i.e.

$$\left|\frac{\Delta\Phi_1}{\Delta\Phi_4}\right| = (N_t/N_r) - (R/N_r)\left|\frac{\int Idt}{\Delta\Phi_4}\right| \ldots\ldots \quad (4)$$

It can be seen from Fig. 2 and 4 that, as the device switches from the Prime state to the Block state, there will be flux switching not only in leg 4 (output leg) but also in leg 1 (input leg). The change of flux in leg 1 of the second device will cause a current to flow in the coupling loop in a direction which will tend to set the first device. This effect, which is known as 'back flow of information', must be prevented by holding the flux in leg 4 of the first device in its upward direction, while the second device is driven. This holding operation is accomplished by wiring the hold winding of one device in series with the block winding of the following device, as shown in Fig. 3.

In an all-magnetic network the information is processed and transmitted step-by-step from device to device. Two clock pulses are used, each of which drives alternate devices and holds the output leg of the previous device in a specific direction.

The OR Operation

The OR operation is obtained by feeding information from a number of input devices into one output device, and if any one of the input devices switches then the output device is set. If the input devices receive information A, B, C, then the output device will represent the function $(A + B + C)$.

In a previous article[4] the author stated that it is impossible to feed information from a number of coupling loops into one input aperture, in order to achieve the OR operation. This is because a second coupling loop would load only leg 1 of the output device and prevent it from switching. It was explained that when setting a device, the low impedance load on the output circuit is isolated from the input circuit. If two or more loops couple into one input leg, then they are not isolated from each other, but load the input leg.

It was shown that when transferring information from one device to another with a single coupling loop, equation (2) holds. If a second coupling loop is placed round leg 1 of the output device, a current I_2 must flow in this second loop. This is on the assumption that leg 1 can be switched by the current I_1 from the first coupling loop, as shown in equation (2). Also, if no other voltage pulse is induced in the second coupling loop, then the second loop is characterised by

$$I_2 R + N_r (d\Phi_1/dt) = 0 \quad \ldots\ldots\ldots\ldots (5)$$

The resistances of both coupling loops in equations (2) and (5) are approximately equal, as is the rate of change of flux in leg 1, although the currents, I_1 and I_2, may be different. By comparing equations (2) and (5), it can be seen that the flux in leg 1 of the output device will be prevented from switching by this extra low-impedance load, and hence no transfer of information is possible. In fact, leg 1, during the input switching, can be regarded as a transformer, where the second coupling acts as a short circuit.

In the previous article[4] the author suggested using a separate input aperture for each coupling loop, i.e. having a device with a number of input apertures, one output aperture and one large control aperture. Although this produces the best results, it does, however, require a specially shaped device to produce the OR function. Instead, the OR function can be obtained by using the transfer device, with only one loop, coupling all the input devices with the output device, as shown in Fig. 5. As only one coupling loop is connected to the input aperture of the final device, there is no problem of loading this device during the input operation.

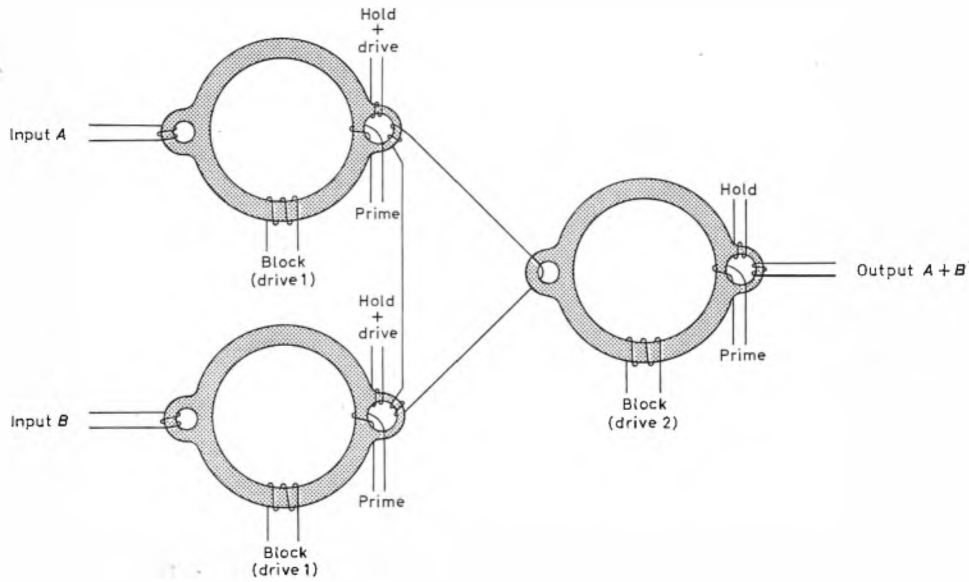


Fig. 5 OR operation

As with the transfer circuit, two clock pulses are used. The two input devices are both blocked at drive 1 time and hence they both may produce an output pulse at the same time. The final device is thus set at the time that drive 1 occurs and is reset by drive 2, with the final OR output being produced at the time that drive 2 occurs.

If only the input device A had been set, then when drive 1 occurs, a voltage pulse would be produced in the output winding of device A but not from device B. The voltage pulse in A induces a current in the coupling loop in a direction which sets the final device. Since one loop couples A and B with the final device, this current also flows in the winding round the output aperture of B. However, there will be no flux changes in B because the field produced by this current tends to block it although it is already in its blocked state. Thus, the output from A only sets the final device, without affecting B.

If both input devices A and B are set, then they will both produce an output voltage as they switch back to the block state by drive 1 pulses. The output voltage pulses from both devices add together in the coupling loop and produce a current which sets the final device.

Thus the final device may be set by any one of three possible input combinations; from switching of A alone, B alone, or from the switching of both input devices A and B together. Hence, leg 1 of the final device switches according to one of the three following equations:

$$N_t (d\Phi_4/dt)_A + I_1 R + N_r (d\Phi_1/dt) = 0 \quad \ldots\ldots (6)$$

$$N_t (d\Phi_4/dt)_B + I_1 R + N_r (d\Phi_1/dt) = 0 \quad \ldots\ldots (7)$$

$$N_t (d\Phi_4/dt)_A + N_t (d\Phi_4/dt)_B + I_3 R + N_r (d\Phi_1/dt) = 0 \ldots (8)$$

where $I_1 < I_3$.

The resistance in all three functions is the same as it refers to the same coupling loop. The currents in equations (6) and (7) are equal but are smaller than that of equation (8). However, this is irrelevant to the operation of the final device, providing that I_1 produces a field larger than the coercive force of the material round the large aperture.

The reason that the extra current I_3, given in equation (8), does not affect the operation is that it is impossible to 'over set' a device. The amount of material that switches is controlled by the minimum cross-sectional area of the path. Any increase of the input field greater than the magnitude of the coercive force will not widen the path, but will reduce the switching time. Hence, when both input

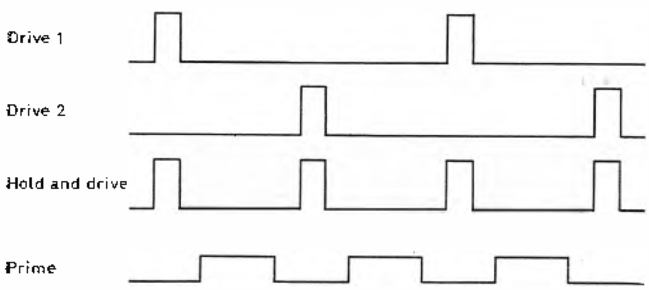


Fig. 6 Modified timing diagram for the OR operation

devices A and B produce an output, the final device will merely switch to its set state faster than if only A or B produce an output.

By having just one coupling loop connecting several devices, inevitably the length of the loop is increased, and hence the resistance of the wire increases. The value of the wire resistance controls the current in the loop and hence, with only one input device producing a pulse, the wire resistance could be too high and not enough current available to set the final device. In order to keep the resistance down, the wire gauge can be increased, but it is not always possible to achieve the optimum resistance value because of the finite diameter of the aperture.

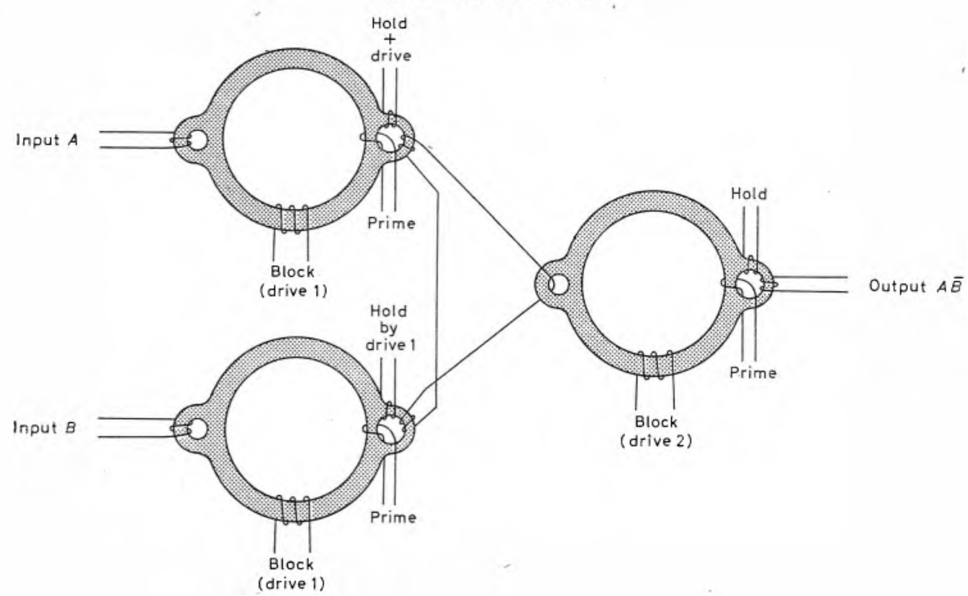
If the resistance of the wire cannot be decreased to its optimum value, it is possible to compensate the loss of energy across the resistance by increasing the voltage induced from the input devices. This can be done by increasing the magnitude of the reset field applied when switching the device to its blocked state. In all-magnetic networks, involving a large number of devices, the drive current usually feeds all of them in series. The magnitude of the reset field is therefore determined by the number of turns in the Block winding on each input device of the OR circuit and it could be increased by adding more turns.

A more practical method of increasing the output voltage induced in the coupling loop is to increase the magnitude of the field switching leg 4 of the input device. The field switching the flux in leg 4 (the output leg) is produced there indirectly from a current in the Block winding round the large aperture. In order to increase the energy available for switching leg 4 the field should be produced directly in the leg by passing an extra current pulse to the winding round leg 4 at the same time as the current pulse is passed in the Block winding. Both these fields assist each other in the leg as they will both tend to switch leg 4 in a downward direction. It should be noted that this extra driving field is also in the same direction as the hold field, applied later in order to prevent back flow of information. The hold winding may then be used for both operations, first to hold the input device while the final device is driven and secondly to assist the driving of the input device. In other words, both drive 1 and drive 2 are applied to the hold winding, as can be seen from the modified timing diagram shown in Fig. 6. A more detailed analysis of the effect of the drive and hold fields on leg 4 of the output aperture was given in a previous publication[3].

Although the OR circuit gives satisfactory results and has proved its usefulness in a number of practical applications, it has a number of possible drawbracks to be overcome. The value of the coupling loop resistance must be as close as possible to its optimum value. As more inputs are required for the OR operation, i.e. more input devices are added to the OR circuit, the length of the coupling loop, and thus its resistance, may increase to a point where no transfer of information is possible. So that, despite the compensating methods described before, the value of resistance may be too high to permit reliable operation and the number of input devices coupled together must be limited to two or three.

When an input device is in its blocked state, the two driving reset fields on the large aperture and the output small aperture tend to drive leg 4 further into saturation but do not cause any remanent switching (reset switching occurs only after the device has been primed). However, there will be some elastic switching caused by the reversible

Fig. 7 INHIBIT operation



domain rotation. Elastic switching will produce a relatively small voltage pulse in the output winding, i.e. noise, which is larger if leg 4 is driven as well as the large aperture. When the drive current is applied to the large aperture alone most of the elastic switching is in leg 3 but by driving leg 4 as well as the large aperture, most of the noise is produced in leg 4. The ratio of the signal output voltage to the noise output voltage from a single device is generally high. However, when a number of input devices are coupled in series, the signal/noise ratio is reduced and the sum of the noise outputs from all the devices may be large enough to cause setting or partial setting of the final device. This is known as 'zero build up into one', where a 'zero' input transmitted from a given device is received by the following device as a 'one'. This effect must be prevented by limiting the number coupled together and also by limiting the extra drive current applied to leg 4. These two precautions will keep the accumulated noise output pulses to a minimum.

The INHIBIT Operation

The INHIBIT circuit is one that produces an output if there is no inhibit pulse and the input pulse is present. If the input pulse is represented by A, the INHIBIT pulse by B, then the output pulse from the circuit is represented by the function $A\overline{B}$.

The method by which the OR function was obtained may be used to obtain the INHIBIT function, that is, using one loop to couple the input devices to the final output device. The whole circuit is similar except that the output winding on the INHIBIT device is reversed, as shown in Fig. 7.

The wiring of the coupling loop is such that any voltages induced in the loop by the two input devices will be of opposite polarity. In order that this circuit may be effective for the INHIBIT operation, the voltages induced by the devices should be of equal magnitude. Hence

$$(d\Phi_4/dt)_A = -(d\Phi_4/dt)_B \quad \ldots\ldots\ldots\ldots \quad (10)$$

Thus, if both A and B are switched to their Blocked state and consequently induce two voltage pulses into the coupling loop, the pulses will tend to cancel each other and no current will be available for setting the final device. If B alone induces an output voltage pulse in the coupling loop, the current produced will tend to drive the final device into its blocked state; since the final device is already in its blocked state, no remanent switching will occur. If device A alone induces an output voltage pulse, then the current in the coupling loop will be able to set the final device.

As with the OR gate, it is advisable to increase the magnitude of the field switching the flux in leg 4 of device A. Thus, this leg is supplied both with drive 1 and drive 2, first to assist the switching and then to prevent the back flow of information.

When A switches alone, the current produced by it tends not only to switch the final device but also to switch the INHIBIT device. In the OR circuit, there is no such danger of setting one input device by the other, but in this circuit the output winding on the INHIBIT device is reversed and the current now flows in such a direction as to set it. To prevent this, leg 4 of the INHIBIT device B must be held while A is switched to its blocked state by drive 1. On the other hand, there is no danger of back flow of information from the final device to the INHIBIT device, and there is no need to hold the output leg of B, while the final device is driven, as was done with A.

It has been stated that the output voltage from device B produces a current in the loop in the direction required to block the final device, but, since the INHIBIT output winding is reversed, this current is also in a direction required to switch A to its set state. Thus, when B switches alone, leg 4 of device A must be prevented from switching and held in its blocked state. The drive pulses applied to the 'Hold' winding on leg 4 of device A have therefore three functions; (a) to assist the transfer of information from A to the final device at drive 1 time, (b) to prevent back flow of information when the final device is reset at drive 2 time and (c) to prevent the output of device B switching A at drive 1 time.

Any number of devices may be wired in series by various coupling loop combinations giving a wide variety of logic functions such as $(A + B)\overline{C}$, $\overline{(A + B)}C$, $AB\overline{C}$, etc. However, care must be taken that the INHIBIT voltage output is always large enough to prevent the switching of the final

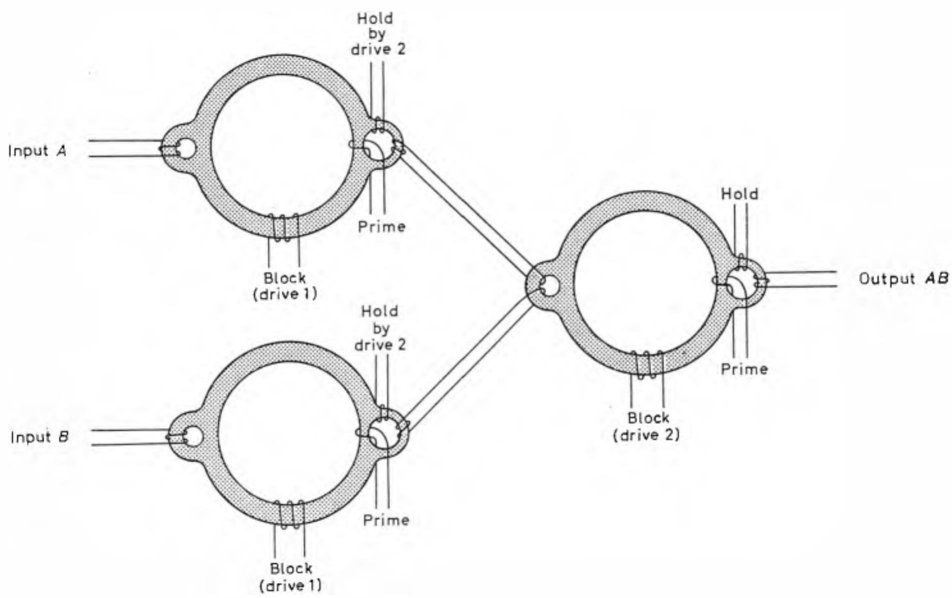


Fig. 8 AND operation

device. The resistance of the wire should also not be increased above the point where no transfer of information is possible.

The AND Operation

To perform AND operation a circuit must produce an output only when all its input pulses are present. If the input pulses are represented by A, B, C, then the output pulse from the circuit will represent the function ABC.

When discussing the OR function it was stated that it is not possible to obtain OR operation by assembling two separate coupling loops on one input aperture. It was also shown that by inserting another coupling loop, the final device is prevented from switching into its Set state. However, this effect, which is a disadvantage in the OR operation, is an advantage for the AND operation.

Two input devices A and B are each connected, *via* a separate coupling loop, to leg 1 of the final device (Fig. 8). If only one input device switches, the second coupling loop presents a partial short circuit, which prevents the final device from setting. If both devices switch, then the two fields produced in the final device by the currents in both coupling loops will add up and switch the final device into its set state.

Since the final device is set by the current supplied from the two input devices, the current produced by each of the two may be less than that required in a single transfer operation. Thus, leg 4 of each input device, need not be driven when they are reset to its blocked state. In the OR operation this was essential for the performance, while here it is irrelevant, although it will be shown later that driving leg 4 is not optional but in fact prohibited. If so, leg 4 is not driven by drive 1 but only held by drive 2 to prevent back flow of information.

When one input device switches alone, the equation for the second coupling loop takes the form of equation (5). Although the wire resistance R is of the order of milliohms, there may be some switching in the final device if enough time is given, as in the case of the Prime operation. Fortunately, such switching is unlikely, as the duration of the output voltage pulse induced from the input device is too short. There is, however, another risk which could be of a more critical nature.

In the Prime operation the magnitude of the field is limited, but in the Block operation magnitude of the drive field is not limited. If the energy transmitted from a single input device is very large, the current in the first loop will also be large. In this undesired case, the excess current can cause some switching in the final device, in spite of the latter being heavily loaded by another coupling loop. The only way of preventing this is by reducing the fields applied to the input devices when driving them to their blocked state. It can be seen that leg 4 of the input device must not be driven while the device is being blocked, and the driving current applied to the large aperture should be decreased by reducing the number of turns on the reset winding. The decrease of current applied to the input devices will not affect the operation as the energy required from each device is less than in a normal transfer operation.

Conclusions

Ideally, in all magnetic logic circuits, a specially shaped device is required for each logical operation. However, if a large number of differently shaped devices are required, then the finished unit would be expensive and not all the required shapes are available.

It has been shown how various logical operations can be performed using one particular shaped device throughout a system with special manipulation of the coupling loops. The number of logic functions is not limited to those suggested, as further manipulation of the coupling loops will enable more complex operations to be performed. Again these coupling loop configurations are not limited to one particular device but could be used with any other shape.

The manipulation of the coupling loops to obtain the same logical functions that could have been obtained from a specially shaped device is achieved at the expense of allowed operational tolerances of the current pulses. Despite this drawback this system can produce good results with reasonable tolerances.

The disadvantages of this system, mainly applying when three or more input devices are used, can be summed up as follows: in the OR circuit there is a possibility that the sum of the noise outputs from all the input devices will be large enough to set the final device; in the INHIBIT circuit there is a possibility that the inhibit outputs will not be large enough to prevent the switching of the final device; and in the AND circuit there is a possibility that the energy transmitted from one input device will be large enough to set the final device although it is heavily loaded by another coupling loop.

Acknowledgments

The author wishes to thank the English Electric Company Limited for permission and facilities to publish this article.

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Radio Control for Overhead Travelling Crane

A Standard-Telemotive remote-control system has been applied to an overhead travelling crane recently commissioned for the seamless-tube-making plant at Stewarts & Lloyds, Corby, Northants. as a production experiment. This is the first application of this type of radio-control equipment to a crane in this country. The system gives full remote control of the crane by means of radio transmission from floor level. A single operator wears a lightweight unit, housing controls which are replicas of the usual cab controls, to actuate the motions by means of appropriate contactors.

Advantages of the system include freedom of movement of the operator at floor level, the absence of pendants, the elimination of hand signals and immediate selection of loads. The system, which is well established in the USA, is immune to interference and several systems can work in the same area without difficulty. Installation time is minimal, and extensive use in the USA with many hundreds of cranes has shown that maintenance time has been considerably reduced.

The system is applicable to all electrically operated overhead travelling cranes—either new or existing—and the necessary capital investment can be fully recovered within eighteen months of normal regular use.

A Variable-Ratio Frequency Divider Using Micrologic Elements

By S. Jannazzo* and G. Rustichelli*

The detailed design of a variable-ratio frequency divider is described and particular stress placed on the use of the RTµL family of integrated circuits. The circuit allows full exploitation of the speed characteristics of this family and the simple extension of the number of division ratios.

(Voir page 475 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 476)

MODERN systems for frequency synthesis are usually designed according to the block diagram in Fig. 1. A voltage-controlled oscillator (block 1) drives a variable-ratio frequency divider (block 3) which can be preceded by a fixed divider (block 2). The output of the variable-ratio frequency divider is compared with a reference frequency by means of a phase and/or frequency discriminator (block 4).

The reference frequency is obtained from a quartz oscillator (block 5) which is generally followed by a fixed divider (block 6). The phase discriminator produces an error voltage used to control the v.c.o. It is evident that the v.c.o. output frequency depends, in steady-state conditions, upon the ratio of the variable divider. In consequence, the system allows the selection, within a certain range, of a number of frequencies equal to the number of frequency ratios that can be set on the variable-ratio divider. The accuracy and stability of these frequencies (apart from secondary effects which, within wide limits, can be minimized at will) is a function only of the accuracy and stability of the quartz oscillator.

This article describes the design of the variable-ratio frequency divider which is the most important part of the system. Division is obtained by means of digital dividers using integrated circuits.

Design

General Considerations

A variable-ratio frequency divider which can be used for the above-mentioned purposes, must have the following characteristics, listed in order of importance:

(a) *high speed.* It is, in fact, obvious that the higher the speed of the divider, the higher the maximum frequency that can be synthesized. This point will be considered further in the next section;

(b) *simple circuitry for setting division ratio.* This characteristic is not of paramount importance from the user's point of view, but a simple solution to this problem allows a decrease in internal delays and, in consequence, increases the circuit speed.

(c) *facilities for simple extension of number of division ratios.* This characteristic increases circuit flexibility, permitting its use in a wide range of different systems.

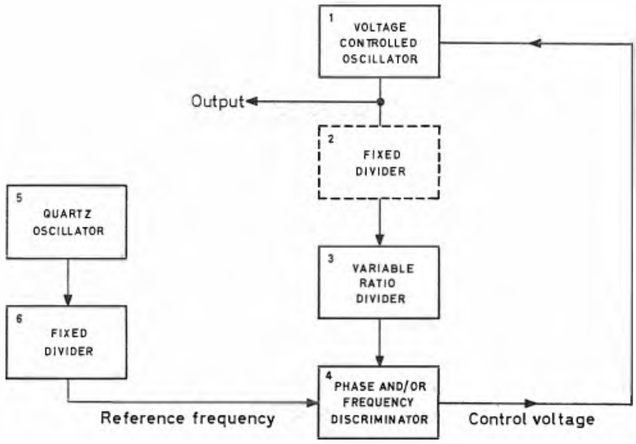
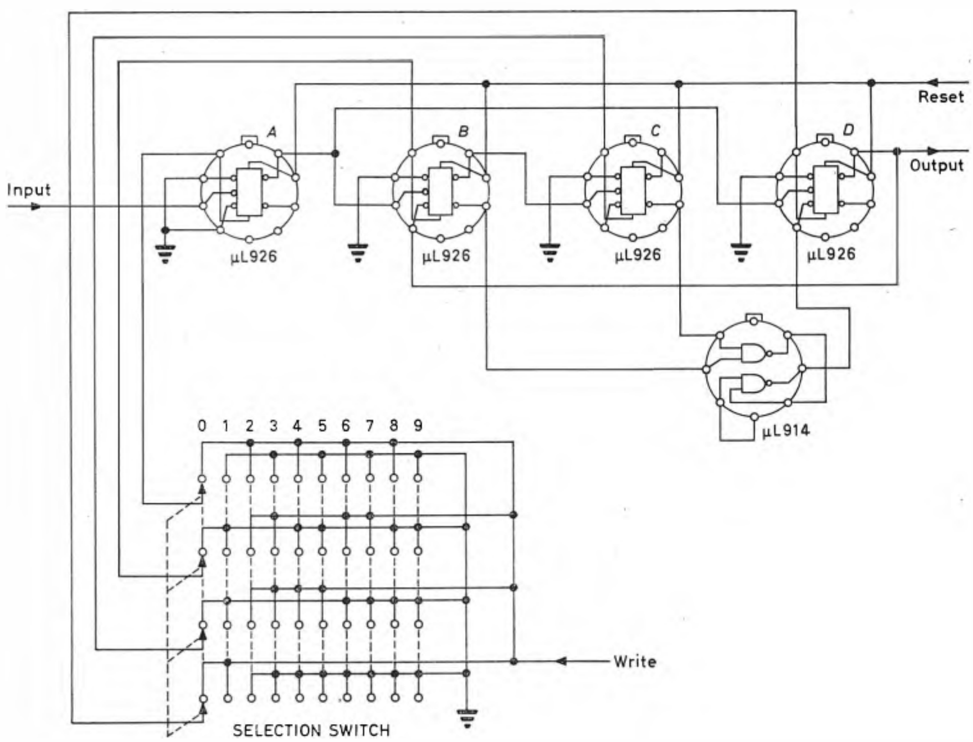


Fig. 1 Frequency synthesizer block diagram

Fig. 2 Decade and selection switch schematic diagram



* *SGS-Fairchild.*

If, as is usually required, the 'channel' division ratio must be selected in decimal form, the most convenient division scheme is that using cascaded decades. This solution enables the division ratio to be programmed by acting separately on each decade, thus avoiding the need of a carry from one decade to another. Furthermore, the extension of the number of channels is obtained simply by adding the necessary decades and a few additional circuits.

The Circuit

Division Ratio Selection

Assuming the use of a convenient number of cascaded decades, the problem concerning the setting of the division ratio desired will now be considered. The solution adopted in the present design is initially to set in the divider a

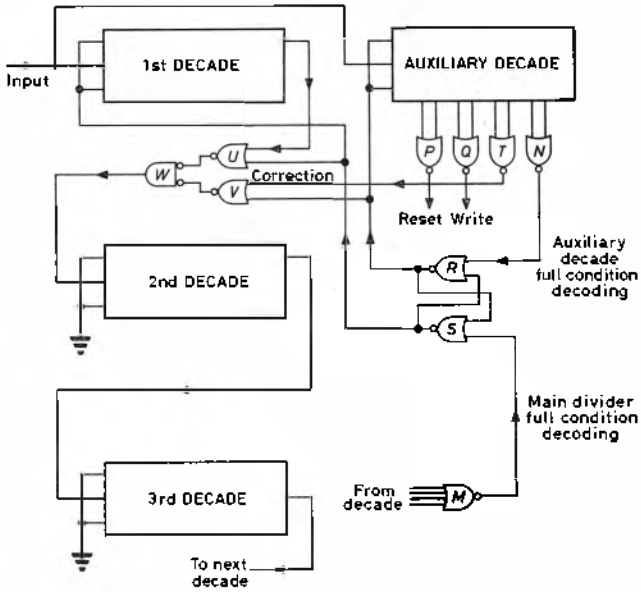
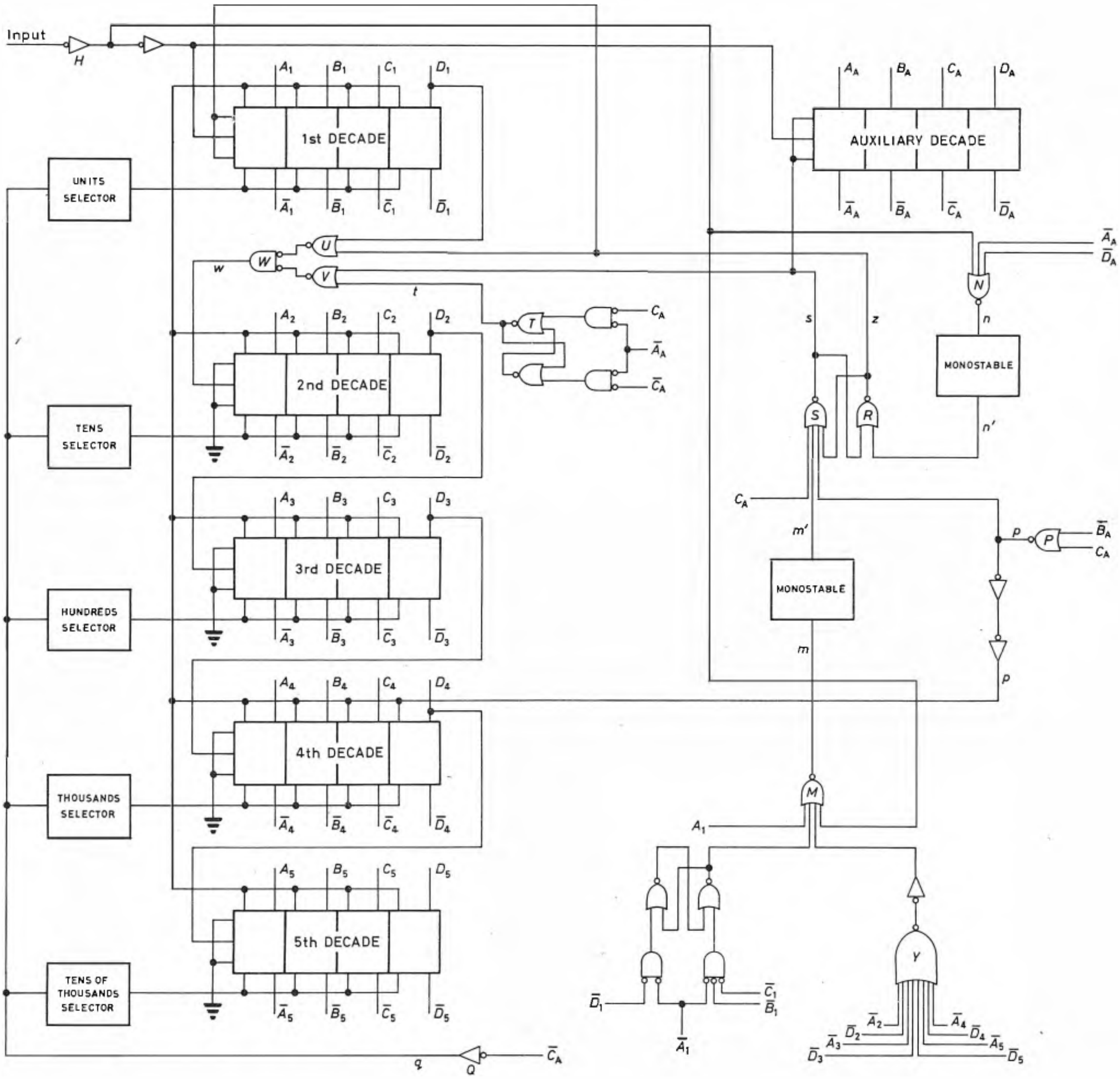


Fig. 3 (right) Divider block diagram

Fig. 4 (below) Five-decade divider block diagram



number equal to the complement of the desired number with respect to the maximum content of the whole divider. Thus, if n is the number of cascaded decades and x the desired division ratio, the number y, necessary to set in the divider, is given by:

$$y = 10^{n} - 1 - x$$

Clearly, once y has been set the number of pulses necessary to fill the divider completely, i.e. to reach the contents $10^{n}-1$, is exactly equal to x. It is, therefore, only sufficient to note the condition in which the divider is completely full, to be sure that x pulses have been inserted. Repeating at each cycle both the setting of y and the reading of the full condition one can obtain the desired division of the input frequency by the number x.

instance, if the switch is positioned at 7, the writing pulse sets the decade as follows:

$$A = 0,\ B = 1,\ C = 0,\ D = 0$$

which corresponds to the decimal number 2 (nine's complement of 7).

An Artifice to Avoid Limitation of Circuit Speed

In order to ensure that the divider speed is not limited by the operation of setting the number y, this operation must be completed within the time elapsing between the active edge of the pulse that causes the divider to reach the full condition, and the active edge of the following pulse, i.e. within the period of the input frequency. It is difficult to fulfil this condition at high frequencies. For instance, when dealing with RTμL circuits, since the μL 926

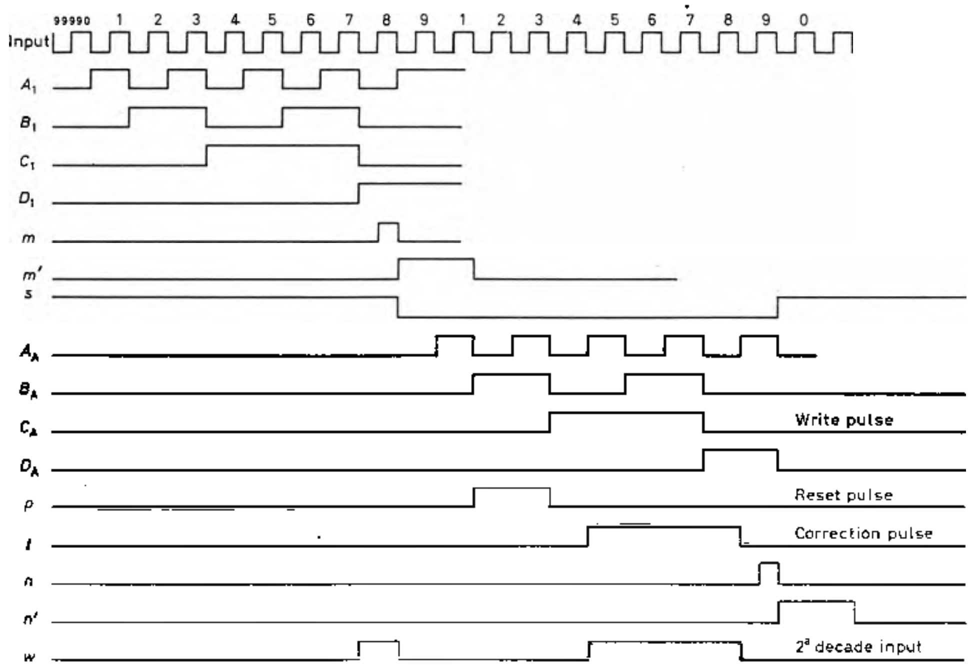


Fig. 5 Timing diagram

For instance to design a divider with a ratio selected between 1 and 999, there will be three cascaded decades. For a division ratio of x, the number to be set on the divider is:

$$y = 999 - x$$

Assume, for instance, that $x = 347$. Therefore $y = 652$. It will consequently be sufficient to decide, by suitable decoding, the condition in which the divider is full, i.e. 999. The number of pulses necessary to change from a content $y = 652$ to the full condition is 347, and the desired result is therefore achieved.

It is important to note that each digit of the number y is the nine's complement of the corresponding digit of the number x. Thus, it is very simple to set the number y. It is only necessary for the wiring of the selection switch (Fig. 2) to be so designed that the nine's complement of the selected number is set in the corresponding decade. In this way interaction between the decades is avoided.

Fig. 2 shows the decade and selection switch schematic diagram. The decade follows the 1—2—4—8 code and, as can be demonstrated, its speed is limited only by the speed of the basic flip-flop. A suitable pulse on the reset line allows the decade points A, B, C, D to be set low; a subsequent pulse on the write line modifies this state according to the position of the selection switch. For

flip-flop can operate at 8MHz (minimum guaranteed frequency) the operation would have to be performed within 125ns. This, due to the inevitable propagation delays through the circuitry, is impossible to achieve unless the maximum operating frequency is considerably reduced.

If the optimum performance of micrologic elements is required it is necessary to use an artifice which enables the time available for the setting of y to be increased. The artifice used in the divider described here is as follows (see Fig. 3). The moment that the full condition of the divider is achieved (detected by gate M), instead of setting y in the divider, the main divider is inhibited and an auxiliary decade, driven by the same frequency as that of main divider, is activated.

This function is obtained by means of a flip-flop consisting of gates S and R; the main divider is activated when a high or low signal, respectively, is fed to the J and K inputs of the first flip-flop in the first decade. This signal is obtained from gate R. Similarly the auxiliary decade is activated or inhibited by the output of gate S.

When the auxiliary decade has counted ten pulses, decoding (gate N) re-establishes the previous situation by acting on the S-R flip-flop i.e. the auxiliary decade is inhibited and the main divider activated; the latter has therefore been inhibited for a time equal to ten periods of

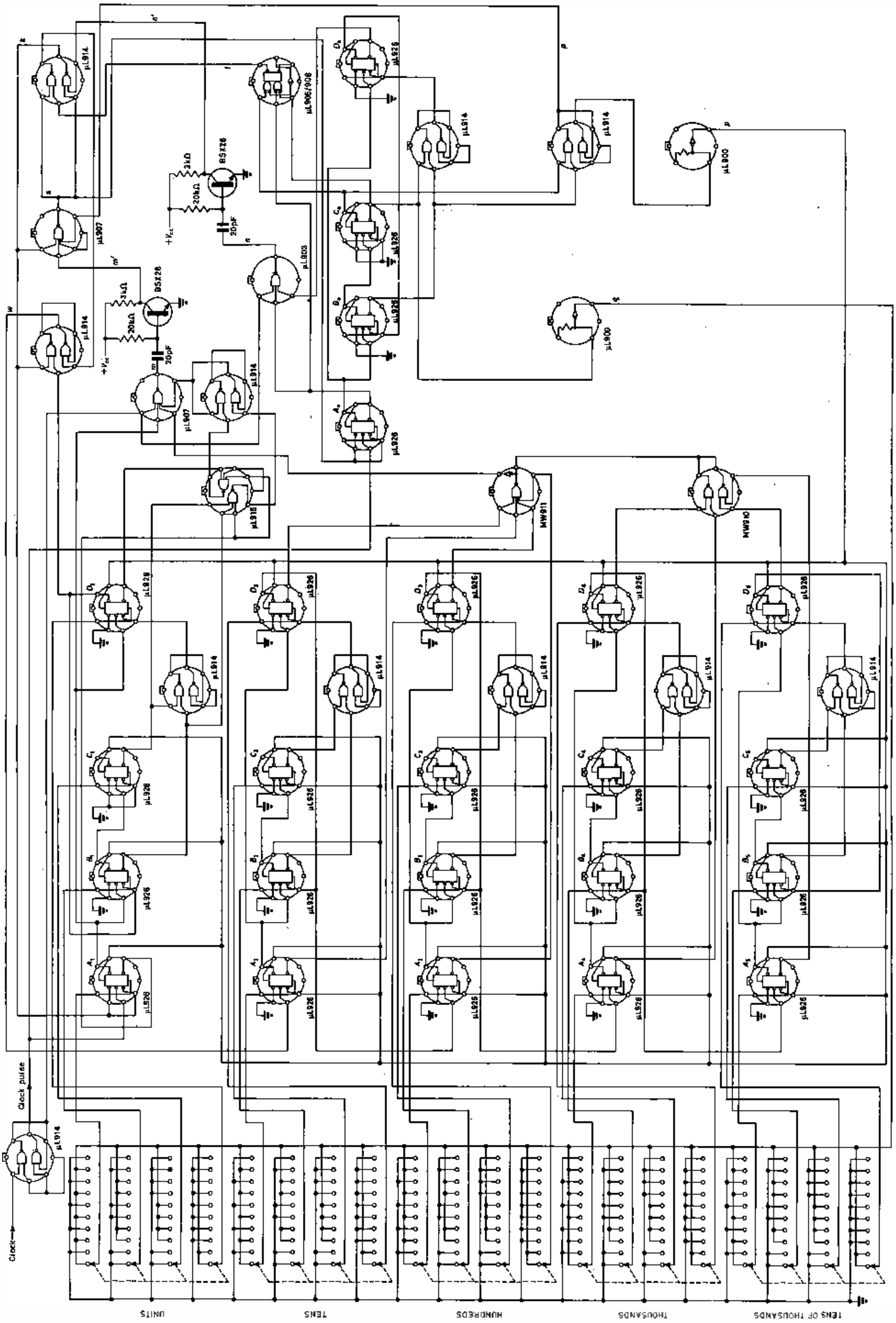


Fig. 6 Complete schematic diagram

the input frequency. This time is sufficient to set the number y. Such a sequence is actuated by suitable decoding of the auxiliary decade (gate P, Q) as will be explained later. Thus, the artifice just described multiplies by ten the time available for the setting of the number y in the main divider, but, since it stops the main divider for a time equal to ten pulses, introduces an error which must be corrected.

Correction of the Error

The contents of the divider, at the instant it is activated, are thus ten times lower than the contents it would have, had the artifice not been used. In consequence, error compensation can be obtained by increasing by ten the content of the divider, immediately after setting of y and immediately before the arrival of the activating signal. The solution is shown in Fig. 3 and consists of sending an additional pulse to the second decade immediately after the reset and write operations. Since such a correction pulse is sent to the input of the second decade, it has the same weight as ten pulses sent to the input of the first one, and consequently fully compensates the error. The correction pulse is obtained by decoding the auxiliary decade (gate T) and is sent to the input of the second decade by means of a logic circuit (gates U, V, W). This allows the second decade to receive either the carry from the first decade, when the main divider is activated, or the correction pulse, when the auxiliary decade is activated.

Complete Unit

Fig. 4 shows the schematic diagram of a five-decade variable-ratio divider. The description already given, and Fig. 2 and 3 will facilitate understanding the operation.

