ELECTRONIC ENGINEERING

VOL. 36 No. 437

JULY 1964

Commentary

THE contribution which Britain is making towards space research is the subject of considerable criticism at the moment. Our failure to compete in space is, according to Sir Bernard Lovell "awful to contemplate".

Speaking at the annual dinner of the Radar and Electronics Association recently, he said that although we had established a pre-eminent place in radio astronomy "our vacillating policy . . . had wrecked the progress of large sections of our industry and had increased the subservience of the country to other nations."

Compared with the stupendous achievements of both America and Russia, particularly in the successful launching of manned satellites, our own efforts are almost negligible.

The fact that other countries in Europe have also found it impossible to sponsor their own individual space research has led them to co-operate and form what is now known as the European Space Research Organization (ESRO) in which twelve European countries, Britain included, have agreed "to provide for, and to promote collaboration among European States in space research and technology exclusively for peaceful purposes".

Already plans involving the expenditure of about £110M spread over the next ten years have been prepared including the development of two small satellites ESRO 1 and 2 within two years, leading to a more ambitious spacecraft for a lunar mission, safe landing of instruments or a lunar satellite, and a "fly by" of a comet or a planet.

Development of EUROPA, a European launching vehicle, too, is well under way. It is to be a three-stage system, the first stage being the British Blue Streak which was successfully fired at Woomera within the last month or so.

The second stage is to be the French Coralie which is due for its first firing at Colomb-Bechar in Algeria towards the end of next year and the final stage is to be constructed in Germany, the first firing of the complete system with an Italian test satellite being planned for 1966.

At first glance all would appear satisfactory, but it seems that there is a lack of complete co-ordination between the European countries and no sense of urgency, which was stressed by Mr. M. N. Golovine, the Vice-president of the European Space Organization. Speaking at a meeting held in London of European delegates connected with space research. Mr. Golovine, who is also President of the British Interplanetary Society and incidentally a director of Hawker Siddeley Aviation, said that unless there was a carefully planned overall programme "there would be a slow, but inescapable down-grading of some of Europe's important industries."

"The progress of decay", he continued, "will begin in the aerospace industry and, eventually through transfer effects, spread to other industries and the danger would be that Western Europe would become a technically underdeveloped area".

Strong criticism indeed, but one hopes that combined with the equally strong criticism from Sir Bernard Lovell it will provide the necessary stimulus to proceed with the European space programme at the greatest possible speed.

What perhaps is worrying some of the European countries is the progress America is making in one particular field of space research, a field which is likely to produce the most handsome dividends in the not too distant future, namely the development of communication satellites, and the fear that space communications will be a complete American monopoly.

It will be recalled that a number of experimental communication satellites such as Telstar, Relay and Syncom have been launched by America and the results obtained from these have shown that communication by satellite on a world wide basis is not only perfectly feasible but a source of considerable revenue as well.

The next important stage is the launching within a month from now of Syncom 3, an attempt to place a satellite at a height of some 22 300 miles in a strictly equatorial orbit, and this is to be followed in the spring of next year by "Early Bird".

"Early Bird", like its successful Syncom predecessors, has been developed by the Hughes Aircraft Company of America.

It will be launched by the American Communications Satellite Corporation (COMSAT) and will be the first regular multi-channel commercial communication satellite in being, and will provide service between Europe and America.

There is no European counterpart to "Early Bird" although the European Launcher Development Organization (ELDO) is considering a European Satellite communication network but the absence of any firm proposal for a communication satellite and for that matter a suitable launching system as well is a matter of concern to which Mr. Golovine again refers.

He fears that the possible ELDO plan "might well suffer a severe set-back if Western European Governments entered into a subservient agreement with the U.S. COMSAT Corporation involving the virtually exclusive use of American operational hardware for years to come."

And that is the position at the moment, the first regular commercial satellite link between Europe and America will be in American hands, but there will be other commercial satellites in the near future and it is highly unlikely that communication satellites will remain an American monopoly for long.

The Shape and Performance of

Multi-Aperture Magnetic Devices

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

Toroidal magnetic cores are being widely used in applications of binary logic. The addition of several apertures arranged in various geometrical configurations results in a range of devices with many significant advantages. These devices are known as transfluxors or multi-aperture devices (m.a.d.). The outstanding features of multi-aperture devices is the provision of isolation of the input circuit from the output circuit, which can give non-destructive read-out and enables the connexions between the devices to be of wire only.

This article describes the performance of the basic multi-aperture device and then shows means of improving its operation by modifying its shape. It then discusses various geometrical configurations for different applications where these devices may be used.

(Voir page 502 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 509)

TOROIDAL cores are being widely used in binary applications. With the addition of several apertures arranged in various geometrical configurations, it is possible to broaden the applications of the core. These devices are known as transfluxors or multi-aperture devices (m.a.d.). As these devices are made of the same ferrite material as toroidal memory cores, they have the same hysteresis characteristics. In addition to these characteristics it is possible to control the flux change round specific apertures. This extra feature of the multi-aperture devices can provide isolation of the input circuit from the output circuits, which provides wide operational tolerances and can permit non-destructive read-out. This isolation of the input from the output circuits also enables the connexions between the devices to be of wire only, making the use of any other components unnecessary.

The application of these devices is still limited, and at this stage they are used in two main fields:

(a) In non-destructive stores.

(b) In all-magnetic logic.

This article discusses the requirements of various shapes of multi-aperture devices for these two applications.

PRINCIPAL SYMBOLS

A	=	Cross-sectional area
$A_{1,2,3,4}$	=	Minimum cross-sectional area of a specific leg
A_L	=	Minimum cross-sectional area of the material round the large aperture
B	=	Flux density
d	=	Diameter of an aperture
F	=	Magnetomotive-force
F_C	=	Threshold m.m.f.
F_S	=	Minimum required m.m.f. to switch all the material
H	=	Magnetic field intensity
H_C	=	Coercive force
h	=	Height of the device
I	=	Current
L	=	Length of a switching path
l	=	Increment of the switching path length
N	=	Number of turns
r	=	Radius of an aperture
t	=	Switching time
Φ	=	Flux

* *English Electric-Leo Computers Ltd.*

The Operation of the Basic Devices

The basic multi-aperture device, which is shown in Fig. 1, contains two apertures of unequal diameter and three distinct legs 1, 2 and 3. An m.m.f. is applied in winding N_1 of the large aperture which produces a field of

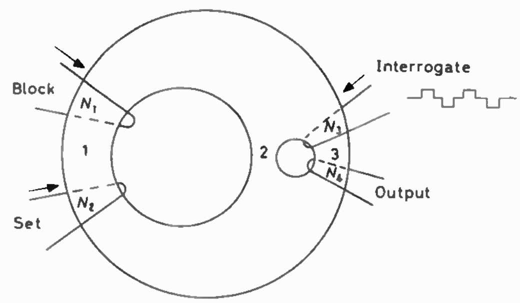


Fig. 1. The basic shaped multi-aperture device

sufficient magnitude to switch all the remanent flux round the device in a cockwise direction and saturate it. The device will remain saturated after the termination of the pulse in winding N_1 and the flux direction will take the pattern as shown in Fig. 2(a). This first field is known as the ' block ' field.

Assume now, that an alternating interrogating m.m.f. is applied to winding N_3 of the minor aperture. This interrogate field, when applied after the block field, will be unable to cause any flux change round the small aperture, since the flux in leg 3 is saturated in an opposite direction to that of leg 2 (in relation to a field round the minor aperture). A positive interrogate field will only tend to drive leg 2 further into saturation and a negative pulse will tend to drive leg 3 further into saturation, thus no change of flux is possible in one of the two legs for each interrogate pulse. Since flux must be along closed loop paths, no flux change may be possible round the minor aperture. The device in this condition is called ' blocked ' as no output voltage is induced in winding N_4 due to the application of the interrogate field.

The flux in a toroidal core, due to an applied field, switches along its shortest available path length, thus it will begin to switch from the inner wall outwards. Consider now the effects of an m.m.f. applied to winding N_2 round the larger aperture, which is in a direction to produce a counter clockwise field in the device. The magnitude of the field produced by this force must be limited so it is sufficient to reverse the flux in leg 2, but insufficient to reverse the flux in leg 3 which will continue to stay saturated in the clockwise direction. At the termination of this second field, known as the set field, the flux pattern in the device will have the form as shown in Fig. 2(b). In this condition of the device, the alternating interrogating pulses which are applied in winding N_3 will be able to produce a flux change round the smaller aperture since the flux round that aperture is now saturated in the same direction, relative to a path round the minor aperture. The first positive interrogate pulse will reverse the flux in both legs 2 and 3, as shown in Fig. 2(c), and the following negative pulse will reverse it back again, and so on indefinitely. Each time the flux is reversed a voltage pulse is induced in winding N_4. The device in this condition is termed 'unblocked' or 'set'.