Timing Diagram

The timing diagram shown in Fig. 5 shows waveforms occurring at points in the circuit in Fig. 4. The diagram shows the transition from one to another counting cycle and this emphasizes the operation of activating the main and auxiliary dividers as well as reset of y and correction operation. $A1$, $B1$, $C1$, $D1$ show the waveforms from the corresponding points of the first decade of the main divider and A_Δ, B_Δ, C_Δ, D_Δ the outputs of the auxiliary decade. Waveform m (output of the gate M) is a decoding of the main divider and, with its negative-going edge indicated that the full condition (i.e. the 99999 content) has been achieved in the main divider. By means of the negative-going edge in the waveform a monostable multivibrator is triggered, which produces a pulse of about 300ns in duration at its output (waveform m'). This pulse drives the S side of the flip-flop S, R. The output of S (waveform s) is negative and activates the auxiliary decade, while output R stops the main divider.

Similarly, waveform n is obtained from the auxiliary decade by means of gate N and its negative-going edge indicates when one full cycle has been completed. This negative-going edge drives a monostable multivibrator identical to the previous one, which resets the S, R flip-flop with S positive and R negative, thus restoring the previous condition. In consequence the negative-going edges of m and n determine the time interval during which the auxiliary decade completes one cycle. During this cycle, following sequence of operations obtains: resetting of the main divider, writing of y, correction of the error. It is therefore evident that the necessary signals can be obtained by suitable decoding of the auxiliary decade.

The reset signal p, obtained from gate P, is amplified and passed to the reset inputs of the main decades. The write signal comes directly from the auxiliary decade C_Δ output and therefore only amplification (obtained by means of the amplifier-inverter Q) is necessary. This signal is sent to the decade selector switches as in Fig. 2.

Finally, the correction signal t is obtained from the auxiliary decade, delaying the signal C_Δ for a time equal to one period of the input frequency; this produces a signal having a negative-going edge (which is required to trigger the μL 926 flip-flop) later than the negative-going edge of the write pulse. Both signal t and output D_1 of the first decade are combined in gates U, V, W with the outputs of flip-flop S, R and produce signal ω which controls the second decade.

It should be noted that in the decoding of the main divider (gate M) and the auxiliary decade (gate N) the inverted input signal (output of H) has been used for the purpose of obtaining signals at the output whose trailing edge, apart from the difference in the delay time of the two gates, is synchronous with the input signal. This, in spite of the need for monostables to increase the pulse duration, virtually eliminates the effect of the μL 926 flip-flop propagation delays and consequently saves time. With this configuration the negative-going edges of the outputs m and n must indicate the transition of the main divider and auxiliary decade into the condition in which they are completely full. Thus, to obtain m and n, the configurations relative to 99998 for the main divider and 9 for the auxiliary decade respectively, are decoded. At the input of gate M a special circuit has been used to obtain a pulse at the output whose duration is not affected by the propagation delay across the decade.

Fig. 6 shows the complete schematic diagram of the divider. Some special points are indicated using the same letters as in Fig. 4 in order to simplify the comparison between the two diagrams.

Performance

The speed characteristic and the problem of the extension of the number of division ratios will now be examined. The speed can be limited by the following two factors:

(1) speed of the basic flip-flop. In the case of RT μL elements, the minimum guaranteed frequency is 8MHz;

(2) ability to trigger the two monostables at the output of gates M and N (Fig. 3).

These monostables are driven by a pulse having a duration of about one-half the period of the input frequency and, if the frequency is very high, the pulse can become so short that it cannot fully charge the capacitances, thus compromising correct operation. In Fig. 4 the monostable recovery time is about 40ns and in consequence, the two limits are roughly the same. In a laboratory prototype reliable operation was obtained up to 16MHz.

The extension of the number of division ratios is quite simple: it is sufficient to add in cascade the proper number of decades (connected like the previous ones) and, for each additional decade, increase the number of inputs of gate Y by two (Fig. 3). A limitation of the circuit is the impossibility of setting division ratios equal to or lower than 11, due to the use of the auxiliary decade (for the first ten ratios) and to the particular decoding scheme used for the main divider (for the eleventh).

This limitation is not, however, too restrictive. The circuit output can be taken from any point where a complete pulse for each division cycle is obtained. This condition obtains at several points in the circuit, e.g. C_Δ, t, P, w, q etc.

A Pulse Circuit for Long Delays

By B. Jovicic*, Dipl.Eng.

Delay with this circuit does not depend on supply voltage. The time-base circuit is, with the exception of the power supply, electrically separated from the remaining part of the circuit. Delays up to 100s/μF with a timing capacitance can be obtained. The delay of one unit can be varied continuously, without changing the basic elements of the time-base, within a three minute duration. Errors in delay due to temperature variations can be compensated. Basic pulse circuits such as monostable and astable multivibrators, delay-time relays, etc., can easily be designed with this circuit.

(Voir page 475 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 476)

IN order to show more readily the difference between the circuit described in this article and other circuits for long delays from which this circuit has been derived, it is useful to give a brief survey of such circuits.

Fig. 1[1,2] and Fig. 2[3] show the solutions that partially satisfy the requirements. Circuit delay from Fig. 1 is obtained from the expression:

$$T = R_t C_t \ln\,[1 + (E/V_{cc})] \quad \dots\dots\dots\dots \;(1)$$

where the effect of the amplitude of the exciting pulse e, and the voltage drop at diode D in the emitter coupling of the transistor in the conducting state, has been neglected. It can be seen that time T depends directly on the absolute values of two supply voltages but does not depend on their equal percentage change. This dependence is, however, very inconvenient if operation is required at high voltage as with this circuit.

This disadvantage is eliminated in Fig. 2 (Iljin circuit). The end of the delay interval is given by the time at which the potentials of points A and B are equalized (if the voltage drop in the diode in the conducting direction is neglected) when an interruption of the current charging the capacitors C_1 and C_2 occurs in the pulse. The disappearance of current through the resistance r ($r \ll R$) creates a corresponding change in the voltage level at point L, indicating that the interval T has ended. If in the above description $R_1 = R_2 = R$ and $C_1 = C_2 = C$, an expression for delay is easily obtained by:

$$T = RC \ln 2 \quad \dots\dots\dots\dots\dots\dots \;(2)$$

This delay circuit is readily inserted into most pulse circuits and their accuracy and operational reliability are greatly improved. Nevertheless, it is best used for shorter time intervals; if a wide range of time delay is needed, it is necessary to change the value of r, and, in addition, to change at least two of four elements of the time base. A particular limitation is introduced by the input impedance of the circuit being connected to point L. The maximum delay that can be obtained with 1μF is therefore considerably reduced.

Suggested Circuit for Long Delays

Fig. 3 shows a suggested circuit where the time base output is *via* capacitor C_s separated from the remaining circuit, and through which, at the moment time T ends, only the exciting pulse e from the generator is transmitted. Potentiometer RV is formed by the voltage divider $R_1 - R_2$ which, together with $R_t - C_t$ and diodes MR_1, makes a bridge circuit. The divider determines the level V_o at which diode MR_1 is directly biased and the trigger pulse is transmitted so that:

$$V_o = \frac{R_1}{R_1 + R_2}\,E \quad \dots\dots\dots\dots \;(3)$$

If, in calculating, the same parameters are ignored as in equation (1), the following expression:

$$T = R_t C_t \ln a \quad \dots\dots\dots\dots\dots\dots \;(4)$$

where $a = \dfrac{R_1 + R_2}{R_2}$

will be obtained for the delay with the circuit of Fig. 3.

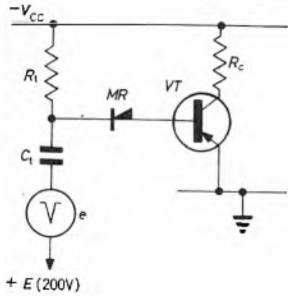


Fig. 1 Pulse circuit in which delay depends on voltage

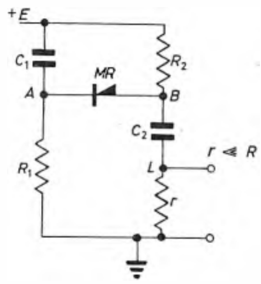


Fig. 2 Iljin's circuit

From equation (4) it can be seen that T does not depend on E. The sum of R_1 and R_2 is always constant, though R_2 can be varied. Theoretically, a may be made arbitrarily large so that the equivalent time T is obtained by multiplying the product $R_t C_t$ by some number considerably greater than 1 ($\ln a > 1$). If the amplitude of the exciting pulse is e and voltage drop across the diode in its conducting state V_o, then:

$$T = R_t C_t \,\ln \frac{a}{1 + ((e - V_d)/E)\,a} \quad \dots\dots \;(5)$$

In order to transmit the pulse, e must be greater than V_d. For E sufficiently great, equation (5) reduces to equation (4) with no loss in accuracy. Thus, T can be continuously changed by varying the ratio a without changing the parameters R_t, C_t or e.

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Capacitor C_s allows voltage E to be not necessarily the sum of the voltages from two sources as in Fig. 1. The sum $(R_1 + R_2)$ does not directly influence the delay but only the relationship $a = 1 + R_1/R_2$. However the sum $(R_1 + R_2)$ should be $> 100\text{k}\Omega$. It is obvious that a depends on temperature only by the same amount as potentiometer RV. If, however, there are temperature variations of other components during time T they can be compensated or considerably reduced by inserting a temperature-variable resistance in series with R_1 or R_2.

The new delay circuit in Fig. 3 is easily applied to pulse circuits as a simple time element. Here it will be described in the circuit of an astable multivibrator (Fig. 4) and of a delay-time relay and monostable multivibrator (Fig. 5). The astable multivibrator consists of two delay circuits that determine independently the duration of one of the stable states of a flip-flop. This flip-flop, shown as a block in the figure, contains a relay. The durations of the stable states are $t_1 = R_{t1}C_{t1} \ln a'$ and $t_2 = R_{t2}C_{t2} \ln a''$.

While C_{t1} is being charged, C_{t2} is discharged. After time t_1, pulse e through C_1 changes the stable state of the flip-flop stage. The relay contact A discharges C_{t1} and contact B allows C_{t2} to charge. After time t_2, pulse e through C_{s2} resets the flip-flop into its initial stable state and the whole cycle is repeated.

If the time element is omitted for time t_2, the delay-time relay in Fig. 5 will be obtained. By opening switch S time interval T commences. After it has terminated, contact A discharges C_t. The relay remains energized until the flip-flop is once again reset into its initial state by means of a switch S through C_o and contact B. The cycle is repeated on an external command.

The value of delay can also include the time of attraction and release of the relay armature. These times are, however, negligible. The change in the voltage level in the flip-flop can be used as an output signal, or the remaining contacts of the relay as control contacts. Control can also be carried out by external pulses through points 1 or 2.

If contact A in Fig. 5 normally short-circuits C_t and the excitation is performed as illustrated by dotted lines through the normally closed contact B, the circuit becomes a monostable multivibrator.

Conclusion

The circuit allows complex automatic control circuits such as time programmers, timers, etc., to be realized easily. The simple structure of this circuit makes it possible to design in a simple way. However, it is necessary to take into account some important points. All other conditions being equal, the repetition accuracy is within the limits of one period of the exciting pulse e. Taking into account the delays for which this circuit is intended, if the frequency of these exciting pulses is 50Hz to 1kHz the actual value of the frequency is not critical.

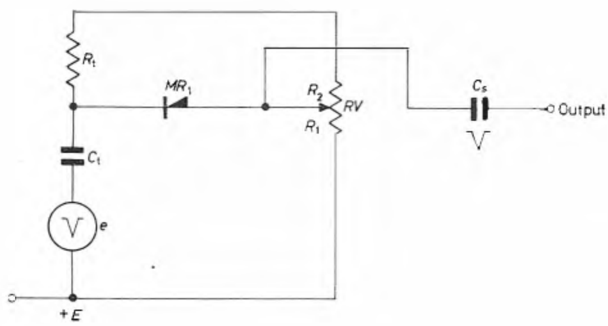
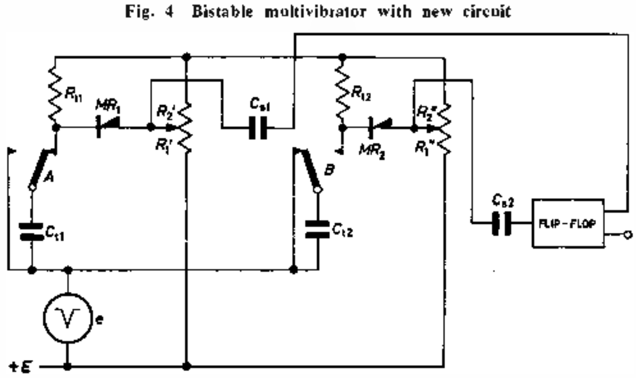


Fig. 3 New pulse circuit for long delays

Fig. 4 Bistable multivibrator with new circuit



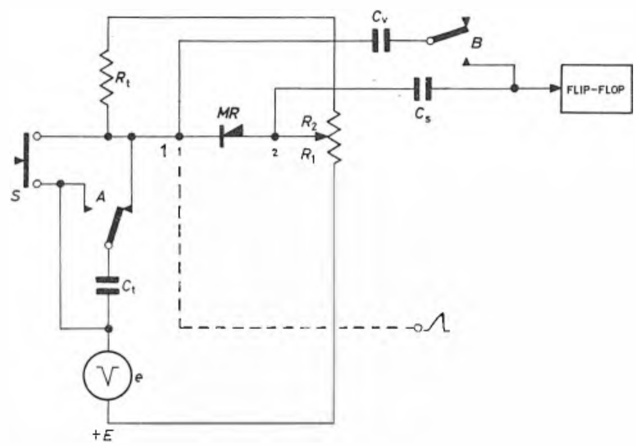


Fig. 5 Time relay or monostable multivibrator with new circuit

The condition for transmitting the trigger pulse, with period T_e and duration t_e, through diode D in the conducting state during the transmission of pulse i_d, is given by:

$$T_e/t_e > \frac{i_d R_t}{E - V_o} \quad \text{.............. (6)}$$

The operational condition of this relationship can easily be checked separately for each case, but in general it may be concluded that equation (6) does not give any practical limitation.

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Calculation of V-I Curves for N.T.C. Thermistors

A Method of Calculating the Steady-State Voltage/Current Characteristics of Directly-Heated Negative-Temperature-Coefficient Thermistors

(Part 2)

By M. R. McCann*, C.Eng., A.M.I.E.R.E.

(Voir page 475 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 476)

Basic Assumptions in the Method of Calculating the Voltage-Current Characteristics

It has been shown that the method described in the text in calculating the voltage-current characteristics agrees with practical measurements on STC thermistors to a sufficient degree of accuracy. However, the small deviations which occur between the theoretical and practical characteristics can be summarized in the following points which are primarily assumptions in the initial mathematical equations:

(a) the equation involving thermistor resistance and temperature, equation (1), assumes that there is no temperature gradient within the thermistor. However, when power is dissipated in the thermistor, a temperature gradient does exist and its value is dependent upon the nature of the medium surrounding the thermistor (thermal conductivity, rate of flow, etc.);

(b) in equation (1) it is assumed that the constant B (characteristic temperature) does not vary over a range of temperature up to 300°C (the highest mean bead temperature which was used in the experiments);

(c) equation (4) assumes that there is no radiation of heat from the surface of the thermistor. In practice, radiation must be present and becomes more significant at high temperatures;

(d) the measurement of the characteristics assumed that the nature of the surrounding medium was such that the heat was transferred away in a free flow of air (according to Newton's law of cooling). This condition was adhered to as closely as possible in the physical arrangements in the practical experiment;

(e) all thermistors used in the experiment were of a material of the same basic composition. No practical experiments were tried on thermistors of other manufacturers to determine whether the former assumptions are more significant with thermistors of different composition. However, calculated characteristics from data quoted by some other manufacturers seem to correlate, with a reasonable degree of accuracy, to their published characteristics of voltage *versus* current;

(f) differences between measured and calculated curves can also be partially accounted for by errors in the measurement of voltage, current and particularly of the ambient temperature of the surrounding medium.

The Effects of Practical Tolerances on the Voltage-Current Characteristic

It has been shown that when the thermistor parameters are given numerical values the shape of the voltage-current characteristic can be accurately determined. In practice there is a tolerance on these parameters due to the slight variations in the composition and size of each thermistor. Given these tolerances, the tolerance on the voltage-current characteristic can be determined. However, the situation is complicated because these quantities are to some extent dependent. For example, if the size of a bead thermistor is increased, its cold resistance R_0 tends to decrease and its dissipation constant increases, the relative variations being dependent upon the geometry of the bead. Thus when calculating the tolerance of the variation in E_{max} one effect might tend to compensate for another.

In practice the major difficulty in calculating the voltage-current characteristic of a thermistor is to predict accurately the transfer of heat away from its surface since its environmental conditions can vary appreciably according to its mounting. If the dissipation constant or E_{max} is known or can be measured under the particular physical conditions in which the themistor is being used, then the problem is much simplified.

Another more obvious conclusion can be obtained from equation (17). If a close tolerance is required on E_{max}, then a thermistor must be chosen having a close tolerance on its cold resistance, R_0.

The Voltage-Current Characteristic of Evacuated Type Thermistors

The majority of the heat dissipated from a thermistor bead mounted in a vacuum (such as in the STC type R thermistor) is by radiation rather than by convection and conduction. The radiant heat is absorbed by the glass envelope, and is finally dissipated in the medium surrounding the envelope by convection and conduction. According to Stefan's law of radiation, the rate of loss of heat from the thermistor is proportional to the difference between the fourth power of the absolute temperature of the thermistor and the fourth power of the absolute temperature of its surroundings. This means that the dissipation constant should be approximately proportional to the cube of absolute temperature of the thermistor. However, the transfer of heat away from the thermistor is modified considerably by the conduction of heat along its leads, which in turn is transferred to the internal medium by radiation and conduction. The heat transferred along the leads of the thermistor tends to reduce the change of the dissipation constant with bead temperature. Hence the solution of the transfer of heat of the surroundings is extremely complex.

It has been found experimentally for the STC type R thermistor (bead diameter approximately 0·01in, lead diameter 0·001in) that the dissipation constant k is approximately proportional to the square of the absolute temperature. The normalized voltage-current characteristics have been calculated assuming that k is directly proportional to the square of the absolute temperature. The

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normalized characteristics are plotted in Fig. 13. Using the curves in this figure the voltage-current curves have been calculated for various R type thermistors and compared with practical measurements as shown in Fig. 14. Measured values of voltage agree within 2 per cent of the calculated values of voltage at given values of current up to a mean bead tempereature of 300°C. This is of sufficient accuracy for most practical purposes. An approximate construction for the voltage-current characteristic at other

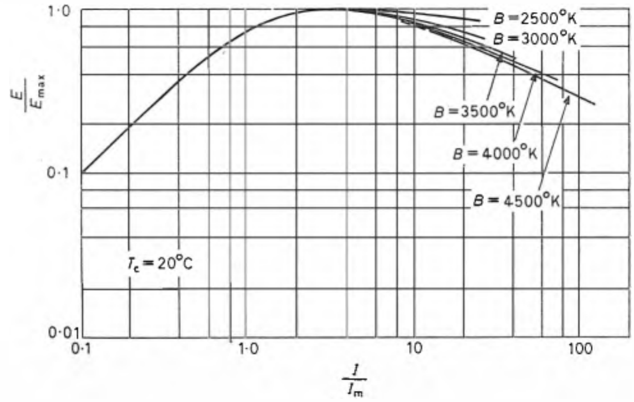


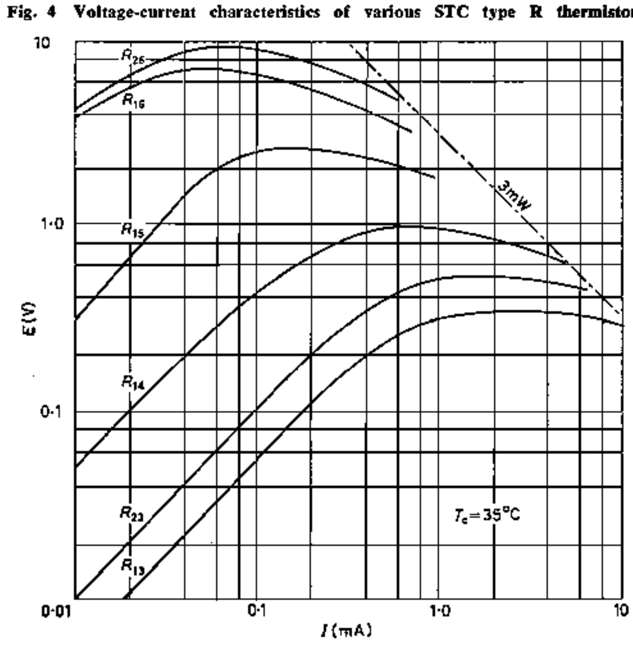
Fig. 13 Relation between the normalized voltage and the normalized current for an evacuated thermistor (STC type R)

ambient temperatures can be derived from the normalized curves shown in Fig. 13 by a method similar to that described earlier in the text for thermistors in air and other media.

Conclusions

Calculation of the voltage-current characteristic of an n.t.c. thermistor can be made by a graphical method accompanied by simple mathematics. The voltage-current characteristic of directly heated thermistors in free air at any ambient tempereature can be calculated with sufficient accuracy for most practical purposes, given the values of resistance at 20°C (power zero), characteristic temperature and dissipation constant. Calculations are simplified if the voltage maximum on the voltage-current curve is known,

Fig. 4 Voltage-current characteristics of various STC type R thermistors



rather than the dissipation constant. There is a greater departure from the predicted characteristic if the medium surrounding the thermistor is a liquid rather than a gas.

Appendix A

Given equations (13) and (16)

$$T_{max} = (B/2)\,[1 - \surd\,(1 - 4T_a/B)] \;\ldots\ldots\ldots\; (A1)$$

$$E_{max} = [R_ok(T_{max} - T_a)]^{\frac{1}{2}}\exp\,[(B/2T_{max})-(B/2T_a)]\ldots(A2)$$

Consider the term $\exp\,(B/2T_{max} - B/2T_a)$ in equation (A2); from equation (A1), this becomes:

$$\exp\left[\frac{1}{1 - \surd(1 - 4T_a/B)} - (B/2T_a)\right]$$
$$= \exp\,\{[(B/4T_a)\,\surd\big(1 - (4T_a/B)\big) - 1]\} \;\ldots\ldots\; (A3)$$

by rationalizing the surd.

The expression in parenthesis can be expanded into an absolutely convergent binomial series in terms of T_a/B provided $4T_a < B$.

Expression (A3) becomes:

$$\exp\,(-1/2)\exp\big(-(T_a/2B) - (T_a^2/B^2) - (5T_a^3/2B^3) - (7T_a^4/B^4) - (21T_a^5/B^5) - \ldots\big)$$

The exponential can be expanded as a series:

$$(1/\surd e)\,[1-(T_a/2B)-(7/8)\,(T_a^2/B^2)-(97/48)\,(T_a^3/B^3) - (2063/384)\,(T_a^4/B^4)-(59681/3840)\,(T_a^5/B^5) - \ldots]$$

Hence, equation (A2) becomes:

$$E_{max} = \sqrt{\left(\frac{R_ok(T_{max} - T_a)}{e}\right)}\,[1 - (T_a/2B) - (7T_a^2/8B^2) - (97/48)\,(T_a^3/B^3) - (2063/384)\,(T_a^4/B^4) - \ldots] \;\ldots\ldots\; (A4)$$

Consider the term $\sqrt{\left(\frac{R_ok(T_{max} - T_a)}{e}\right)}$ in equation (A4); this becomes:

$$(R_ok/e)^{\frac{1}{2}}\,\{(B/2)\,[1 - \surd\,(1 - 4T_a/B)] - T_a\}^{\frac{1}{2}}$$

Expressing this in the binomial expansion:

$$= \big[(R_ok/e)\,\big(T_a+(T_a^2/B)+(2T_a^3/B^2)+(5T_a^4/B^3)+ (14T_a/B^4)+(42T_a^6/B^5)+(132T_a^7/B^6) + \ldots - T_a\big)\big]^{\frac{1}{2}}$$

$$= \big[(R_okT_a^2/eB)\,\big(1 + (2T_a/B) + (5T_a^2/B^2) + (14T_a^3/B^3) + (42T_a^4/B^4) + \ldots\big)\big]^{\frac{1}{2}}$$

Again, using the binomial expansion:

$$\sqrt{\left(\frac{R_ok(T_{max} - T_a)}{e}\right)} = T_a\,\surd\,(R_ok/eB)\,[1 + (T_a/B) + (2T_a^2/B^2)+(5T_a^3/B^3)+(14T_a^4/B^4)+(42T_a^5/B^5)+\ldots] \;\ldots\ldots\ldots\ldots\; (A5)$$

From equations (A4) and (A5):

$$E_{max} = T_a\,\surd\,(R_ck/eB)\,[1+(T_a/2B)+(5/8)\,(T_a^2/B^2) + (53/48)\,(T_a^3/B^3) + (903/384)\,(T_a^4/B^4) + (21769/3840)\,(T_a^5/B^5) + \ldots] \;\ldots\ldots\; (A6)$$

Let:

$$D=(T_a/\surd e)[1+(T_a/2B)+(5/8)(T_a^2/B^2)+(53/48)(T_a^3/B^3) + (903/384)\,(T_a^4/B^4)+(21769/3840)\,(T_a^5/B^5) + \ldots]$$

Then:

$$E_{max} = D\,\surd\,(R_ok/B) \;\ldots\ldots\ldots\ldots\; (A7)$$

Appendix B

Let $E_{max}' = D'\,\surd\,(kR_o'/B)$ when the ambient temperature is T_a', and $E_{max}'' = D''\,\surd\,(kR_o''/B)$ when the ambient temperature is T_a''.

Then:

$$\frac{E_{max}''}{E_{max}'} = (D''/D')\,\surd\,(R_o''/R_o')$$

Now:

$$D'' = (T''_a/\surd e)\,[1 + (T_a''/2B) + (5/8)(T_a''^2/B^2) + \ldots]$$

and:

$$D' = (T_a'/\surd e)\,[1 + (T_a'/2B) + (5/8)(T_a'^2/B^2) + \ldots]$$

Then:

$$D''/D' = (T_a''/T_a')\left[1 + \frac{T_a'' - T_a'}{2B} + \ldots\right]$$

But $\left[1 + \frac{T_a'' - T_a'}{2B} + \ldots\right]$ is a rapidly convergent series, since in practice $T_a'' - T_a' \ll 2B$.

Therefore:

$$D''/D' \simeq (T_a''/T_a')\left[1 + \frac{T_a'' - T_a'}{2B}\right]$$

and:

$$\frac{E_{max}''}{E_{max}'} = (T_a''/T_a')\left[1 + \frac{T_a'' - T_a'}{2B}\right]\surd(R_0''/R_0')$$

Thus:

$$\frac{E_{max}''}{E_{max}'} = (T_a''/T_a')\left[1 + \frac{T_a'' - T_a'}{2B}\right] \exp\{(B/2)\,[(1/T_a'') - (1/T_a')]\}$$

If $\Delta T_a = T_a'' - T_a'$, then

$$\ln\left(\frac{E_{max}''}{E_{max}'}\right) = \ln\,[(1+\Delta T_a/T_a')(1+\Delta T_a/2B)] - \frac{B\Delta T_a}{2T_a''T_a'}$$

Expanding $(1 + \Delta T_a/T_a')(1 + \Delta T_a/2B)$ in a logarithmic series, $\Delta T_a/T_a' + \Delta T_a/2B - \Delta T_a'^2/2T_a'^2$ are the most significant terms, since $\Delta T_a/T_a \gg \Delta T_a/2B$.

Thus:

$$\ln\frac{E_{max}''}{E_{max}'} \simeq (\Delta T_a/T_a') + (\Delta T_a/2B) - (\Delta T_a^2/2T_a'^2) - \frac{B\Delta T_a}{2T_a''T_a'}$$

$$= -\frac{\Delta T_a}{T_a''T_a'}\left[(B/2) - T_a'' - \frac{T_a''T_a'}{2B} + \frac{\Delta T_a T_a''}{2T_a'}\right]$$

$$= -\frac{\Delta T_a}{T_a''T_a'}\,[(B/2) - T_a''(1 + (T_a'/2B) - (\Delta T_a/2T_a'))]$$

i.e.

$$\ln\left(\frac{E_{max}''}{E_{max}'}\right) = -\frac{\Delta T_a}{T_a''T_a'}[(B/2) - T_a'(1 + (T_a'/2B) + (\Delta T_a/2T_a') + (\Delta T_a/2B) - (\Delta T_a^2/2T_a'^2)]$$

If T_a is 293°K and B is 3 000°K, then:

$$T_a'(1 + (T_a'/2B) + (\Delta T_a/2T_a') + (\Delta T_a/2B) - (\Delta T_a^2/2T_a'^2))$$

$= 307$°K when $\Delta T_a = 0$°C ($T_a'' = 20$°C)

$= 310$°K when $\Delta T_a = 5$°C ($T_a'' = 25$°C)

$= 340$°K when $\Delta T_a = 100$°C ($T_a'' = 120$°C)

$= 284$°K when $\Delta T_a = -40$°C ($T_a'' = -20$°C)

Hence:

$$\ln\left(\frac{E_{max}''}{E_{max}'}\right) \simeq -\frac{\Delta T_a}{T_a'T_a''}\,[(B/2) - 310]$$

Appendix C

If Fig. 3 is replotted as lines of equal I/I_m as a function of F on one axis and E/E_{max} on the other, then more precise values of E/E_{max} can be obtained for intermediate values of F. This is particularly useful in the preparation of voltage-current curves for data sheets on thermistors. This graph would also allow a universal table of numerical values of E/E_{max}, I/I_{max}, and F to be constructed.

Appendix D

A resistive load line for a graph with logarithmic scales can be constructed by a rectangular hyperbola whose asymptotes are the open-circuit voltage and the short-circuit current. The asymptotes are made perpendicular to the axes of the voltage-current characteristics, and their intersections with the axes are set to the open-circuit voltage and the short-circuit current. The equivalent circuit of such an arrangement is shown in Fig. 8. An example of a load line is shown in Fig. 7.

This hyperbolic load line is of a shape which is only dependent on the size of logarithmic scales of the graph on to which it is to be drawn. Its shape is entirely independent of the voltages, currents and resistances associated with the thermistor and the d.c. supply. The position of the load line on the graph is the only variable required to define the operating conditions.

The basic shape of the load line is constructed by plotting a line on the logarithmic graph such that the sum of the numerical values from the unity asymptotes to any point on the line is unity. A value of 0·7 on one axis corresponds to a value of 0·3 on the other; 0·6 on one axis corresponds to 0·4 on the other, etc. It must be noted that the above relationship for determining the shape of the load line does not hold when it is moved to other positions on the graph.

Acknowledgment

The author is indebted to Standard Telephones and Cables Ltd for permission to publish this article. He also wishes to thank his colleagues at STC for their valuable assistance.

New Electromagnetic Lens for Television Displays

A new electromagnetic lens has been introduced which can produce on a c.r.t. screen a spot of light so small that, on a 12in screen, it can write a line finer than a human hair. It enables high concentrations of data to be presented on cathode-ray-tube displays of the type used in radar and computer systems.

The device—Marconi type F1500—acts as an electromagnetic lens and it can focus the electron beam to produce a spot a fifth of a millimetre in diameter in the central region of the screen and no more than twice this diameter in other areas.

The focus-coil unit is made up of seven annular-shaped sections bolted together to form a compact cylinder which fits over the neck of a c.r.t. Furthest from the screen is a coil which corrects for any slight misalignment of the aperture plate. This coil is separated from the remaining sections of the unit by a blank section, which is followed by two more alignment coils which deflect the beam along the axis of the focus-coil unit. The focus coil itself is next in line and contains static and dynamic coils, layer-wound to ensure accurate focusing. The remaining sections (nearest to the screen) correct for astigmatism arising from the effects of the previous coil units.

Throughout the assembly, the magnetic parts are made of mumetal, which minimizes remanence or hysteresis effects that would otherwise complicate the adjustment of the unit when it is being set up. Thermistors are employed with the alignment and astigmatism-correction coils to compensate for changes in temperature over the range from −10°C to +70°C. The assembly can be used with all types of magnetically deflected c.r.t.s with neck diameters up to 35mm. Where the neck of the tube cannot accommodate the whole of the unit one or two of the correction coils can be omitted.

Automatic Regulation of Line Level in Long Carrier Links

By J. Korn*

Line regulation in carrier systems compensates for the temperature-dependent losses of the cable. Two methods are employed to achieve this: a feedback control system with a pilot voltage as reference and an open-loop control with the cable temperature as reference.

A stepwise digital line regulator is employed as pilot regulation. The dynamic behaviour of a great number of such line regulators in tandem connexion is excellent. The overshoot amplitude of the transient oscillation after a sudden level change does not exceed 40-50 per cent of the initial level change.

Open-loop control uses the temperature-dependent remote power current to the underground repeaters, transmitted via the inner conductors the cable, as control current. The control element consists of two indirectly heated thermistors in cascade. This open-loop control ensures that pilot regulators are needed only at long intervals, e.g. at every remote power supply station.

(Voir page 475 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 476)

IN carrier communication systems the losses in the transmission line and of other parts of the link change with frequency; however, at a fixed temperature, the loss-frequency characteristic is constant, except for the changes caused by ageing. This mean loss of a repeater section can be compensated and equalized by fixed amplifiers.

Variations in loss are caused by changes in ambient temperature; these changes influence mostly the ohmic resistance of the line. In coaxial cables, the temperature change has the same effect as if the cable were extended or shortened at a constant temperature. The loss in the cable as a function of frequency is thereby changed approximately in accordance with a $\sqrt{f}$ law. An idea of the magnitude of loss to be overcome may be obtained from the fact that 2 500km of small-diameter coaxial cable, with a carrier system for 300 telephone channels on that cable, result in a total loss of 15 220dB at the highest transmitted frequency; the variation in loss through temperature is 30dB per degree of temperature change. This means that for cables in our climate, with a temperature variation of ± 10 degC over a year, a loss variation of ± 300dB exists when the laying depth is about 1m. The maximum loss variation in the course of a day is 30dB. In view of the fact that only a certain level deviation (for instance ± 2dB) is permitted throughout the system, loss compensation is absolutely necessary at intervals along the link.

For economic reasons surface repeater stations are provided, in long-distance carrier communication links, by coaxial cables only at rather long intervals (about 100km) and between these stations underground repeaters are mounted in manholes or in buried repeater boxes. These underground repeaters are powered by d.c. current from the surface stations *via* the inner conductors of the coaxial pairs. The temperature-dependent loss variation of the cables is so large that provisions for compensating these may be necessary, as for instance, after each second repeater.

As cable temperature varies in a matter of hours, it is advisable to correct the resultant loss variation automatically. This may be effected either by closed-loop control (feedback control) or by open-loop control method. For regulation of the line level by feedback control (regulation that compensates for the temperature-dependent cable loss) a pilot voltage of suitable frequency is used. This is added to the transmission band at the input of the system and is attenuated by the same amount as the communication signal. At intervals, for instance after the line amplifier at a surface station, the pilot voltage is picked off and is used for setting the amplifier. If the pilot level deviates from the nominal value, the amplifier gain is changed such that this value is maintained. In a coaxial cable in which, as mentioned before, the frequency response is a function of $\sqrt{f}$, a single pilot frequency is sufficient for setting an equalizer to have a proper loss characteristic.

The requirements with regard to line regulation are: good dynamic behaviour, high regulation speed, and economy. The regulator must be inhibited if the pilot suddenly fails so that traffic on the link can continue, even if the regulation system is out of order.

Automatic line level regulation by feedback control through a pilot is capable of maintaining the level very accurately to the nominal value, but it is expensive. Furthermore, the rather complicated circuitry increases the danger of faults, and it is difficult to supervise pilot-controlled regulators in underground repeaters. For these reasons, efforts have been made to regulate the gain in underground repeaters by an open-loop control method, using temperature sensing elements and to correct accumulating errors by pilot-controlled regulators at longer intervals. It is possible to use the cable itself as the temperature-sensing element and the remote power current on its inner conductor as the control current, because if the remote power voltage is constant, the power current is a measure of the total cable resistance and thus of cable loss. Pilot-controlled regulators for the elimination of residual errors are required in this case at the surface repeater stations only when they are also the power-feed stations. The pilot-controlled line regulators are designed for step-by-step operation. This combination of open-loop and closed-loop control to compensate for the temperature-dependent loss variations will be called line level regulation. The pilot-controlled line regulator will be described first and then the open-loop control circuits.

Line Level Regulation by Feedback Control

The length of the intervals between gain-controlled line amplifiers must be chosen such that the difference between the nominal value and the actual value of the pilot level does not exceed a set value (for instance 0·2 Np) at any point on the transmission route. Thus, on long routes

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many similar pilot-controlled line regulators are connected in tanden (in the L1-system of Bell Telephones, for instance, up to 200 gain-controlled amplifiers are so connected). Each one of these regulators maintains the respective line amplifier at the nominal value.

If a fluctuation in the pilot level occurs at the beginning of this tandem connexion, all pilot-controlled line regulators will react to it and return the level everywhere to its nominal value. The transient behaviour of such a tandem connexion is one of the central problems of line regulation.

In conventional types of line regulation transients appear at the end of a long tandem connexion from gain enhancement caused by compensation of sudden level changes. The overshoot amplitudes of such a transient oscillation increase as more regulators are involved in the tandem connexion[1]. The application of a stepwise digital line regulator favourably influences transient behaviour in such a tandem network[2].

Closed-loop control circuits will first be explained and the transient behaviour will then be discussed. With the digital line regulator the same level change and the same regulation speed is assigned to each regulating step. Prior to each step the level is checked so that a new step can only be made if there is still a level deviation. If the steps are made sufficiently small, the process will be almost continuous. Such a regulator must contain the following (see Fig. 1):

(a) a level discriminator which determines that the level has deviated and in which direction;

(b) a pulse generator which maintains constant regulation timing;

(c) a regulating unit which controls the manipulated variable and retains its state indefinitely if the level does not change (memory);

(d) a control element which rapidly follows any changes in the manipulated variable.

In order that the regulation may reach a steady state, level discriminator must have two response thresholds—one for

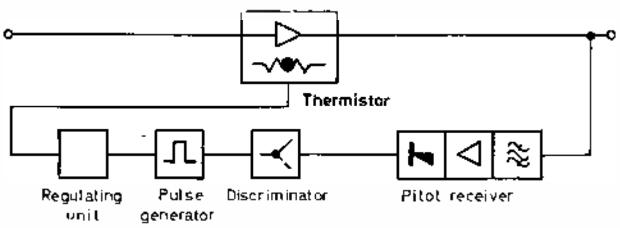
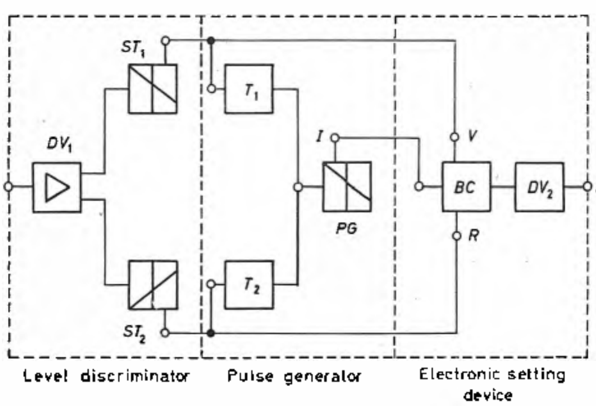


Fig. 1 Block schematic of a digital line regulator

Fig. 2 Scheme of a digital pilot regulator



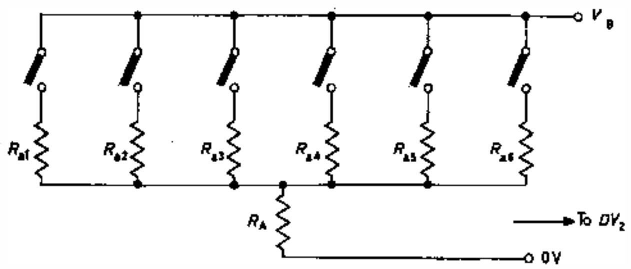


Fig. 3 Principle of regulating unit

positive and one for negative level changes—providing a tolerance range greater than one regulation step.

The discriminator (Fig. 2) consists of two Schmitt triggers (ST_1, ST_2) in front of which a differential amplifier DV_1 is connected. This amplifier compares the rectified diode (nominal value) with the voltage of a Zener. The Schmitt triggers are connected to the two amplifier outputs (collector contacts of the transistors). They operate only if the response thresholds are exceeded and then a voltage jump occurs at the input of the pulse generator. One of the Schmitt triggers responds to levels which are too high and initiates regulation in one direction and the other one operates on levels which are too low and initiates regulation in the other direction.

The differential amplifier DV_1 reduces errors in the level discriminator resulting from the temperature dependence of the Schmitt triggers and from their hysteresis. The function of the amplifier can be compared to that of a magnifying lens; it increases the immediate range about the nominal value of the pilot voltage by the amplification factor. A small deviation in the input voltage of the differential amplifier causes a large change in the voltage supplied to the Schmitt triggers, and, conversely, any error (effects of hysteresis, temperature dependence etc.) caused by the thresholds appears greatly reduced at the input of the differential amplifier.

The pulse generator PG—preferably a somewhat modified Schmitt trigger which does not need a regeneration period like an astable multivibrator—is controlled by the Schmitt triggers of the level discriminator *via* one of the two coupling networks T_1 or T_2. The generator only operates as long as one of the Schmitt trigger thresholds is exceeded and remains quiescent when the level is within the tolerance. The coupling networks cause a small delay during which the direction of regulation is determined. The pulse period which, together with the height of one regulation step, determines the regulation speed, is about 300ms. The pulse generator drives the regulating unit.

The principle of the regulating unit is shown in Fig. 3. The diagram shows that n controlled switches are connected in parallel, each being series-connected with a resistor (R_a to R_{an}), and this circuit is connected in series with a weighting resistor R_A to the voltage U_B. If a resistances ratio of $R_{a1}:R_{a2}:\ldots:R_{an} = 1:2:\ldots:2^n$ is chosen, the ratio of the partial currents $I_{an} : I_{a(n-1)} : \ldots : I_{a1}$ will also be $1 : 2 \ldots : 2^n$, providing the resistance R_A, compared with the resistances R_{a1} to R_{an}, is small and does not influence the partial currents I_{a1} to I_{an}. In the case of n switches, 2^n combinations of their switching states are possible. Each of these combinations supplies a total current which is an integral multiple of I_{an}. By adequate control of the switches, the total current is varied *via* R_A in equal steps.