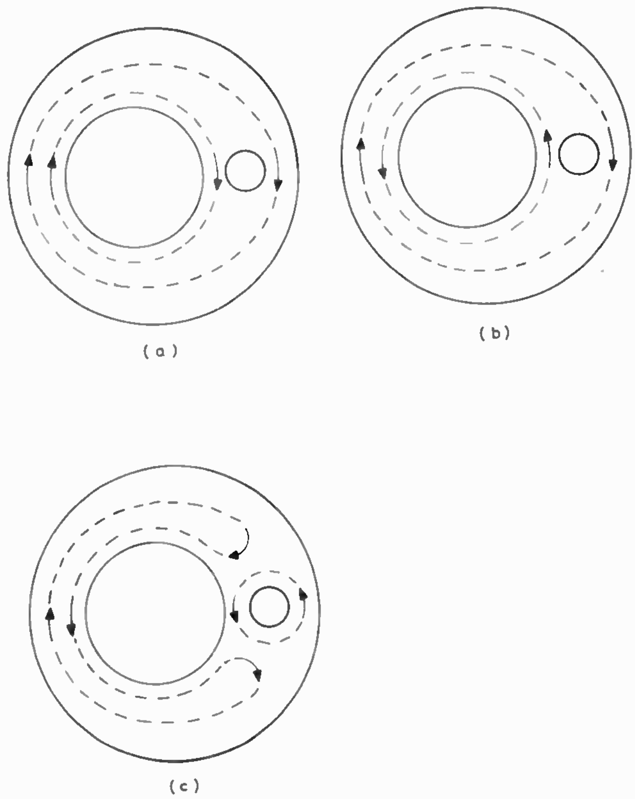


Fig. 2. Flux patterns in the basic multi-aperture device
(a) Blocked state
(b) Set state
(c) Set state, after a positive interrogate pulse

The Read-Out Pulses

By applying the 'block' and 'set' fields round the large aperture, it is possible to control the output pulses in winding N_4 of the minor aperture. The advantage of the multi-aperture device over cores is that information stored in the device is not destroyed on read-out. If binary information '1' is represented by the 'set' state and '0' by the 'block' state, then it is possible to read out the stored information indefinitely.

The output from a multi-aperture device can be obtained in two forms:

(a) Non-destructive read-out, obtained when interrogating the device, which can be produced at any given time and may be repeated indefinitely.

(b) Destructive read-out, where only a single output pulse is obtained when the device is reset and switched to its blocked condition. There are two destructive read-out pulses, one pulse each time the device is switched from one state to another. That is, one pulse when the device is set, and another pulse when the device is blocked, thus these destructive pulses are synchronous with the clock pulses. When transferring the information from one device to another it is usual to take the output at the time the device is reset to its blocked state.

The non-destructive output, due to the interrogate field, indicates a '1' when there are continuous output pulses in winding N_4, and when there is no output pulse, it indicates a '0'. Fig. 3 shows the non-destructive read-out pulses obtained from a multi-aperture device in its 'set' and 'blocked' states. The 'set' and 'block' driving pulses, shown in Fig. 3, are superimposed showing the first pulse setting the device and the second pulse blocking the device. The device is continually interrogated, thus the signal-to-noise ratio of the 'set' and 'blocked' conditions is easily detected.

The destructive read-out pulses are similar to the output pulses from a ferrite core, but, in multi-aperture devices it is possible to load the output winding heavily without affecting or loading the input circuit. This is possible because of the isolation capabilities of the multi-aperture devices which are not found with ferrite cores. The destructive read-out pulse can be taken from the larger aperture or from leg 2, but a low impedance output circuit on

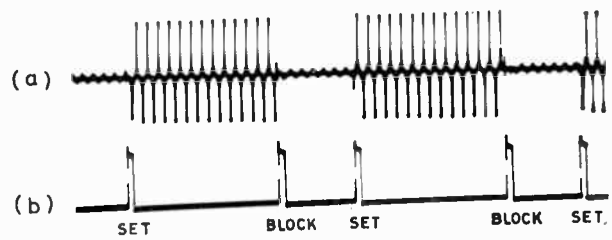


Fig. 3 (a). Non-destructive read-out pulses
(b). 'Set' and 'block' pulses

either leg 2 or on the larger aperture will only tend to load the device. In order to isolate the output circuit from the input circuit, the destructive pulse should be taken from leg 3.