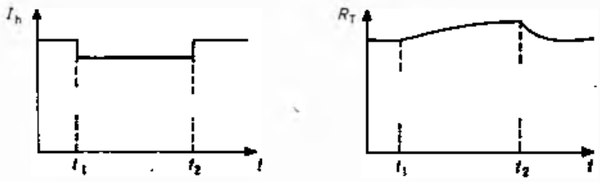


Fig. 4 Current and resistance of a thermistor with stepwise regulation, without acceleration

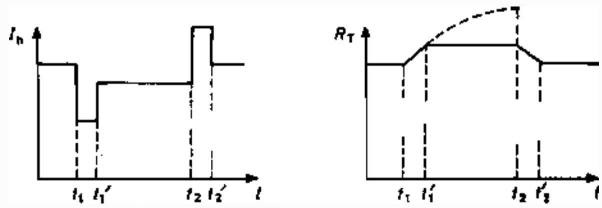


Fig. 5 Current and resistance of a thermistor with stepwise regulation, with acceleration

In the final regulating unit, the switches are transistor circuits and are made to function as bistable elements interconnected so that they form a reversible binary counter[3]. Each pulse from the generator causes a variation of the control current by one step. The Schmitt triggers in the level discriminator apply voltage either to terminal R (backward) or to terminal V (forward) and determine thereby whether the control current decreases or increases, depending on whether the level is too high or too low. The output current, variable in steps, of the binary counter BC controls a d.c. amplifier (R_A is the input resistance of this amplifier) which functions differentially (DV_2). The latter supplies the heating current for an indirectly heated thermistor (the control element). If a large level deviation has to be corrected if may happen that the binary counter reaches one of its two end positions. In this case the regulator must be stopped and an alarm must be given. If the level deviation changes its direction so that the binary counter can leave the end position, the regulator is automatically unblocked. If no level deviation occurs, no switching pulses are supplied to the counter and it retains its setting indefinitely. In the case of a sudden pilot failure (rapid level decreases), the pulse generator stops and the setting of the regulator does not change.

Indirectly heated thermistors are used as control elements through which a part of the signal current flows. The large thermal time-constant of the thermistor has the advantage of good harmonic suppression but, on the other hand, a limit is set to regulation speed. Owing to the large time-constant, the thermistor resistance follows a sudden change in the heating current only slowly, and it varies approximately in accordance with an e-function. Fig. 4 shows the time characteristics of the heating current I_H decreased or increased by one step at instants t_1 and t_2, respectively. It also shows the corresponding characteristic of the thermistor resistance. Such a behaviour requires a long pulsing period and leads consequently to low regulation speed. This disadvantage can be eliminated with stepwise regulation by the use of an acceleration circuit which supplies, to the thermistor heater from the regulating unit, an additional current pulse of high amplitude and short duration at each step of current. The thermistor resistance approaches, in accordance with an e-function, a steady-state value corresponding to the high current pulse. The amplitude and the duration of the acceleration pulses are chosen so that at the end of each pulse the change of thermistor resistance corresponds to the current step of the regulating unit. The maximum duration of the acceleration pulse may be the same as the pulse period of the switching pulse generator.

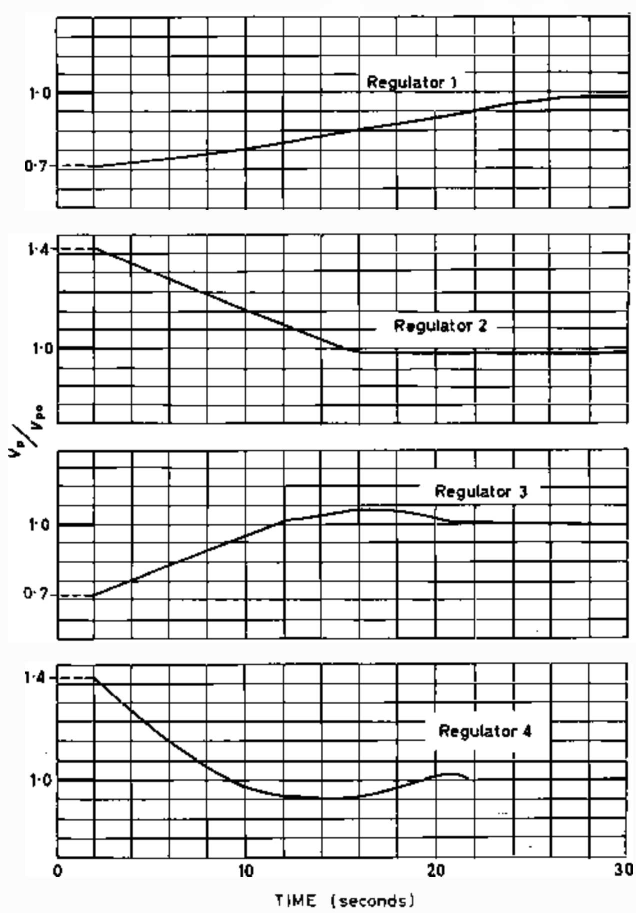
For heating the thermistor, a positive acceleration pulse is supplied to the heater and for cooling down a negative acceleration pulse is used. The change of heating current and of the thermistor resistance with respect to time, when an acceleration circuit is used, is shown in Fig. 5.

If, as in the case of line regulation discussed here, the pulse period of the switching pulse generator is of the same duration as the acceleration pulses, the thermistor resistance changes continually during a period of regulation. The acceleration is by means of the level discriminator of Schmitt triggers which, when they are actuated, apply an additional control voltage of adequate magnitude to the differential amplifier DV_2.

Testing the regulator and system is made easier if manual adjustment to any desired position is possible. For this the level discriminator must be disconnected from the pilot receiver and coupled to the output of the binary counter *via* a potentiometer. The regulator then reaches a state of equilibrium, during which the output voltage of the binary counter, mutiplied by the voltage divider ratio of the potentiometer, is equal to the response threshold of the level discriminator. By varying the setting of the potentiometer, the regulator can be forced into the desired position[4].

Theoretical investigations and experiments on models have shown[2] that the digital line regulator guarantees

Fig. 6 Level characteristics after the 1st, 2nd, 3rd and 4th regulator after a sudden level change of 3dB, measured on a model. V_p = pilot voltage; V_{po} = nominal value of the pilot voltage



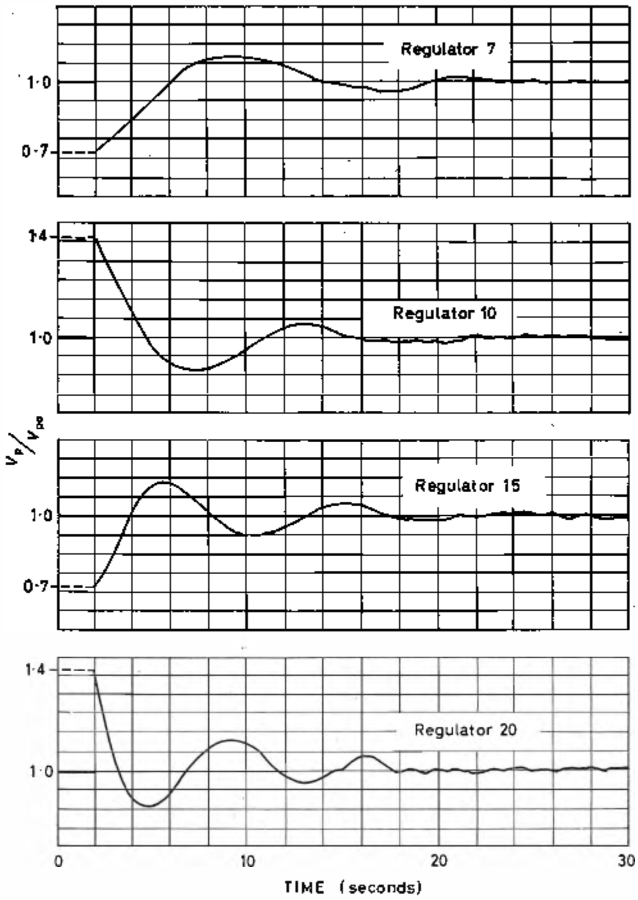


Fig. 7 Level characteristics after the 7th, 10th, 15th and 20th regulator after a sudden level change of 3dB, measured on a model. V_p = pilot voltage: V_{po} = nominal value of the pilot voltage

good transient behaviour when compensation of sudden level changes occurs. The analysis leads to a differential equation which is not solvable in a closed form; the solution for *n* regulators connected in tandem has however been evaluated by means of a computer. It was found that the first maximum value of transient does not exceed 40 per cent of the level change to be compensated, even when the number of regulators is infinite. The subsequent maxima decrease rapidly.

For verification of the theoretical results, measurements of the dynamic behaviour of a tandem network were made on a test model comprising digital regulators connected in tandem, supplemented by simulated line amplifiers, to form a complete closed-loop circuit. For this purpose it was sufficient to transmit one pilot frequency only, and the pilot filters could be dispensed with. The test arrangement consisted of 20 similar line regulators connected in tandem. Figs. 6 and 7 show the level values measured on the test model after the 1st, 2nd, 3rd, 4th, 7th, 10th, 15th and 20th regulator when compensating for a sudden level change of 3dB. The level change is alternately negative and positive. The scales are linear, and after conversion into dB, the first maximum of 40 per cent is obtained after the 10th regulator and 43 per cent of the level change after the 20th regulator. This shows clearly that the first maximum approaches an asymptotic value of slightly more than 40 per cent of the level change, as can be expected from the theoretical analysis.

The level change of 3dB is compensated after the 1st regulator within a period of 25s. Subsequently regulators reach the nominal value before the 1st regulator. For better comparison, the graph of the theoretical level calculated by assuming ideal conditions and the graph of the level measured at the test model after the 10th regulator, have been superimposed in Fig. 8. The discrepancy between the calculated and the measured levels results partly from the fact that a level discriminator without hysteresis has been assumed for the calculation, but the agreement is quite good.

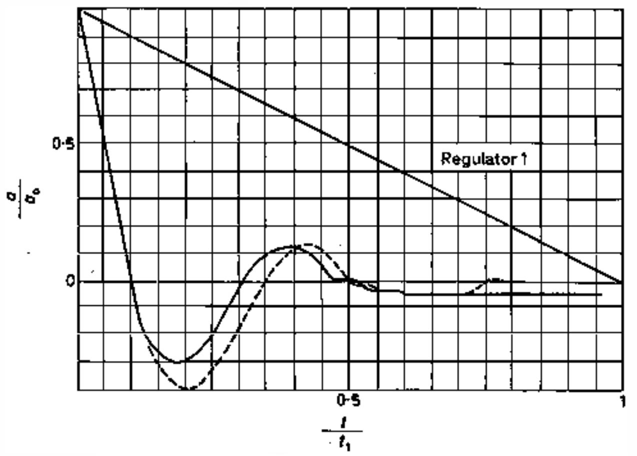
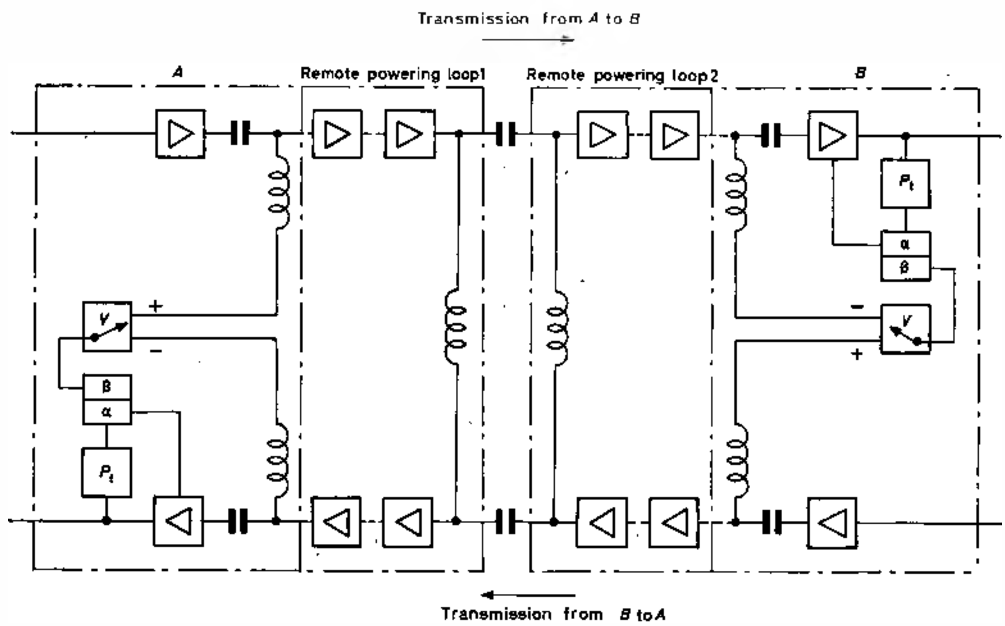


Fig. 8 Level characteristic after the 10th regulator. Continuous curve with zero hysteresis is calculated; dashed curve is measured

Fig. 9 Combination of forward and feedback control



Level Regulation by Open-Loop Control through the Cable Temperature

As has already been stated, it is advisable to combine closed-loop feedback control with a less expensive and less complicated open-loop control to regulate line level. The open-loop control system discussed in the following uses the cables themselves as temperature sensing elements and the remote power current sent along their inner conductors as control current. In Fig. 9 the remote power supply between two surface repeater stations (*A* and *B*), which are power-supply stations, is shown. Normally, a remote power loop extends over half the transmission section between two neighbouring power-supply stations. In these stations, remote power equipment (G_1 and G_2) produces a stabilized d.c. voltage U. The inductances L and the capacitances C indicate filters which separate the remote power d.c. from the carrier signals at the points shown.

If the operating voltages of the remotely powered repeaters are stabilized by Zener diodes, the remote power equipment voltage at the power-supply station is kept constant; the remote power current will be mainly determined by the temperature-dependent resistance of the inner conductors of the coaxial pairs. It will thus be a measure of the temperature-dependent cable loss variation.

Thermistors are used here as control elements. However, they are very sensitive to temperature changes and must be temperature-compensated so that they are affected by the control current only. Temperature compensation, in turn, must not affect the remote power current and must not affect the total resistance of the remote power loop. This means that none of those conventional temperature compensation methods can be used which have a network with a directly heated thermistor in parallel with the heater of the indirectly heated control thermistor.

A further difficulty arises from the fact that gain regulation is often in the negative feedback path of the amplifier. Networks have been found suitable here in which the gain decreases with increasing resistance of the control element and *vice versa*. Remote power current cannot therefore be used directly for control of the regulating thermistor as a temperature and loss increase would decrease the remote powering current and the thermistor resistance would increase and the amplifier gain would be reduced.

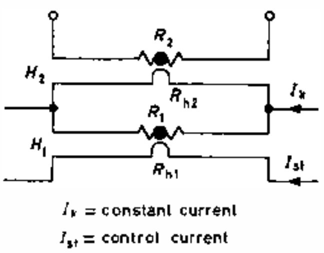
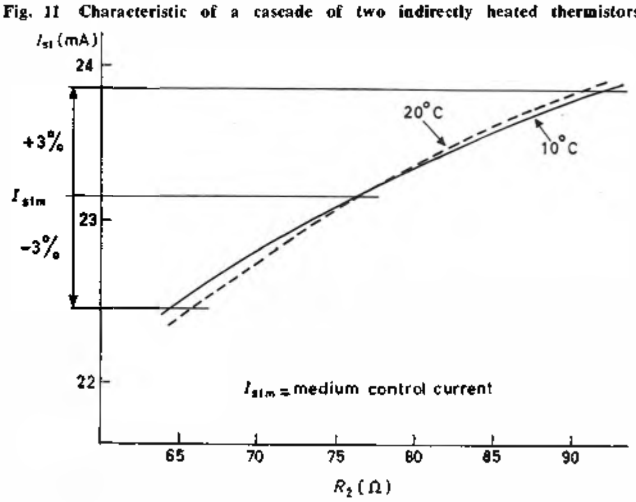


Fig. 10 Cascade of two indirectly heated thermistors

Fig. 11 Characteristic of a cascade of two indirectly heated thermistors



A solution has been found with a cascade of two indirectly heated thermistors[2] as shown in Fig. 10. The heater R_{h1} and the bead resistor R_1 belong to thermistor 1 and heater R_{h2} and the bead resistor R_2 to thermistor 2. The variable control current I_{st}, which is a portion of the remote power current, flows through R_{h1} while a constant auxiliary current I_k flows through $R_1 \| R_{h2}$. If I_{st} increases, the resistance of R_1 decreases at constant ambient temperature; a larger portion of the auxiliary current I_k flows through R_1 and R_2 is heated less and its resistance rises. With increasing ambient temperature and constant control current I_{st} the resistance of R_1 decreases, and again a larger portion of the auxiliary current I_k flows through R_1; R_2 is thus heated less so that the decrease in its resistance caused by the temperature rise is compensated. Thermistor 1 consequently not only causes an inversion of the temperature coefficient but also a compensation for changes in ambient temperature; thermistor 2 is the actual control element. The power loss in the bead of thermistor 2 is negligible.

Fig. 11 shows the resistance characteristic of a cascade network of two indirectly heated thermistors with small current changes and at ambient temperatures of 10°C and 20°C. If the control current varies by ± 3 per cent (this corresponds approximately to a temperature change of ± 10 degC in the cable), the resistance of thermistor bead 2 changes by about ± 19 per cent. This change in resistance is sufficient to vary the gain of the line amplifiers. It should be noted that there is no great difference between the characteristics for 10°C and 20°C.

In the regulation system described a network of cascaded thermistors is used in each underground repeater. Fig. 12 is an equivalent circuit diagram of the d.c. path in the underground repeater. Remote power is supplied by a series-parallel feed and the amplifiers of the go and the return direction in each repeater are in parallel with the d.c. circuit. All underground repeaters are series connected in the remote power loop. The power supply voltage of both amplifiers is stabilized by a common Zener diode Z. The heaters R_{h1} of the two thermistor networks are connected in parallel and the $R_1 \| R_{h2}$ circuits (see Fig. 10) are in series with a common limiting resistor R_v. A parallel resistor R_p bypasses the remote power current. The supply voltage for the auxiliary current I_k is likewise stabilized by Zener diode Z.

Open-loop control of level in all underground repeaters has the advantage that all repeaters can be of the same design and that each need compensate only for the loss variation of one repeater section, so that the necessary regulation range is small. The $\sqrt{f}$-networks can be made more easily and with less error. By the choice of suitable Zener diodes (temperature coefficient, differential resistance) the total voltage drop at an underground repeater will remain constant, irrespective of variation in the remote power current due to climatic changes, provided that the temperature variation in the cable and in the underground repeater is the same. This means that the variation in transmission level caused by the temperature-dependent copper loss of the cable can be exactly compensated by the change of the power-feed current.

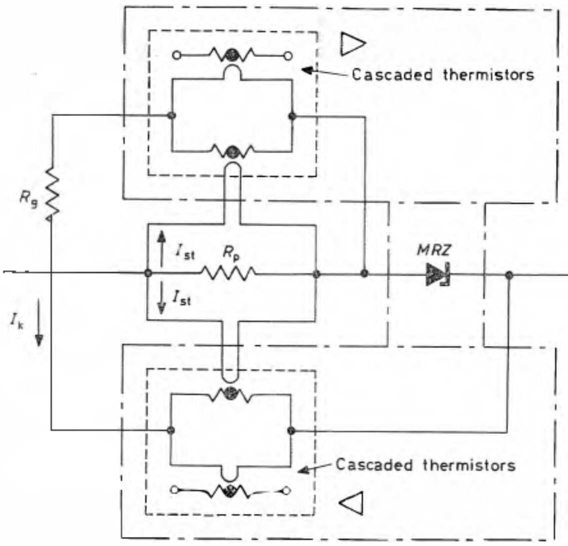


Fig. 12 D.C. circuit of underground repeater

As, however, the temperature changes of the cable and of the repeaters are different[5] and for other reasons, open-loop control alone is not capable of compensating for all loss variations; level errors are left which have to be eliminated by the pilot-controlled regulators. Assuming a maximum difference of 5 degC between the temperature of the cable and that of the repeater housing and assuming further that this temperature difference is the same at all 20 underground repeaters of a section between two adjacent power supply stations, then the maximum error that accumulates at the end of this section is (in the V300 system) 1·4dB. If the temperature difference of the individual repeater sections has different signs, the total error is, of course, smaller.

It is also possible to obtain the auxiliary current I_k in parallel with resistor R_p instead of in parallel with the Zener diode Z and remote supply power is saved. In this case resistor R_g is small and the auxiliary circuit shows a noticeable negative temperature coefficient. As a result, the control current I_{st} of the thermistor cascades depends more on the ambient temperature, i.e. their characteristic becomes steeper. This may be of advantage with longer repeater sections where larger loss variations have to be compensated. Another advantage of open-loop level regulation by remote power is that during lining up of new cable routes, systematic errors from changes of the mean remote power current can be easily eliminated.

In densely populated and highly industrialized areas it often happens that long power lines and carrier frequency cables are arranged in parallel. The electromagnetic field surrounding the power lines induces a.c. into coaxial cables and the magnitude depends on many factors. With normal operation it is smallest in the inner conductor. The interfering frequencies are 50 or 16·66Hz. It is extremely difficult to suppress them in the amplifiers with the aid of smoothing circuits. Again, remote supply current for the amplifiers must not be too strongly influenced by induced current, because of disturbance to the amplifier. Hence induction in inner conductors should be minimized and a possible way is to operate at a float-in potential, where the outer conductors of the cables are not grounded. Measurements at different test routes have shown that the induction currents in inner conductors are only a few milliamps with floating potentials and a short circuit on the power line.

No additional efforts are required to render the effect of induction currents on level regulation negligible. The r.m.s. value of the control current for level regulation is obtained from the remote supply current and the induced a.c. Let I_g be the remote supply current and $I_b = aI_g$ the induction current, where $a \ll 1$. The resulting r.m.s. control current is given by:

$$I_r = \sqrt{(I_g^2 + I_b^2)} = I_g \cdot \sqrt{(1 + a^2)} \simeq I_g(1 + 0{\cdot}5a^2)$$

In the V300 system (300 telephone channels on small-core coaxial cables) $I_g = 125$mA. Assuming an induced current of 5mA, then $a = 0{\cdot}04$ and $I_r = 1{\cdot}0008\,I_g$. The resulting level error per amplifier amounts to 0·016dB for this system even with continuous interference and such errors are negligible.

Level Regulation by Co-ordination of Feedback and Open-Loop Control

For reducing level errors left by open-loop control, feedback and open-loop control are co-ordinated so that the pilot-controlled regulator at a power-feed station controls not only the line amplifier at that station but also the voltage U of the remote power loop of the preceding power supply section[2]. The remote power current is so changed that, as shown in Fig. 9, all thermistor networks of individual repeaters in that power loop are equally adjusted to reduce the level error. This means that the pilot-controlled regulator assigned to one direction of transmission controls a remote power loop for both directions of transmission.

This correlation of the control function of the pilot-controlled regulators, provided at two successive power supply stations and assigned to different directions of transmission, gives a closed loop. A gain change initiated by the pilot-controlled regulator in direction A-B at station B causes, through the gain change of the amplifiers in direction B-A, a level change at station A. This change, again, is sensed by a pilot-controlled regulator at station A and causes a level change at station B. In order to keep this mutual influence to non-critical limits, compensation of level deviation must be shared between the pilot-controlled amplifiers at the power-supply stations and the amplifiers controlled by the d.c. in the remote power loop. If β is the amount by which the regulating networks in underground repeaters contribute to the total level regulation then it is necessary[2] that $\beta \leqslant 0{\cdot}5$.

By the combination of feedback and open-loop control, the error of open-loop control is reduced[2] by $\frac{1}{2}$ to $\frac{1}{3}$. It is only this small residual that must be compensated by the pilot-controlled regulator. Since over one section the temperature changes of the two coaxial lines and the remaining system errors in the go and return direction are equal, the mutual effect in the two directions causes an equal correction of the remaining errors in all underground repeaters. Errors occurring at one point only and in one direction of transmission will also influence the other direction of transmission, but this influence is not critical when a proper choice of β is made.

In order to avoid temperature changes occurring at the power-feeding station affecting regulation through the remote power loop, temperature compensation of thermistor control elements in line amplifiers at the power supply stations is necessary[5].

The cascade networks of thermistors in the remotely powered repeaters only slowly follow a change in current because of their very large time constant, while regulation at the power supply station takes place very rapidly due to acceleration. This means that a sudden change in level will be corrected first by the regulators at the power-feeding stations and the remotely powered repeaters will take over part of the compensation subsequently. Level regulation by open-loop control will thus not affect the dynamic behaviour of the system.

The factor α (see Fig. 9) in pilot-controlled regulation on all surface repeaters is made the same in all such repeaters. The factor β in pilot-controlled regulation at all remotely powered repeaters in the appropriate remote power loop must therefore be varied in accordance with the length of the section between the surface stations. If the section is long an $\alpha : \beta$ ratio of 60 : 40 is chosen; in the case of short sections and remote power loops the factor β will be correspondingly decreased. The system errors in shorter sections will in this case be reduced to a lesser extent. However, since errors in the shorter sections are smaller, the degree of regulation will remain the same, independent of the route length. This arrangement involves no difficult design problem, as all pilot-controlled regulators must have the same characteristics. Independently of the number of remotely powered repeaters, setting up adjustments are made so that in all cases the effect of the pilot-controlled regulator on the remote power current is the same. A certain change in gain in the fully regulated amplifier at the power-feeding station always causes a fixed change in the remote power current.

The effect of the pilot-controlled regulator on the remote power equipment is produced by a voltage from the regulating unit proportional to the state of this unit; the voltage is applied to a comparator in the remote power unit.

In order to ensure that the regulator will retain its instantaneous state in case of a pilot failure or a large level drop, a Schmitt trigger is included after the pilot receiver. The threshold values of this trigger circuit are about 3dB lower than the nominal pilot level. If the pilot fails, the Schmitt trigger initiates an alarm and blocks the regulator. When the regulator is put in service initially or when it is restarted after an operating voltage failure, it may happen that the binary counter stops at a position which corresponds to a very low level and the regulator is blocked. In order to avoid permanent blocking after restoration of the operating voltage, for a short time the regulator is driven so that the associated line amplifier is at maximum gain. Initially, the thermistor is cold and has a high resistance, the gain is low and the regulator is blocked. After a short time, the thermistor heats up to such an extent that the pilot level changes by less than 3dB from the nominal value; the regulator is then unblocked and adjusts itself in a short time and the regulated line amplifier output returns to the nominal level.

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Electroluminescent Panels for Display of Complex Systems

Electroluminescent panels are already well known and widely used for such purposes as instrument lighting, background illumination in aircraft cockpits, signs and simple indicators, particularly where safety, light weight, small size and low power consumption are important. Electroluminescent mimic diagrams offer a solution to the severe data-display problems presented by industrial process control and the monitoring of elaborate systems, and the electroluminescent-tile production unit in the Chemical and Metallurgical Division of Plessey Components Group, Towcester, has been established for the production of elaborate mimic diagrams in the form of tiles. Each tile consists essentially of a parallel-plate capacitor of which the luminescent layer forms the dielectric, the whole being sandwiched between protective sheets of glass. The tile is built up in layers on the (front) glass substrate and in its simplest form consists of a uniform square of emitting surface. However, it is possible to produce a pattern in the transparent conducting electrode before applying the phosphor layer, although in the majority of tiles used on mimic diagrams it is normal to produce the pattern in the rear electrode using masking techniques.

By changing the phosphor, it is possible to produce luminescence in any of five colours: red, green, blue, yellow and orange. It is thus possible to produce several independently switchable patterns in different colours on a single tile. The simplest kind of sign can be made by lettering or drawing on the front glass of the tile, and the same principle is used in even the most elaborate mimic diagrams. Artwork is prepared on a glass or Perspex panel and electroluminescent tiles are fitted in a frame behind it, shining through the overlay. The use of multipattern tiles enables adjacent areas of the diagram to be discretely illuminated, and the range of colours allows separate streams in a complex system to be clearly distinguished.

The illuminated areas can be as narrow as 0·02in, spaced 0·01in apart: so that very large and elaborate systems can be represented on a panel of modest dimensions, without confusing the viewer or making it difficult to extract data.

The system admits the display of information in several forms—graphs, schematics and pictures of sections of plant or equipment. Alphanumeric displays can be built up by illuminating selected elements of a matrix to form the character. Although the complexity of the necessary logic at present limits the type of usage, development of solid-state switching is expected to open the way to large-scale exploitation of the technique within a few years.

An important feature of mimic diagrams is their ability to show movement and flow—the line or area is divided into a series of small and independently switched sections. These sections are connected to, and sequentially switched by, an animator unit comprising a 4-stage ring counter operating four relays for direct switching. The animator is a self-contained unit that runs on 250V 50Hz and provides for the speed to be varied over a wide range.

The tiles are normally 9in square and operate on 115-400V, 50-400Hz, with an average drain of 0·8mA/in^2 of illuminated area at 400V, 400Hz. Theoretically, a sinewave should be applied, but in practice a squarewave is normally used, since this is easier to generate. A range of static invertors enables operation from low-voltage d.c. sources. Several switching methods are available, making the mimic diagram compatible with any logic system and any type of transducer. Illumination is uniform over the entire lit area, and frequent switching has no effect on brightness or life.

The panels are typically less than 9·5mm thick, are extremely reliable and produce no heat. They present no hazard from explosion or implosion, nor are any noxious substances released if the tiles are accidentally broken. These characteristics make electroluminescent mimic diagrams particularly suitable for high-density control centres.

Analysis of Feedback Amplifiers

by G. J. Pridham*

A common approach for series and parallel feedback systems is illustrated. Voltage and impedance are used for the series, and current admittance for the parallel case.

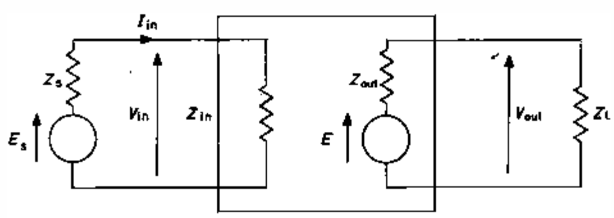
(Voir page 475 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 476)

THE established theory of feedback deals with a voltage proportional to the output voltage or current being fed back in series with the input of a voltage amplifier. With the advent of transistors this should be modified to include the case of a current proportional to the output voltage or current being fed back in parallel with the input of a current amplifier. If voltage gain and impedances are considered in the first case and current gain and admittances in the second case, similar expressions may be easily derived.

SYMBOLS

$A\angle\theta$	= Overall gain of voltage amplifier without feedback
A'	= Overall gain of voltage amplifier with feedback
$A_I\angle\theta$	= Overall gain of current amplifier without feedback
A_I'	= Overall gain of current amplifier with feedback
β	= Feedback factor
E	= Output voltage with load open-circuited
E_s	= Source e.m.f.
I	= Output current with load short-circuited
I_s	= Source current
I_{in}	= Input current
I_{out}	= Output current
$\mu\angle\alpha$	= E/V_{in}
$\mu_I\angle\alpha$	= I/I_{in}
V_{in}	= Input voltage
V_{out}	= Output voltage
Y	= Part of voltage feedback network
Y_f	= Feedback admittance
Y_s	= Source admittance
Y_L	= Load admittance
Y_{in}	= Input admittance without feedback
Y_{in}'	= Input admitance with feedback
Y_{out}	= Output admittance without feedback
Y_{out}'	= Output admittance with feedback
Z	= Part of current feedback network
Z_1	= Part of voltage feedback network
Z_2	= Part of voltage feedback network
Z_s	= Source impedance
Z_{in}	= Input impedance without feedback
Z_{in}'	= Input impedance with feedback
Z_{out}	= Output impedance without feedback
Z_{out}'	= Output impedance with feedback
Z_L	= Load impedance

Voltage Amplifier Without Feedback



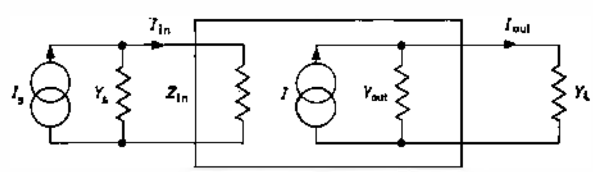
(a)

Fig. 1(a) Voltage amplifier without feedback

For the simple voltage amplifier shown in Fig. 1(a), the input voltage is given by:

$$V_{in} = \frac{E_s Z_{in}}{(Z_{in} + Z_s)} = \frac{E_s}{(1 + Z_s/Z_{in})}$$

Current Amplifier Without Feedback



(b)

Fig. 1(b) Current amplifier without feedback

For the simple current amplifier shown in Fig. 1(b) the input current is given by:

$$I_{in} = \frac{I_s Y_{in}}{(Y_{in} + Y_s)} = \frac{I_s}{(1 + Y_s/Y_{in})}$$

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The output e.m.f. is then

$$E = \frac{\mu \angle\alpha\, E_s}{(1 + Z_s/Z_{in})}$$

and the output voltage is given by

$$V_{out} = \frac{EZ_L}{(Z_{out} + Z_L)}$$

$$= \frac{\mu \angle\alpha\, E_s}{(1 + Z_s/Z_{in})\,(1 + Z_{out}/Z_L)}$$

Hence the overall voltage gain without feedback is

$$A \angle\theta = V_{out}/E_s = \frac{\mu \angle\alpha}{(1 + Z_s/Z_{in})\,(1 + Z_{out}/Z_L)}$$

If the input impedance is very high, as in the case of valves or field-effect resistors, this expression represents the voltage gain of the amplifier itself.

Voltage Amplifier With Series Feedback

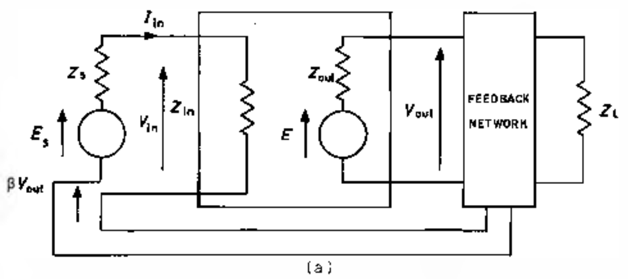


Fig. 2(a) Voltage amplifier with feedback

Fig. 2(a) shows a voltage amplifier with a proportion of the output voltage fed back in series with the input.

The input voltage is then

$$V_{in} = \frac{E_s + \beta V_{out}}{(1 + Z_s/Z_{in})}$$

and the output voltage may easily be shown to be

$$V_{out} = \frac{\mu \angle\alpha\,(E_s + \beta V_{out})}{(1 + Z_s/Z_{in})\,(1 + Z_{out}/Z_L)}$$

i.e.

$$V_{out} = A \angle\theta\,(E_s + \beta V_{out})$$

and the overall voltage gain

$$A' = V_{out}/E_s$$

or

$$A' = \frac{A \angle\theta}{1 - \beta A \angle\theta}$$

Input Impedance With Series Feedback

From Fig. 2(a)

$$E_s + \beta V_{out} = I_{in}\,(Z_s + Z_{in})$$

i.e.

$$E_s + \frac{\beta A \angle\theta\, E_s}{1 - \beta A \angle\theta} = I_{in}\,(Z_s + Z_{in})$$

or

$$\frac{E_s}{1 - \beta A \angle\theta} = I_{in}\,(Z_s + Z_{in})$$

therefore $E_s/I_{in} = (1 - \beta A \angle\theta)\,(Z_s + Z_{in})$

The input impedance with series feedback is therefore given by

$$Z_{in}' = E_s/I_{in} - Z_s$$

$$= Z_{in}\,(1 - \beta A \angle\theta) - Z_s \beta A \angle\theta$$

Hence the input impedance is increased for negative series voltage feedback.

The output current with load short-circuited is then

$$I = \frac{\mu_I \angle\alpha\, I_s}{(1 + Y_s/Y_{in})}$$

and the output current under normal conditons is given by

$$I_{out} = \frac{IY_L}{(Y_{out} + Y_L)}$$

$$= \frac{\mu_I \angle\alpha\, I_s}{(1 + Y_s/Y_{in})\,(1 + Y_{out}/Y_L)}$$

Hence the overall current gain without feedback is

$$A_I \angle\theta = I_{out}/I_s = \frac{\mu_I \angle\alpha}{(1 + Y_s/Y_{in})\,(1 + Y_{out}/Y_L)}$$

If the input admittance is very high, as in the case of bipolar transistors, this expression represents the current gain of the amplifier itself.

Current Amplifier With Parallel Feedback

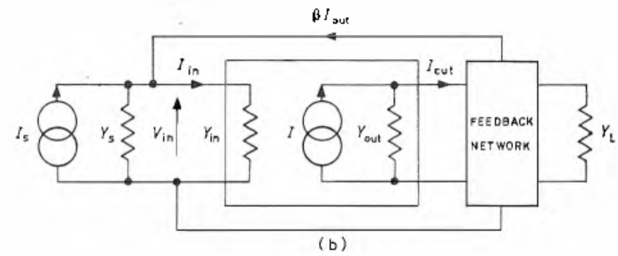


Fig. 2(b) Current amplifier with feedback

Fig. 2(b) shows a current amplifier with a proportion of the output current fed back in parallel with the input.

The input current is then

$$I_{in} = \frac{I_s + \beta I_{out}}{(1 + Y_s/Y_{in})}$$

and the output current may easily be shown to be

$$I_{out} = \frac{\mu_I \angle\alpha\,(I_s + \beta I_{out})}{(1 + Y_s/Y_{in})\,(1 + Y_{out}/Y_L)}$$

i.e.

$$I_{out} = A_I \angle\theta\,(I_s + \beta I_{out})$$

and the overall current gain

$$A_I' = I_{out}/I_s$$

or

$$A_I' = \frac{A_I \angle\theta}{1 - \beta A_I \angle\theta}$$

Input Admittance With Parallel Feedback

From Fig. 2(b)

$$I_s + \beta I_{out} = V_{in}\,(Y_s + Y_{in})$$

i.e.

$$I_s + \frac{\beta A_I \angle\theta\, I_s}{1 - \beta A_I \angle\theta} = V_{in}\,(Y_s + Y_{in})$$

or

$$\frac{I_s}{1 - \beta A_I \angle\theta} = V_{in}\,(Y_s + Y_{in})$$

therefore $I_s/V_{in} = (1 - \beta A_I \angle\theta)\,(Y_s + Y_{in})$

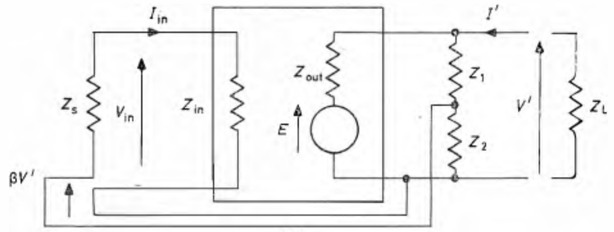
The input admittance with parallel feedback is therefore given by

$$Y_{in}' = I_s/V_{in} - Y_s$$

$$= Y_{in}\,(1 - \beta A_I \angle\theta) - Y_s \beta A_I \angle\theta$$

Hence the input admittance is increased for negative parallel current feedback.

Output Impedance With Series Voltage Feedback



(a)

Fig. 3(a) Circuit to determine output impedance with series voltage feedback

Fig. 3(a) illustrates a circuit where a voltage V' is applied to the output terminals of a voltage amplifier with series voltage feedback.

To a close approximation the feedback ratio is given by

$$\beta = \frac{Z_2}{Z_1 + Z_2}$$

and a proportion $\beta V'$ of the applied voltage is fed back to the input circuit.

Then output e.m.f.

$$E = \frac{\beta\mu \angle\alpha\, V'}{(1 + Z_s/Z_{in})}$$

and the current from the voltage source is given by

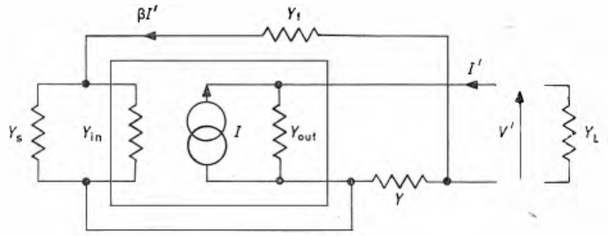
$$I' = 1/Z_{out}\left[V' - \frac{\beta\mu \angle\alpha\, V'}{(1 + Z_s/Z_{in})}\right]$$

The output impedance with series voltage feedback is

$$Z_{out}' = V'/I' = \frac{Z_{out}}{1 - \dfrac{\beta\mu \angle\alpha}{(1 + Z_s/Z_{in})}}$$

The output impedance is thus reduced for negative series voltage feedback.

Output Admittance With Parallel Voltage Feedback



(b)

Fig. 3(b) Circuit to determine output admittance with parallel voltage feedback

Fig. 3(b) illustrates a circuit where a current I' is applied to the output terminals of a current amplifier with parallel voltage feedback.

To a close approximation the feedback ratio is given by

$$\beta = \frac{Y_f}{Y + Y_f}$$

and a proportion $-\beta I'$ of the applied current is fed back to the input circuit.

Then the generated current

$$I = -\frac{\beta\mu_I \angle\alpha\, I'}{(1 + Y_s/Y_{in})}$$

and the voltage required to produce this current is given by

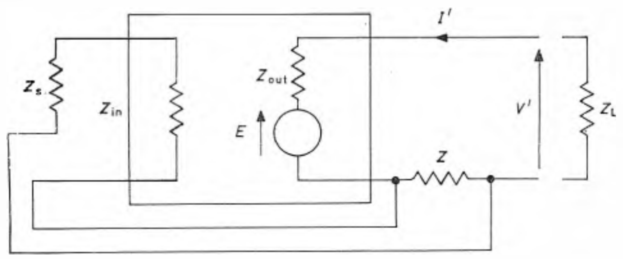
$$V' = 1/Y_{out}\left[I' - \frac{\beta\mu_I \angle\alpha\, I'}{(1 + Y_s/Y_{in})}\right]$$

The output admittance with parallel voltage feedback is

$$Y_{out}' = I'/V' = = \frac{Y_{out}}{1 - \dfrac{\beta\mu_I \angle\alpha}{(1 + Y_s/Y_{in})}}$$

The output admittance is thus reduced for negative parallel voltage feedback.

Output Impedance with Series Current Feedback



(a)

Fig. 4(a) Circuit to determine output impedance with series current feedback

Fig. 4(a) illustrates a circuit where a voltage V' is applied to the output terminals of a voltage amplifier with series current feedback.

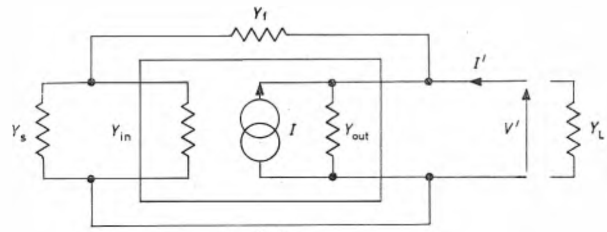
To a close approximation the feedback ratio is given by

$$\beta = \frac{Z}{Z + Z_L}$$

and a voltage $(-I'Z)$ is fed back to the input circuit. Hence the current I' flowing from the source is given by

$$I' = \frac{V' - \dfrac{\mu \angle\alpha\,(-I'Z)}{(1 + Z_s/Z_{in})}}{Z_{out}}$$

Output Admittance with Parallel Current Feedback



(b)

Fig. 4(b) Circuit to determine output admittance with parrallel current feedback

Fig. 4(b) illustrates a circuit where current I' is applied to the output terminals of a current amplifier with parallel current feedback.

To a close approximation the feedback ratio is given by

$$\beta = \frac{Y_f}{Y_f + Y_L}$$

and a current $V'Y_f$ is fed back to the input circuit. Hence the voltage V' required to produce the current I' is given by

$$V' = \frac{I' + \dfrac{\mu_I \angle\alpha\, V'Y_f}{(1 + Y_s/Y_{in})}}{Y_{out}}$$

and the output impedance with series current feedback is

$$Z_{out}' = V'/I'$$
$$= Z_{out} - \frac{\mu \angle\alpha Z}{(1 + Z_s/Z_{in})}$$

i.e. the output impedance is increased for negative series current feedback.

and the output admittance with parallel current feedback is

$$Y_{out}' = I'/V'$$
$$= Y_{out} - \frac{\mu_I \angle\alpha Y_f}{(1 + Y_s/Y_{in})}$$

i.e. the output admittance is increased for negative parallel current feedback.

The Plessey Product-Assessment Laboratories

In June 1963, the Environmental Laboratory, which until then had been sited at Ilford, moved to Titchfield in Hampshire. From that time the laboratory's work expanded rapidly, not only in volume but also in scope. As a result it soon became clear that the name had become a complete misnomer. Late in 1966, a decision was taken to change the name. From a number of possibilities, Product Assessment Laboratories was finally chosen as being the name most representative of the work involved.

The Product Assessment Laboratories are approved by the Ministry of Aviation and the Air Registration Board. From the start, the management recognised the importance of independence in promoting customer confidence at all levels both inside and outside the parent company. As a result, the laboratories are independently controlled and enjoy the respect of approval authorities for the accuracy and impartiality of their reports. A service in the product-evaluation field is provided to The Plessey Co. as a whole; this broadly based service is also available to British and overseas industry on a confidential and exclusive basis.

The Laboratories have been testing components for 14 years. During this period they have handled nearly all types of electronic component and tested them to the majority of ruling specifications. The work covers both electrical and environmental evaluation at all stages from the birth of a new product. Included are development and approval tests to assess intrinsic capability, acceptance sampling as an aid to the manufacture of consistent and reliable products, and help with the diagnosis of component failure.

Perhaps the most useful service to the equipment designer is advice on mechanical construction that affords maximum protection against environmental stress. Of particular interest is the use as a design aid of photostress techniques, to show both transient and permanent deformation of structures under mechanical stress. An example of the laboratories' successful use of this technique is the detailed design optimization for a range of instrument cases. A manufacturer sometimes finds that his product is damaged and unusable on reaching the customer at home or in a valuable export market. It is not uncommon for such catastrophes to be caused by failure to recognise the importance of environment-resistant design. Here, considerable help can be given to rectify the conditions without the introduction of complex and costly modifications; the laboratories are usually able, not only to pinpoint the cause, but also to design, fit and prove modifications.

A comprehensive range of environmental chambers allows simulation of any climate from the Equator to the Poles, at sea level or at many tens of thousands of feet above the Earth's surface. Special-purpose plant is held which can, for example, reproduce vibration levels and frequencies met in an intercontinental ballistic missile, the shocks experienced during railway shunting or the sustained accelerations resulting from high-speed manoeuvres in supersonic aircraft.

A new $27ft^3$ panclimatic chamber will soon be installed which can combine a pressure range from ambient to the equivalent of 100 000ft altitude with temperature in the range -75 to 150°C and humidity from 15% R.H. to saturation. Its integrated programming and recording arrangements for each parameter are believed to be unique.

The need to carry out development work in the laboratories springs from the constant stream of new specifications. New documents often carry novel test requirements which cannot be met by existing plant or test equipment; neither can suitable apparatus necessarily be bought out as a standard item. The only solution, to prove a product to the specification in the minimum time, is to develop equipment. The small development section is mainly concerned with the design and prototype construction of special-purpose environmental plant. Developments include shock and bump machines, centrifuges, humidity chambers, leakage equipment and thermal-cycling cabinets. This concentration on environmental equipment has not prevented contributions in other areas. In the field of precision electrical measurements, the laboratories have developed a dielectric test set that allows measurement up to 200°C at 100-500MHz. A small number of these instruments have been made and sold in Britain and abroad.

Project engineers make full use of specialized services provided within the Laboratories as an adjunct to economic operation. Mention has already been made of the various environments which are provided by the largest of the specialist groups. Another section provides photographic and reprographic services, not only to the Laboratories but also to a number of other Plessey establishments over a large area in the south of England. Black-and-white or colour photography, in cine or still, is widely used in a variety of applications. Among the more sophisticated equipment is a high-speed cine camera which allows the detailed behaviour of high-speed mechanisms to be examined and analysed. This equipment has many applications in design, development and performance investigation. Microsectioning and microhardness testing are also undertaken as an internal service. The main uses are as an aid to quality control and in failure investigation.

A major reliability hazard in the life of any equipment is the 'rogue' component. Its existence is a costly nuisance to the equipment manufacturer who, even if the direct cost is covered by guarantee, is still faced with considerable expense for diagnosis and replacement. The rogues which survive until after delivery, besides affecting the early reliability of the system, become even more expensive to replace, imply an unacceptable level of serviceability and tend to destroy the confidence of the user in the equipment and its maker. The Laboratories are engaged in a programme which seeks to stop the rogues from ever leaving the component-manufacturer's premises. The method is in the design and application of 100% nondestructive tests at the end of the production line.

A computer-controlled and automated test equipment was installed at Titchfield in January 1967. Component parameters are measured on digital instruments whose output is transferred to the computer for checking, correlation and statistical analysis. Input to the measuring instruments is via a multiway switch which is stepped under computer-program control. The computer feeds a teleprinter providing both printed format and corresponding punched paper tape. In addition to any parameters which can be converted to a voltage analogue, the system gives precise measurements of resistance deviation, capacitance and dissipation factor. The primary function is measurement and analysis during long-term reliability trials of electronic components. However, the ability to make rapid accurate measurements, combined with real-time checking and correlation capacity, makes it a forceful tool in many other applications. It is hoped to use the equipment for parameter measurements and correlation during component development programmes and as a support to the reliability studies carried out in the Laboratories. Adequate capacity exists for the immediate future.

A Simple Recorder Integrator Using a Unijunction Transistor

By A. R. Atkins*

An integrator suitable for determining the mean of a varying recorded quantity is described. The accuracy derives from the very stable characteristics of a unijunction transistor as a relaxation oscillator. The integrator in this particular case was designed to determine the mean recorded radiation and convection heat losses from a human body in a climatic chamber.

(Voir page 475 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 476)

IT is often necessary to obtain the mean of a varying signal, which in many cases is recorded. This signal may represent the weight of substance in a process, a temperature of a furnace, water flow or power input to a system. One method, assuming there is a linear relationship between the variable to be measured and the signal, is to determine the area under the recorded value then divide by the length of the base-line. If there are large and rapid variations it is not easy to obtain an accurate result and in any case it is tedious.

A simple integrator is described, with an accuracy of the same order as a potentiometer recorder, which will give in digital form an integrated value of the signal for any given time. If the time is fixed then the integrated scale can be suited to the input or recorder scale. Although this integrator was designed primarily for a potentiometric recorder, it does not have to be used with one and in many cases could replace it if only the mean is of interest.

Principle of Operation

Let V_i (Fig. 1) be the potential which has to be integrated. It is derived from a slave potentiometer on a recorder where the output V_i is proportional to the recorder indicator. The integrator will then produce a rising output voltage V_o, the slope of which is directly proportional to V_i. The integrator is coupled to a voltage-sensitive switch which triggers when V_o reaches a certain value V_s, and then rapidly resets the integrator output to zero. The cycle is repeated and V_o restored to zero each time it rises to a value V_s. The frequency of the cycling integrator will therefore depend on the slope of the integrator output waveform which is directly proportional to the input voltage V_i. The total number of cycles, measured by the counter, will therefore be the integral of V_i. This technique is sometimes used for digital voltmeters. If the time of integrating is fixed then the frequency count is proportional to the mean input voltage.

When the input voltage to the integrator is small, the cycling frequency is very slow. A longer integrating time is therefore required to obtain a reasonable resolution. This is arranged by biasing the input voltage. For example, when the recorder is at zero the integrator count is set at 100 and when the recorder is at full-scale the integrator count is 200.

* *Chamber of Mines Research Laboratories, South Africa.*

The accuracy of the integrator depends on three factors; (a) a linear relation between the integrator output and the input voltage V_i, (b) the stability of the voltage-sensitive switch and (c) the time taken for the switch to restore the integrator to zero.

There are two ways of producing such an integrator. The most common technique is the Miller integrator illus-

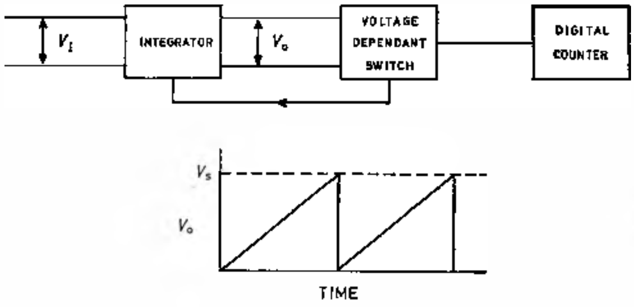


Fig. 1 **Integrator principle**

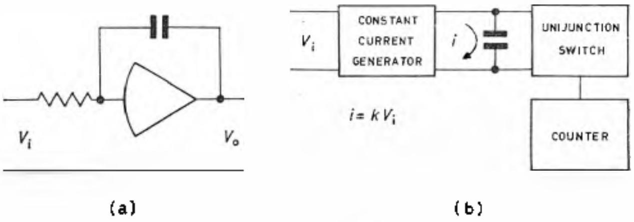


Fig. 2 **Miller integrator**

trated in Fig. 2a. An alternative method is to charge a capacitor with a current i which is proportional to V_i, Fig. 2b, so that

$$i = k\,V_i \quad (k \text{ is a constant})$$

The voltage across the capacitor is given by

$$V_o = (1/C) \int idt$$

and substituting for i

$$V_o = (1/C) \int k\,V_i\,dt.$$

The output voltage is therefore proportional to the integral of the input.

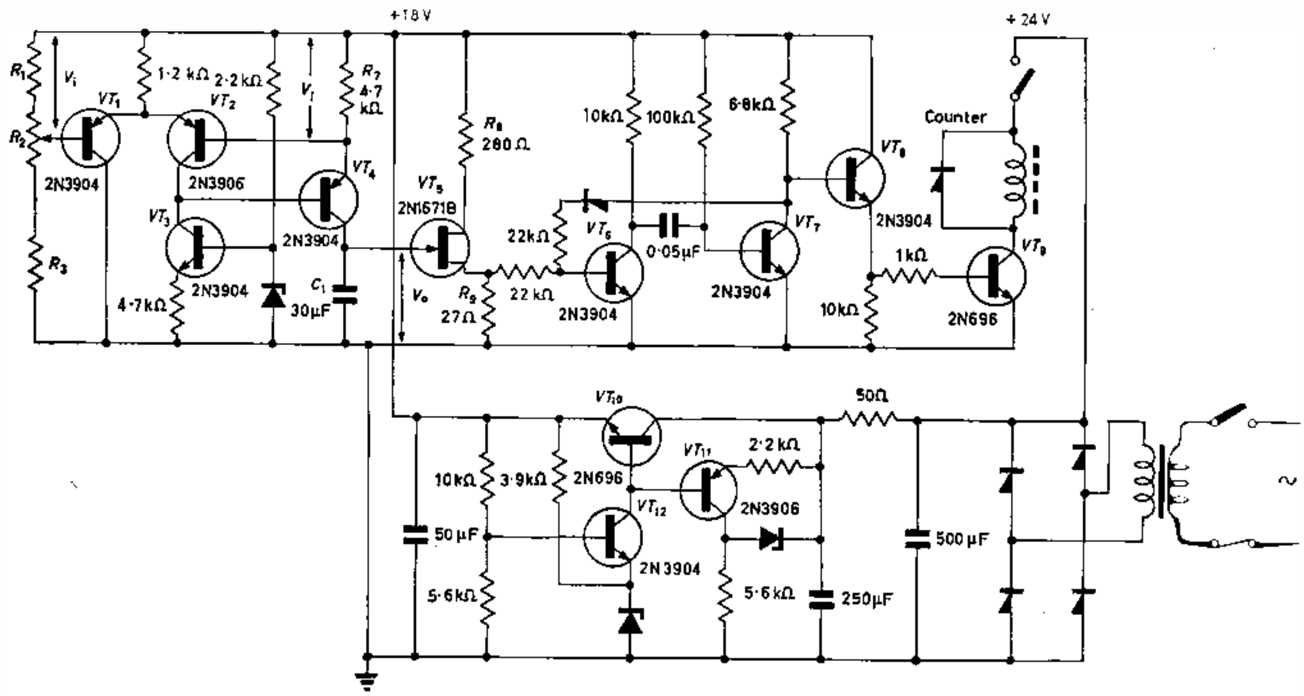


Fig. 3 Integrator circuit

The voltage-sensitive switch and counter are included in the illustration to demonstrate the entire principle.

Circuit Detail

A slave potentiometer R_2 (Fig. 3), on the recorder provides the input e.m.f. V_i which has to be integrated. The integrating capacitor C_1 is charged by a current flowing through R_7 and T_4. The current, which must be proportional to V_i, is measured by the voltage drop V_i' across R_7. The difference between V_i and V_i' is determined by VT_1 and VT_2, amplified and used to control the current generator VT_4. The transistor VT_3 in the collector circuit of VT_2 is for additional gain.

When C_1 has charged to a certain value the unijunction transistor, VT_5, conducts and C_1 is then rapidly discharged through R_9. A simple relaxation oscillator is made up of VT_5, C_1 and R_7 with VT_4 in series. Each time the capacitor discharges a positive pulse appears across R_9 which triggers the monostable multi-vibrator $VT_6 - VT_7$ which in turn provides a pulse of adequate length to switch on VT_9 and operate the electromechanical counter. The stabilized power supply comprises VT_{10}, VT_{11}, VT_{12}.

The speed of the whole instrument is limited by that of the counter to 25Hz. If a faster count is required for short period measurements then an electronic counter may be used. The values of C_1, R_1 and R_3 are adjusted to suit the integrator scale.

TABLE 1

Integrator Calculation

Recorder Position	Frequency	Error %
0	99·97	−·03
1	109·87	−·12
2	119·83	−·14
3	129·84	−·13
4	139·89	−·09
5	149·93	−·05
6	159·96	−·03
7	169·97	−·02
8	179·96	−·02
9	190·05	+·03
10	200·00	·00

Errors

The linearity of the charging characteristics will depend on the accuracy of the difference amplifier and the total gain of the current regulator circuit. This particular circuit was linear to within 0·1% and had an overall maximum error of 0·3% for the temperature range 20°C to 40°C. If the integrator it attached to a recorder it is important that the linearity of the slave potentiometer is also checked.

If the interbase resistance and the intrinsic stand-off ratio of the unijunction transistor are measured it is possible to select the value of R_9, for a given supply voltage, to produce a very stable relaxation oscillator[1]. The frequency drift over a temperature range of 20 deg C can be as little as 0·05%.

The discharge time has to be small in comparison with the charging time of the capacitor. This needs to be considered only with fast integrating cycles. It is important, however, that the peak discharge current does not exceed the specified value for the transistor, which is generally about 2A.

Table 1 gives the results of a typical calibration test. The calibration was made with an accurate frequency counter by measuring the period between individual pulses. The instrument is suitable for use with potentiometer recorders which have errors of ±0·3% of full-scale.

REFERENCE

1. Transistor Manual. General Electric Company. Seventh Edition, 1964.

An Electronically Swept Frequency Synthesizer

By E. Kantarizis*

A synthesizer has been designed to produce 666 coherent frequencies in the frequency band from 2 to 12MHz, in steps of 15kHz. It was desired that the entire synthesized frequency range should be electronically swept and that, at each successive frequency step, the electronic sweep could be halted for a given time interval before selecting the next 15kHz step. Additional specifications were: the adjacent frequencies at either side of any selected frequency should be attenuated by a minimum of 45dB, and the design should remain simple and the entire equipment should cost as little as possible.

(Voir page 475 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 476)

THERE are several methods by means of which the frequency synthesis of the required stepped oscillator could have been achieved. These methods can be broadly classified as: (a) the pure synthesis technique or direct method, and (b) the indirect method.

The pure synthesis technique can be either the method of addition or the method of subtraction. Both methods utilize balanced mixers, in the first case positive mixers and in the second case negative mixers. Harmonics and subharmonics of a crystal-controlled oscillator are mixed to produce the desired output, the synthesized output frequencies being coherent and having only one reference standard.

The indirect method of synthesis is that of division: a crystal-controlled oscillator feeds a chain of dividers and the frequency of the standard is thus reduced to the desired step frequency of the final synthesized output. The output from the dividers is compared with the output of a voltage-controlled slave oscillator, the frequency or phase of which is corrected and locked to that of the standard by means of an error voltage resulting from the comparison. This method depends on digital techniques for the precise control of the slave-oscillator.

In general, for both pure and indirect methods of synthesis, the long-term stability of the synthesized output is the same as that of the crystal-controlled oscillator[1].

In the present case, the output of the synthesizer was required to be electronically swept and each synthesized frequency in the 2-12MHz range, in steps of 15kHz, was to be selected in succession and not at random. For this reason the method of subtractive direct synthesis was used.

The output of a variable-frequency oscillator of 17-27MHz range, is mixed with harmonics of a 15kHz crystal-controlled oscillator, the resulting waveform being filtered to pass only frequencies within 5kHz of 15MHz. The output of the filter is then mixed once more with the variable-frequency oscillator, the difference being extracted from a balanced mixer, thus giving the desired synthesized frequencies within the range of 2-12MHz. The frequency synthesizer is shown as a block diagram in Fig. 1.

The harmonic generator consists of a 15kHz crystal-controlled oscillator, the output of which is converted into a square wave with very fast rise time, of the order of 5ns, in order to obtain a harmonically rich output. The square waveform is then differentiated, and the positive pulses removed to produce harmonics with as equal amplitude as possible[2].

The output of the harmonic generator is passed through a low-pass filter, allowing only harmonics lying below 12MHz to pass through and is then mixed with the v.f.o. frequency of 17-27MHz, the difference being extracted from a balanced mixer by the use of a band-pass crystal filter.

If $\sin 2\pi f_1 t$ represents the waveform of the harmonic generator, and $\sin 2\pi(f_2 t + \phi)$ represents the waveform of the variable-frequency oscillator, where ϕ is some un-

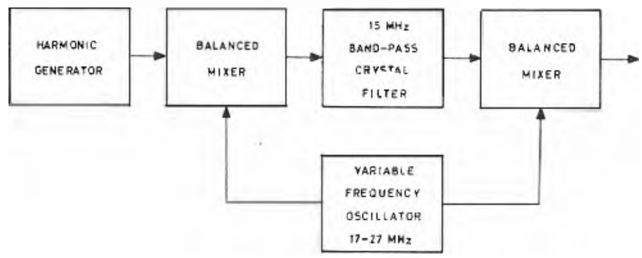


Fig. 1 Block diagram of the frequency synthesizer

specified phase difference between the two waveforms, then the output from the first balanced mixer will be:

$$2\sin 2\pi f_1 t \,.\, \sin 2\pi(f_2 t + \phi) = -\cos 2\pi(f_1 t + f_2 t + \phi) + \cos 2\pi(f_1 t - f_2 t - \phi)$$

The term $[-\cos 2\pi(f_1 t + f_2 t + \phi)]$ will be above 19MHz at all times and therefore will be rejected by a 15MHz band-pass crystal filter.

The crystal filter, of 15MHz central frequency and 10kHz bandwidth, will select the mixed output of the variable-frequency oscillator with one harmonic from the harmonic generator at a time, thus giving the effective difference of the original waveform at 15MHz and simultaneously rejecting all other combinations from the balanced mixer. Thus the term $[\cos 2\pi(f_1 t - f_2 t - \phi)]$ will be allowed through the crystal filter, and it will then be mixed once more with the variable-frequency oscillator to give:

$$\begin{aligned}2\cos 2\pi(f_1 t - f_2 t - \phi) \,.\, \sin 2\pi(f_2 t + \phi) &= \sin 2\pi(f_1 t - f_2 t - \phi + f_2 t + \phi) \\ &\quad - \sin 2\pi(f_1 t - f_2 t - \phi - f_2 t - \phi) \\ &= \sin 2\pi f_1 t - \sin 2\pi(f_1 t - 2f_2 t - 2\phi)\end{aligned}$$

The term $[-\sin 2\pi(f_1 t - 2f_2 t - 2\phi)]$ remains at all times above 32MHz and the term $\sin 2\pi f_1 t$ can be extracted quite easily from the output of the second balanced mixer by a combination of band-pass and low-pass filter.

By mixing twice, phase information in the synthesized 2-12MHz output relating to the 15kHz crystal-controlled oscillator is retained and thus all the synthesized output frequencies are coherent.

Harmonic Generator

The harmonic generator is designed to give as many harmonics of the fundamental frequency of 15kHz as

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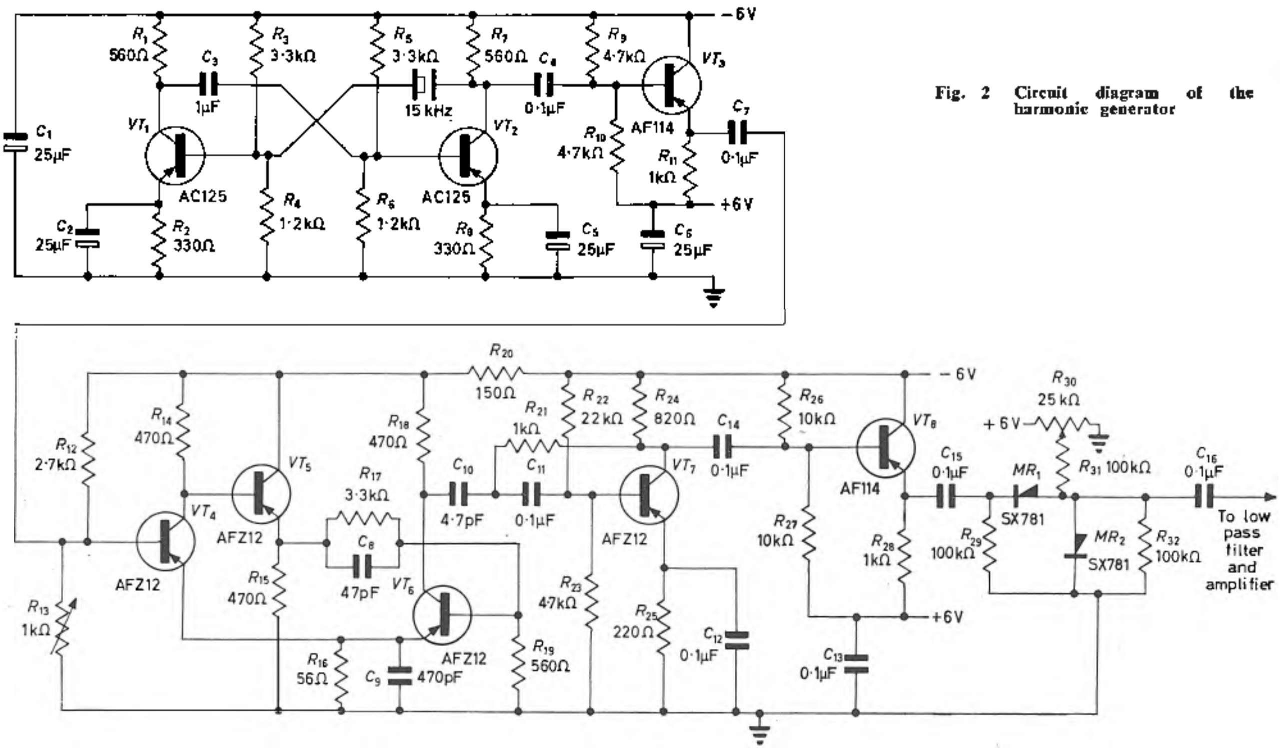


Fig. 2 Circuit diagram of the harmonic generator

possible. This is achieved by converting the output of a 15kHz crystal-controlled oscillator into a square wave with very fast rise time and then differentiating the square waveform. The Fourier analysis of the resulting positive and negative pulses is given by:

$$y_{(t)} = 2A_{av}\,[\tfrac{1}{2} + (2/\pi)\cos\theta - (2/3\pi)\cos 3\theta + \ldots]$$

where $\theta = 2t/T$, and thus they contain only the odd harmonics of the 15kHz crystal-controlled oscillator. If, however, either the positive or negative pulses are eliminated from the output of the differentiator, then the Fourier analysis of the resulting waveform is given by:

$$y_{(t)} = 2A_{av}\,[\tfrac{1}{2} - (1/\pi)\sin\theta - (1/2\pi)\sin 2\theta - \ldots]$$

which contains both the odd and even harmonics of the 15kHz crystal-controlled oscillator. The circuit diagram of the harmonic generator is shown in Fig. 2.

The crystal-controlled oscillator, consisting of two AC125 transistors, is connected in an astable saturated multivibrator arrangement, giving an output of 3½V. The crystal-controlled oscillator operates at −6V supply, drawing a total current of 8mA, the collector dissipation of each transistor being 2·4mW.

The output of the crystal-controlled oscillator is subsequently fed into a conventional Schmitt trigger through an emitter follower. The Schmitt trigger contains two high-frequency transistors, type AFZ12, with a third transistor connected between the two stages to act as a buffer amplifier. The Schmitt trigger operates at −4½V supply, drawing a total current of 15mA, the collector dissipation of each transistor being 12mW.

The square waveform at the output of the trigger has an amplitude of 2V peak-to-peak and a rise time of 5ns. This is fed into an operational amplifier of unity gain which acted as a differentiator. The operational amplifier without feedback has a gain of 200 at 5MHz, reduced to unity by appropriate choice of the feedback resistor. The capacitance connecting the Schmitt trigger to the operational amplifier, together with the input impedance of the operational amplifier (of the order of a few ohms), differentiates the square waveform into pulses of equal rise and fall times.

The output of the operational amplifier is then fed *via* an emitter follower into an acceptance circuit for negative-going pulses, employing two silicon junction diodes of type SX781, giving an output waveform of pulses at 15kHz repetition rate, of ½V amplitude and 10ns width, matched to a resistive load of 75Ω. The waveform of one such pulse is shown in Fig. 3.

The signal output from the two silicon diodes is passed through a low-pass filter, in order to remove all harmonics lying above 12MHz. It is imperative that the 15MHz harmonic should be removed from the output of the harmonic generator, since this exact frequency is also the band-pass frequency of the crystal filter to be used later. The output waveform of the harmonic generator, after filtering, is shown in Fig. 4.

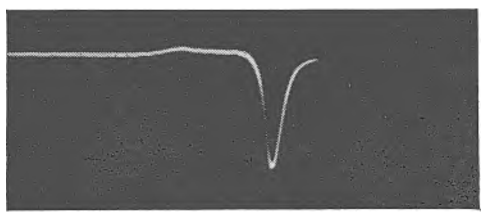


Fig. 3 Output waveform from the harmonic generator

An *m*-derived low-pass filter was designed with $m = 0{\cdot}55$ and a cut-off frequency at 12·5MHz. The notch frequency is set on 15MHz in order to minimize the signal at that frequency. Since the values of the capacitors resulting from the design are not identical with those available commercially, the capacitors associated with the position of the notch frequency are made variable. In this way it is possible to set the notch frequency with considerable accuracy at the exact band-pass frequency of the crystal filter.

The output from the low-pass filter is fed into a wideband, two-alternate-stage feedback, amplifier[3]. In order to achieve maximally flat response, the transfer-function of the amplifier is chosen so that two poles are at 45° to the negative real axis on the pole-zero diagram. The output stage of the amplifier has two poles and a zero and, for maximally flat response, these poles are set at 45° and the zero is arranged to coincide with the interstage pole on the negative real axis, by means of inductive peaking, so that the zero cancels. For optimum stability, the remaining input stage pole is set on the negative real axis at approximately three times the cut-off frequency of the amplifier.

The amplifier is designed to have a bandwidth of 12MHz maximally flat, with a gain of 20dB and with an input and output impedance of 75Ω. The transistor used is type AF114.

Balanced Mixer and AGC

The output from the harmonic generator, containing all the harmonics within the band 2-12MHz of the 15kHz crystal-controlled oscillator, is mixed with the output from a variable-frequency oscillator within the range of 17 to 27MHz. The balanced mixer employs two transistors, type AF114, with their collectors connected together.

The input signals to be mixed are applied each to the base of a transistor which operates in the grounded emitter configuration, giving phase inversion. At the same time the transistor is arranged to operate as a pseudo-grounded base, giving no phase inversion. The latter condition demands low output impedance from the preceding stage. Thus, because of the self-balancing nature of the circuit used, the output from the balanced mixer is practically free of the input frequencies and presents only the effective addition and subtraction of these, being identical in action with a balanced modulator. The circuit diagram of the balanced mixer and the automatic gain control amplifier is shown in Fig. 5.

Since only the difference of the input frequencies to the balanced mixer is required, and specifically the difference at 15MHz, a tuned circuit is connected between the common collectors of the two transistors and the supply, in order to aid selection of the required frequency. The output from the balanced mixer is taken by inductive coupling from the inductor of the tuned circuit and fed into a conventional grounded emitter amplifier tuned at 15MHz. The amplifier has a gain of 18dB, giving further aid to the selection of the required frequency. Both the

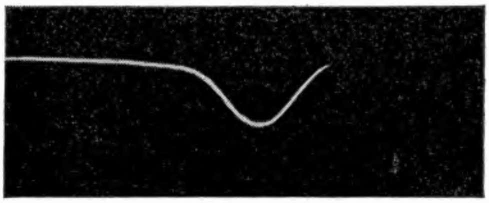
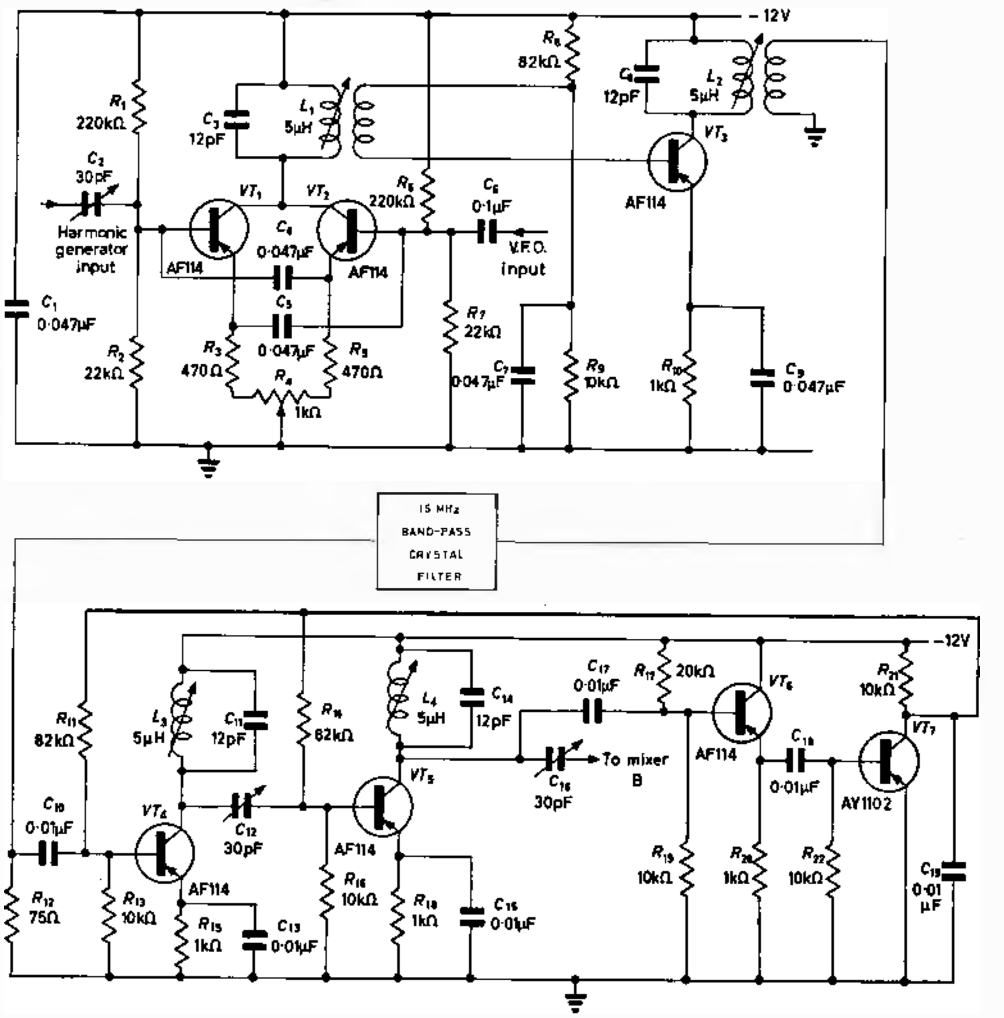


Fig. 4 Pulse containing harmonics within 2-12MHz

Fig. 5 Circuit diagram of the balanced mixer and the automatic gain control



balanced mixer and the tuned amplifier operate on −12V supply, drawing a total current of 2·4mA; the collector dissipation of each transistor of the balanced mixer is 3mW and that of the tuned amplifier transistor, 7mW.

The signal output from the tuned amplifier is fed into a 15MHz band-pass crystal filter by inductive matched coupling. The crystal filter employed has a bandwidth of 10kHz at −6dB centred on 15MHz, the output from which is fed into a conventional two transistor cascaded tuned amplifier at 15MHz, designed to give a total gain of 60dB. The output waveform from the crystal filter, after amplification, is shown in Fig. 6.

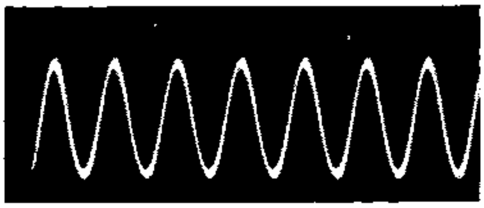


Fig. 6 Output waveform from crystal filter at 15MHz after amplification

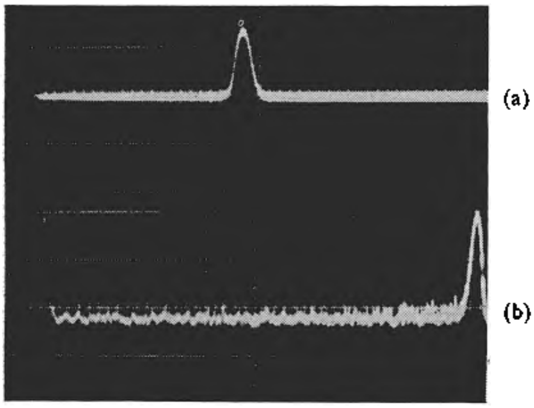


Fig. 7 (a) Display of spectrum analyser of the output from the Harmonic Generator at 6MHz
(b) Two adjacent harmonics at 6MHz

Each time the v.f.o. frequency increases by 15kHz, the output from the crystal filter gives a maximum when the v.f.o. frequency is a pure harmonic of 15kHz and falls to a minimum when the frequency is between 15kHz harmonics. By amplifying this output and feeding it, *via* an emitter follower, into an a.g.c. amplifier, it is possible to obtain the electronic control needed fully to automate the frequency synthesizer.

To obtain a.g.c. action, use is made of the forward voltage-delay characteristic of the emitter junction of a silicon transistor. The transistor used is type AY1102, which needs 0·6V input to the base before overcoming the forward voltage delay and allowing current to flow. Thus, normally the transistor is arranged to be in the 'off' condition, with the full supply voltage appearing on the collector. However, as soon as the input to the base has reached 0·6V, it turns the transistor hard 'on' and the collector voltage drops rapidly towards ground.

The output from the band-pass crystal filter is arranged to turn the a.g.c. transistor hard 'on' whenever the output frequency from the balanced mixer is almost exactly 15MHz. This gives a rectified signal output from the collector of the a.g.c. transistor, which is then smoothed with an *RC* network of appropriate time constant. A d.c. variation is produced, the level of which is used eventually to stop the onward sweep of the variable frequency oscillator, at frequency intervals of almost exactly 15kHz.

The d.c. output variation from the a.g.c. amplifier is then d.c. coupled to the base of each transistor of the cascaded tuned amplifier, thus varying its bias and ensuring a constant-amplitude waveform of 0·8V at the output of the amplifier. This output is then mixed once more with the variable-frequency oscillator, the filtered output of which gives the desired synthesized frequencies.

The variable-frequency oscillator is designed to give an output of constant amplitude within the frequency range of 17-27MHz. The v.f.o. consists of a conventional grounded base Colpitts oscillator, the output of which is fed to two emitter followers, to give two outputs. The tuned circuit of the oscillator employs a Varicap, biased in the reverse direction, the capacitance being varied by applying the voltage output from a ramp generator across it. The oscillator is then forced to change its frequency.

The ramp generator is designed to increase its amplitude in small steps, so that it covers the equivalent frequency range of the v.f.o. in approximately 10^4 steps. A command from the a.g.c. stops the sweep of the ramp when the output of the synthesizer is coherent, thus holding it at that frequency. After a suitable time interval, the ramp generator is commanded to restart its onward sweep, until the output of the synthesizer is coherent once again.

This cycle of events continues until the output of the synthesizer has covered successively the entire range of synthesized frequencies. A zero-sensing circuit within the ramp generator allows the automatic restarting of the ramp, thus permitting the output of the synthesizer to be swept electronically for as long as desired.

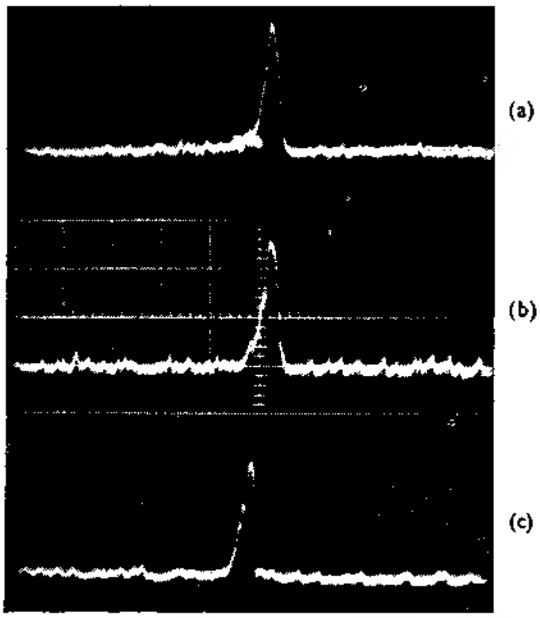


Fig. 8 Display of spectrum analyser of the output from the 15MHz crystal filter after amplification

Performance Analysis

Balanced mixers tend to produce not only the sum and difference of the original frequencies, but also those of the harmonics of the input frequencies. This tendency results from overloading the mixer, and/or imperfect phase inversion of the input signals, the latter being an inherent fault of the balanced configuration of mixers. Thus the purity of the output synthesized frequencies depends on the careful selection of the band of frequencies to be mixed. If this is observed, the respective positions in the frequency spectrum of the sum and difference of the original frequencies will not coincide. It will then be relatively easy to isolate the required frequencies by simple filtering.

In the design of this frequency synthesizer consideration was given to the above mentioned facts, and the input frequencies to both mixers are chosen to minimize spurious modulation products.

Table 1 shows the possible subtractive action of mixer *A* with inputs f_1 and f_2, as well as their harmonics, resulting in the extraction of 15MHz at the output of the 15MHz band-pass crystal filter. The frequencies nearest to 15MHz, resulting from mixing are also shown in the table. Input f_1 consists of the harmonics of the 15kHz crystal-controlled oscillator, which have been restricted within the frequency band 2-12MHz. Signal f_2 is the output of the variable frequency oscillator within the range of 17-27MHz.

As $(f_1 + f_2)$, and also the sum of the harmonics of f_1 and f_2, lie at all times above 19MHz, only the subtracted output of the mixer need be considered. It can be seen that only the difference between f_2 and f_1 or f_2 and the harmonics of f_1 can produce 15MHz. But as the harmonics of f_1 are also harmonics of the 15kHz crystal-controlled oscillator, this is not important in this case.

The band-pass crystal filter used has a bandwidth of 10kHz at -6dB and 16kHz at -60dB, giving a substantially pure sinusoidal output at 15MHz, as seen in Fig. 6. This output is then mixed once more with that of the variable frequency oscillator: the difference 2-12MHz is extracted from the output of the second balanced mixer. Since the sum of f_1 and f_2, or its harmonics, lies at all times above 32MHz, only the subtractive action of mixer *B* is considered in Table 1. The difference $f_2 - f_1$ is sufficiently isolated, in the frequency spectrum, from $2f_2 - f_1$ or $3f_2 - f_1$ to be extracted quite easily.

Fig. 7 shows the output of the harmonic generator as displayed by a spectrum analyser. Fig. 7(a) shows one harmonic at 6MHz with a dispersion of 100Hz/cm and Fig. 7(b) two adjacent harmonics, at the same frequency, spaced by 15kHz, with a dispersion of 2kHz/cm. The latter figure illustrates that adjacent harmonics from the output of the harmonic generator, have equal amplitudes. The calibration of the vertical scale in both cases is 12dB/cm.