To obtain an output pulse on leg 3 when the device is reset and switched to its blocked state, the flux in leg 3 must be first switched from the downward direction to the upward direction (Fig. 2(b)). In other words, the flux round the minor aperture, after the device has been switched to its set condition but before it is switched back to its blocked condition, must be reversed, thus 'transferring' the upward direction flux from leg 2 to leg 3. This could be done by a single positive interrogate pulse, which induces a field in a direction to switch the flux in a counter clockwise direction round the minor aperture (Fig. 2(c)). This field, which is known as the prime field, could conveniently be supplied by a continuous unidirectional m.m.f., thus reducing the requirement of an extra clock-pulse.

Both the interrogate and prime m.m.f.s must be limited so as to induce a field of sufficient magnitude to switch the flux round the minor aperture but insufficient to switch the flux round the large aperture.

The Switching of Flux

The switching properties of multi-aperture devices are similar to those of toroidal ferrite cores, as they are made of the same ferrite material. The ferrite material is characterized by its approximately rectangular hysteresis loop with two well defined states of magnetization.

The switching of ferrite material is in two forms. The first is the remanent switching (known also as inelastic switching) which is caused by completely reversing the domain orientation through motion of 180°. That is, switching the flux in a counter clockwise direction round the hysteresis loop, from one remanent state to another. The second is caused by the reversible domain rotation, hence when the applied field is removed the domains return to their original direction: this is known as elastic switching. Elastic switching may be caused by two factors, (a) the application of a field strength which is less than the coercive force of the material, $H < H_C$, (b) the application of a field which tends to switch the flux in a clockwise direction in relation to the hysteresis loop. In both these cases the orientation of the domains is not reversed but is merely 'disturbed'.

The output pulse obtained when the material is fully switched from one remanent state to another, is a function of both the elastic and inelastic switching. When an m.m.f. is applied to a ferrite core, which induces a magnetizing field round the core, the direction of the domains which are reversible, will reorientate in the right-hand screw turning in relation to the applied m.m.f. This change in the orientation of the domains will cause a small change of flux which produces a small output voltage spike. As the induced field exceeds the coercive force, $H > H_C$, there will be a complete reversal of the domain orientation which will cause a secondary output pulse of wider duration and larger amplitude than the first output spike. The more the applied field exceeds the coercive force the faster is the domain motion. Thus as the applied field exceeds the coercive force, the elastic effect is reduced.

In multi-aperture devices there are several switching paths whereas in the standard ferrite core there is only one switching path. The selection of a switching path in a multi-aperture device, that is, guiding the switching flux into a desired path, enables the broadening of the application of these devices in comparison to ferrite cores. Furthermore, it can provide isolation of the input from the output circuits. Thus, due to the additional features of the multi-aperture devices, the switching of the flux requires further considerations.

In order to simplify the flux patterns in the multi-aperture devices an arrow model is used, where the arrowhead represents the direction of magnetization and each arrow represents a predetermined amount of flux. The actual flux patterns are considerably more complicated but, nevertheless, the arrow model is convenient and helpful in understanding the switching properties of multi-aperture devices.

For the operation of these devices and the selection of one switching path in preference to other paths, the following simple principles are applied:

(1) Flux continuity must be preserved when no external field is applied, so the flux lines must be closed on themselves.

(2) The switching of flux may be only along closed loops perpendicular to the applied m.m.f.

(3) The switching of flux will seek a path with a minimum energy state, i.e. a path which presents the lowest reluctance.

(4) With some exceptions (e.g. loading on a path), the lowest reluctance path is also the shortest path which has flux available for reversal.

(5) In any closed path, it is impossible to reverse the direction of the flux around the path, if any part of the path is already saturated in the direction of the induced field.

(6) The applied m.m.f. must be larger than the threshold m.m.f. of the material along a selected path. The threshold m.m.f. (F_C) is the line integral of the coercive force of the material (H_C) along the shortest available path, $F_C = \oint H_C dl$. In the toroidal path, shown in Fig. 4, the threshold m.m.f. is $F_C = H_C L_{min}$ where the minimum path length, L_{min}, is the inner circumference of the path, $L_{min} = 2r_1\pi$.

(7) The total cross-sectional area of a switching path is constant and is equal to the minimum cross-sectional area of the material which can be switched.

(8) The amount of flux (Φ) that is reversed along a path is given by $\Phi = 2B A_{min}$, where B is the saturation flux density and A_{min} is the minimum cross-sectional area of the path. $2B$ is taken since the flux traverses round two quadrants of the hysteresis loop.

(9) Further domain orientation takes place round the boundary of the switched path so as to preserve flux continuity outside the switched path. An example of this is given in Fig. 2(c).