Fig. 8 shows the output of the 15MHz band-pass

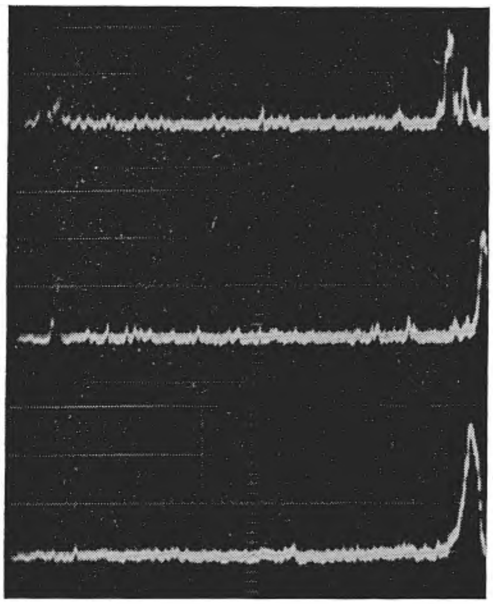


Fig. 10 Display of spectrum analyser of the output from the frequency synthesizer, showing two adjacent frequencies at 6MHz for an a.g.c. direct voltage level of: (a) 10V, (b) 9·3V, (c) 8·5V

TABLE 1
Subtractive action of mixers employed in synthesizer

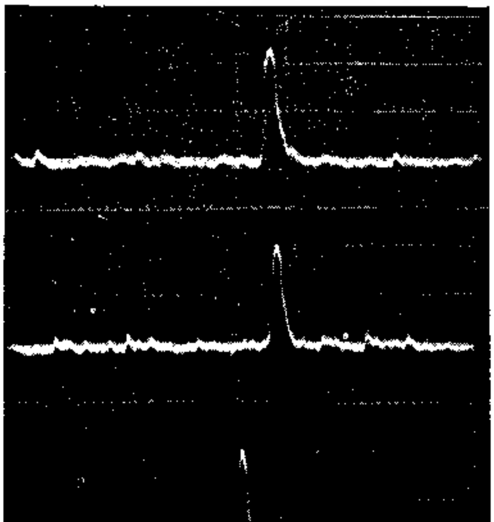
MIXER A. To extract 15 MHz

	f_2 (17 – 27)	$2f_2$ (34 – 54)	$3f_2$ (51 – 81)
f_1 (2 – 12)	$f_2 - f_1$ (15)	$2f_2 - f_1$ (22)	$3f_2 - f_1$ (39)
$2f_1$ (4 – 12)	$f_2 - 2f_1$ (15)	$2f_2 - 2f_1$ (22)	$3f_2 - 2f_1$ (39)
$3f_1$ (6 – 12)	$f_2 - 3f_1$ (15)	$2f_2 - 3f_1$ (22)	$3f_2 - 3f_1$ (39)

MIXER B. To extract 2–12 MHz

	f_2 (17 – 27)	$2f_2$ (34 – 54)	$3f_2$ (51 – 81)
f_1 (15)	$f_2 - f_1$ (2 – 12)	$2f_2 - f_1$ (19 – 39)	$3f_2 - f_1$ (36 – 66)

Fig. 9 Display of spectrum analyser of the output from the frequency synthesizer at: (a) 3MHz, (b) 7MHz, (c) 11MHz



crystal filter, after amplification, with the v.f.o. frequency at the input of the first balanced mixer at 18, 22 and 26MHz for (a), (b) and (c) respectively, with a dispersion of 2kHz/cm. The vertical calibration for all three cases is 26dB/cm.

Fig. 9 shows the output waveform from the frequency synthesizer displayed by the spectrum analyser. The frequency setting of the v.f.o. is the same as that used in Fig. 8, with the same dispersion and vertical calibration. It can be seen that the required frequencies are at least 50dB above noise level.

Fig. 10 shows the output of the frequency synthesizer at 6MHz, with a dispersion of 2kHz/cm, for a d.c. level from the a.g.c. amplifier of 10, 9·3 and 8·5V for (a), (b) and (c) respectively. Because of the incorporation of the a.g.c. amplifier system into the design of the frequency synthesizer, it is necessary that locking the sweep of the v.f.o. be effected only when the d.c. level from the a.g.c. amplifier is at 8·5V or less, the minimum d.c. level being 7·8V. Under this condition, the output from the synthesizer at any required frequency is at least 45dB above the adjacent frequencies, as illustrated in Fig. 10(c). The calibration of the vertical scale for all three cases is 20dB/cm.

Acknowledgments

The author would like to thank Dr. J. D. Whitehead for his help and supervision, also D. Horgan for his advice and for designing the ramp generator and associated electronic logic.

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Radio & Space Research field station at Chilbolton

The new field station at Chilbolton, Hants, with its 82ft (25m) dish-type fully steerable aerial, will give the Radio & Space Research Station of the Science Research Council a new facility required for research on radio propagation, including study of problems concerned with satellite-communication systems.

Work on the aerial and field station started in 1964, when the Ministry of Public Building and Works placed a contract with AEI Electronics. The site—on the disused airfield at Chilbolton—was chosen because it is not near industrial areas or main air routes, nor is it too far from the main Radio & Space Research Station at Ditton Park, near Slough, Bucks. The overall cost of the field station was £428 000. The station will be manned by a team of about six under the direction of the officer-in-charge. Experiments will be carried out by teams visiting the station from the parent establishment.

The aerial specifications are similar to those of the Post Office aerial at Goonhilly, Cornwall. It is also similar to some aerials used in Britain for radio astronomy, but the reflector is built to closer tolerances, which must be preserved when the aerial is being moved rapidly to follow satellites.

The 25m-diameter aerial is designed for use at wavelengths down to 3cm. One requirement, a true paraboloid reflector surface adjusted to an accuracy of 0·1in, was achieved with the aid of a special optical rangefinder system called a Parabscan. The honeycomb construction of the reflector panels and the stiffness of the backing structure (the main part of which weighed 42tons when lifted onto the support arms in its prefabricated form) ensure that the parabolic shape is maintained under working conditions. The aerial, which can be driven in azimuth through 540°, and in elevation from 5° below the horizontal to 30° beyond the zenith, is steerable at rates up to 3° per second in azimuth and 1° per second in elevation in order to follow transmissions from Earth satellites. The aluminium reflector surface is painted matt white to avoid overheating of the equipment at the focus when the aerial is being used to track the Sun. With its cabin the reflector is supported on a rigid steel platform rotating on a reinforced concrete tower.

The large aluminium cabin surrounds the top of the tower. The lower floor of the cabin will house experimental radio equipment, and the upper floor the motor drives and contactor gear.

The control room contains the control desk, from which there is a good view of the aerial. The computer and other equipment in this room provides three methods of control—programmed movement from punched paper tape, automatic tracking of a radio source and manual operation. The main receiving equipment and displays will be in this room.

An important part of the Radio & Space Research Station's work is the study of how radio waves travel through the Earth's atmosphere and through space. Though mainly of a fundamental character, this research is designed to provide information needed in the practical exploitation of radio. Many valuable contributions in this field have been made by R&SRE. Work carried out at the station (when it was part of the Department of Scientific and Industrial Research) has led to such developments as radar, improved direction finding and the use of ionospheric sounding to increase the effectiveness of long-distance high-frequency communications.

Line-of-sight radio-relay links and satellite communications at centimetre wavelengths have introduced new requirements for propagation data. Experiments at these wavelengths have been in progress for some years, using aerials of moderate size. However, some of these investigations require an aerial similar to those being used at communication satellite stations to achieve high gain and directivity. The Chilbolton aerial will fill this need—though it is regarded as a general purpose research instrument and will be used in addition for other experiments concerned with the structure of the troposphere and properties of the ionosphere. It will also be used for some radio-astronomical work.

The reflector profile is now sufficiently accurate to proceed with radio tests and some of the equipment has been installed. Measurements of pointing accuracy, gain and beamwidth will be made using a link from a transmitter at a wavelength of 3cm, and also using celestial sources. Receiving equipment has been built for several other wavelengths for use in the experimental programme. Some early work will be concerned with an evaluation of the behaviour of the aerial as a mechanical structure, carrying out measurements of the surface accuracy when subjected to the influence of acceleration, wind and temperature changes. Much of this work will be undertaken in collaboration with the National Physical Laboratory. A radar method of measuring the profile with precision has been developed for comparison with the optical technique.

In the initial stages the research programme will be concerned largely with the propagation of radio waves through the troposphere, and the ways in which the waves are influenced by meteorological factors. Particular attention will be paid to mechanisms such as scatter from atmospheric irregularities and rain, which can lead to interference between radio services sharing the same frequencies—for example interference by a radio-relay transmitter to a satellite Earth station. Although these phenomena have been studied in the past, the Chilbolton aerial, similar to those used in satellite communications, will enable new aspects to be investigated. The structure of the troposphere will also be examined by radar techniques.

Other experiments will be concerned with the ionosphere; these include measurements of the scintillation of radio stars, to study the composition of the ionosphere and interplanetary plasma. An experiment to determine ionospheric characteristics by a radar-scatter technique is being planned in co-operation with the Royal Radar Establishment.

Dynamic Measurements on Narrow-Band Receivers

By J. F. Golding*

The accurate tuning of a signal generator to the centre frequency of a narrow-band receiver often presents a problem of frequency stability. For certain measurements—including sensitivity—the most economical way of solving this problem is by continuously sweeping the signal generator through the receiver passband. This, however, sometimes calls for somewhat abnormal modulating facilities, which are not included in the majority of commercially available signal generators but are present in the Marconi Instruments Type TF 2002.

(Voir page 475 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 476)

IT is not uncommon for h.f. a.m. and s.s.b. receivers to have acceptance bandwidths as small as 3kHz, and in those used for c.w. telegraphy it is often as low as 100Hz. Selectivity and sensitivity measurements on such receivers by conventional methods can demand extremely high standards of frequency stability from the signal generator and very careful tuning on the part of the operator. The The stringency of these requirements can, however, be relieved by the use of a frequency-swept test signal synchronized to the time-base of an oscilloscope. This method can provide a graphic display of the receiver response at all frequencies in its passband; and, from this display, accurate assessment of both sensitivity and selectivity can be made.

Swept-frequency bandwidth measurements are fairly common at v.h.f. and above, while in the u.h.f. and microwave regions the technique is also applied to noise-limited sensivity measurements. At the lower radio frequencies, however, the test gear requirements are, in certain respects, more exacting because the measurements must necessarily bear interpretation in similar terms to those obtained by conventional methods.

The generally accepted test methods are, of course, based on a close simulation of actual operating conditions which, in the case of a.m. receivers, implies the use of an amplitude-modulated test signal. Furthermore, as the receiver bandwidth is likely to be very narrow, frequency-swept tests must be conducted at low sweep velocities, in order to avoid errors due to ringing in the receiver resonant circuits. Two unusual requirements of the signal generator are therefore that:

(a) it must be possible to apply amplitude modulation and frequency sweep simultaneously

(b) the signal generator voltage-sensitive reactor (which produces the frequency change) must be d.c. coupled to the input terminal in order to avoid phase error at the low sweep speeds.

In many of the commercially available sweep generators (wobbulators), a heterodyne system is used to produce a wide sweep at low centre frequencies. Such systems are unsuitable for swept-frequency signal generators owing to the greater likelihood of frequency drift. The wide sweep range is, however, not merely unnecessary but undesirable, so that the most suitable arrangement for the signal generator is that of applying the sweep directly to an oscillator operating at the output frequency.

All the above requirements are met in the Marconi Instruments TF 2002 signal generator, which is the only suitable one known to the author for covering the h.f. band.

The Test Set Up

The basic test-gear arrangement for bandwidth and sensitivity measurement is shown in Fig. 1. This differs from the usual arrangement for dynamic display of the receiver frequency/response characteristic in that an external detector rectifies the a.f. output from the receiver to provide the vertical deflexion voltage. The more usual system of utilizing the d.c. component. of an a.m. receiver demodulator output is certainly very suitable for display of the overall bandwidth characteristic of the tuned amplifiers; but this d.c. component is not necessarily directly proportional to the final output of the receiver, and it is often inconvenient to make the internal connexion. Moreover, in s.s.b. receivers the final filtering is sometimes applied after the demodulation, which may well be such that no d.c. component is available anyway.

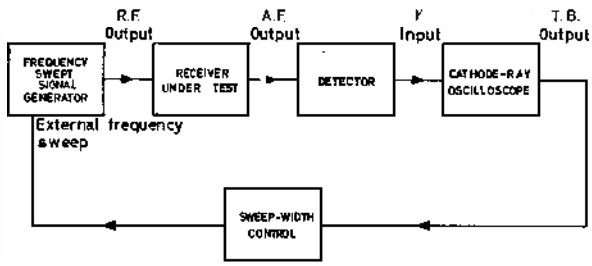


Fig. 1 The basic test-gear arrangement

S.S.B. and Telegraphy Receivers

The test procedure for s.s.b. receivers is similar to that for c.w. receivers, because a c.w. output from the signal generator is utilized to simulate the wanted sideband.

The frequency/response characteristic is displayed on the oscilloscope screen using the conventional sweep-generator technique. A low-speed sawtooth voltage from the time-base output terminal of the oscilloscope is fed to the signal generator voltage-sensitive reactor. This causes the c.w. test signal to be swept through the receiver pass band in unison with the traverse of the spot across the display screen. As the vertical deflexion voltage, obtained by rectifying the a.f. output, is directly proportional to the receiver response, the spot traces out the shape of the frequency/response characteristic.

A graticule marked with a vertical decibel scale assists interpretation in the usual terms for relative amplitude measurement. Calibration of the horizontal scale in terms of frequency, however, can present something of a problem.

When using the Marconi TF 2002 signal generator, the most satisfactory method utilises the incremental frequency

* *Marconi Instruments Ltd.*

control. This control carries a dial divided linearly into 100 divisions. Its tuning range can be varied by operation of a pair of secondary controls, and there is provision for calibrating it against an internal crystal oscillator which provides calibration 'pips' at intervals as small as 10kHz.

Changes in the centre frequency have the effect of moving the displayed curve horizontally by a distance proportional to the frequency change. It is, thus, a simple matter to adjust the sweep width so that a 10kHz movement of the incremental frequency control shifts the displayed curve through 10 divisions of a linear horizontal graticule scale to give 1kHz per division.

Fig. 2 shows the frequency/response characteristic of a single-sideband receiver obtained by the method described. In the case of an s.s.b. receiver, it is desirable to know the horizontal location of the suppressed carrier frequency (i.e., the a.f. zero) on the displayed characteristic. With some receivers this can be obtained by beating the output of the receiver final detector with the internally generated carrier. An external mixer is used, and the resulting output is fed to the z-modulation terminal of the oscilloscope to produce a bright spot as shown in Fig. 2.

Use of the frequency-swept test signal for bandwidth measurement is common enough. Its application to sensitivity measurement, however, is less familiar; but, in the case of s.s.b. and c.w. receivers, all the necessary information for calculating signal-to-noise ratio can be presented in a single display on the oscilloscope screen.

For high-level bandwidth measurements the baseline of the display is, for all practical purposes, equivalent to zero output from the receiver. At input signal levels approaching the receiver noise-limited sensitivity, however, it is necessary to use maximum gain in order to obtain a display; and the off-tune output from the receiver is, of course, the internally generated noise. This is represented by the baseline of the displayed response curve.

In order to measure the noise level it is first necessary to establish the deflexion for zero input in the position coincidence with the baseline of the oscilloscope graticule. This is very easily done by disconnecting the y-input and setting the trace to the graticule base line by means of the y-shift control. When the oscilloscope is reconnected, the noise level can be measured as the distance between the graticule baseline and baseline of the display. The distance between the graticule baseline and the peak of the displayed response curve is, of course, proportional to the signal-plus-noise voltage.

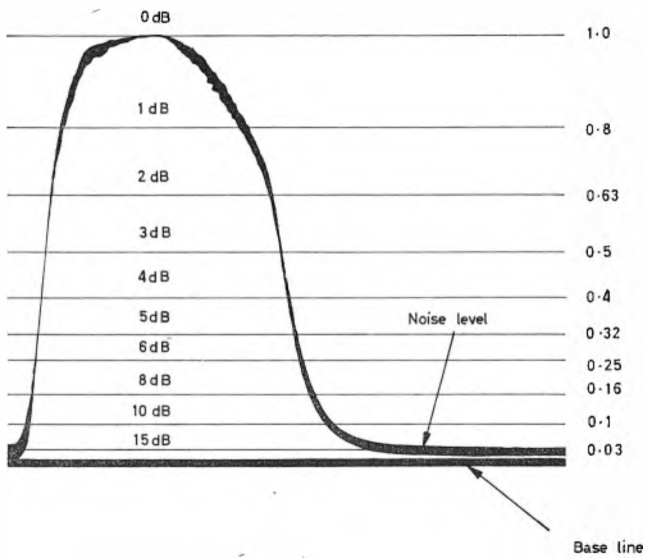


Fig. 3 The sensitivity-measurement display for a single-sideband receiver; the baseline with the Y input short-circuited is also shown in the Figure

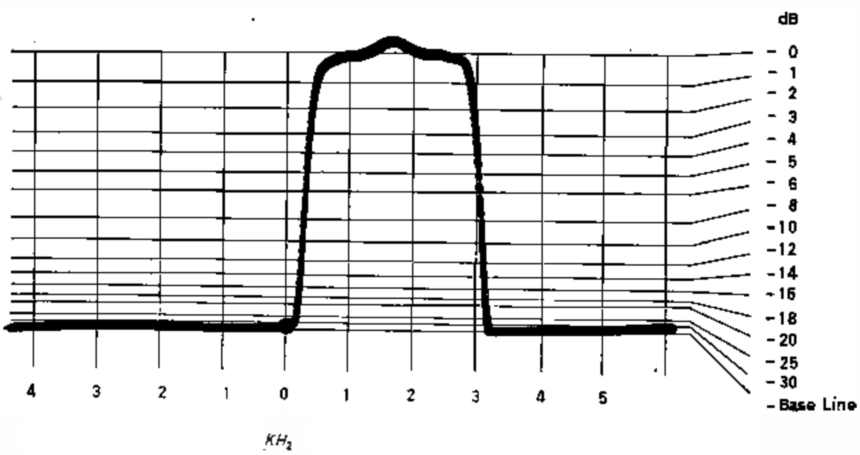
At this point, the argument is open to the very valid objection that signal-to-noise ratio is essentially a power ratio, whereas, with the usual type of detector, the oscilloscope deflexion is proportional to peak voltage. As the output impedance of the receiver is likely to be low, however, it is not difficult to construct a true square-law detector using a pair of germanium diodes, the basic design of this item of the test gear being dealt with later in the article.

Fig. 3 shows the displayed response curve of an s.s.b. receiver operating close to its 15dB noise-limited sensitivity input level. It is immediately apparent that the discrimination for measurement of the noise level leaves something to be desired—especially as most s.s.b. receiver sensitivity measurements are made at 20 to 40dB signal/noise ratios. The difficulty can be overcome by inserting a variable attenuator between the output of the receiver and the external detector.

To establish the noise level, the attenuator is set to zero and the gain of the system is adjusted to bring the base line of the display to some convenient graticule line. Attenuation is then increased until the peak of the displayed curved is coincident with the same line; and the attenuator setting in this final condition is equal to the signal/noise ratio. Theoretically, there is an error, owing to the fact that the overall height of the curve includes the noise as well as the signal; but this error is negligible in practice.

When the noise-level datum is being established (zero attenuation) the spot is deflected right off the screen during the in-tune part of the sweep, but it was found experimentally that this produces no adverse effect on the measurement.

Fig. 2 Oscillogram of frequency/response characteristic of a single-sideband receiver



D.S.B. A.M. Receivers

The usual swept-frequency method of displaying the response characteristic of an a.m. receiver is by connecting the y-input of the oscilloscope directly to the receiver demodulator. Theoretically, the use of a swept c.w. signal produces a more accurate curve than measurement with an amplitude modulated test signal. In practice, however, the only advantage lies in the freedom from the need to apply the a.m.

It is often inconvenient to make the internal connection to the receiver demodulator; and the arrangement shown in Fig. 1 is usually preferable when a suitable signal generator is available. The amplitude modulating frequency must, of course, be low compared with the pass-band of the receiver, and the modulation depth should be restricted to about 20%. Providing these conditions are fulfilled, the displayed response characteristic is normally indistinguishable from that obtained by d.c.-coupling the receiver demodulator to the oscilloscope.

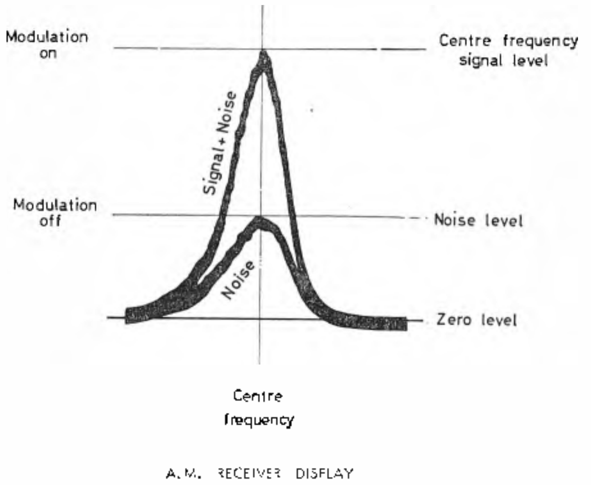


Fig. 4 Oscillogram of mod-on/mod-off displays for noise-limited sensitivity on an a.m. receiver

Noise-limited sensitivity of an a.m. receiver is usually measured with 30% modulation applied to the carrier; but the swept-frequency method of measurement is otherwise similar to that described for s.s.b. receivers. The most important difference lies in effect of the receiver internally generated noise.

The noise level in the output of an s.s.b. receiver remains constant regardless of whether the low-amplitude r.f. input signal is applied or not; but this is not so with an a.m. receiver. When this type of receiver is tuned to an unmodulated r.f. signal which is below the a.g.c. threshold, the noise output level is noticeably higher than the noise for zero input. With no input signal, the internally generated noise is not sufficient to drive the demodulator to the linear part of its characteristic; but, when an externally generated carrier is applied, the noise voltages add to it and effectively produce random modulation, which is linearly detected.

Signal/noise ratio is, therefore, always measured with the carrier, applied throughout the test—normally at a level close to the specified sensitivity—the ratio being calculated from the receiver output levels with modulation on and off.

When a low-level frequency-swept test signal carries no amplitude modulation, the curve displayed on the oscilloscope screen is the input-frequency/noise characteristic of the receiver. When amplitude modulation is applied the curve represents the signal-plus-noise as for the s.s.b. receiver. Relative heights of the two curves at the centre-frequency datum thus provide the information for calculating the signal/noise ratio.

The true amplitudes are, of course, the heights of the peaks of the curves from the zero-input line on the oscilloscope screen, as already described. In practice, however, the noise output of most communication receivers is so small when the test carrier is off-tune that the base line of the display can usually be regarded as the zero datum.

Fig. 4 is a composite oscillogram showing the noise curve and the signal-plus-noise curve superimposed. The test signal level was reduced below the specified noise-limited sensitivity in order to illustrate the shape of the curves. In practice, the noise curve would normally be much shallower, and it may be advisable to use the variable attenuator interposed between the receiver output and the square-law detector in the manner recommended for s.s.b. receivers.

Sweep Speeds

Most engineers are quite conscious of the need to use low sweep speed when displaying the frequency/response characteristic in order to prevent the deformity of the displayed curve that results from ringing in the resonant stages of the receiver.

For measurements which are not concerned with the true shape of the response characteristic, however, there is a temptation to ignore the ringing effect and to increase the speed so that a brighter or more easily observed display is obtained. This can lead to amplitude error and should be avoided. As a guide to the maximum permissible sweep speed note that there is negligible significant deformation so long as the highest-frequency component of the dis-

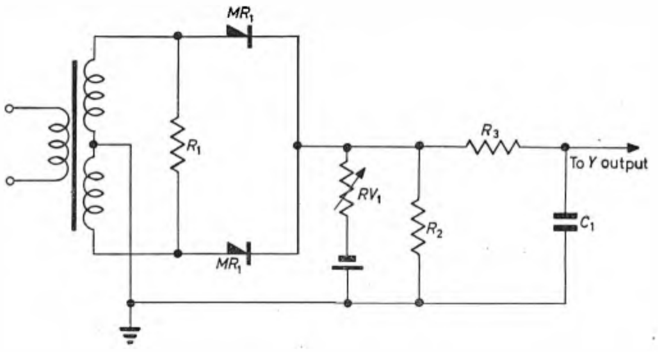
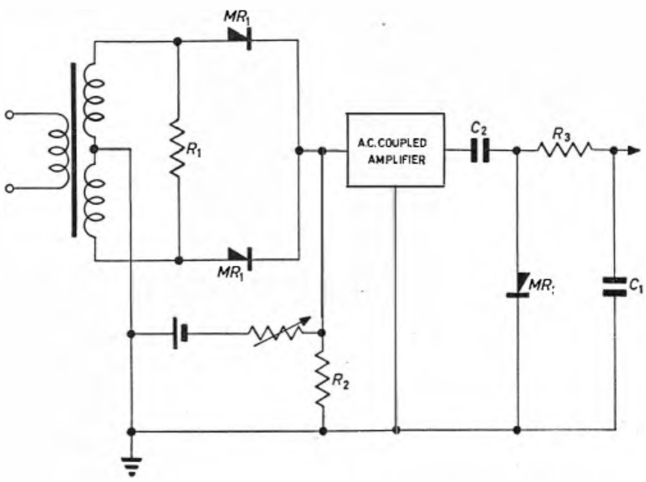


Fig. 5 Basic circuit of square-law detector

Fig. 6 Use of an a.c.-coupled amplifier with d.c. restorer to increase the deflexion voltage



played response curve is less than 1/100 of the receiver 3dB bandwidth.

The Square Law Detector

In order to obtain an oscilloscope deflexion which is proportional to relative power, the external detector must deliver an output voltage which is proportional to the square of the input voltage. A suitable convertor utilises the square-law characteristic of point-contact germanium diodes in the circuit shown in Fig. 5.

For low-level inputs, the forward current/voltage characteristic of almost any point-contact diode follows the law $I = k(V - v_0)^2$, where k is a transfer constant, v_0 is the zero-current voltage of the diode, and V is the applied voltage. It follows, then, that true square-law a.c. response can be obtained very easily by the application of a d.c. bias equal to v_0. The effective law of the diode then becomes $I = kV^2$; and, by connecting a pair of the diodes in a full-wave rectifier configuration, the full involute square-law voltage/current characteristic can be obtained.

Provided that the total series impedance of the associated circuit is kept low compared with the forward resistance of the diode, the instantaneous output voltage will be proportional to the square of the input voltage. It follows, therefore, that the d.c. voltage component of the output will be a direct linear function of the mean power in R_1.

Fig. 5 shows the detector followed by a simple low-pass filter. In this form it can be used directly connected to a sensitive d.c.-coupled oscilloscope. But, as the impedance of the circuit is necessarily low, the overall sensitivity of the system is likely to be insufficient for many of the practical applications. An a.c. input of 1V peak-to-peak across R_1 produces a deflexion of about 2·5 cm on an oscilloscope having a y-sensitivity of 10mV/cm.

Fig. 6 shows a simple method of increasing the output voltage of the detector by means of an a.c.-coupled amplifier. In this arrangement the diode system becomes a square-law convertor rather than a detector. Only the a.c. component of its output voltage is amplified, the d.c. component being re-inserted by the action of a d.c. restorer diode. This d.c. signal is then applied to the oscilloscope via a low-pass filter as before.

Using an amplifier having gain of 100, 5cm deflexion is obtainable on a 50mV/cm oscilloscope with a 1V peak-to-peak input signal to the convertor.

Combined Communication System for Light Aircraft

A new combined h.f. and v.h.f. communication system has been developed by The Marconi Co., primarily for use in light aircraft and helicopters. This single system (type AD1400) provides all the communication facilities that are necessary for a wide range of light aircraft. The equipment is completely transistorized, and employs solid-state techniques throughout, eliminating all moving parts in the tuned stages of the transmitter/receiver unit. The standard unit has an output of 40W on the h.f. range and 10W on the v.h.f. range.

In its standard form, the complete AD1400 system in two units weighs only 20kg (44lb). It provides 28 000 channels in the high-frequency band (2-29·999MHz) and 2 800 frequency-modulated channels in the very-high-frequency band (30-99·975MHz). Each frequency is crystal-controlled and selected from a simple decade-switch controller. Tuning to the correct frequency is completely automatic. Alternative units are available to provide a high-power output of 400W for the h.f. transmitter and a homing system on a range of v.h.f. frequencies. This homing unit produces a simple left/right indication, enabling the pilot to fly directly towards a v.h.f. beacon transmitter on the ground.

The complete equipment has been designed for extreme reliability, and uses many comparatively new techniques which have proved themselves in practice in the current range of airborne equipment. A very large part of the circuitry is contained in sealed modules, to provide protection from atmospheric dust and moisture. These modules are normally filled with dry nitrogen, to provide the additional protection of a completely inert atmosphere.

The AD1400 system is designed to provide h.f. and v.h.f. communications in lightweight units suitable for installation in the growing number of light aircraft or helicopters. Common circuits are used for the majority of the h.f. and v.h.f. sections of the system, and also for both transmit and receive modes wherever possible. The complete transmitting and receiving circuits are contained in the first unit, which weighs 12kg. The final stage of power amplification, together with the aerial-tuning unit and the power supplies for the complete system are contained in the second unit, which weighs 6·1kg.

The high-power version of the system (type AD1410) employs a separately mounted 400W amplifier, together with the power supplies for the complete system; the aerial-tuning unit is contained in another case, and these two would normally be mounted together, close to the aerial in the aircraft.

The homing system provided with the equipment, to enable the pilot to fly directly towards a v.h.f. beacon on the ground, has two aerials, and it is designed to accept and compare the inputs from these two aerials, providing left/right instructions on an indicator on the flight panel. This indicator can be shared with the VOR/ILS systems if required.

Control for the complete system is exercised from a single unit which provides digital frequency selection in the h.f. band in 1kHz steps from 2 to 29·999MHz, and in the v.h.f. band in 25kHz (narrow bandwidth) or 50kHz (broad bandwidth) steps from 30 to 99·975MHz. A single 'systems' switch provides for the selection of double- or single-sideband working in the h.f. band, together with narrowband, broadband or homing facilities in the v.h.f. range. An r.f. gain control is also mounted on the controller, and an indicator lamp shows when the tuning cycle of the system is complete.

A Hall-Effect Measuring System for Large Direct and Alternating Currents

By D. J. Lloyd*, B.Sc.(Eng.), Ph.D., A.M.I.E.E. and C. W. Taylor*

A system is described which provides (a) a direct indication of r.m.s. value, and (b) a voltage proportional to the instantaneous value, of currents of the order 1 to 1000A over the frequency range zero to 5kHz. The system uses the Hall effect in indium arsenide to detect a magnetic flux density produced by the current to be measured.

(Voir page 475 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 476)

IT is some years since the Hall effect was proposed as the basis of an ammeter for large currents[1]. In this application, the current to be measured produces a proportional magnetic field which is in turn detected by a Hall-effect probe, and the resulting signal is amplified. The method has the advantages of a response to zero frequency, electrical isolation between the instrument and the conductor carrying the current to be measured, and the possibility of producing, as an analogue of this current, a voltage output capable of operating recorders or amplifiers.

The system described is based on this principle, and was developed for measuring and recording currents of some hundreds of amperes supplied to large electromechanical vibrators. For this purpose, the required frequency response was 5Hz to 2kHz, but in fact a wider range has been achieved; the system is flat from zero frequency to about 5kHz.

The Hall Effect

The Hall effect in a conductor or semiconductor is the interaction between an electric current and a magnetic field at right angles to it, producing an e.m.f. across the conductor in a direction perpendicular to the other two. This is illustrated in Fig. 1. The value of the e.m.f. across X-X, moreover, is strictly proportional to the product of current and magnetic flux density:

$$E \propto BI \quad \text{.................... (1)}$$

A Hall coefficient may be defined which, together with the linear dimensions of the conductor, can be inserted into expression (1) to convert it from a proportionality to an equation. For a Hall plate of given dimensions, and for a given flux density and input current, the coefficient should be a maximum in order to produce the greatest open-circuit output voltage. Maximum open-circuit output voltage is not the only criterion, however. In practice, the input and output circuits of the Hall plate will be matched, especially when it is possible to use transformer coupling. If matching for maximum power transfer is assumed, then it can be shown that the important factor for the Hall material is not the coefficient, but the *mobility*.

Materials with the highest known mobility are the semiconductor compounds indium antimonide and indium arsenide, and, of these, the former has the higher mobility. Indium arsenide is usually chosen, however, since its properties are less temperature-dependant than those of indium antimonide. From expression (1) above; if B and I are made proportional to two other quantities, then a Hall voltage is produced which is in turn proportional to their instantaneous product. In this instrument, the large current to be measured produces a proportional flux in a ferromagnetic ring, and the flux density forms one of the inputs to the Hall plate. The current I, which is the other input, is sinusoidal with a constant amplitude, and of a frequency well above that of the highest-frequency component to be measured. Thus the Hall voltage consists of a high-frequency waveform modulated by a signal proportional to the current to be measured.

* United Kingdom Atomic Energy Authority

Summary of Operation of Current Measuring System

A simplified block diagram is shown in Fig. 2. The conductor carrying the current to be measured is fed through the centre of a gapped magnetic circuit. The Hall plate is mounted in the gap, so that any current flowing in the main conductor produces a proportional flux density in the Hall plate. The other input to the Hall plate is a current of constant amplitude at a frequency of 60kHz. This is fed *via* a matching transformer from an oscillator.

The output from the Hall plate consists of a 60kHz carrier, modulated by a signal proportional to the main current. It is passed through a matching transformer to

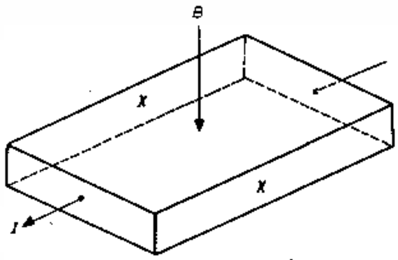


Fig. 1 Illustration of Hall effect

an amplifier whose gain can be switched to different values in order to change the sensitivity of the system, and hence the full-scale current.

The amplifier is followed by a synchronous demodulator which is switched by the 60kHz local oscillator. Following this is a filter which removes the 60kHz component, leaving an amplified signal proportional to the original main current.

The output from the system takes three forms:

(a) an output of 1V for the full-scale main current of the particular range in use

(b) an output of 1mV/A irrespective of the range

(c) a meter reading of r.m.s. current.

Output (a) comes directly from an emitter-follower circuit succeeding the demodulator. Output (b) comes from the same emitter follower *via* an attenuator controlled by the same relays which switch the amplifier gain. The r.m.s. meter in (c) is a thermocouple type, and has an overload-protection device.

The system is calibrated by establishing a known magnetomotive force in the magnetic core, by means of a measured current flowing in a coil, the number of turns of which is known.

The current ranges in amperes r.m.s. of the monitor are as follows:

0 — 30
0 — 100
0 — 300
0 — 1000

The monitor has been designed for a greatest peak/r.m.s. ratio of 3:1 at full-scale deflection.

Detailed Description of the System

60kHz Oscillator and Amplifier

The carrier frequency of 60kHz was chosen because it gives more than ten sampling points per cycle, even at the highest input frequency envisaged. The carrier supply circuit in Fig. 3 contains a Wien-bridge oscillator with its output stabilized at 1V r.m.s. by a thermistor in a feedback loop. This is passed through an emitter-follower buffer amplifier to the power amplifier.

The power amplifier operates in class C, and has a resonant output circuit with a Q of 5. It has two outputs the Hall-plate supply current of 100mA r.m.s. is taken from a winding coupled to the tuning inductor, and the demodulator switching voltage of about 20V peak-to-peak comes from the collector *via* a capacitor.

For proper operation of the Hall plate, a constant alternating current input is required and the power amplifier gives this, being substantially a constant-current device. Although this means that any changes in Hall-plate resistance are reflected as changes in demodulator switching voltage, this is unimportant since the demodulator is relatively insensitive to them.

Magnetic Circuit

In the magnetic circuit there are, broadly, two points to consider. The first is the choice of magnetic material which is to concentrate the flux in the Hall plate, and the second is the determination of the width of the gap in the magnetic circuit which accommodates the Hall plate.

The choice of magnetic material is a compromise between two requirements. Stray signals, such as those due to unbalance of the Hall plate or to coupling between its

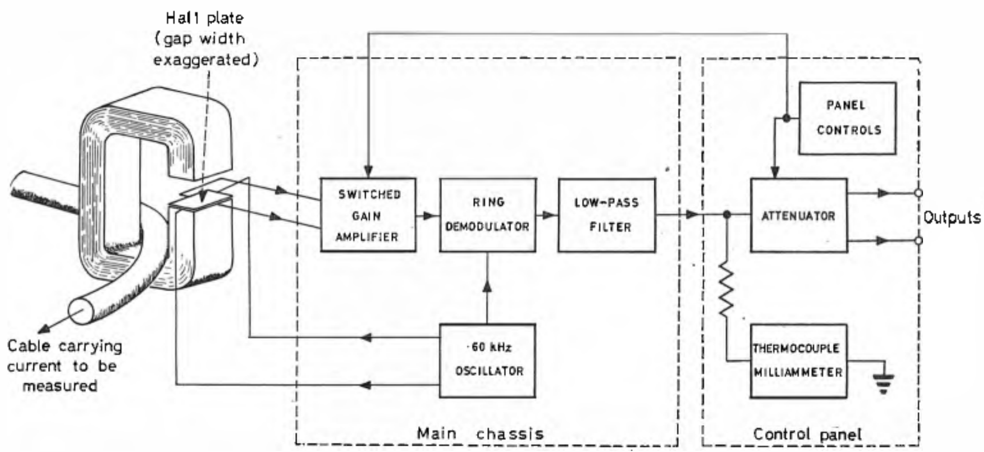
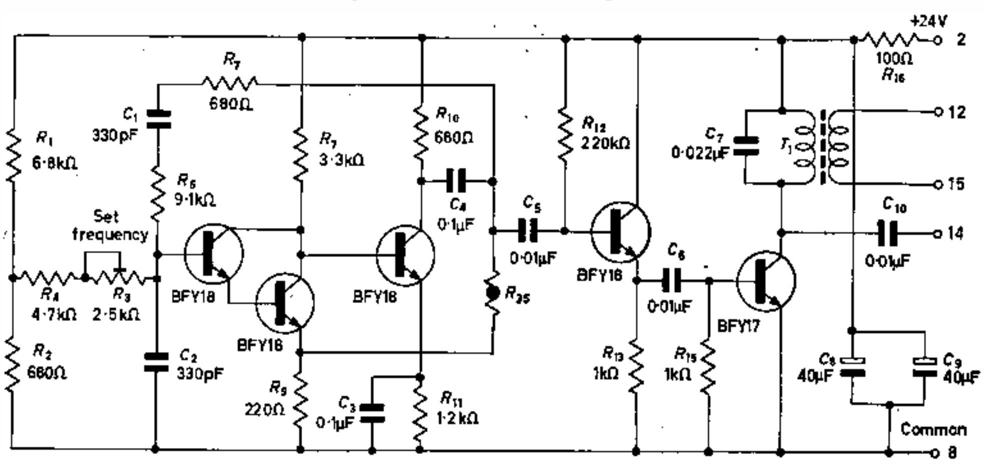


Fig. 2 Simplified block diagram

Fig. 3 60kHz oscillator and amplifier



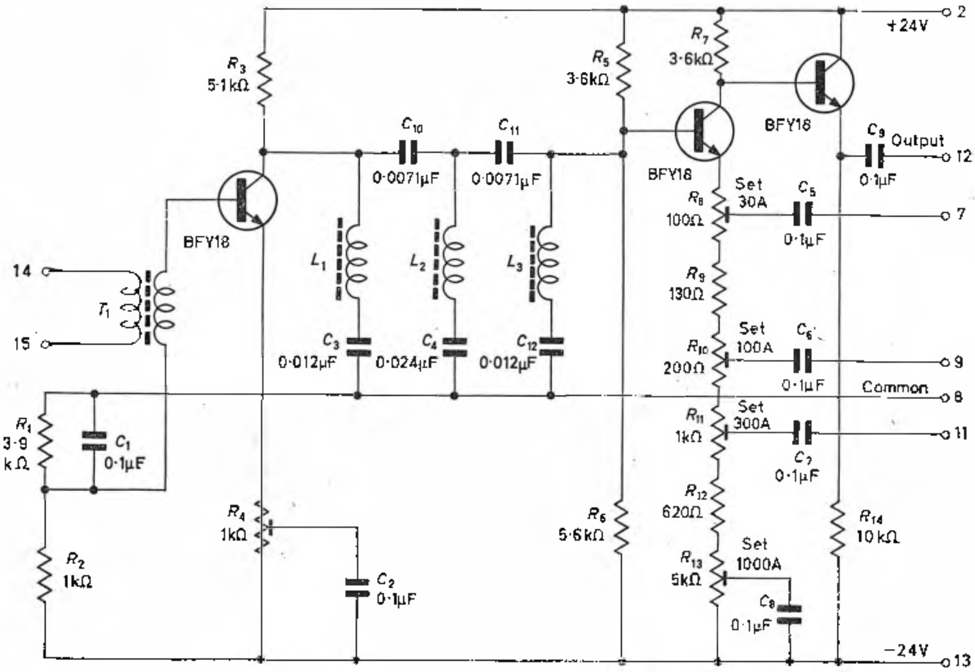


Fig. 4 Switched gain amplifier

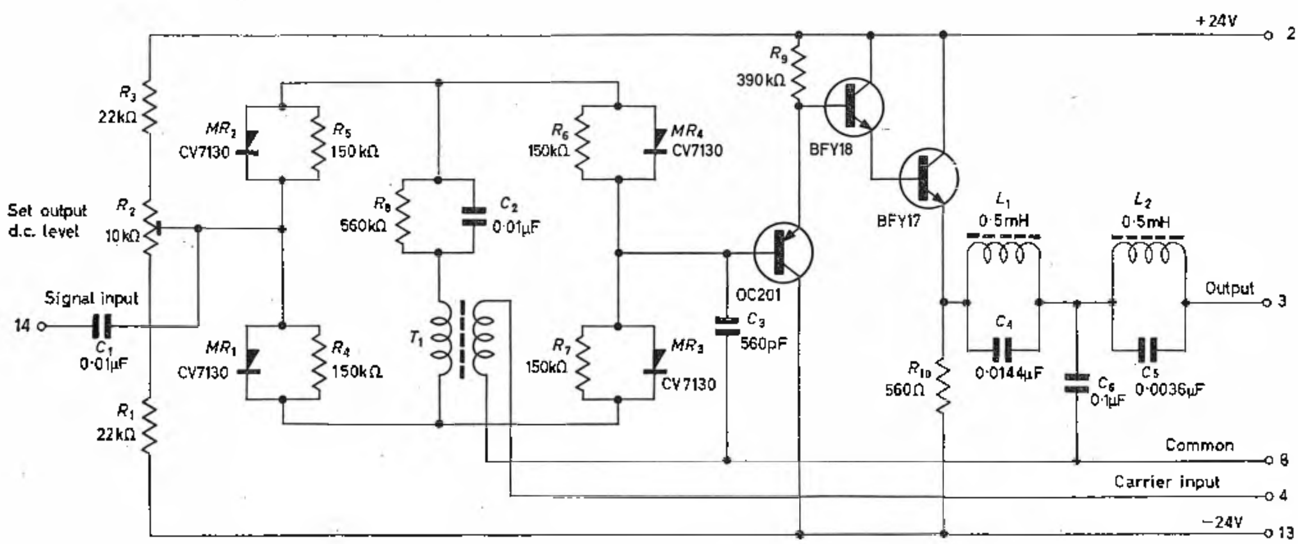
input and output leads, will always be present in the output of the plate and will cause various types of error in the output circuit. To minimize the effects of these, it is necessary to keep the wanted signal amplitude, and hence the flux density, as high as possible. With any magnetic material, distortion of the flux waveform will increase as the peak flux density increases.

A material must be used, therefore, in which a high maximum flux density is obtainable with low distortion. Silicon iron was chosen for this reason.

Three factors influence the choice of width of air gap. First, the width cannot be less than that of plate and clamp. Secondly, in order to keep the flux density high, the gap width should be as small as possible. Lastly, to minimize distortion from hysteresis in the iron, the variable reluctance should be swamped by as large a fixed reluctance as possible by having as large a gap as possible. This third consideration applies particularly to saturation distortion, and is the main factor in the choice of gap width.

The maximum current of the unit is, as given above, 1000A r.m.s., and the maximum peak/r.m.s. ratio is 3:1 This means that a peak current of 3000A must not cause excessive distortion. An arbitrary choice of 10% instantaneous error at this peak was made, and the gap width to give this was determined. (It should be noted that the error in r.m.s. reading in these circumstances will be much less than 10%, the actual error depending on the wave shape.) This optimum width was 0·108in, but has been rounded off to $\frac{1}{8}$in so that material of standard thickness can be used for mounting the Hall plate and as a spacing element between the pole faces.

Fig. 5 Demodulator and filter



Mechanical Design of Measuring Head

The silicon-iron magnetic circuit and the Hall plate are rigidly supported in the measuring head in such a way that the cable(s) carrying the current to be measured may be easily passed through the magnetic circuit. The magnetic material is in two C-cores; the lower of these is cast in epoxy resin and the Hall plate is thinly coated with silicone grease and held in contact with one of its faces. The upper C-core is clamped down against the casting and the gap in the magnetic circuit automatically takes the correct width.

The entire head is mounted on a right-angled baseplate, which provides a choice of two mounting positions.

Switched-Gain Amplifier

The output from the Hall plate is amplified in order to drive the demodulator. Since the full-scale demodulator output is the same on each range, whereas the Hall voltage is proportional to current being measured, the amplifier must have its gain switched to a different value on each range. A circuit diagram is shown in Fig. 4.

The Hall plate is transformer-coupled to the first stage of the amplifier. The gain of this stage is made variable by means of a preset emitter resistor to compensate for Hall-plate sensitivities liable to differ by as much as 3:1.

The second stage is fed from the first *via* a derived-*m* high-pass filter, which is included to minimize interfering signals at the frequency of the current to be measured. (These occur as a result of coupling between the magnetic field and the Hall plate output path). The four preset values of gain are obtained by switching the emitter feed-back in the second stage of the amplifier.

The final stage of the amplifier is an emitter follower which provides a low impedance output to drive the ring demodulator.

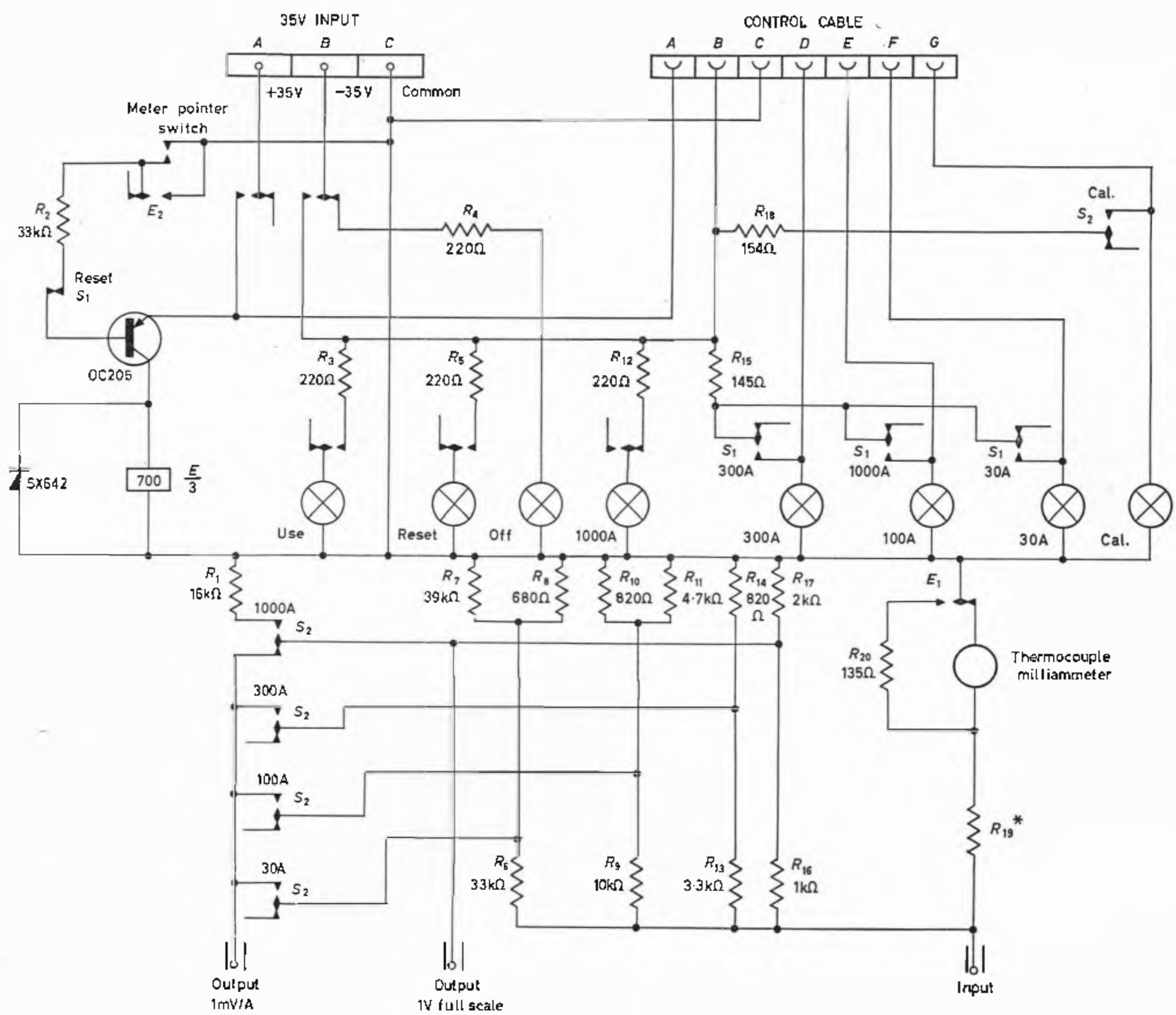
Typical gain figures of the complete amplifier on the four ranges are:

Range	Gain
30A	400
100A	122
300A	40
1000A	12·2

Demodulator

The circuit diagram of the demodulator is included in

Fig. 6 Control panel with output arrangements



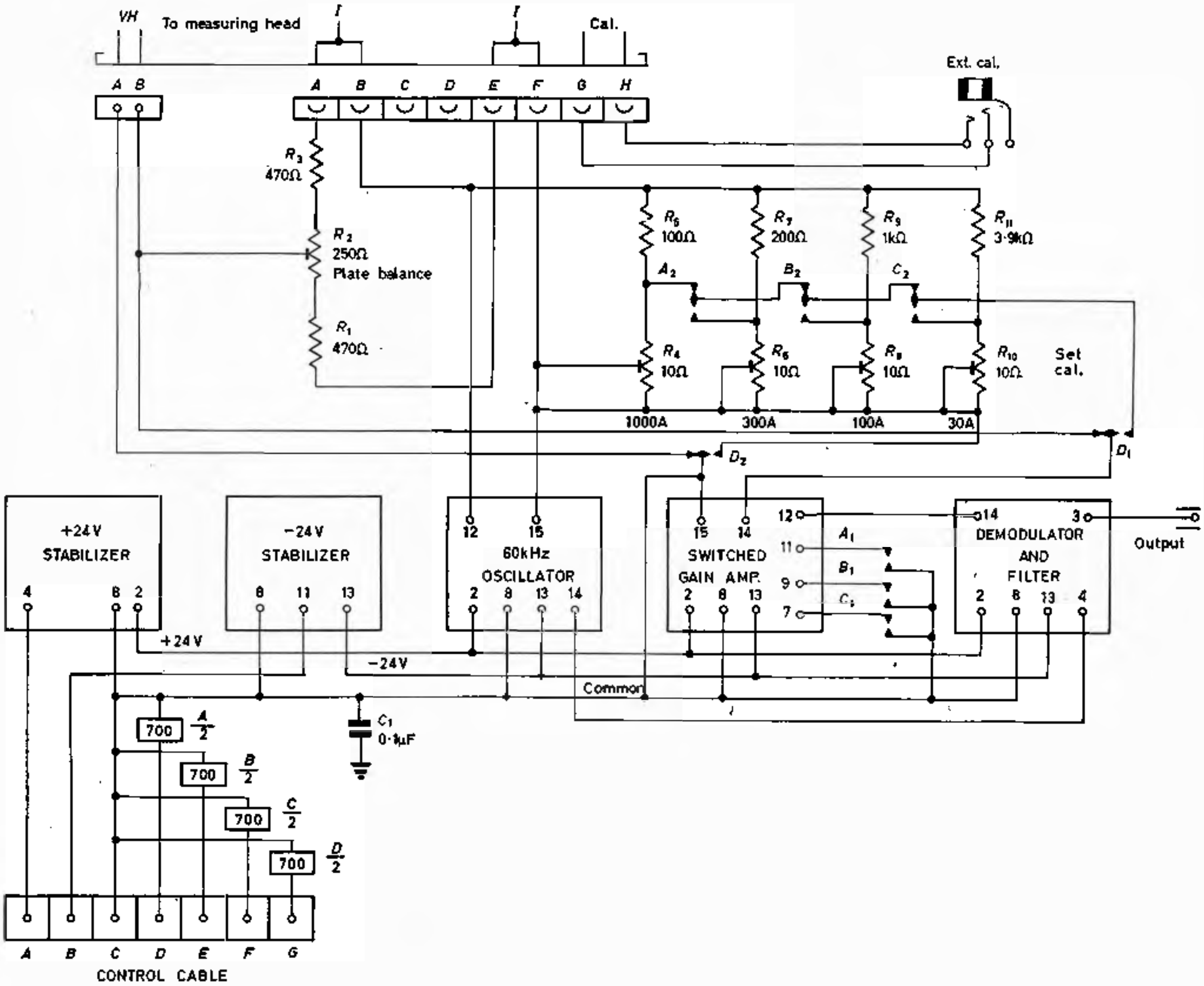


Fig. 7 Main chassis interconnexion diagram

Fig. 5, and is based on the circuit described in reference 2. On positive peaks of the switching voltage, the capacitor C_3 is rapidly charged to the value of the incoming modulated voltage then obtaining. In the intervening halfcycles, the diode switch is open, and C_3 only discharges slowly into the input impedance of the succeeding amplifier. Thus the output of the demodulator consists of a series of steps approximating to the original waveform.

This demodulator has the following advantages: the carrier component in the output is comparatively low, so that a primitive filter may be used to remove it; the output of the demodulator is single-ended, and the demodulator has a rapid response to signals of falling amplitude, as well as to those of rising amplitude.

The demodulator proper is succeeded by a triple emitter-follower stage, in which high input impedance is necessary to avoid discharging C_3 too rapidly, and which is substantially drift-free.

Following the demodulator, and built into the same unit, is the filter. Its main function is to remove the 60kHz component and its harmonics from the demodulator output. It has two parallel tuned circuits in series with the output, one resonant at 60kHz and the other at 120kHz. This type of filter is used in order to keep the phase shift below 5kHz as small as possible. An additional capacitor across the output gives attenuation at high frequencies, and effectively removes harmonics of the carrier beyond the second.

Output Circuits

The output circuits consist of switched attenuators for the various outputs, and the thermocouple r.m.s. meter with its overload protection circuit. They are shown in Fig. 6.

The output attenuators are sited in the control panel. The thermocouple meter gives full scale deflexion for 10mA, and at this current has 1·35V across its terminals. The input to the control panel (from the demodulator and filter) at full-scale is 1·5V r.m.s. and a 15Ω resistor in series with the meter sets the current to the correct value. The '1V full scale' output is fed from the 1·5V line *via* a potential divider network, as also is the '1mV per ampere' output when the '1000A' button is pressed. On any other range the output to this latter position is attenuated to the correct level by one of the three potentiometers, automatic selection being made by the range switch.

The impedance at both output sockets is 666Ω. The '1mV per ampere' output is normally connected to a tape recorder with an input impedance of 16kΩ. This has a negligible effect on all except the 1000A range, where the output is reduced by about 3%. To compensate for this error, a 16kΩ resistor is switched across the '1V full scale' socket on all except the 1 000A range, and the input to the control panel is raised 3% above 1·5V.

Meter protection is provided by a switch actuated by the pointer. When full-scale deflexion is exceeded, the switch turns on a transistor which operates a relay. This switches out the meter and replaces it with a 135Ω resistor, giving a constant load to the output circuits and maintaining a correct voltage to the tape recorder. The meter is not replaced in circuit until the 'reset' button on the control panel is depressed.

Power Supplies

The various circuits are supplied from the 230V, 50Hz mains from rectifiers and stabilizers. The standards adopted are positive and negative 24V, and a separate stabilizer is used for each of these lines. They are derived from circuits given by Marshall[3]. In each stabilizer the output voltage is amplified and used to control the base current of a transistor whose collector delivers the current to the load.

Calibration

There are two degrees of calibration of the current-measuring system; a full calibration in which a known magnetomotive force is applied to the magnetic circuit, and a partial calibration in which the oscillator output is switched to the input of the 60kHz amplifier.

The full calibration is normally only done on the initial testing of the system, and is as follows. A coil of 1000 turns is placed round the upper C-core, and currents of 500, 150, 50 and 15mA are passed through the coil from the supply mains from a Variac and an isolating transformer. These currents produce one-half of the value of magnetomotive force required for full-scale deflexion on each range, and therefore allow each corresponding value of amplifier gain to be correctly set. The calibration coil is removed, and the system is now ready for use.

The partial calibration or 'check', however, is provided for the use of the operator at any time. When the 'cal' button on the control panel is depressed, a relay in the main chassis closes to switch the oscillator output into the switched gain amplifier *via* one of four potentiometers. Each of these potentiometers is pre-set to give a half-scale reading. The correct potentiometer for each range is selected automatically by using the *B* contacts of the range-change relays. The difference between the full calibration and the partial calibration, therefore, is that in the latter the Hall plate is not directly checked. There is an indirect check, however, since in the event of an open- or short-circuited Hall plate, the amplitude of the 60kHz oscillator will change, and this will show in the partial calibration.

Test of Prototype

Testing of the Hall-effect system was approached from two different angles. The first of these was the same as that of the setting-up procedure, in which a known magnetomotive force was applied to the magnetic circuit containing the Hall plate. The second involved checking the system against equipment containing a conventional current transformer and ammeter. Neither of these tests was exhaustive, but the results from the two gave good confidence in the overall accuracy of the system.

Test Using the Setting-up Procedure

This method is described in the calibration section above. It is open to two possible criticisms. The first is that only one frequency of the calibrating m.m.f. was used, namely 50Hz. The problem of whether the same m.m.f. at other frequencies would give different output readings was solved as follows: by passing a simulated Hall voltage into the switched gain amplifier (i.e., a 60kHz carrier modulated by any frequency up to 5kHz) it was shown that the output signal remained constant. The frequency response of the electronic circuitry was thus flat from zero to 5kHz. The only other possible sources of frequency-dependent error, i.e., the magnetic coupling and the Hall effect itself, are above suspicion since they are both strictly independent of frequency. The second criticism of this first method is that the m.m.f. produced, for instance, by a current of 0·5A flowing in a coil of 1000 turns may not be the same as that produced by a single conductor threading the magnetic circuit and carrying a current of 500A, owing to different leakage in the two cases. This possibility was investigated by passing a constant current in a single wire through the magnetic circuit, and noting the output reading when the wire was moved back and forth. The current registered varied by less than 2% of the mean value when the wire was moved between extreme positions, suggesting that the leakage was small.

A further check on magnetic leakage was made as follows. Two of these systems were set up with their measuring heads close together (almost in physical contact). The cable carrying the current to be measured was passed through one of them, and the other was not energized except possibly by leakage flux. Currents of up to 350A were passed in the cable at various frequencies and readings on the two were noted. At no time did the non-energized instrument show any reading other than zero (even on its most sensitive range), and at no time did the proximity of the non-energized measuring head to the energized one affect the current reading from the latter. This result confirms that the errors due to leakage flux are small, and indirectly supports the contention that the setting-up calibration is an accurate one.

Test Against Other Equipment

Currents in the range 100 to 350A having frequencies from 5 to 2000Hz were fed to a resistive load through the current measuring system and also through a current transformer and ammeter. The two readings agreed to within 5% of the full-scale value at all currents used.

Conclusions

A measuring and monitoring system for large currents has been described, which is based on the Hall effect in indium arsenide. It is, in fact, a versatile, high-current ammeter of wide frequency range. It can display the r.m.s. value of current, or produce a proportional output voltage for recording or other purposes at a distant point. Its absolute accuracy is hard to assess owing to the difficulty of producing accurately known currents of some hundreds of amperes. It is, however, thought that accuracy is within 5% of full-scale deflexion on all ranges and at all frequencies up to the design limit of 5kHz.

Acknowledgments

The authors are grateful to the Director, AWRE, for permission to publish this paper.

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LETTERS TO THE EDITOR

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

Cryogenic Regenerative Hall-Effect Devices

SIR,—I think that this is a most interesting device, and of the original papers written by Corbino some years ago, perhaps the worst thing that might be said about Corbino is that, although he discovered something new, he was unable to describe it accurately, and, perhaps plagued by an inadequate supply of laboratory or research funds, he published his accounts inaccurately. His mistakes, as it turned out, were easily detected by investigators in well equipped laboratories and it is my impression that Corbino's good work was equally discredited along with some of his unfortunate errors. To this day there is still some question about the total value of Corbino's work, because of the unfortunate measurement errors. However, this point is now only academic, except that Corbino might have become more famous had he been luckier.

For reasons of historic interest as well as from a natural curiosity, I wonder if Dr. Midgley might be able to answer two questions as they relate to the Corbino disk. Could he draw a diagram of the direction of the current in the Corbino disk and also the direction of the self magnetic field?

Yours sincerely,

EDWARD HALAS,
Windsor, Ontario.

The author replies:

DEAR SIR,—There are four possibilities for the directions of the current and magnetic field, determined by (1) the radial source polarity, (2) the p or n nature of the material. For each of these possibilities the axial magnetic field may initially have either polarity; whichever polarity is initiated, the axial field will increase if (1) and (2) are correctly combined (correct combination always has majority carriers flowing inwards).

It is easier and more sensible to illustrate components of the fields and currents rather than the complex resultants. If only a uniform radial current is present in the disk and if this current divides equally in a central axial wire, then the magnetic field is entirely circumferential and falls inversely with radius, as for an infinite wire; the directions of this field change on each side of the disk; inside the disk the field is zero on a central plane and increases in proportion to axial displacement from that plane.

The additional components of fields and currents for an n-type material with

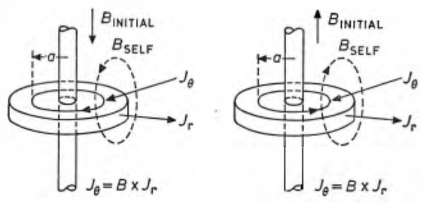


Fig. 1 Components of currents and fields in an n-type material

outward radial source current are illustrated in Fig. 1.

Yours sincerely,

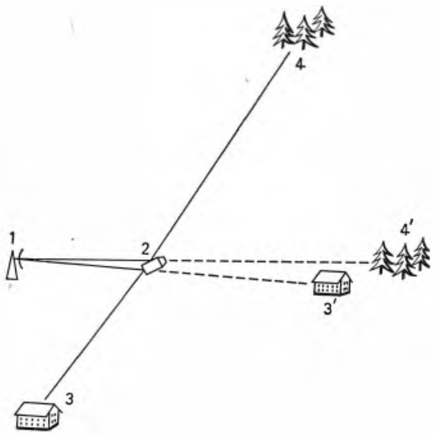
D. MIDGLEY,
University of St. Andrews.
Edinburgh.

The Causes of 'Angel' Echoes on the c.r.t. of Radar

DEAR SIR,—During functioning of a radar device which is operated on u.h.f. and which is equipped with m.t.i. circuits, various echoes have been observed; these echoes have sometimes been at points where no targets were found and have not been suppressed in the m.t.i. circuits. If displays with various scanning periods are compared, then movements of these angel echoes in various directions are observed.

After several observations the probable causes of these effects have been found: these angel echoes are ground objects which are being reflected from mirror-like plane surfaces of vehicles moving in the neighbourhood of the radar. If the reflector is moving, then the ground target behaves like a moving target. It changes its distance from the radar and also its direction relative to it. The angel echo is shown on the c.r.t. in a direction according to where the reflecting surface is to be found. The echo may, however, be displayed at a point at which there is no actual target. Moving objects such as automobiles, or others capable of reflecting electromagnetic waves, can be in such a position with respect to the radar aerial that they are not displayed on the c.r.t. and thus they are not visible *per se*.

Fig. 2 A typical situation where angel echoes can arise. 1 Radar device; 2 Moving object; 3 Building; 3′ Building angel echo; 4 Copse; 4′ Copse and echo



A sketch of this situation is shown in Fig. 2. It is obvious that the angel echoes can show 'movement' not possible with real vehicles. This explanation may, perhaps, solve the problem of several 'mystery' targets that have been observed on the c.r.t. screens of radar devices.

Yours faithfully,

E. KVÍTEK,
Pardubice,
Czechoslovakia.

A Binary-to-Decimal Convertor

DEAR SIR,—I feel that Mr. Hanna and Mr. Dean did less than justice to each other in their correspondence in your March issue.

If Mr. Hanna had developed a truth table for adding 6 after a decade boundary had been crossed as well as when a digit value exceeded 9, he would have arrived at the same result as given in Table 1 of Mr. Dean's article.

Clearly adding a filler of 3 before a shift pulse is the same as adding a filler of 6 afterwards. The only difference between the two systems is that the rules for adding the filler 3 cope automatically with the last two cases on the truth table. Like Mr. Hanna I find it difficult to understand why this should be and even more difficult to see how the system of adding fillers 3 is deduced to be correct.

I have arrived at both sets of rules in a slightly different way which will I hope help to explain the method of conversion.

The normal way of converting the radix of a number is to divide successively by the new radix, writing down successive remainders.

Thus if
$N = a_n p^n + a_{n-1} p^{n-1} \ldots . a_2 p^2 + a_1 p^1 + a_0$
dividing N by p we get a remainder $R = a_0$ and a quotient Q_1;
dividing Q_1 by p we get a remainder $R = a_1$ and a quotient Q_2;
and so on.

Now consider $N = Q_1 p + R$;
shifting left gives $2Q_1 p + 2R + b$, where b is the next bit shifted in from the right.

The new remainder will be:

STATE BEFORE SHIFT PULSE	STATE AFTER SHIFT PULSE			STATE AFTER 6 ADDED IF REQUIRED	
$QRST$	P	$QRST$		P	$QRST$
0000	0	$000X$		0	$000X$
0001	0	$001X$		0	$001X$
0010	0	$010X$		0	$010X$
0011	0	$011X$		0	$011X$
0100	0	$100X$		0	$100X$
0101	0	$101X$	exceeds nine	1	$000X$
0110	0	$110X$	exceeds nine	1	$001X$
0111	0	$111X$	exceeds nine	1	$010X$
1000	1	$000X$	boundary crossed	1	$011X$
1001	1	$001X$	boundary crossed	1	$100X$

$2R + b$ if $2R + b < p$,
or $2R + b - p$ if $2R + b \geqslant p$,
and in the latter case 1 must be added to the next decade.

For a p of 10 we get the rule:
if $2R < 10$ shift right
if $2R \geqslant 10$ shift right, subtract 10 and add 1 to the next stage.

As 6 is the 16's complement of 10 this is the same as shifting and adding 6, the 1 being added automatically to the next stage.

The above rule is the same as:
if $R < 5$ shift right,
if $R \geqslant 5$ subtract 5, shift right and add 1 to the next stage.

As 3 is the 8's complement of 5 this is the same as adding 3 and shifting, the 1 being added automatically to the next stage.

Note that because 10 is even b does not affect the logic.

In Mr. Dean's original article he stated that b.c.d.-to-binary conversion can be performed by a similar method to the one described for binary-to-b.c.d. conversion. If he meant to do this by shifting left and subtracting 3, the b.c.d. digits would have to be presented four at a time, which would complicate the logic.

On the other hand if conversion is performed by shifting from left to right the digits may still be presented singly.

If N is a b.c.d. number in the form

$$(a_n . 2^3 + b_n . 2^2 + c_n . 2 + d_n) . 10^n$$
$$+(a_{n-1} . 2^3 + b_{n-1} . 2^2 + c_{n-1} . 2 + d_{n-1}) . 10^{n-1}$$
$$+ \cdots$$
$$+ (a_1 . 2^3 + b_1 . 2^2 + c_1 . 2^1 + d_1) . 10$$
$$+a_0 . 2^3 + b_0 . 2^2 + c_0 . 2 + d_0$$

then dividing by 2, the remainder is d_0, and

$$Q_1 = (a_n . 2^2 + b_n . 2^1 + c_n) . 10^n$$
$$+(a_{n-1} . 2^2 + b_{n-1} . 2 + c_{n-1} + d_n . 5) . 10^{n-1}$$
$$+ \cdots$$
$$+ (a_1 . 2^2 + b_1 . 2 + c_1 + d_2 . 5) . 10$$
$$+a_0 . 2^2 + b_0 . 2 + c_0 + d_1 . 5$$

This gives the rule:

After shifting right, 5 must be added to a stage if, before shifting, the least significant digit of the next more significant decade was 1.

Not surprisingly this gives a truth table exactly the reverse of that given above and a certain amount of simplification in the logic is possible if the same circuit is used to perform both conversions.

I hope that the above analysis will help to remove any lingering doubts about the practicability of Mr. Dean's system of conversion and of the validity of Mr. Hanna's comments if not of his conclusions drawn from them. I am most impressed by the ingenuity and economy of the system and would like to congratulate Mr. Dean on devising it.

Yours faithfully,

C. P. E. Brown,
Royal Naval Engineering College,
Manadon, Plymouth.

Dear Sir,—Having read Mr. Hanna's comments in your March issue, and re-read Mr. Dean's original article (October 1966) I feel that a further comment may be useful.

The standard method of radix conversion is by repeated division by the new radix, successive remainders giving the required digits in order of increasing significance. This is what Mr. Dean's circuit does. For binary-decimal conversion one has to divide by 10, which may be done by dividing first by 5, then by 2. The second step is trivial: one merely omits shifting one place left. To divide by 5 one examines the first four digits (or in the first step, the first three) Q,R,S,T, and if they represent a number greater than 5, one subtracts 5 and records the digit 1 in the quotient (P), otherwise 0 is recorded. Q thus becomes 0, and one shifts the digits in the R,S and T columns one place left to form the new digits Q,R,S, and continues the process (Table 1). Adding 3 to a four-digit binary number greater than 5 and less than 9 has the same effect on the three least significant digits as subtracting 5. (Numbers greater than 9 will not occur, as 5 would have been subtracted in the previous step.) In Mr. Dean's circuit division by 10 for the second time (to obtain the tens digit) is begun and continued as the required digits become available, and so on for the higher decimal digits.

Yours faithfully,

K. S. Hall,
The City University,
London.

The author replies:

Dear Sir,—I am indebted to Lieutenant Brown for his careful analysis of the method I had proposed for binary-decimal conversion. The essence of the method is the point which he makes that successive division can be used so that the resulting b.c.d. number consists of the remainders from these divisions.

It is perhaps interesting to point out that the method is quite general for division by any even integer. In the case of division by 10 the number is first divided by 16, by introducing four extra digits to the left of the most significant digit and then modifying this during the subsequent shifts. In the same way, to divide by 6, one must first divide by 8 and then add in a filler to compensate for the 'over division' by 2.

Thus, for example 141 ÷ 6 is as follows:

```
        0 0 0   1 0 0 0 1 1 0 1
        0 0 1   0 0 0 1 1 0 1
        0 1 0   0 0 1 1 0 1
        1 0 0   0 1 1 0 1
      1 0 1 0   1 1 0 1
    1 0 1 0 1   1 0 1
  1 0 1 1 0 1   0 1
1 0 1 1 1 0 0   1
1 0 1 1 1 0 1 1
```

Result 23 remainder 3

In this example the filler of 1 was added whenever the three digits examined had a value of 3 or more. In general, to divide a number m by p, where p is even and can be contained in q bits, first divide by 2^q and shift the number through a register, examining q bits at a time, and when these represent a number $p/2$ or greater, then $(2^q - p)/2$ is added.

When the divisor is odd, division by $2p$ can be used, followed by a single shift, so that the final division is by p. Thus to divide by 9, one must first divide by 18. This involves five stages of the register and examining the logic for five literals, using a filler of 7. This is still within the scope of the method I suggested in the original article.

As might be expected, this same method can be used for successive doubling of a b.c.d. number. In this case also the filler must be added. This is shown in the following example, where the number, 79, is sucessively doubled to 158, 316 and 632.

```
        0 1 1 1   1 0 0 1
            add 3 to each decade and shift
      1 0 1 0 1   1 0 0 0
            add 3 to each decade and shift
    1 1 0 0 0 1   0 1 1 0
            add 3 to units decade and shift
  1 1 0 0 0 1 1   0 0 1 0
```

Regarding the method for a decimal-binary encoder, I would rather leave this for a separate article.

Yours faithfully,

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

NEW BOOKS

Transformers for Electronic Circuits

By N. R. Grossner. 321 pp. Med. 8vo. McGraw-Hill Publishing Co. Ltd. 1967. Price £5 12s. 6d.

THIS has been written by a transformer design engineer for all users of transformers in electronic circuits. It is not a design manual. The discussion of design considerations is planned to give the user an understanding of the logic and methods of the transformer specialist. With this book the practising engineer will be better able to obtain limit-of-the-art performance from his transformers, to write specifications with greater confidence of obtaining a desired performance and to consult certain charts and tables directly useful in his work.

The book concentrates on basic principles, fundamental relationships and design considerations avoiding complicated equations (or arranging them in a form which allows the basic relationships to emerge) and masses of design data (which can be found elsewhere), on the assumption that most users today select their transformers rather than design them.

Each chapter stresses topics and concepts of major importance in modern electronic circuits. These include the following, some of which are discussed in depth: the basic and the newer functions of the transformer, the iron-copper area product and ratings of the power transformer, rectifier circuits, the d.c. invertor, reliability and environment, temperature rise and techniques of thermal design, miniaturization, the μQf merit factor, harmonic distortion, polarized inductor gap design, the band-span parameters and the application of concepts from network synthesis to the design of wide-band and pulse transformers having optimum performance characteristics. Recent advances such as conductive heat shields, evaporative cooling and the use of newer core materials (e.g. Supermendur, ferrites) are also included.

Certain types of transformer are not treated (e.g. the constant-voltage transformer and the magnetic amplifier). Some specialised considerations, including shock, corona, thermal ageing, transient loading and heating and precision transformers, have been omitted because of space limitations. The author has written elsewhere on these topics and references to these articles are given.

This is a clearly written and well-illustrated comprehensive guide, containing many references, for everyone using electronic transformers. In spite of the almost universal acceptance of the MKS system of units by engineers and teaching bodies, however, the book does not use it.

F. A. BENSON.

Antenna Theory and Design Volume II Second Edition

By H. Paul Williams. 795 pp. Demy 8vo. Isaac Pitman & Sons Ltd. 1967. Price 105s.

VOLUME I deals with fundamental theory, and in the 1950 edition, remains applicable today. Volume II, Electrical Design of Antennae, now re-appears in a revised and enlarged form. As before, it is a complete, independent book.

The introductory chapter is a general survey of aerials and propagation. The next two deal with l.f. and m.f., h.f. and v.h.f. aerials respectively. Chapter 4, Microwave Antennae, is more than doubled in size and occupies one sixth of the book. After receiving antennae comes directional antennae and arrays, with a communications emphasis. The chapters on d.f. and mobile aerials are followed by a new chapter, Antenna Measurements, which includes site reflections. The last chapter deals with transmission lines.

Propagation is now dealt with in detail in an appendix, the earlier appendices on mechanical design, units and constants being retained. The greatly augmented bibliography lists 1064 references. There are numerous diagrams and graphs and 22 plates.

The long section on noise is new, and new topics such as buried aerials, space vehicles and plasma are included. Many new aerials are described, such as the log-periodic family. Smaller additions deal with subjects such as sidelobes, hybrids and baluns.

Dr. Williams writes clearly and keeps his mathematics straightforward, so that appreciation of his book will not be limited to the graduate engineer. Although ostensibly general, the book is better on communications and wire aerials than on radar and microwaves.

It is a pity that advantage was not taken of this opportunity to regroup the contents under more convenient and up-to-date chapter headings. Practically all of the first edition is retained, including early aerials which are omitted from most other modern texts. For a 50% price increase, the revised edition offers over 50% more pages, and is still reasonable value by today's standards.

M. F. RADFORD.

Nachrichtenübertragung

By Hölzler and Thierbach. 931 pp. Med. 8vo. Springer-Verlag, Berlin. 1966. Price DM.86

THIS book on communication techniques is a comprehensive work containing contributions from twenty-nine authors, coordinated by four editors. All the contributors were members of the Siemens Central Communications Laboratory when the project was started. The book is intended both as a textbook for teaching purposes, and as a reference work on communication techniques. This dual role is assisted by the division of the book into two parts, the first on the basic principles and methods of communication, and the second on communication systems.

There are four chapters in the first section. In the first chapter the types of signal, continuous or pulsed, to be transmitted are considered and analysed. Methods of communication, amplitude, frequency and phase modulation, pulse modulation and multiplexing, are discussed in the next chapter. In chapter 3 the media of communication are given—transmission lines, waveguides and free space. In this chapter antennae are examined in some detail. Finally system and network planning is introduced.

The first chapter in the second section deals with low-frequency communication. Then telegraphy and data-transmission are considered, followed by a description of techniques for carrier and pulse transmission. Chapter 5 is devoted to a detailed study of cables, transmission lines and waveguides. Short wave radio techniques and equipment are considered, and there is an interesting chapter on mobile radio communication. The final chapter deals with communication by means of satellites. This is a fascinating chapter giving a brief history of satellite communication, the methods of transmission including Doppler-effect problems and details of the ground station antenna system. The means for achieving amplification of low-power space signals, masers, tunnel diodes, and parametric amplifiers are compared.

The book is well written and clearly

presented. At the end of each chapter a long list of references is given. The book should achieve its objective; it would perhaps have made a better reference book if some of the circuits were given in more detail.

G. D. BERGMAN.

Nachrichtentechnik—eine einführende Darstellung

By K. Steinbuch and W. Rupprecht. 462 pp. Crown 4to. Springer-Verlag, Berlin. 1967. Price DM.48

WHILE the history of electrical communication and data transmission goes back well over a century, the processing of such data by electrical machines is relatively new.

Prof. Steinbuch and his colleague at the T. H. Karlsruhe have broken new ground by treating the whole field, electrical transmission *and* processing of data, in one volume.

The book is divided into three main chapters of approximately equal length. The first discusses the hardware involved, i.e. the unit bricks required to make up the complex circuitry, switching elements, passive networks, linear amplifiers and pulse technology. Chapter 2 on data transmission starts with an interesting section on speech and transducers followed by two sections on transmission paths and modulation and selection which are treated in a more conventional manner. The final chapter, 'Data processing', makes most interesting reading; it deals with codes, information theory, switching algebra, logic circuits and finally with data-processing systems, including the fundamentals of digital computers.

The authors have spared no effort to be as concise as possible, yet to give sufficient details for full comprehension. There are not many books available dealing so fully with all the aspects involved in this new technology, which is likely to be regarded by future generations as perhaps the most revolutionary technological contribution of the century. It is a pity that there is still only a minority of scientists and engineers in the English-speaking world who can read German well enough to make full use of a book of this type. To them, it can be highly recommended, on account of the width of its range, the depth of its treatment and the clarity of its representation.

Printing of text and figures are excellent, typical of Springer publications.

E. BILLIG.

Electronic Circuit Analysis—Volume 2—Active Networks

By Phillip Cutler. 628 pp. Med. 8vo. McGraw-Hill Publishing Co. Ltd 1967. Price 84s.

CHAPTER 1, occupying roughly one-sixth of the book, is entirely concerned with the small-signal valve amplifier and is an excellent chapter in which important aspects of performance, such as gain-frequency dependence, transient response, impedance levels etc., are fully and clearly discussed, and in which the worked examples are practical design problems.

Then follow two chapters of analysis theory, namely four-terminal network parameters and piecewise linearization, which are prerequisite chapters for the treatment of junction diodes and transistors given in Chapter 4. In these the treatment is less sure, the use throughout of electron current rather than conventional current has nothing to recommend it, particularly in the four-terminal network theory, and a number of editing errors are evident. For example, equation (2-46) should read

$$h_o = h_{22} = \frac{i_2|}{e_2|_{i_1=0}};$$

in Fig. 3-3(*d*) the current source polarity is wrong, in Fig. 4-24(a) the scale is wrong by a factor of 10, and in a number of places the polarities of v_{BC} and v_{BE} are wrongly drawn. One wonders, too, whether or not too much emphasis has been placed on the techniques of piecewise linearization and particularly on their application to the derivation of models whose characteristics approximate those of transistors, in a text which is aimed at the training of technicians.

Chapter 5 is a short, definitive chapter on the theory of signal-flow-graph analysis, leading into a very full chapter on feedback principles and applications. This is very well presented, and includes a comprehensive survey of the effects of feedback, a section on the operational amplifier, a good discussion of stability criteria and physical methods of stabilization and finally the analysis of a number of oscillator designs. Signal-flow-graph methods of analysis are used throughout and there are a number of worked examples in both valve and transistor circuits.

The last three chapters of the book are devoted to transistor power amplifiers, rectifiers and controlled rectifiers and power-supply regulators. These, too, are well presented, the design examples chosen use typical manufacturers' data and there are a number of very useful worked examples. Again editing errors are evident, probably the worst of which is equation (7-21) which as written defines I^2_{rms}, not I_{rms} as stated.

To sum up, the book is ambitious in scope for its intended readership of technicians for, in fact, university-course students would find it very useful as a handbook of electronic-design principles. It is very comprehensive and the only serious adverse criticism that one can make is on the use throughout the book of electron flow rather than conventional current flow. In the first place, this has led to many errors in sign and secondly in an instructional work it is certainly not a good thing to present analysis in a form which is different from that of the generally accepted practice.

C. S. GLEDHILL.

Les Machines Électriques en Automatique Appliquée

By J. Henry Baudot. 310 pp. Royal 8vo. Dunod, Paris. 1967. Price 68F.

HERE is a book intended for those with an electronic bias who want to keep abreast of the more recent developments in the field of electric machines used in automation and control engineering at present.

By way of an introduction, the first chapter sketches the evolution and scope of electrical engineering in general terms. The various electrical materials, including some recent additions, which are used in electric machines, are described in Chapter 2. Reference is made in Chapter 3 to the new techniques which have improved the performance of modern machines.

The fundamental principles of motor theory are summarized in Chapter 4 and the basic requirements for an electric machine are developed in the following chapter. From the functional and theoretical viewpoints, Chapter 6 serves as a brief introduction to the various types of electric machines described in more detail in subsequent chapters.

After describing the ring and drum windings in Chapter 7, the on-load characteristics of the drum winding in motors and generators are discussed in Chapter 8. Synchronous motors and alternators are dealt with in Chapter 9 and d.c. motors and generators in the following chapter.

Two short chapters on modern synchronous machines and development from d.c. machines precede the introduction to induction motors. The various types of induction motor, including the linear induction motor, are described in Chapter 13.

Various motors and other devices used in control systems are treated in the following three chapters. Low-power synchronous motors, stepping motors, magnetic clutches, tachometers, synchros and power amplifiers are given careful consideration. Servometers and methods of controlling them are analysed in Chapters 17 and 18. The last chapter is concerned with the various units of measurement; some methods of measuring servomechanism characteristics and calculating certain parameters are also given.

The author has used a straightforward approach to the theory and has avoided over-complex mathematical expressions. This book could go a long way towards relieving the anxieties of electronic engineers, among others, when faced with the problem of choosing an electric machine to suit their particular requirements.

D. K. HILLS.

SHORT NEWS ITEMS

Exports of valves, tubes and semiconductor devices during the first quarter of 1967 amounted to £3 589 484, an increase of just under 3% over the same period of 1966, according to a statement issued by the British Radio Valve Manufacturers' Association (BVA) and the Electronic Valve and Semiconductor Manufacturers' Association (VASCA).

Exports of valves and cathode-ray tubes for the first quarter of this year amounted to £2 987 833 (as compared with £2 769 637 for the same period in 1966) and the exports of semiconductors amounted to £601 651 (as compared with £689 793 for the same period in 1966).

British computer manufacturers exported 44% of their output last year. This achievement was well in excess of anything previously attained—the earlier record being for 1962, when almost 22% of the output was exported. The Ministry of Technology statistics for manufacturers' deliveries of computers and peripherals show record sales in 1966 in both the home and export markets. Total deliveries reached £70M, of which almost £39M were home sales and £31M were for export.

The exports were almost equally divided between peripheral equipment and digital-computing systems, both amounting to £15·2M. Exports of analogue and hybrid computers and data-transmission equipment accounted for the small residue.

A British Cryogenics Council has recently been formed by the following bodies:

Society for Low Temperature Biology
The Institution of Chemical Engineers
Society of Chemical Industry
The Royal Institute of Chemistry
The Institution of Electrical Engineers
The Institution of Gas Engineers
The Institution of Mechanical Engineers
The Institute of Petroleum
The Institute of Physics and The Physical Society
The Institute of Refrigeration

The following bodies were also represented at the first meeting of the governing body:

The Chemical Society
The Institution of Electronic and Radio Engineers
The Institute of Metals
The Iron and Steel Institute

The objects of the Council are as follows:

(a) To stimulate interest in, spread knowledge of, and foster the development and applications of cryogenics.
(b) To afford a common meeting ground for the adhering organizations, whereby such of their activities as fall within the purview of the Cryogenics Council can, if they so desire, be co-ordinated and extended, and generally to stimulate dissemination of knowledge in this field.
(c) To advise on education and research needs in the field of cryogenics.
(d) To promote international co-operation in cryogenics, to maintain liaison with appropriate bodies in other countries, to support international conferences on aspects of the subject and to encourage adequate contribution from the UK to such conferences.

The secretariat for the British Cryogenics Council is being provided by the Institution of Chemical Engineers, 16 Belgrave Square, London SW1 (telephone: 01-235-3647).

A 15-year agreement has been signed between The General Electric Company Ltd and the Plessey Company Ltd in the field of public-telephone-exchange equipment. The agreement provides for:

1. collaboration in research and development
2. facilities to offer countries throughout the world total communications networks, comprising exchange switching equipment, transmission and other telephone plant
3. the manufacture and sale by both companies of the 5005 Crossbar system developed by Plessey.

The agreement is seen as an important measure of rationalization in the UK telecommunications industry. It is expected to provide the conditions to meet the technical and production needs of the British Post Office and to greatly strengthen the position of the UK industry in overseas markets.

The negotiations between The General Electric Company and Elliott-Automation, under which Elliotts were to take over the GEC Computer and Automation Ltd, have been discontinued. It has now been decided to integrate GEC Computers and Automation with GEC (Electronics) Ltd at Stanmore and Portsmouth.