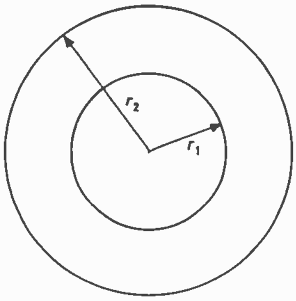


Fig. 4. Simple toroid core

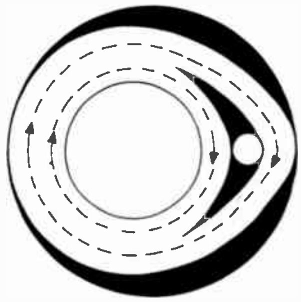


Fig. 5. Shadowed regions

(10) Increases of the applied m.m.f. above the m.m.f. required to switch all the material within the path controlled by the minimum cross-sectional area, will reduce the switching time but will not increase the amount of material which can be switched.

The minimum required m.m.f., F_S, to switch all the flux material will be $F_S = H_C L_{max}$ where L_{max} is the maximum available path length controlled by the boundary of the switched path. In the example shown in Fig. 4 $L_{max} = 2r_2\pi$.

The Shadow Effect

The operation of the multi-aperture device depends greatly on the geometrical structure of the device. Although it is possible to use the device shown in Fig. 1 for all-magnetic logic and for non-destructive stores, the operation will be very marginal, requiring very close operational tolerances and would give poor signal-to-noise ratio of the read-out pulse. Means for improving the structure must be found.

The multi-aperture device described shows that the flux is in either a clockwise direction or a counter clockwise direction round the large aperture. This shows the ideal flux pattern, however, in practice there are many variations from this pattern caused by material imperfections, flux continuity and stress conditions.

The amount of flux switched from one direction to the other is limited by the narrowest cross-sectional area of the material. Fig. 5 shows the flux distribution round the device after the application of the block field. As the cross-sectional area is not constant round the device, there will be some regions which might not follow the required 'block' clockwise flux pattern. These isolated regions are shown in Fig. 5 as shadows.

As the flux distribution must be continuous within the material and not through air-gaps, the flux pattern within the shadowed regions will be continuous. That is, the flux pattern within the shadowed regions will be of local flux closures.

There are various theories of the flux patterns within the shadowed regions and its actual pattern is still an unknown factor. Two possible shadow patterns round the minor aperture, when the device is in its blocked state, are shown in Fig. 6. The regions are divided into zones which include several domains. The direction of the zones shows the average orientation of the domains, thus, the direction of individual domains may deviate slightly from the zone direction.

It can be seen in Fig. 6(a) that the direction of the zones,

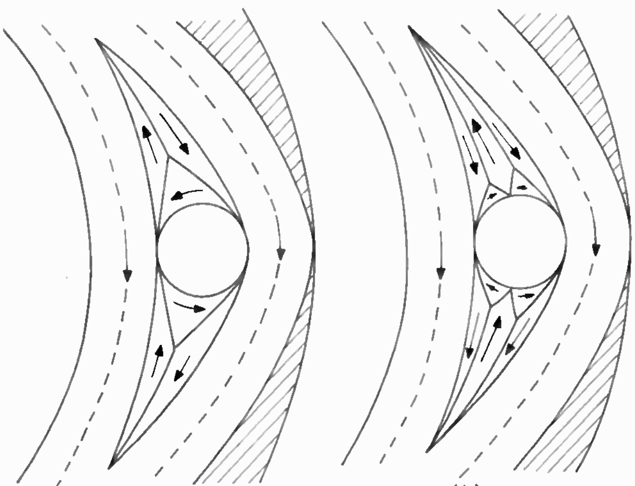


Fig. 6. Alternate shadow patterns round the minor aperture in the 'block' state

along the boundary of the shadow region, is of opposite direction to that of leg 2, while it is in the same direction with leg 3. Thus if the theory shown in Fig. 6(a) is correct, then there could be two possible shadowed patterns depending on whether leg 2 or leg 3 had been previously switched. The shadowed pattern shown in Fig. 6(b) does not change if leg 2 or leg 3 had been previously switched.

Due to the shadowed regions, there will be some flux switching round the minor aperture when the device is interrogated while it is in the blocked condition. Hence, the flux switching between the two shadowed regions round the minor aperture will give a relatively large '0' output pulse, which will reduce the discrimination of the 1/0. These shadowed regions may cause difficulties in logic applications where a high discrimination ratio is required between the blocked ('0') condition and the set ('1') condition.

In the example shown in Figs. 5 and 6, the cross-sectional area of leg 1 is equal to the sum of that of leg 2 and leg 3, $A_1 = A_2 + A_3$. However, some devices which are manufactured have $A_1 \neq A_2 + A_3$ and the shadowed region in these devices would be far more critical and the discrimination ratio worse.

As explained before, the shadowed regions which cause most concern are those round the minor aperture. In principle the shadowed regions could be removed completely, but this would complicate the manufacturing process. To overcome the 'shadow effect', multi-aperture devices can be designed having an ear-shape round the minor aperture, thus ensuring approximately equal cross-sectional area round the device. An example of such a device is shown in Fig. 7. This shape does not completely eliminate the shadow effect but only reduces it.

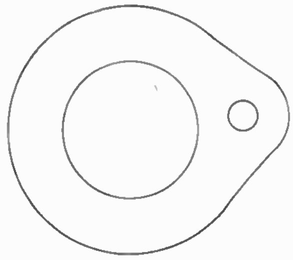


Fig. 7 (left). Ear-shape multi-aperture device

Setting the Multi-Aperture Devices

Another disadvantage of the multi-aperture device which was shown in Fig. 1, is that the 'set' m.m.f. applied to the large aperture must be accurately controlled, since the amount of material switched is directly proportional to the applied set m.m.f. The amount of flux that could be interchanged between leg 2 and leg 3, due to the alternating—interrogate field, will depend on the degree by which the device is set. That is, the amount of flux that can be interchanged will depend on the minimum cross-sectional area of the flux which is in one of the two directions. If leg 2 is not fully switched in the counter-clockwise direction, i.e. the device is under-set as shown in Fig. 8(a), the interrogate field may only switch, round the minor aperture, that amount of flux controlled by the cross-sectional area which is in counter clockwise direction. If the device is over-set, as shown in Fig. 8(b), that is leg 2 and part of leg 3 are in a counter clockwise direction, the amount of flux which would be switched round the minor aperture will be controlled by the minimum cross sectional area of the flux which remained in a clockwise direction. When setting a multi-aperture device it is required that the applied field will switch the leg 2 in a counter clockwise direction while leg 3 remains in the clockwise direction. To avoid this problem, an extra small aperture could be added and the device may then be set on this small aperture. Fig. 9 shows such a device which has one large aperture and two smaller apertures, hence. having four legs.

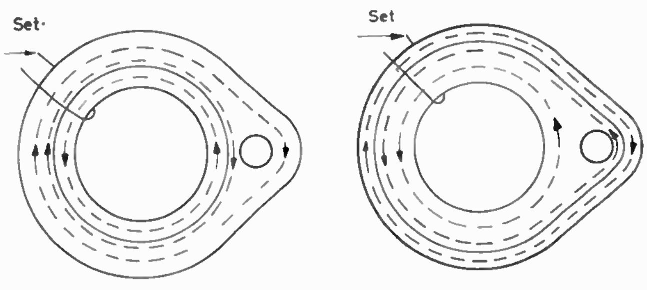


Fig. 8. Flux patterns of the set state

An m.m.f. applied to leg 2 of the new device shown in Fig. 9 can switch only that material controlled by the cross-sectional area of leg 2. Increasing the applied m.m.f. to leg 2 will not affect the amount of material switched, and consequently a close tolerance of the set m.m.f. will not be further required. The only limit is its lower value which must be sufficient to switch all the flux in leg 2. The blocking field applied to this device will still be round the larger aperture.

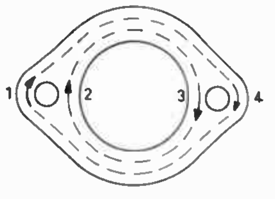


Fig. 9. Ear-shaped three aperture device in its blocked state

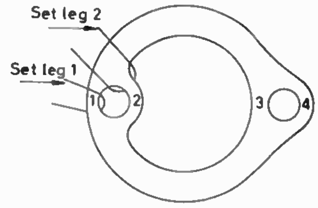


Fig. 10. Shaped input aperture

The setting of this device can be performed by applying the set field either on leg 1 or on leg 2. In both cases the flux switches round the large aperture and reverses the flux in leg 3 from a downward direction to an upward direction. The set m.m.f. applied to leg 1 cannot switch round the input minor aperture, although the path provided by leg 2 is the shortest available path. This is because leg 2 is saturated in the opposite direction to that of leg 1, relative to a path round the input aperture, as shown in Fig. 9. The shortest available switching path for an m.m.f. applied on leg 1 or leg 2, where the device is in its blocked condition, will be round the larger aperture and not round the input minor aperture.