The Electronic Engineering Association, with Government support, is to sponsor an international symposium on electronics for civil aviation on 15-19 September 1969 at the Imperial College of Science and Technology, London.

The aim of the symposium will be to stimulate international interest in the techniques of civil-aviation electronics by illustrating how British civil aviation is solving current and future problems, and to highlight the achievements of British industry in research, development and application of advanced techniques to meet the most stringent operational requirements.

Britain's first commercially available computer-time-sharing service using the public-telephone dialling system will be inaugurated later this year by **De La Rue Bull**. Time sharing allows many people to use a large computer simultaneously *via* directly connected remote terminals, regardless of where they are located. Time sharing of computers in Britain and Europe has so far been either experimental or limited to relatively few users *via* special private lines. It has not yet become generally available to anyone who wants to install a data-transmission terminal in his premises.

Four new submarine cables linking the United Kingdom with Belgium, the Netherlands, the Federal Republic of Germany and Scandinavia have been recommended by a conference of seven European countries meeting in London.

These cables will be of a new design, with a capacity of 1140 telephone circuits, which is more than double the capacity of any cable system so far laid. The Post Office Research Station at Dollis Hill are designing a suitable system for this capacity.

The first cable—to the Federal Republic of Germany—is expected to be in service before the end of 1970, and the others will follow at yearly intervals.

Subject to confirmation by the administrations concerned, tenders will shortly be invited. The seven countries represented at the conferenec were the United Kingdom, Belgium, the Netherlands, Federal Republic of Germany, Denmark, Sweden and Norway.

AEG-Telefunken have undertaken the full development and production of the current-supply system and the telemetric transmitter and receiver, and the construction of the tape recorder, for the first West German exploration satellite 'Azur', which is expected to be launched in the autumn of next year.

This first German exploration satellite is intended to carry out experiments in near-Earth missions in co-operation with the American Space Agency (NASA), who will make available free of charge the fourstage Scout booster rocket, the launching pad and the tracking stations.

As examples of the missions envisaged during the next period of solar activity, beginning in the autumn of 1968, it is proposed to carry out measurements on the energy spectra of electrons and protons and their flux density in the inner Van Allen radiation belt, in addition to investigations of the influence of solar eruptions on the magnetic field of the Earth, at the same time measuring the intensity of various auroral spectral lines.

The scientific and technical management, as well as the financing, of the development and production of the first German exploration satellite is in the hands of the Gesellschaft für Weltraumforschung (Company for Exploration in Outer Space), acting on behalf of the Federal Ministry for Scientific Research.

Ferranti Ltd, Edinburgh, have been awarded a contract by the Ministry of Technology for the supply of Inertial Navigation and Attack Systems (INAS) for prototype models of the strike version of the Anglo-French Jaguar aircraft for the RAF.

This INAS installation, which is a derivative of those ordered for the Phantom and Harrier, gives RAF aircraft an unrivalled capability in weapon delivery.

It was designed by Ferranti for incorporation in a wide range of aircraft and is the only one of its type to be designed and manufactured in Europe. It will continue to form the basis for military-aircraft navigation and weapon-delivery techniques for many years.

The Wayne Kerr Co. Ltd has been awarded a contract worth more than £100 000 by the Ministry of Technology, for the development of a precision attenuator calibrator. The specification requires that the measurement uncertainty should be less than 0·005dB up to 100dB from 200MHz to over 10GHz, and calls for an overall capability of 150dB over the same frequency range.

Measurements are referred to an internal standard piston operating at 60MHz, whose physical dimensions are of critical importance, since ovality and taper must not exceed 1μm over an effective length of 16cm. Furthermore, in order to calculate absolute attenuation with the required high precision, surface microstructure must be taken into account.

A number of frequency changers will be employed to cover the full range, and all of these must operate with a very high order of linearity, in order to maintain the full measurement accuracy.

Elliott-Automation have been awarded the contract for the automatic-test equipment for the F-111K swing-wing fighters for the RAF. This order, worth over £2M, is the largest-ever single order placed for automatic test equipment in Europe. A large proportion of the equipment is common to the automatic-test installations recently ordered for the HS801 Nimrod maritime aircraft. Each test-equipment set will be controlled by three Elliott 900-series computers: one will select and route the automatic-test programs and the others will run the test sequences and generate written results and instructions on cathode-ray-tube displays; they will also control the printers recording the results of the tests.

These automatic-test installations are versatile and can test all the different types of electronic 'black box' found in a modern aircraft, including the radio and radar, in a fraction of the time required by manual methods, thus greatly increasing operational availability and the speed of turn-round.

The Irish Department of Posts & Telegraphs has awarded a contract to Philips Electrical (Ireland) Ltd for the supply and installation of a radio-relay network which will link Dublin to Portlaoise, Athlone, Cork and Belfast. The project has been designed to expand the number of existing telephone circuits between these centres. The first stage of the project consists in linking Dublin with Portlaoise, via a radio terminal at Capard. This will enable a maximum of 960 simultaneous telephone calls to be made between Dublin and Portlaoise. A protection channel will be switched into operation automatically in the event of failure or degradation of the main radio channel.

The second stage of the project will link Athlone to the network by a radio channel between Athlone and Capard. Cork will then be connected to the radio terminal at Capard; this will give 960 further telephone circuits between Dublin and Cork. Together with this network to the south of Ireland, a radio channel will be beamed from Dublin via three repeater stations to a terminal outside Belfast. This will expand the existing number of telephone circuits to Belfast and the rest of Britain.

The radio equipment for this telephone network will be manufactured by N.V. Philips Telecommunicatie Industries, Hilversum, Holland.

A satellite-communication terminal has been installed for trials in HMS *Wakeful*, the 2200-ton frigate which is currently employed as a trials and training ship. The project is part of the research programme at the Admiralty Surface Weapons Establishment near Portsmouth. At present, *Wakeful* is the only RN ship at sea fitted with the facility to participate in the Interim Defence Communication Satellite Program IDCSP. This is a US trials programme in which the UK and NATO are collaborating in the use of military satellite communications. The 'space segment' is provided and financed by the US Department of Defense and managed by the Defense Communications Agency in Washington. An agreement concluded between the US Department of Defense and the UK Ministry of Defence allows a limited number of British earth stations to operate within the network; three other British stations are in operation, one at Christchurch and two overseas. These stations have larger aerials but occupy much less space than conventional terminals and can be moved more readily.

The provision of a shipborne system presents special problems, and the Naval terminal has been built to explore them. In particular, the terminal is being used to determine the communication-traffic capacity of a small terminal in a satellite network and the procedures necessary to provide an effective mobile communication centre.

Plessey Radar Ltd built the transmitter-receiver system installed in HMS *Wakeful* at their Space Division at Cowes, Isle of Wight.

The first publications in the Burghard scheme—the scheme for electronic parts of assessed quality, which is now recognised as the first major attempt to produce common standards covering both military and civil purposes—are now available from the British Standards Institution. These are BS 9000:1967 'General requirements for electronic parts of assessed quality' and BS 9001:1967 'Sampling procedure and tables for inspection by attributes for electronic parts of assessed quality'.

Optimism is being shown over readings on current and tide patterns in the Singapore area, where ambitious new industrial harbour complexes are envisaged. To aid them in their 6-month survey, which began in February, a Canadian team working in close co-operation with the Singapore Economic Development Board, is using some of the latest British-built electronic equipment. MO21 Recording-Current Meters manufactured by the Marine Systems Division of the Plessey Electronics Group, have been submerged deep in the sea to the south-west of Singapore, and are now providing information which will influence the location and size of the new harbour. One of the meters is situated between Sisters Island and St. Johns, another is to the south of Blakang Mati and a third is at Tanjoug Teritip, off Jurong. Because the water to the south of Blakang Mati is believed to be the deepest in the region, the reading from the second meter is considered to be the most important.

As the new ports will be using electronic methods to handle super oil tankers and other large ships of the future, accurate current and tide readings are essential.

Home disposals to the trade of radio and television receivers during March 1967 were below those for the same month of 1966 and 1965, according to the Economic & Statistical Division of the British Radio Equipment Manufacturers' Association. At 96 000, television despatches to the trade showed an increase of 2 000 over February of this year, but were well below the 126 000 and 207 000 of March 1966 and 1965, respectively.

Radiograms, at 12 000, showed a deficit of 1000 compared with last year and a drop of 4000 compared with the previous year. Home disposals of radio receivers showed an increase of 19 000 over last month but were 6000 below those for March 1966 and 54 000 lower than the comparable figures for March 1965.

These estimates are net figures of deliveries by manufacturers to the home market on firm and other accounts, including those to specialist rental and relay companies. Radio receivers include car radios.

Standard recommendations for the abbreviation of periodical titles have been published recently as BS 4148: 1967 'Recommendations for the abbreviation of titles of periodicals.' The purpose of abbreviating periodical titles is to save time and space in writing, typing or printing such titles for bibliographical references and title lists. The recommended abbreviations deviate as little as possible from the usage of the World List and other major lists of serial publications. Copies may be obtained from the BSI Sales Office, 101-113 Pentonville Road, London N1. Price £2 each (postage 2s. extra to nonsubscribers).

The Electronics Division of the Institution of Electrical Engineers, in collaboration with the Institution of Electronic and Radio Engineers, has arranged a conference on computer technology to be held in the Renold Building, the University of Manchester Institute of Science and Technology, on 18-20 July 1967. In addition to sessions on automatic methods, storage, central-processor techniques, semi-conductor devices, transmission-line circuit techniques, communications and displays, the programme will include a series of informal discussions, in which it is hoped that speakers from eastern Europe will take part. Technical visits on Thursday, 20 July to ICT Ltd, West Gorton, Ferranti Ltd, Wythenshawe, and English Electric Computers Ltd, Kidsgrove, will complete the conference programme.

Further details and registration forms are available from the Conference Department, IEE, Savoy Place, London, WC2.

The Post Office will continue to be the name of the new public corporation that will be established instead of the present Government department by legislation to be introduced in the 1967-68 Session of Parliament, according to a White Paper published recently. The Corporation will be given statutory powers to run Post Office services, including Giro. The Bill will give the Corporation various general powers, including the power to manufacture anything it uses. It will also have power to form subsidiaries and to engage in joint organizations. It will be expected to consult the Minister concerned if it proposes to undertake on a large scale any manufacturing work which the Post Office had not done previously.

The Corporation will have the Postmaster General's existing telecommunications monopoly. The position of private telecommunications networks operated at present outside the Postmaster General's monopoly will not be affected.

Among the functions of the single Minister who will take over from the Postmaster General will be the appointment of the chairman and members of the Corporation. He will assume the Postmaster General's present functions in wireless telegraphy, and his associated functions in the broadcasting field, including general control of radio transmission and reception and the issue of licences. He will also assume the responsibility for broadcasting policy and constitutional functions in relation to the BBC and ITA.

The first all-British satellite, Aerial 3, which was launched from the Western Test Range, California, on 5 May is successfully relaying back to Earth the data collected by the experimental equipment which is on board. This equipment, and the telemetry and control systems, are powered by batteries which are kept charged by 7392 solar cells mounted on the body and booms of the satellite.

The solar cells, of which 38 000 were ordered for the Ariel 3 project, are blue-sensitive *n*-on-*p*-type cells specially developed by Ferranti Ltd in conjunction with the Royal Aircraft Establishment, Farnborough, for electric-power generation on spacecraft.

Development work is continuing in order to achieve an increase in efficiency and a decrease in weight of these cells for future UK and international space projects.

EMI Electronics is currently engaged in developing scanning equipment to produce infra-red television pictures. Results from the equipment, which is being developed for the Royal Aircraft Establishment, were shown recently at the Infra-red Symposium at the Royal Radar Establishment, Malvern. Whereas a normal television system utilizes the reflexion of light, infra-red television uses the heat, or infra-red radiation, emitted by an object. With this system, temperature variations of a few hundredths of a degree can be detected and, since the method does not require light, pictures of acceptable quality can be obtained in complete darkness.

Infra-red-television systems find uses in both military and commercial fields, wherever contrasts between hot and cold can yield useful information. In many applications, phenomena can be detected that would be difficult if not impossible to discover by any other known means. Examples in the field of medicine include the early diagnosis of cancerous growths and areas of poor blood circulation. In the military field of activities, one example of its use is in a night-surveillance system without disclosure of the presence of an observer.

One of the first practical applications of a new technique which enables both analogue and digital circuitry to be combined on a single chip of silicon and contained in one microminiature integrated-component package has been achieved by Elliott-Automation at their new Microelectronics Research and Development Laboratory at Glenrothes, Fife.

A complete audio switch, of the type used in many forms of communication equipment, has been produced using metal-oxide/semiconductor integrated circuit technology. The significance of this is that the ability to produce a truly integrated combination of digital and analogue circuitry on one silicon chip can lead to important reductions in the cost of electronic components and to major increases in reliability. Previous attempts to produce microminiature components of this type have consisted of separate digital and analogue circuits wired together within a single component package.

A new edition of PD 5686, 'The use of SI units' has just been published by the British Standards Institution. This booklet was first issued 18 months ago to provide a simple explanation of the units of the Système International, and has subsequently been in great demand among industrial and educational bodies. Some 60 000 copies have been sold to date. The new revision takes account of recent discussions and agreements reached at the international level since December 1965.

Copies of PD 5686 may be obtained from the BSI Sales Office, 101-113 Pentonville Road, London N1. Price 2s. each (postage 6d. extra will be charged to nonsubscribers).

A Marconidata data-transmission system has been sold to Czechoslovakia by The Marconi Co. It will be used by the Ministry of Transport (Railways) Research Institute, SUDOP, in the course of a complete evaluation of the use of the telephone network for data transmission. This new system, type H6010, forms part of the Marconidata range of medium-speed data-transmission equipment. By making a simple telephone call, an operator can transfer information from punched paper tape into a computer, simply by pressing a button on the transmitting Marconidata terminal. No manual intervention is necessary at the receiving end.

Data can be transmitted at up to 50 000 characters per second, more than ten times faster than with a conventional telegraph or teleprinter circuit, and with a very high degree of protection from errors caused by noise and interference on the telephone circuits. It is claimed that less than one character in ten million will be transmitted incorrectly.

NEW EQUIPMENT

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

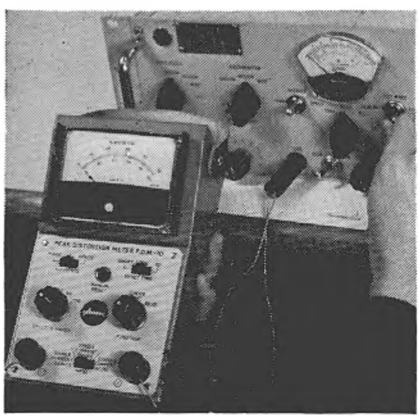
PEAK-READING DISTORTION METER

The Plessey Automation Group, Data Handling Division, Ilford, Essex

(Illustrated below)

A transistorized, battery-powered, portable peak-reading distortion meter, measuring 8in × 4½in × 3in (20·4cm × 11·4cm × 7·6cm) is available from the Data Handling Division of the Plessey Automation Group. The meter, designated PDM 10, measures peak distortion on d.c. telegraph circuits.

The equipment operates on the shortest-pulse principle by which only elements shorter than unit length are measured. The shortest pulse—being the peak distortion—is indicated on a meter display, which may be reset automatically at either of two time constants, or may be retained for a period with manual reset. Using the shortest reset time, the PDM 10 can be used to observe distortion variation while aligning terminal or v.f. equipments.



A secondary meter scale is provided for line voltage and line current measurements for shunt and series monitoring respectively. Provision is also made for checking the internal batteries, which provide a minimum life of 75 hours continuous operation at 20°C.

The equipment has switch selection for monitoring mark elements only, space elements only or all elements, to assist in identifying the type of distortion.

The meter may be used with signals of any unit code and with start-stop or synchronous systems. Five fixed speeds may be provided within the range 45 to 200 baud.

For more information circle **No. 291**

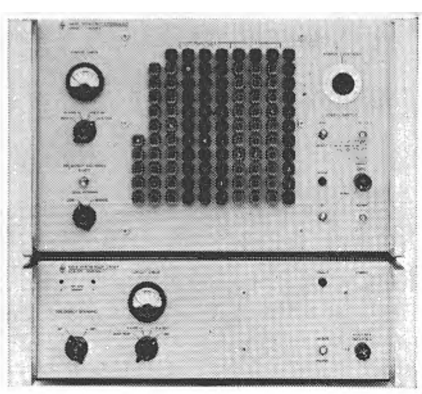
U.H.F. VARIABLE-FREQUENCY SYNTHESIZER

Hewlett-Packard Ltd, 224 Bath Road, Slough, Buckinghamshire

(Illustrated below)

A frequency synthesizer from Hewlett-Packard supplies any frequency from 0·1 to 500MHz and can change frequency typically within 20μs. The frequencies are spectrally pure, crystal-controlled for high stability, and are selectable in steps down to 0·1Hz from either remote electrical signals or the 10-column front panel switchboard.

The specifications for the model 5105A frequency synthesizer include a long-term frequency stability claimed to be better than ±3 parts in 10^9 per day and r.m.s. fractional frequency deviations, for a 10ms averaging time and 30kHz noise bandwidth, of less than



2×10^{-9} at 50MHz and 6×10^{-10} at 500MHz. Output amplitude is 0dBm ±1dB over the entire frequency range, or it may be varied. A continuously variable frequency control (manual or remote) may be submitted for any keyboard column except the highest two. The instrument also has an input that gives phase modulation of the output.

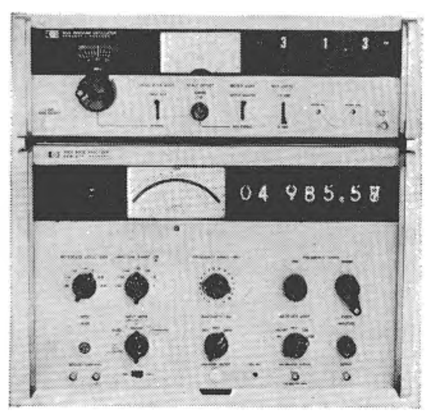
For more information circle **No. 292**

HIGH-FREQUENCY WAVE ANALYZER

Hewlett-Packard Ltd, 224 Bath Road, Slough, Buckinghamshire

(Illustrated below)

An analyser, model HP-312A, for the frequency range 10kHz to 18MHz has been introduced by Hewlett-Packard for communications use. It has a 7-digit readout with automatic decimal-point and units indicator. Signals of amplitude down to −110dBm can be measured and the stability is claimed to be 2 parts in 10^6 per week. A drift-tracking a.f.c. mode of operation gives signal locking which remains within a range of ±3kHz; the digital readout then displays drift continuously. Many types of input can be accommodated including terminated, bridging, balanced or unbalanced or via a wideband zero-loss probe. Inputs are displayed on a taut-band meter, calibrated in dBm and voltage. The reference level can be selected in 10dB steps from −40 to +20dBm and the amplitude range can be similarly set in the range 0 to −60dB.



There are three bandwidths for analysis, 200, 1000 and 3000Hz and a notch filter with a bandpass of 4Hz allows the centre frequency to be found and indicated. Calibrated measurements can be made on systems with 50, 60, 75, 124, 135, 150 and 600Ω impedances.

Model HOI-312A is a similar analyzer for telephone transmission systems. The frequency range is continuous up to 22MHz and the input is for 75Ω only.

For automatic tracking of centre frequencies, Hewlett-Packard manufacture a tracking oscillator (model 313A), tuning continuously from 10kHz to 22MHz. The meter reading of the HP-312A analyzer can be expanded by the oscillator to 2dB full-scale; frequency response may be measured with amplitude resolution of nearly 0·02dB and frequency resolution to within ±10Hz. Model 313A may be used separately as a wideband signal generator.

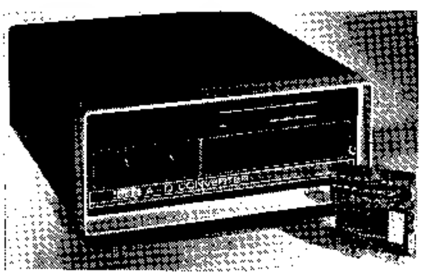
For more information circle **No. 293**

ANALOGUE/DIGITAL AND DIGITAL/ANALOGUE CONVERTORS

Digital Equipment Corporation (U.K.) Ltd, 3 Arkwright Road, Reading, Berkshire

(Illustrated below)

Three convertors have been introduced by Digital Equipment for use as separate units or in conjunction with the company's computers. Type ADC-I is a general purpose analogue-to-digital unit with a variable word length of 12 bits maximum. A rotary switch selects word length, conversion accuracy and conversion time; a pushbutton on the control panel initiates the conversion. A multiplexer control unit (type AMX-I) may be used in conjunction with the ADC-I, the whole forming a combined convertor-multiplexer type CMX-I. A 10-bit integrated circuit analogue/digital convertor (A801) is combined on a flip-chip module; several of these may be stacked in parallel to increase the system bandwidth. The third convertor, type A-611 is a 10-bit integrated circuit



digital/analogue convertor and has two buffer registers, level convertors, a precision divider network and a current summer to drive large loads. Double-buffering and bipolar output can be provided if required. The convertor output can be offset 0 to −10V by a coarse and fine control.

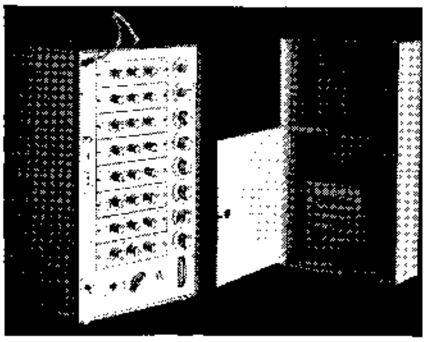
For more information circle **No. 294**

PORTABLE ELECTRO-ENCEPHALOGRAPH

Distributed by: T.E.M. Sales Ltd, Gatwick Road, Crawley, Sussex

(Illustrated below)

An eight-channel portable e.e.g. machine is manufactured by Van Gogh in the Netherlands and marketed in the U.K. by T.E.M. Sales. The Ahrend-van Gogh type-EP machine is fully transistorized and the channel preamplifiers



and most of the printed circuits are of the plug-in type. The instrument is in two sections—the recorder and the amplifier unit. Maximum sensitivity is better than 10μV/cm pen deflexion, and noise and drift together are less than 1μV. Time constants are adjustable in five steps from 0·035 to 1s; optionally they may be 3s or 6s. Gain is adjustable in each channel and there are 9 steps of $\sqrt{2}$ per step. There is a 26-position electrode selection switch. Paper speeds extend in eight ranges from 0·25 to 6cm/s and the marker pen has a time base of 1s. Transient overloads causing circuit blocking are reduced by a special circuit; for example, an overload of 1000 times gives a blocking time of less than 2s.

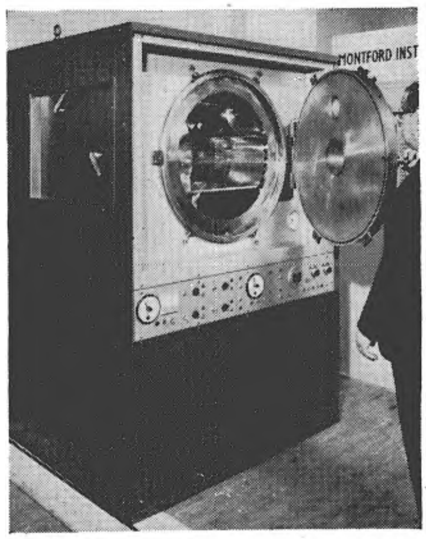
For more information circle **No. 295**

ENVIRONMENTAL TEST CHAMBER

Montford Instruments Ltd, 79 Lots Road, Chelsea, London, S.W.10

(Illustrated above, right)

Designed to simulate high-altitude conditions, the space chamber THV 1085 manufactured by Montford Instruments, produces test conditions for humidity, temperature, pressure and vibration simultaneously. It has an internal diameter of 24in and a depth of the same amount; cooling is with liquid nitrogen or liquid carbon dioxide. The temperature range attainable is from −65°C to +150°C, with a maximum rate of change of 20degC/min. Temperature sensitivity is ±0·degC. Relative humidity can be controlled in the range 20 to 100 per cent, with a sensitivity of ±2 per cent r.h. The possible degree of



vacuum extends from 760 to 5 torr and the maximum evacuation rate is 10 000ft/min. The chamber may be optionally fitted with an automatic program controller.

For more information circle **No. 296**

ENCAPSULATED SILICON DIODES

International Rectifier Company (Great Britain), Hurst Green, Oxted, Surrey

Voltage ratings from 100 to 1 200 p.t.r.v. are obtainable in a range of 600mA subminiature encapsulated silicon diodes introduced by International Rectifier. The range, 5DO5-5DI0, with a top frequency of 50kHz, has been designed for private circuit use and particularly for radio and television power supplies. The surge-current rating of the diodes is 40A for 10ms. Maximum reverse currents at 100°C ambient and at rated load conditions are 200μA f.c.a. for 5D1 reducing to 50μA for 5D10. At 3A d.c. and 25°C ambient the maximum forward voltage drop of all types is 1·2V. Operating or storage temperature range is −40°C to +150°C.

For more information circle **No. 297**

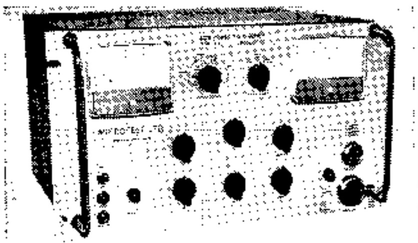
LOW POWER T.W.T. SUPPLY

Microtest Ltd, 9 Old Bridge Street, Kingston-upon-Thames, Surrey

(Illustrated on page 467)

The type-615 power supply is intended as a bench supply capable of energizing a wide variety of low-power, low-noise, travelling-wave tubes.

The power supply has a continuously variable, well regulated 600 to 1000V collector (beam) voltage, a 300 to 800V variable helix voltage. 150, zero to −650V variable grid—4 supply voltage, a 50 to 300V variable grid-3 voltage, a 20V to −50V variable grid-2 voltage, a zero to −20V variable grid-1



voltage and solenoid current supply 10A max. at 12 to 26V.

Interlocking ensures correct switch-on sequence and prevents accidental application of voltages to the collector and other electrodes unless the solenoid current is passing.

All voltages may be monitored on one of the internal 4in meters, while the collector, helix and solenoid currents can be read on the second internal meter. The output to the t.w.t. is *via* a mark-IV 18-pin connector, and the connexion to the solenoid is *via* a mark-IV 4-pin socket.

For more information circle **No. 298**

DIFFERENTIAL DISCRIMINATOR

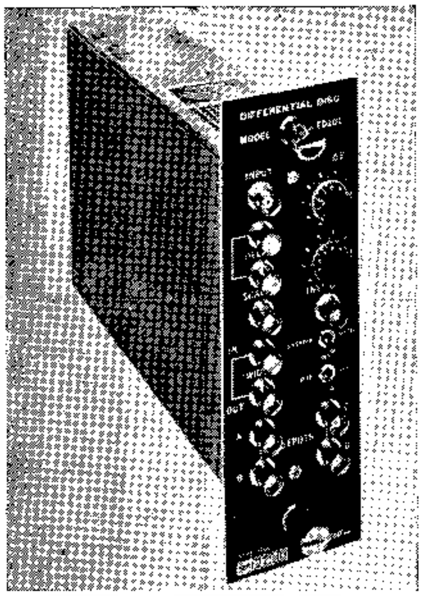
E.G. & G. Nuclear Ltd, Nuclear Measurements, 67 Dudley Street, Luton, Bedfordshire

(*Illustrated above, right*)

A direct-coupled, thermally stable differential discriminator, which operates in three difference modes, has been developed by E.G.&G. Inc. and is now available from Nuclear Measurements.

Designated TD101, the module may be used as a differential discriminator or single-channel pulse-height analyzer. It produces an output signal if the input signal lies within a 'window' δE which is placed on a baseline E. It will function as an integral discriminator, producing an output if the input exceeds a threshold E. Finally, the TD101 can operate as a 'fast/slow' discriminator, producing an output if the input exceeds a threshold (E + δE), but with output timing taken from a lower-level point E on the input signal.

The E potentiometer always determines the lower threshold in all three modes while the δE potentiometer determines the upper threshold in the DIFF and LLT modes (δE is not used in the INT mode). The range of E is −100 to −1 000mV and the δE range is 0 to −900mV. (E + δE) may not exceed −1 000mV in any mode.



The TD101, part of the E.G.&G. M100 modular counting system, is direct-coupled and thermally stable so that the output signal width, leading-edge timing and E and δE thresholds are stable as instantaneous burst rate, temperature and line voltage vary. Since it is direct-coupled, the module will discriminate for slow signals. Threshold interactions are minimized, so that window width δE changes very little as the baseline E is varied.

For more information circle **No. 299**

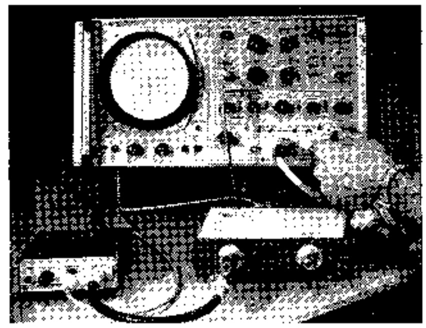
12·4GHz SAMPLING OSCILLOSCOPES

Hewlett-Packard Ltd, 224 Bath Road, Slough, Buckinghamshire

(*Illustrated above, right*)

Sampling oscilloscopes from Hewlett-Packard allow observation of waveforms with bandwidth beyond 12·4GHz, and direct observation is possible for signals through X-band. Sensitivity from d.c. through 12·4GHz is 1mV/cm. Pulse carriers can be viewed directly, without degradation of pulse waveshape. Time-domain reflectrometry is comparably extended; with a Hewlett-Packard 20ps pulse generator, and with the 28ps response of the sampling oscilloscope, a combined measuring-system risetime of less than 40ps is attained. Discontinuities millimetres apart can be resolved.

Any of six possible solid-state microwave oscilloscopes may be formed by choosing one of two main frames—HP model 140A or the variable-persistence/storage model 141A, one vertical amplifier, and one of three microwave samplers. One may choose response beyond 12·4GHz, optimized either for pulse display or for most uniform frequency response and minimum s.w.r. Two difference sweep timebases are offered; one of these makes available delayed sweep for calibrated magnification and identification of waveform



details and for elimination of pulse-train jitter. The other, without delay, has automatic triggering beyond 5GHz.

A two-diode feedthrough sampler is employed to extract waveform information without requiring that the circuit be terminated at the sampler. The input line passes through a modified biconical cavity whose impedance is maintained constant at 50Ω throughout. Within the cavity, a fast-rise pulse from a step-recovery diode turns the sampler on, the switch-off arising from a controlled reflection, within the cavity, of the switch-on pulse. There are two sampler channels.

Any one of the microwave oscilloscopes will consist of an oscilloscope main frame, an external sampler chassis which is brought directly to the circuit under test, a two-channel vertical amplifier, and a sweep timebase. Either the conventional model 140A or the variable-persistence/storage model 141A may be chosen as main frame. There are three samplers: model 1430A is optimized for flat pulse response with 28ps risetime; model 1431A is optimized for flat frequency response and minimum s.w.r.; model 1432A is the 4GHz (90ps risetime) model. The two timebases are model 1425A, with delayed sweep and automatic triggering to 1GHz, and model 1424A, without the delay feature, but with automatic triggering to 5GHz. A single vertical two-channel amplifier (model 1411A) functions with any of the three samplers. It has front-panel recorder outputs to drive x-y or strip-chart recorders, and makes available also an A versus B mode for x-y oscilloscope presentations. A control for risetime adjusts for best trade-off between overshoot, noise and risetime.

The 20ps pulse generator, model 1105A, is a separate accessory which generates a step-function pulse, extending time-domain-reflectometry applications. Accessories include the 1106A tunnel diode mount, which may be triggered to produce a 20ps rise-time step-function. The 1105A pulse generator supply triggers the 1106A through a cable, so the pulse may be generated as near the circuit as may be desired. The 1104A countdown unit functions with the 1106A, and, with inputs up to 18·0GHz, produces a 100MHz, 100mV counted-down signal, suitable for synchronized triggering of the sampling base up to 18·0GHz.

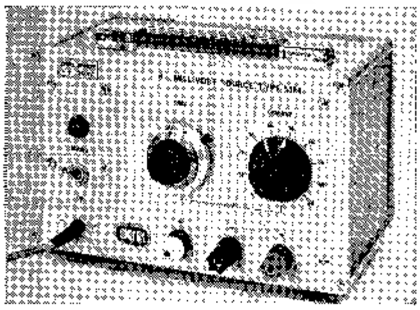
For more information circle **No. 300**

D.C. MILLIVOLT SOURCE

Smiths Industries Ltd, Kelvin House, Wembley Park Drive, Wembley, Middlesex

(Illustrated below)

The Industrial Instrument Division of Smiths Industries have announced a



millivolt source, designed specifically as an accessory for their Servoscribe potentiometric recorder. It may, however, be used separately as a calibrated reference source.

The output is continuously variable from zero to 100mV d.c., the minimum increment being 10μV. The output terminals are isolated from earth and may be used in circuits with a potential difference to earth of up to 500V. The insulation resistance is not less than 250MΩ.

The accuracy claimed, after deduction of the zero error, is $\pm$10μV of dial setting over the range 50mV/degC—100mV. The temperature coefficient is 2μV/degC in 100mV over the temperature range -10°C to 45°C. The instrument is suitable for operation from 110-200/240V, 50Hz supply.

For more information circle **No. 301**

OPERATIONAL AMPLIFIER

Analog Devices Ltd, 38-40 Fife Road, Kingston-upon-Thames, Surrey

A differential operational amplifier, model 180, without chopper stabilizing, has been marketed by Analog Devices. The maximum voltage drift is claimed to be 1·5μV/degC and the differential offset current rate between -25°C and $+85$°C is 50 $\times$ 10^{-12}A/degC maximum. While the differential input impedance is 4MΩ, the common-mode impedance is 500Ω. Open-loop gain is 200 000. The amplifier has a consumption of 5mA at $\pm$15/15V.

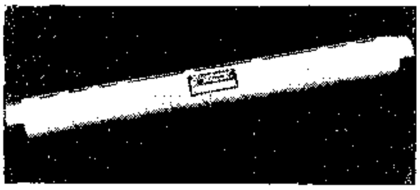
For more information circle **No. 302**

BROADBAND VARIABLE DELAY LINE

Microwave Associates Ltd, Cradock Road, Luton, Bedfordshire

(Illustrated below)

An O-type electron beam variable-delay line has been developed by Microwave Associates. Derived from a travelling-wave tube it has a dynamic bandwidth of approximately 2 octaves,



making possible low pulse distortion. Its design allows rapid variation of delay, handling of multiple signals without crossmodulation and with moderate insertion loss.

Specific variable delays of up to 250ns have been obtained with a minimum residual delay of approximately 30ns. Over the frequency range 1100 to 2300MHz the insertion loss is approximately 40dB at 150ns of delay. Experiments have shown that this approach will yield variable delays of several tenths of a microsecond with wide bandwidths.

The MA2021 delay line has been designed for applications including target simulation in radar testing and phased arrays.

For more information circle **No. 303**

SHORT OSCILLOSCOPE TUBE

Brimar, Thorn-AEI Radio Valves & Tubes Ltd, 7 Soho Square, London, W.1

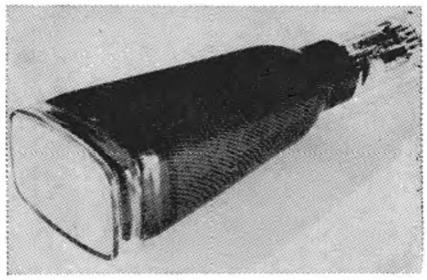
(Illustrated above, right)

The D13-47GH oscilloscope tube manufactured by Brimar has been designed to meet the requirement of a 5in-diagonal rectangular-face oscilloscope tube of short overall length, capable of being deflected by transistored x and y deflexion circuits. It may be regarded as a development of the earlier D13-33 and D13-46 tubes, retaining the same face dimensions of 122mm $\times$ 86mm and minimum scanned area of 100mm $\times$ 60mm but with an overall length reduced from 386mm to 368mm.

Beam blanking at anode potential removes the disadvantage of having to provide a beam-blanking circuit coupled to the grid and therefore operated at a high mean negative potential.

The tube is designed to operate at fourth anode (i.e. screen) potentials of up to 7kV with a post-deflexion acceleration potential of 4 : 1.

At a fourth-anode (screen) operating voltage of 4kV with respect to cathode, the y-deflexion sensitivity is typically 7·5V/cm and at 6kV this is reduced to



about 11·5V/cm. The deflexion assembly incorporates a specially designed x-deflexion plate system, permitting a reduction in deflexion defocusing.

The beam-blanking arrangement used in D13-47GH is similar to that featured in previous Brimar tubes. The first anode consists effectively of two apertured plates connected at the same potential. Between these two plates is interposed a third apertured plate connected to the beam-blanking pin P5. The plates at first-anode potential and the beam-blanking plate carry short deflexion electrodes parrallel to the tube axis. These are disposed so that, when the beam blanking electrode G_2 is taken to a potential different from that of the first anode, the beam is deflected sequentially, first away from and then towards the tube axis.

The overall effect is almost that of a shift of the beam sideways parallel to itself. The aperture in the last plate of the first-anode assembly being small, it requires only a small shift for the beam to be entirely intercepted by this plate, so extinguishing the image on the screen.

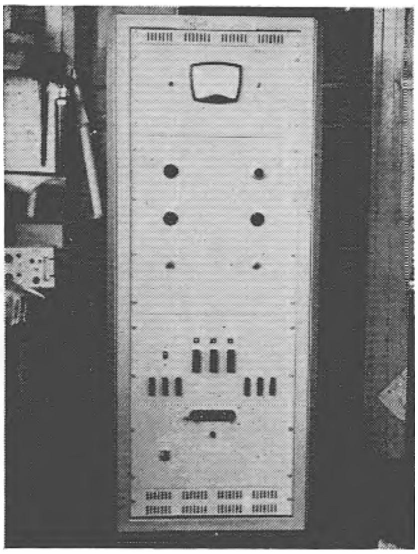
For more information circle **No. 304**

HIGH-CURRENT D.C. STABILIZED POWER SUPPLIES

K.S.M. Electronics Ltd, 139-149 Fonthill Road, Finsbury Park, London, N.4

(Illustrated on page 469)

K.S.M. Electronics have now added to their range of d.c. stabilized power supplies two models of constant-current stabilized power supplies. The first, SC1 200, has a constant-current output in the range 0-200A and the output-voltage range is 30V. The second, type SC1 500, gives output current stabilized in the range 0-500A with a voltage range of 30V. The current in both these models is stabilized to within 0·02% and the ripple current is less than 0·2%. Both units use silicon semiconductors and the current control is continuously variable by a 10-turn helical potentiometer. One application of these units is for a highly stable constant current in a superconducting magnet



coil. Under these conditions the resistance of the coil is for all practical purposes zero, and the units supply current, of possibly 500A, into a short circuit. The units are capable of maintaining a constant current under these conditions. Other units with higher power are in the course of preparation.

For more information circle **No. 305**

SEMICONDUCTOR GO/NO-GO TEST SYSTEM

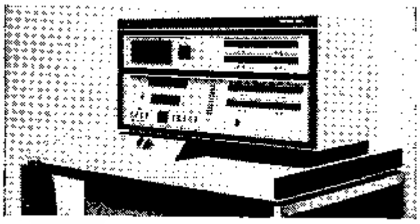
Fairchild Instrumentation Ltd. Grove House, 551 London Road, Isleworth, Middlesex

(Illustrated below)

Fairchild Instrumentation has announced a semiconductor-test system which performs high-speed, digital readout and go/no-go decisions on transistors, diodes and s.c.r.s.

The basic series-600 semiconductor-test system tests leakage, B_V, L_V, h_{FE}, V_{SAT}, and $V_{BE\ ON}$ on transistors and reverse leakage I_{GF}, V_{GF}, V_{BO}, V_{RDO} and V_P on s.c.r.s.

Digital readout and go/no-go decisions are made at the rate of 100 tests per second. Ranging and placement of the decimal point is automatic.



The system is programmed by five switch registers on the front panel, one for each test. It may be expanded to 21 tests in increments of four switch registers. Sorting capability is increased with the addition of an optional classifier.

For more information circle **No. 306**

V.H.F. TUNER DIODES

S.T.C. Semiconductors Ltd. Footscray, Sidcup, Kent

(Illustrated below)

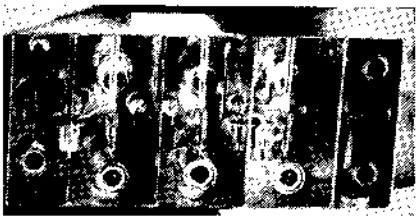
S.T.C. Semiconductors has announced two new variable-capacitance diodes, the BA 141 and BA 142, for use at frequencies up to 1000MHz.

Mainly intended for t.v./u.h.f. and v.h.f./f.m. tuners, they are Jedec DO-7 devices which feature a capacitance range of 4·5 :1.

The BA 141 is suitable for continuous tuning over each of the television bands I, II and III, and over bands IV and V combined; the BA 142 is suitable for bands I, II and III only. Tracking of both diodes is claimed to be better than 3%; Q varies from 300 at 47MHz to 75 (minimum) at 800MHz.

The diodes allow television receivers to have electronic fine-tuning, so that, when a given channel is selected, the set automatically tunes itself to the channel frequencies.

Both devices are suitable for a.f.c. and remote tuning circuits in v.h.f. communications equipment, and can also be employed in frequency-modulated systems.



For television use, the BA 141 is supplied in matched sets of four, and the BA 142 in matched sets of three.

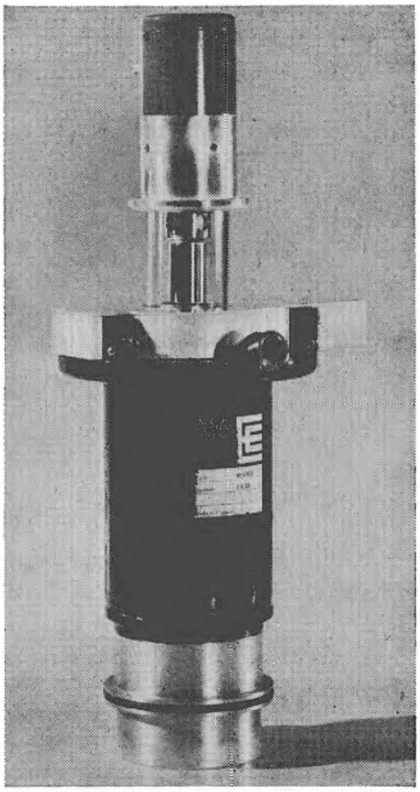
For more information circle **No. 307**

MAGNETRON FOR LINEAR ACCELERATORS

The English Electric Valve Co. Ltd. Chelmsford, Essex

(Illustrated above, right)

English Electric Valve announces a magnetron, type-M5028, designed for linear accelerators. It is a 5MW, tunable, S-band, long-anode valve and has a number of special features: integral water jacket to ensure low-frequency drift; low-torque tuning mechanism, incorporating a balanced bellows system, which allows automatic frequency control with low servo effort, and sturdy construction, to minimize both frequency and power variations when the magnetron is orbited about a horizontal axis.