As the read-out pulses are taken from the output aperture, there is no difference whether the device is set on leg 1 or leg 2, since the resultant effect will be the same, i.e. switching the flux in leg 3 in an upward direction. However, the path length along legs 1 and 3 will be slightly longer than that along legs 2 and 3. This will result in more m.m.f. being required to set the device on leg 1 than on leg 2. In some applications, the m.m.f. required to set the device on leg 1 must be equal to that of leg 2. For example, a two input OR gate or a bi-directional shift register, where in both cases the device may be set either on leg 1 or leg 2. To achieve this, the path length round the larger aperture and along leg 1 may be made equal to that along leg 2 by reversing the ear shape of the input aperture as shown in Fig. 10.

The Diameter Ratio of the Apertures

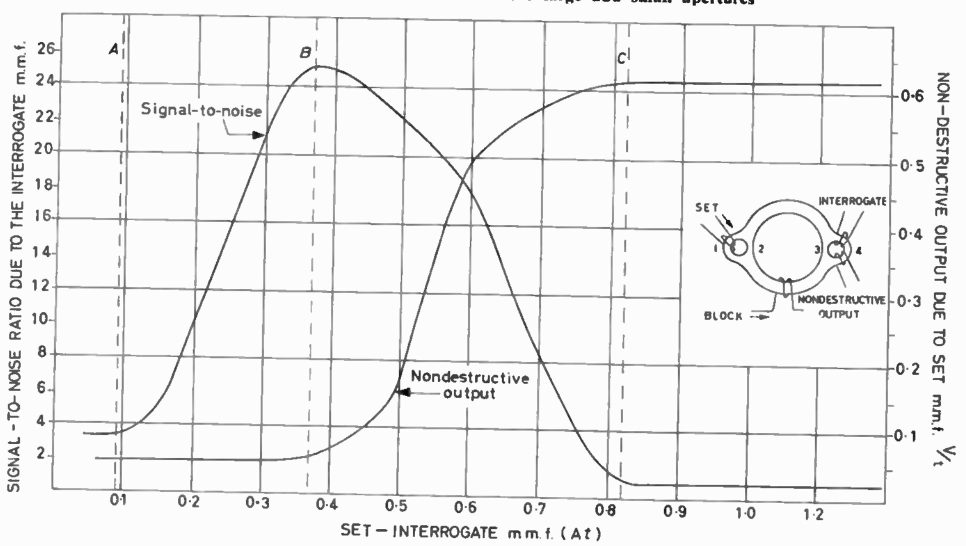
It has been discussed that both the prime and interrogate m.m.f.'s must be limited so as to induce a field of sufficient magnitude to switch the flux round the output minor aperture but insufficient to switch the flux round the large aperture. As the flux switching paths are a direct function of the diameter of the apertures, the range of the prime and interrogate m.m.f.'s will depend on the diameter of the apertures. Hence, for further improvement of the operation of these devices, it is necessary that the ratio of the diameter of the larger and smaller apertures should be as large as possible.

The relation between the switching paths round the larger and smaller apertures can be seen from the curves in Fig. 11. Two separate curves are given in Fig. 11 which represent two different operations. The first curve shows the signal-to-noise ratio of the non-destructive read-out pulses as a function of the interrogate m.m.f. The signal represents the output pulses when the device is fully set and the noise represents the output when the device is fully blocked. The second curve shows the non-destructive read-out pulses from the device as a function of the set m.m.f., for a constant interrogate m.m.f. In both these experiments the read-out pulses were taken from the output aperture.

It is interesting to compare the two separate curves shown in Fig. 11. The signal-to-noise ratio is almost equal to unity if the interrogate m.m.f. is below the threshold m.m.f. required to switch the material round the minor output aperture. Point *A* indicates this threshold m.m.f., $F_{c(small)}$, of the smaller aperture. The signal-to-noise ratio rises with the increase of the interrogate m.m.f. above the magnitude of the threshold m.m.f., till it reaches a peak point, and then decreases till it is equal to unity. In the second curve, the output pulses are small for set m.m.f.'s. below the magnitude of the threshold m.m.f. required for switching the flux round the larger aperture. Point *B* shows the threshold field of the large aperture, $F_{c(large)}$, which also corresponds to the peak point shown in the first curve. This peak point indicates where the interrogate m.m.f. starts to switch the flux round the large aperture and thus where the signal-to-noise ratio starts decreasing. As the set m.m.f. increases, above the threshold m.m.f. round the large aperture, the output will increase till it reaches point *C* and then remains constant for any further increase of the set m.m.f. Point *C* shows the minimum required set m.m.f. to fully set the device, $F_{s(large)}$. This point also corresponds to the point where the signal-to-noise ratio is equal to unity, since here the interrogate m.m.f. fully switches the flux round the large aperture.