The pulse length is 2μs and the repetition is 200 pulses/s. Typical performance figures are: anode voltage 48kV, peak output power 5MW, mean output power 2kW, frequency tunable from 2851 to 2861MHz.

For more information circle **No. 308**

MINIATURE H.F. OSCILLATOR

The Marconi Co. Ltd. Chelmsford, Essex

(Illustrated below)

The second of the oscillators announced by the Specialized Components Division of Marconi incorporates a microelectronic circuit which acts as a heating element and which is enclosed in a transistor can, together with a temperature-sensing device, and the quartz crystal which controls the frequency.

The complete assembly has been designed to have maximum thermal insulation, to reduce the amount of heating necessary to stabilize the oscillator. The oscillator, type F3181, is designed primarily for use in portable test equipment. A complete unit occupies 2in^3, and weighs less than 1¾oz. Preset frequencies are available in the range 10 to 20MHz. A long-term frequency stability of 1 part in 1 000 000 is claimed over the temperature range 0 to 60°C without the use of conventional ovens. Short-term stability is 1 part in 10^8.

For more information circle **No. 309**

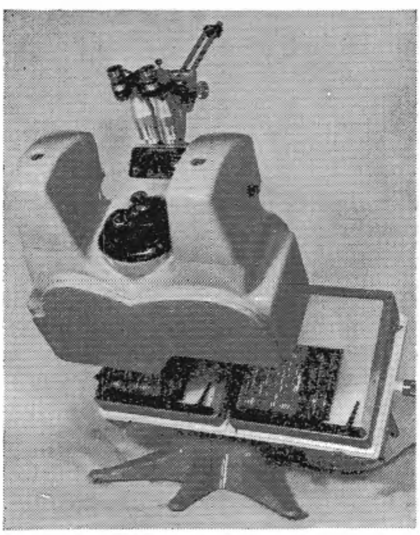
VISUAL INSPECTION COMPARATOR

Vision Engineering Ltd, Send Road, Send, Woking, Surrey

(Illustrated below)

The Vision Comparascope introduced by Vision Engineering has been developed primarily as a rapid method of inspecting assembled printed circuit boards for missing, reversed, misplaced or incorrectly assembled components. In addition, the instrument can be used for inspecting a wide range of items, which includes the checking of components where it is required to establish quickly that all of a large number of holes have been correctly drilled.

Images are observed by the operator, from both the master component and the work to be inspected, through a twin optical path. These images pass through polarizing filters which are placed at right angles to each other. They are then passed through a rotating analyzing disk, the speed of rotation of which is controlled by the operator. An image is received by the operator first of the master component and then of the workpiece in repeating sequence; any difference between the two com-



ponents can be seen without the need of successive eye transfer from master to component.

Illumination is provided by fan-cooled quartz iodine lamps giving approximately 1000h life. The light is collimated through condensing lenses and placed in such a manner as to provide shadow-free illumination. Each lamp unit is fitted with an independent rheostat, so that the lights may be balanced to suit individual operator requirements.

A binocular attachment is available to provide magnifications of the order of 5 : 1. This may be used for examination of miniature and microminiature components. In addition it may also be employed to check badly marked components. This attachment is mounted above the main control panel on sliders and may be brought rapidly into use when required.

The image sequence rate is variable between stop and persistence-of-image speed. A manual override is fitted for master-stop/component-stop. The field size at any table position is 6in × 6in.

For more information circle **No. 310**

CLOSED CIRCUIT TELEVISION CAMERA

Associated Electrical Industries Ltd, New Parks, Leicester

AEI Electronics now markets a self-contained, low-cost, silicon-transistor closed-circuit television camera. It is capable of reliable operation at temperatures from −35°C to +55°C.

Any one of four line standards—525, 625, 735 or 875—can be selected by changing one of the three basic modules. High resolution is provided by a 10MHz bandwidth.

A built-in automatic sensitivity circuit handles light variations in the ratio 10 000:1. A stabilized focus current and aperture and phase correction circuits are also included. Optional features for the FA 42 camera include remote control of lenses for focus, iris and zoom; automatic iris control (for operation beyond the normal wide sensitivity range): remote control of power from built-in power supplies in the camera; remote stand-by switching and an input for a synchronizing pulse to suit studio operations.

A radio-frequency output, 500mV in band I (into 75Ω) or 250mV in band III (into 75Ω), is also available.

For more information circle **No. 311**

REVERSIBLE SCALER

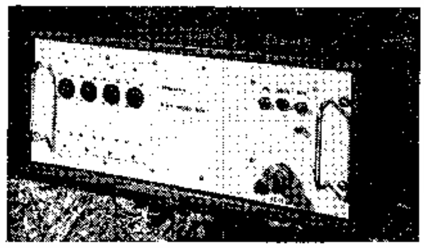
R.G.F. Electronics Ltd, 131 High Street, Teddington, Middlesex

(Illustrated above, right)

A reversible scale by R.G.F. Electronics comprises a 120kHz four-decade scaling circuit with number tube display and operates from a pair of phased square-wave inputs. A 10-line decimal output is available at the rear of the unit for print-out and control purposes. A 4-line b.c.d. output is available as an alternative.

The input facilities are a count input with positive-going signal counting +1 and a negative-going signal counting −1 and a gate input where negative gives no count and positive gives count as with the count input.

In both cases a positive signal is defined as greater than +1V into 10kΩ and a negative signal as less than +3V



into 10kΩ. The gate input must be set at least 1·5μs before a count pulse is received. There is no limit to the rise-time of either input signal. In use, the gate signal is so phased with the count signal that, for a positive count, negative-going count signals are suppressed and *vice versa*.

This input arrangement allows any form of d.c.-coupled sensing device which will give a phased pulse input to be used for bidirectional counting applications. Photodiodes or mechanical contacts may be connected between the input lines and the +6V supply line.

The unit operates from 100-115 and 200-250V, 50 or 60Hz supply. The total consumption is 24W. A 6V d.c. supply is available at the input socket of the unit for the lamp, when using photocells.

For more information circle **No. 312**

PROGRAMMABLE SEMICONDUCTOR CURVE TRACER

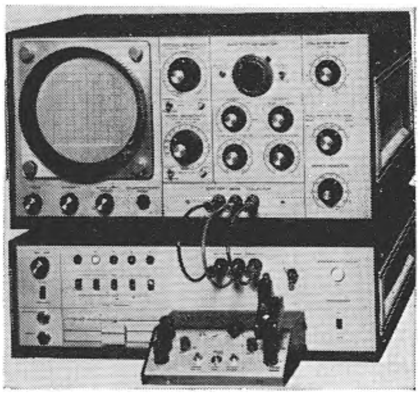
Fairchild Instrumentation Ltd, Grove House, London Road, Isleworth, Middlesex

(Illustrated on page 471)

The model 6200 B/P programmable curve tracer announced by Fairchild Instrumentation performs up to five separate tests automatically on a semiconductor device at a single insertion.

The model 6200B/P and model 3509B programmer combine to make a complete semiautomatic test station, which can provide information on almost every type of two and three terminal semiconductor device. Silicon solid-state circuits are employed.

The 3509B programmer has two plug-in printed-circuit boards containing a diode matrix. The matrix selects all functions needed to establish up to five test parameters for each device.



Combining a programmer with the model 6200B/P allows the unit to complete five individual tests automatically during a period ranging from 0·1s to greater than 3s (20 to 750ms/test). The tests can also be manually stepped if desired. By programming the testing limits, operator error is eliminated when performing single tests repeatedly.

Functions which can be programmed on the model 6200 B/P are: vertical and horizontal sensitivity, collector sweep range and polarity, two values of collector dissipation resistance and all base ranges. Polarity and manual 1-2 or 5 multiplier as well as normal and pulsed-mode operations are also possible. By use of the reed-relay interconnexion matrix, a 3-lead device can be mechanically connected in any possible configuration.

The model 6200B/P includes: 1000V, 100mA collector generator; pulse-mode operation; continuously variable base drive; independent selection of first and last step and base-voltage source to 35V. It also has a base-current source down to 100nA, display inverting switches, detachable test fixture and remote programming.

For more information circle **No. 313**

MINIATURE CERMET POTENTIOMETERS

Bourns (Trimpot) Ltd, Hodford House, 17/27 High Street, Hounslow, Middlesex

Two Trimpots have been added to the series manufactured by Bourns (Trimpot). Model-3282 element has a resistance range from 10Ω to 1MΩ and a power rating of 1W at 70°C. The maximum operating temperature is 175°C. The component is totally enclosed in a moulded plastic case and can be supplied with insulated stranded leads or printed-circuit pins. Model 3052 Trimpot has similar characteristics to those of model 3282 but is proofed against humidity. It can be supplied with solder-lug terminations, printed-circuit pins or insulated stranded leads. Both potentiometers have infinite resolution.

For more information circle **No. 314**

SUBMINIATURE COAXIAL SWITCH

Microwave Associates Ltd, Cradock Road, Luton, Bedfordshire

Microwave Associates are now marketing their subminiature coaxial switch, which operates over a range of frequencies from d.c. to 12·4GHz. The MA7530 has an overall volume of less than 12cm^3 and weighs 35g. It has a 60dB minimum isolation, a v.s.w.r. of 1·3 : 1 and a switching time of 20μs; its r.f. power rating is 15W cw. The switch is suitable for channel selection applications in military and commercial communications and radar systems where size and weight are important factors.

Actuation is remote and connectors are type OSm or equivalent.

For more information circle **No. 315**

TRANSISTORIZED CCTV MONITORS

Associated Electrical Industries Ltd, New Parks, Leicester

Two silicon-transistorized closed-circuit television monitors are now marketed by AEI Electronics for use in studios or as part of industrial and commercial closed-circuit t.v. systems.

Model BG 21 is a high-performance monitor with an 8½in screen; model BG36T has a 14in screen. Both use wide angle (110°) c.r.t.'s. There is a choice of 525, 625, 735 or 875 line operation with a bandwidth from ±1dB at 7MHz and 3dB at 10MHz. The displayed picture geometry is within 1 per cent.

Simple plug-in connexion of external synchronizing pulses is available and other options include provision for rack-mounting, remote control unit for brightness, contrast and power, external light shield and provision for battery operation. There is also a power supply unit available to operate a 60Hz field signal from a 50Hz mains supply.

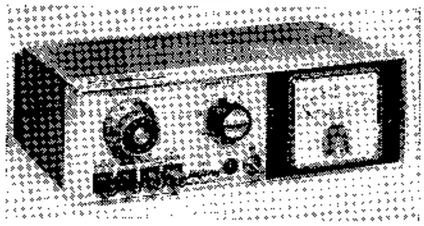
For more information circle **No. 316**

VARIABLE OUTPUT POWER SUPPLY UNITS

Weir Electronics Ltd, Durban Road, Bognor Regis, Sussex

(Illustrated above, right)

The 440/460 series of power packs from Weir Electronics have a single range continuously variable from 0-60V. Ripple and noise are less than 220μV. Four versions are available: standard type-441, 40W output and type-461, 50W, both with a nominal 1 per cent setting accuracy; the calibrated units,



type-442, 40W and type-462, 50W, have a nominal 0·3 per cent accuracy and a digital readout. All units have protective thermal cutouts with automatic reset and their internal impedances are less than 10MΩ at 1kHz. They may be remotely programmed (external resistor) and operated in series or parallel, grounded or ungrounded. Accidental short-circuit loads can be tolerated and an automatic electronic reset operates.

For more information circle **No. 317**

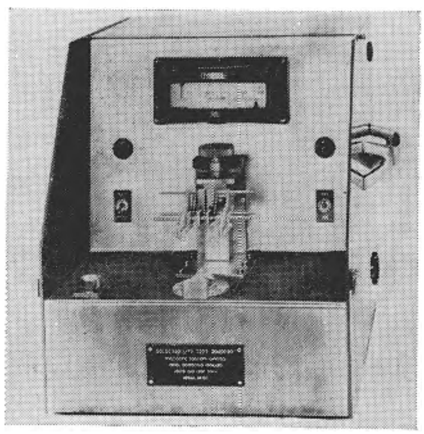
THE NEW MARK II MULTICORE SOLDERABILITY TEST MACHINE

Multicore Solders Ltd, Multicore Works, Maylands Avenue, Hemel Hempstead, Hertfordshire

(Illustrated below)

Multicore Solders have introduced the model-S Multicore Solderability Test Machine for assessing the solderability of component wires.

Operation consists in lowering a specimen of the wire, previously fluxed, horizontally onto a molten globule of solder, which is then cut in half. The time in seconds for the solder to flow round the wire and unit above it is an inverse measure of the solderability of the wire. A fresh pellet of solder is used for each test, the size of the pellet being determined by the diameter of the specimen wire.



The equipment is built strictly to comply with BS 2011, part 2T: 1966. The model-S incorporates several refinements, including semiautomatic timing and a remote-controlled, constant-speed specimen-lowering system. The meter is accurate to 0·25 per cent f.s.d. at test temperature. The model-S operates on 200/240V and model-T on 110/130V.

For more information circle **No. 318**

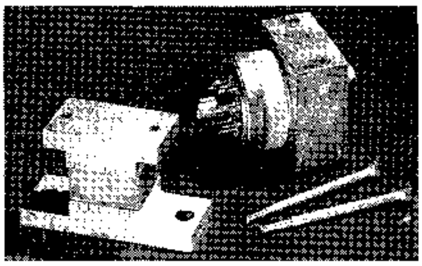
U.H.F. TETRODE

S.T.C. Components Group, Footscray, Sidcup, Kent

(Illustrated below)

A u.h.f. tetrode announced by the Valve Division of Standard Telephones and Cables, designated the 4KC/160M, does not require forced-air cooling and eliminates the need for blowers and motors.

The valve uses a heat sink that can be either water or convection cooled. In applications where the anode must be isolated electrically from earth, the 4KC/160M has an insulating block (HS10A) of beryllia ceramic.



The combination of the 4KC/160M and the HS10A constitutes a low-output-capacitance (8pF) valve assembly suitable for mobile transmitters, as it can be mounted directly onto the transmitter outer case, which acts as a heat sink.

The 4KC/160M is electrically similar to the 4CX250B and is suitable for use up to 500MHz.

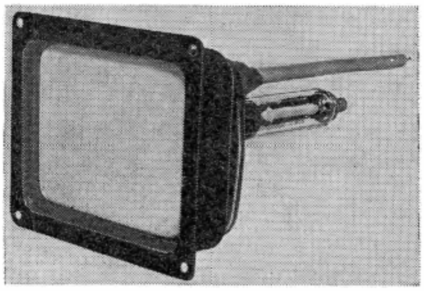
For more information circle **No. 319**

SMALL C.R.T. FOR MULTIPLE DISPLAYS

Brimar, Thorn-AEI Radio Valves & Tubes Ltd, 7 Soho Square, London, W.1

(Illustrated above, right)

The Brimar V3188 c.r.t. has been designed primarily for a small camera viewfinder but can be used for multiple displays. It is a 4½in diagonal rectangular tube with a flat face, giving 98 × 78mm usable screen area. Transistor scanning circuits may be employed, as the neck diameter is 20mm and the deflexion angle 90°. Magnetic deflexion is used and the gun has low-voltage electrostatic focusing. A typical operating voltage is 10kV, with grid voltage for cut-off of the raster at −30 to −50V with respect to the cathode. The heater requires 0·16A at 11V. A flying-lead e.h.t. connector



is integral with the tube. Four V3188 c.r.t.s may be mounted side-by-side in a standard 19in rack.

For more information circle **No. 320**

NEW 7in-SCALE LOW-COST VALVE VOLTMETER

Bach-Simpson Ltd, 19 Nortoft Road, Chalfont St. Peter, Buckinghamshire

Bach-Simpson has introduced their valve voltmeter model 312. The instrument has a 7in scale, with a tracking error of less than 1% and provides accuracies of ±3% on all a.c. and d.c. ranges.

The instrument has an input resistance of 16Ω, and when used with a probe, has a frequency response up to 250MHz. The low-voltage d.c. range permits accurate measurements below 0·5V to be made. Both a.c. and d.c. voltages up to 1500V can be measured; d.c. measurements up to 30 000V can be made by use of an external high-voltage probe.

The model 312 has a meter protection circuit and is built into a rugged phenolic case.

For more information circle **No. 321**

LOGIC CONTROL SYSTEM

The Vosper Thornycroft Group, Industrial & Marine Controls Division, Castle Trading Estate, Portchester, Fareham, Hampshire

The Industrial and Marine Control Division of the Vosper Thornycroft Group has introduced a complete semiconductor control system for digital sequence control. As well as sequence control the system can be used for position control, or embrace a counting facility. These three functions can also be in combination, if required.

Basis of the system is a selection of plug-in logic and control modules containing semiconductor devices, and in some cases reed relays, housed in the standard Vosper cans, and encapsulated. The cans are colour-coded according to function, and in many cases contained a number of logic circuits. The basic module functions are OR (dark blue can), AND (light green can), POWER GATE (dark green can), MEMORY (black can), REED RELAY (red can), TRIGGER TIMER (gold can), and DIODE MATRIX (pink can). These modules can be built up into systems of any degree of complexity, housed in a cabinet or panel as required.

For more information circle **No. 322**

SUBMINIATURE POTENTIOMETERS

W. Greenwood (London) Ltd, 21 Germain Street, Chesham, Buckinghamshire

A series of wirewound trimming potentiometers manufactured by Contelec of Switzerland is marketed by W. Greenwood (London).

Resistance values are in the range 10Ω to 20kΩ, all with power ratings of 0·5W at 50°C. Nominal percentage resolution varies from 1·89 for 10Ω down to 0·21 at 20kΩ. Operating temperature range of all potentiometers is from −55°C to +150°C and the total temperature coefficient is 70 parts per 10^6 for each deg C. Maximum permissible wiper current is 50mA and the maximum working voltage is 250V a.c. Insulation resistance at 500V d.c. is better than 1000MΩ. The components will withstand a 50*g* shock for 11ms or 20*g* at 10-2 000Hz (MIL-STD 202B). Equivalent resistance is 100Ω maximum.

For more information circle **No. 323**

MICROMINIATURE IN-LINE DIGITAL DISPLAY

Counting Instruments Ltd, Elstree Way, Borehamwood, Hertfordshire

Counting Instruments announce an addition of a microminiature display to their range of in-line digital displays.

The display is a multiple assembly of five microminiature units. Each separate unit is capable of displaying up to 11 different numbers, letters, words, symbols, colours, or any special characters which can be reproduced photographically. It contains an assembly of 11 miniature incandescent lamps at the back, a negative with 11 message displays, a series of lenses, and a front viewing screen. On lighting one or more of the 11 lamps, the corresponding part of the negative is illuminated and focused through the lens system on to the viewing screen.

The displays, as single units or in assemblies, plug into a stainless steel frame mounted behind the equipment panel. This is permanently wired with the connexions for the lamps. The display can be removed from the housing through the front of the panel for easy lamp replacement, or it may be changed for another display.

Lamp voltage is 6V and the rating is 0·75V. The display maximum viewing distance is 20ft and the included viewing angle is 160° in both vertical and horizontal directions.

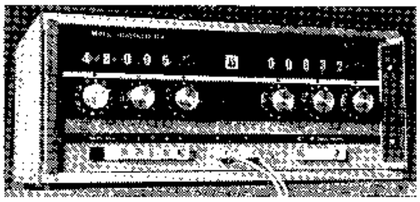
For more information circle **No. 324**

AUTOMATIC BRIDGE

Mark Instruments Ltd, Portsmouth Road, Esher, Surrey

(Illustrated below)

The model K1 manufactured by Mark Instruments is a high-speed, three-



terminal bridge for the measurement of capacitance, conductance, and dissipation factor. It can be used for analogue recording of temperature coefficients in the laboratory or fast component sorting on the production line.

Range: Capitance 0·001pF to 1·1μF; conductance 10pmho to 100mmho; dissipation 0·00001 to 1·00000.

Accuracy: C and G 0·025 %; D 0·25%. The test frequency is 1kHz and d.c. bias can be applied to the unknown component. Both frequency and bias are variable so that semiconductors can be measured. The K1 has a measuring speed of 6 per s. Integrated circuits are used.

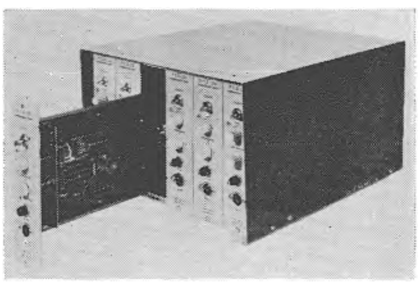
For more information circle **No. 325**

GALVANOMETER INPUT-ATTENUATOR MODULE

P.C.D. Ltd., 219 Sycamore Road, Farnborough Hampshire

(Illustrated below)

The ZAT-6 galvanometer input-attenuator module, available from P.C.D., reduces the sensitivity of galvanometers and recorders by inserting series resistance which may be adjusted in 23 steps from 130MΩ to 30Ω; each step halves the series resistance. At the same time, a suitable shunt resistor is connected across the galvanometer to maintain correct (250Ω) matching and thereby ensure proper damping of the movement. A 25mA fuse provides protection against overload.



In addition, a 6-position (250Ω) T attenuator may be inserted between the matching resistor and the movement to give output levels of 0, 0·2, 0·4, 0·6, 0·8 and 1, thus providing a fine control of the galvanometer sensitivity. When the T attenuator is set to zero, a short circuit is connected across the movement, permitting zero setting without disconnecting the input signal.

Printed-circuit techniques are employed for switches and wiring and the switch contacts and wipers are hard gold plated. The unit is constructed as a 2in wide module which may have alternative panel heights from 7in to 8¾in; it may be supplied unmounted, in a box, or in a nest unit for rack mounting. The switch is protected by a transparent plastic cover.

For more information circle **No. 326**

MARK 4 CRIMP CONNECTOR

The Plessey Company Ltd, Surrey House, Temple Place, London, W.C.2

New internal features are embodied in the Mark-4 Crimp Connector range available from the Plessey Components Group. Designed for crimp termination, the range is an adaptation of the Plessey Mark-4 aluminium and brass solder range, and complete interchangeability can be achieved between both types. The effects of moisture ingress have been eliminated in the crimp range by the introduction of contacts individually insulated along their length, and high insulation figures have been maintained after humidity testing. Tracking paths inherent in the Mark-4 'sandwich' construction have also been considerably extended. Four shell styles are available in three sizes, with contact arrangements for from 2 to 25 poles.

For more information circle **No. 327**

DIGITAL ELECTRONIC TACHOMETER

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

Available with 4 or 5 digits, the 19in rack-mounting versions of the 668 series of Digital Electronic tachometers can be supplied with either an oven-temperature-controlled crystal timebase or a mains-supply-frequency-derived timebase.

The tachometers have four gate times of 1·0, 6·0, 10·0 and 60·0s, enabling direct rev/min indication with 60, 10, 6 or 1 pulse per revolution. Alternative gate times are available to meet special applications and the input range between 100mV and 300V allows use with most types of pick-up transducers.

The tachometer has an in-line Mullard 13mm numerical display, and display time is automatic over a continuously variable range of approximately 0·5 to 10·0s controlled by front-panel push-button and remotely by closing external contacts.

Inputs are a.c.-coupled 100mV to 20V r.m.s.; d.c.-coupled 4V to 300V peak, positive-going from earth; a.c.-coupled input impedance approximately 1kΩ and d.c.-coupled input resistance approximately 50kΩ. The input rates can be at least 10 000 pulses per second, equal to 10 000 rev/min at 60 input pulses per revolution.

For more information circle **No. 328**

TRANSISTORIZED CHARGE AMPLIFIER

Kistler Instrumente AG, Ch 8408 Winterthur, Switzerland

The charge amplifier type 504 by Kistler Instrumente is fully transitorized (with a field-effect transistor at the input). It has 12 measuring ranges with the sequence 1 : 2 : 5, and allows a continuous setting of the calibration factor on a 10-turn potentiometer (to adapt any piezoelectric transducer).

Type 504 is suitable for measuring small charge signals (10× higher sensitivity) of high frequency (nominal up to over 100kHz), when long input cables are required.

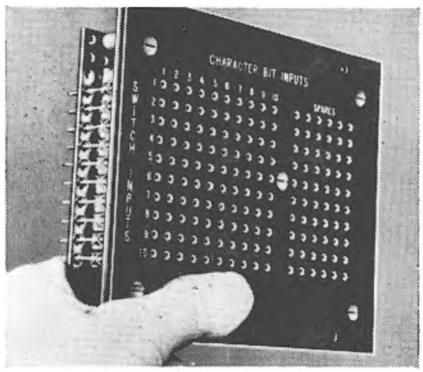
For more information circle **No. 329**

MATRIX BOARD FOR COMPUTER MEMORY

Sealectro Ltd, Farlington, Portsmouth, Hampshire

(Illustrated below)

Sealectro have announced the availability of a 10 × 16 *x-y* matrix Sealectroboard with 10 × 6 spares to provide low-cost preselection of up to 10 addresses for data-block assignment. Each word utilizes up to 10 bits.



The programmer provides non-destructive logic in changeable form. Program pins employed to set up the logic permit simple visual readout. When a matrix board is used, computer logic may be released for other purposes.

The matrix programmer consists of

vertically and horizontally bussed contact strips. Diode pins inserted at the intersections of these strips form the word bits.

Sealectroboard programmers with greater address capacity and larger word ability can be supplied to meet specific requirements.

For more information circle **No. 330**

DIFFERENTIAL OPERATIONAL AMPLIFIER

Fenlow Electronics Ltd, Springfield Lane, Weybridge, Surrey

The Fenlow differential operational amplifier type AD 4000 features very low input offset (total) current and, with open loop gain of 500 000 at d.c. and 10 000 at 10kHz, the amplifier is fully stable as an integrator, without the need for additional stabilizing networks. Suitable for use with line voltages from 10-0-10 to 24-0-24, output swings up to 38V peak-to-peak can be obtained, and a swing of 20V peak-to-peak is maintained to 40kHz on 15-0-15V lines. The output is 5mA and the amplifier is protected against short circuits, overloads and incorrect pin connexions. It is encapsulated in a unit 3½in × 2½in × ½in with pin connexions for installation on 0·2in modular printed board.

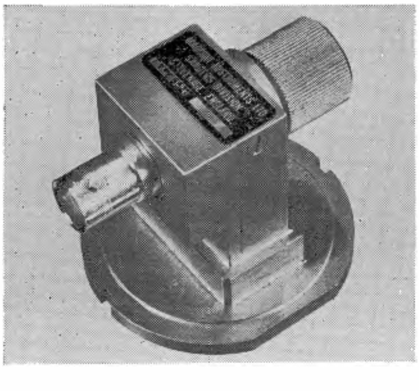
For more information circle **No. 331**

X-BAND MIXER

Marconi Instruments Ltd, Sanders Division, Gunnels Wood Road, Stevenage, Hertfordshire

(*Illustrated below*)

The Sanders Division of Marconi



Instruments announce the latest addition to their microwave components. X-band mixer type 6521 is specifically designed to accept the new Mullard crystal AAY39—a subminiature germanium point-contact mixer diode with a noise figure of 6 dB. For typical diodes, a v.s.w.r. of better than 2·5 : 1 is achieved over the frequency range 9·0-11·8GHz, and the output capacitance is less than 10pF. The output connector is a type BNC and both square- and round-flange versions are available.

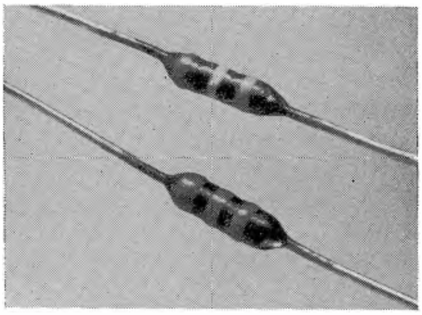
For more information circle **No. 332**

MICROMINIATURE RESISTORS

G. A. Stanley Palmer Ltd, Elmbridge Works, Island Farm Avenue, West Molesey Trading Estate East Molesey, Surrey

(*Illustrated below*)

Microminiature resistors type Rsx 00 are being marketed by G. A. Stanley Palmer in ohmic values up to 4·7MΩ instead of the previous maximum value of 470kΩ. The insulated high-stability film resistor, 2·5mm long and 0·8mm in diameter, is available in I.E.C. preferred ohmic values series E6 (20% tolerance range) and in I.E.C. series E12 (10% range). The Rsx 00 is rated at 0·05W at 40°C and is derated to 0·02W at 70°C, and the noise figure for all values in the range is better than 3μFV/V. The



tinned copper leads are soldered into cavities in the ends of the metallized ceramic rods. This technique ensures low noise and long-term stability.

For more information circle **No. 333**

SILICON INTEGRATED CHOPPER TRANSISTOR

Transitron Electronic Ltd, Gardner Road, Maidenhead, Berkshire

The 3N120 from Transitron Electronic is a dual emitter chopper developed specifically for applications which require the lowest possible offset voltage, e.g. chopper-stabilized amplifiers, multiplexors and other low-level switching circuits. Emitter-emitter channelling is eliminated by channel-stopping guard rings around each emitter in the base region; ion-drift on the emitter oxide surface is avoided by extending the emitter-contact metallization over the oxide and ultrasonic aluminium-wire bonding prevents purple plague at the chip. The offset voltage is typically less than 4μv over a dynamic base-current range of 100μA to 10mA, with a specification limit of 10μV maximum, tracking to 0·2μV/degC from 0 to 100°C. The dynamic resistance is specified as 25Ω maximum, and the emitter-emitter breakdown voltage (BV_{EEO}) as 20V minimum. The device is available in a 4-lead TO-18 package.

For more information circle **No. 334**

ANALOG MULTIPLIER

Intronics Incorporated, 57 Chapel Street, Newton, Mass., U.S.A.

Intronics model M101 4-quadrant, solid-state encapsulated d.c. voltage multiplier performs multiplication, squaring, division and square-rooting without the use of special or external amplifiers. Mode selection is by short-circuiting pins. Specifications include ±10V differential inputs with common-mode capability; 75kΩ minimum input impedance; output ±10W at 5mA maximum, short-circuit protected; output impedance less than 1Ω; full-scale linearity claimed to be better than 0·25%; offset error ±10mV maximum; operating temperature −25°C to +85°C; frequency response in multiplication mode d.c. to 1kHz. Power requirements are ±15V d.c. at 50mA maximum.

For more information circle **No. 335**

Résumés des Principaux Articles

Réseaux entièrement magnétiques comportant des montages nouveaux de circuits de couplage par D. J. Morris

Résumé de l'article aux pages 412 à 418

Les circuits logiques entièrement magnétiques, ne comprenant que des dispositifs à plusieurs ouvertures et leurs fils de connexion, nécessitent habituellement un dispositif de forme spéciale pour chacune des fonctions logiques ou pour une série de fonctions logiques. Cet article décrit la manière dont un seul type de dispositifs peut être utilisé pour exécuter les différentes fonctions logiques, telles que ET, OU et INHIBIT, par une disposition appropriée des cadres de couplage qui relient les dispositifs.

Un diviseur de fréquence à rapport variable à dispositifs micrologiques par S. Jannazzo et G. Rustichelli

Résumé de l'article aux pages 419 à 423

Il s'agit de la réalisation d'un diviseur de fréquence à rapport variable avec l'utilisation de la famille micrologique $RT\mu L$. Le circuit permet l'exploitation complète de l'opération rapide de cette famille et de l'addition de plusieurs des rapports de division.

Un circuit à impulsions pour retards prolongés par B. Jovicic

Résumé de l'article aux pages 424 à 425

Le retard dans ce circuit ne dépend pas de la tension d'alimentation. Le circuit de la base de temps est, à l'exception de la source de puissance, séparé électriquement du reste du circuit. On peut obtenir avec une capacitance de minutage des retards allant jusqu'à 100μF. On peut varier le retard d'un élément de façon continue et sans changer les éléments principaux de la base de temps en l'espace de trois minutes. Les erreurs de retard dues aux variations de température peuvent être compensées. Des circuits à impulsions de base tels que les multivibrateurs monostables et astables, les relais à temps de retard, etc peuvent être réalisés facilement au moyen de ce circuit.

La régulation automatique du niveau de ligne dans les liaisons à grande distance par J. Korn

Résumé de l'article aux pages 429 à 435

La régulation de ligne dans les systèmes porteurs compense les pertes du câble résultant de la température. Deux méthodes sont utilisées pour réaliser cette régulation, soit un système de commande à réaction avec tension-pilote de référence et une commande à circuit ouvert pour laquelle la température du câble sert de référence. Un régulateur de ligne numérique en échelon est utilisé pour la régulation-pilote. Le comportement dynamique d'un grand nombre de ces régulateurs de ligne en andem est excellent. L'amplitude de suroscillation de l'oscillation transitoire après un brusque changement de niveau ne dépasse pas 40 à 50% du changement de niveau initial.

La commande à deux fils utilise le courant de puissance éloigné dépendant de la température qui va vers les répétiteurs souterrains et qui est transmis à travers les conducteurs intérieurs du câble sous forme de courant de commande. L'élément de commande comprend deux thermistors chauffés indirectement en cascade. La commande à deux fils permet de n'employer les régulateurs-pilotes qu'à des intervalles espacés, c'est à dire à chacune des centrales de puissance éloignées.

Analyse des amplificateurs à réaction par G. J. Pridham

Résumé de l'article aux pages 436 à 439

Une méthode d'approche commune pour les systèmes à réaction en parallèle et en série est exposée. La tension et l'impédance sont utilisées pour le système et le courant, l'admittance étant utilisée pour le système en parallèle.

Un intégrateur enregistreur simple à transistor à une jonction par A. R. Atkins

Résumé de l'article aux pages 440 à 441

L'auteur décrit un intégrateur permettant de déterminer le moyen de faire varier les quantités enregistrées. La précision est due aux caractéristiques très stables d'un transistor à une jonction utilisé comme oscillateur de relaxation. L'intégrateur dans ce pas particulier a été conçu pour déterminer la radiation moyenne enregistrée et les pertes de chaleur de convection du corps humain dans une chambre climatique.

Un synthétiseur à fréquence de balayage électronique par E. Kantarizis

Résumé de l'article aux pages 442 à 447

Le synthétiseur dont il est question ici a été réalisé pour produire 666 fréquences cohérentes dans la bande de fréquence de 2 à 12MHz, en plots de 15kHz. Le but envisagé était d'effecteur le balayage électronique de l'ensemble de la gamme de fréquences et, à chaque degré successif de fréquence, d'arrêter le balayage électronique pendant un intervalle de temps déterminé avant de choisir le plot de 15kHz suivant. Les spécifications supplémentaire prévoyaient l'atténuation des fréquences adjacentes sur les deux côtés de toute fréquence choisie par un minimum de 45dB, le maintien de la simplicité du dessin et le prix le plus réduit possible pour l'équipment complet.

Mesures dynamiques de récepteurs à bande étroite par J. F. Golding

Résumé de l'article aux pages 448 à 451

L'accord précis d'un générateur de signaux avec la fréquence centrale d'un récepteur à bande étroite présente souvent un problème de stabilité de fréquence. Pour certaines mesures, dont celle de la sensibilité, le moyen le plus économique de résoudre le problème est d'effecteur le balayage continu du générateur de signaux par la passe-bande du récepteur. Toutefois, ce moyen exige parfois un dispositif de modulation peu habituel, c'est à dire absent dans la plupart des générateurs de signaux en vente courante mais qu'on trouve, cependant, dans les instruments Marconi Type 2002.

Un système de mesure à effet Hall pour grands courants continus et alternatifs par D. J. Lloyd et C. W. Taylor

Résumé de l'article aux pages 452 à 457

Cet article décrit un système qui fournit une indication directe de la valeur efficace et une tension proportionelle à la valeur instantanée de courant de l'ordre de 1 à 1000A dans la gamme de fréquence de 0 à 5kHz. Ce système utilise l'effet Hall dans l'arsenure d'indium pour détecter une densité de flux magnétique produite par le courant devant être mesuré.

Zusammenfassung der wichtigsten Beiträge

Vollmagnetische Schaltungen mit neuer Kopplungsschleifenanordnung von D. J. Morris

Zusammenfassung des Beitrages auf Seite 412-418

Nur aus Viellochkernen und den sie verbindenden Drähten bestehende vollmagnetische logische Schaltungen erforderten normalerweise speziell geformte Kerne für jede logische Funktion oder Gruppe von logischen Funktionen. Der Beitrag beschreibt, wie durch geeignete Anordnung der die Kerne verbindenden Kopplungsschleifen nur ein Kerntyp für die Verrichtung der verschiedenen logischen Funktionen wie beispielsweise UND, ODER und SPERREN benötigt wird.

Regelbarer Frequenzteiler mit Mikrologik-Elementen von S. Jannazzo und G. Rustichelli

Zusammenfassung des Beitrages auf Seite 419-423

Der Entwurf eines regelbaren Frequenzteilers wird eingehend beschrieben und die Anwendung von integrierten Schaltungen der Familie RTμL besonders hervorgehoben. Die Schaltung erlaubt volle Ausnutzung der Geschwindigkeit, die dieser Familie eigen ist, und gestattet einfachen Ausbau der Anzahl der Teilungsverhältnisse.

Eine Impulsschaltung für lange Verzögerungen von B. Jovicic

Zusammenfassung des Beitrages auf Seite 424-425

In dieser Schaltung hängt die Verzögerung nicht von der Speisespannung ab. Die Zeitbasisschaltung ist mit Ausnahme der Stromversorgung elektrisch von der übrigen Schaltung getrennt. Mit einer Verzögerungskapazität können Verzögerungen bis zu 100 s/μF erreicht werden. Die Verzögerung eines Gerätes lässt sich ohne Änderung der Grundelemente der Zeitbasisschaltung innerhalb eines Drei-Minuten-Intervalls kontinuierlich variieren. Durch Temperaturschwankungen hervorgerufene Verzögerungszeitfehler können kompensiert werden. Mit Hilfe dieser Schaltung ist es möglich, Impulsgrundschaltungen wie beispielsweise monostabile oder astabile Multivibratoren, Verzögerungszeitrelais usw. ohne Schwierigkeiten zu entwerfen.

Automatische Regelung des Leitungspegels in langen Trägerverbindungen von J. Korn

Zusammenfassung des Beitrages auf Seite 429-435

Die Leitungsregelung in Trägersystemen kompensiert für die temperaturabhängigen Verluste des Kabels. Sie kann mit zwei Methoden erreicht werden: einer auf eine Pilotspannung bezogenen selbsttätigen Regelung und einer auf die Kabeltemperatur bezogenen Steuerung. Ein stufenweiser Digital-Leitungsregler wird als Pilotregler eingesetzt. Das dynamische Verhalten einer grossen Anzahl dieser Leitungsregler in Reihenschaltung ist ausgezeichnet. Die Übersteuerungsamplitude des Einschwingens nach einer plötzlichen Pegeländerung überschreitet nicht 40 . . . 50 Prozent der ursprünglichen Pegeländerung.

Der durch die Innenleiter des Kabels zugeführte temperaturabhängige Fernspeisestrom für die unterirdischen Verstärker dient der Steuerung als Bezugsquelle. Das Steuerelement besteht aus zwei indirekt beheizten Thermistoren in Kaskade. Diese Steuerung benötigt Pilotregler nur in langen Abständen, zum Beispiel in jeder Fernstromversorgung.

Analyse eines rückgekoppelten Verstärkers von G. J. Pridham

Zusammenfassung des Beitrages auf Seite 436-439

Es wird eine gemeinsame Methode für Serien- und Parallelrückkopplungssysteme erläutert. Für den Fall der Serienrückkopplung werden Spannung und Impedanz, für Parallelrückkopplung Strom und Admittanz eingesetzt.

Ein einfacher Integrator mit Unijunctiontransistor für Registriergeräte von A. R. Atkins

Zusammenfassung des Beitrages auf Seite 440-441

Beschrieben wird ein Integrator für die Bestimmung des Mittelwertes schwankender registrierter Mengen. Die Genauigkeit beruht auf den sehr konstanten Eigenschaften eines Unijunctiontransistors als Kippgenerator. Der Integrator wurde zur Bestimmung des Mittelwertes der registrierten Wärmeverluste eines menschlichen Körpers durch Strahlung und Konvektion in einer Klimakammer entwickelt.

Ein elektronisch gewobbelter Frequenzgenerator mit Frequenzsynthese von E. Kantarizis

Zusammenfassung des Beitrages auf Seite 442-447

Ein Generator mit Frequenzsynthese wird beschrieben, der im Frequenzband 2 . . . 12 MHz 666 kohärente Frequenzen in 15-kHz-Schritten erzeugt. Aufgabenziel war elektronisches Wobbeln des gesamten synthesierten Frequenzbereiches und die Möglichkeit, den elektronischen Durchlauf bei jedem aufeinanderfolgenden Frequenzschritt—und vor Selektion des nächsten Frequenzschrittes—für eine vorgegebene Zeit zu halten. Weitere erwünschte Eigenschaften waren: Nachbarfrequenzen zu beiden Seiten jeder selektierten Frequenz sollten mindestens 45 dB gedämpft sein; der Entwurf sollte unkompliziert bleiben, und das ganze Gerät sollte so wenig wie möglich kosten.

Dynamisches Messen von Schmalbandempfängern von J. F. Golding

Zusammenfassung des Beitrages auf Seite 448-451

Das genaue Abstimmen eines Messenders auf die Mittenfrequenz eines Schmalbandempfängers erweist sich oft als ein Problem der Frequenzkonstanz. Zur wirtschaftlichsten Lösung dieses Problems kann man für bestimmte Messungen einschliesslich der Empfindlichkeitsmessung den Messender durch den Durchlassbereich des Empfängers wobbeln. Dafür benötigt man jedoch etwas ungewöhnliche Modulationseinrichtungen, die es in der Mehrzahl der kommerziell greifbaren Messender nicht gibt, die aber im Modell TF2002 der Marconi Instruments vorgesehen sind.

Ein Hall-Effekt-Messystem für grosse Gleich- und Wechselströme von D. J. Lloyd und C. W. Taylor

Zusammenfassung des Beitrages auf Seite 451-457

Ein beschriebenes System gibt für Ströme in der Grössenordnung von 1 . . . 1000 A im Frequenzbereich 0 . . . 5 kHz (a) Direktanzeige des Effektivwertes und (b) eine dem Momentanwert proportionale Spannung. In diesem System findet der Hall-Effekt im Indiumarsenid Anwendung, um eine durch den zu messenden Strom erzeugte magnetische Flussdichte nachzuweisen.