By comparing the two curves, shown in Fig. 11, it can be deduced that as the ratio of the diameters of the larger and smaller apertures is increased the range of the interrogate m.m.f. is also increased. The increase of the diameter ratio may be achieved either by decreasing the diameter of the small aperture, which reduces point *A*, or by increasing the diameter of the large aperture, which will increase points *B* and *C*. The selection of the diameter of the apertures is a practical problem since the minor aperture must be large enough to accommodate the windings and the large aperture must not be increased indis-

Fig. 11. The threshold m.m.f.s. of the large and small apertures



criminately, in case the required set m.m.f. becomes too large. A ratio of the diameters of about 5/1 would give reasonably good results.

The Cross-Sectional Area of the Legs

In ferrite cores the signal-to-noise ratio of the output pulses depends on the squareness of the hysteresis loop, where the signal is the output pulse due to the remanent switching and the noise output is due to the elastic switching. In multi-aperture devices the signal-to-noise ratio depends not only on the squareness of the hysteresis loop but also depends on how well the material is blocked and on the degree of setting.

As mentioned before, the amount of flux switched in the device in one direction or the other, is limited by the minimum cross-sectional area of the switched path. When the devices are manufactured it may be difficult to maintain very close tolerances of the leg cross-sectional area. If the sum of the cross-sectional areas of legs 1 and 2

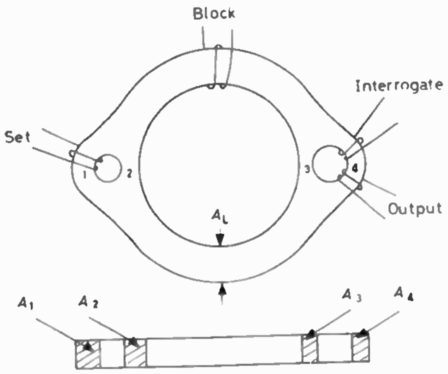


Fig. 12. Apertures and legs dimensions

$A_1 + A_2 > A_3 + A_4 \qquad A_L > A_3 + A_4$

(the input aperture) is smaller than the sum of the cross-sectional areas of legs 3 and 4 (the output aperture), then it will be impossible to fully block the output minor aperture. This will cause some shadowed regions round the output apertures and will reduce the signal-to-noise ratio of the output pulses. The cross-sectional area of the shadowed region will be $(A_3 + A_4) - (A_1 + A_2)$. The shadowed region may also be caused if the cross-sectional area of the material round the larger aperture, A_L, is smaller than the sum of the cross-sectional areas of leg 3 and leg 4. To obtain a good signal-to-noise ratio at the output aperture, the output aperture must be fully blocked with no shadowed regions. Thus the requirements of the multi-aperture device are as follows:

$$(A_1 + A_2) - (A_3 + A_4) > 0$$

$$(A_3 + A_4) < A_L$$

These requirements will cause the device to have the diameter of the input aperture smaller than that of the output aperture, as shown in Fig. 12.

The maximum signal output due to a positive interrogate field is obtained when the flux in leg 3 is in a counter clockwise direction and the flux in leg 4 in a clockwise direction. Thus it is required that the cross-section of leg 1 (where the set field is applied) will be equal or slightly larger than the cross-sectional area of leg 3, so as to fully switch leg 3 when the set field is applied.

$$A_1 \geqslant A_3$$

The amount of flux switched will depend on the minimum cross-sectional area of leg. 1. Hence a further requirement must be fulfilled so as not to over-set the device. That is, the cross-sectional areas of legs 3 and 4 must be much larger than that of leg. 1.

$$A_1 \ll A_3 + A_4$$

The Input Aperture

In most applications the input-set m.m.f. is applied to leg 1. However, in applications where the device may be set on leg 1 or leg 2, it is essential that the cross-sectional areas of these two legs should be equal to each other.

$$A_1 = A_2$$

This equation does not apply to all other applications, especially in the all-magnetic logical circuits where the connexion between the devices is of wire only. In these circuits any spurious flux changes in the previous device may cause the following device to switch into its set condition. This effect is known as zero built up into ones, where a zero (noise) output pulse from one device is large enough to set or partially set the following device.

In order to avoid the zero build up effect an alternative switching path must be added to the device into which the noise input pulse may be guided. If the device is set on leg 1 it is possible to artificially create this extra path in leg 2 by having the cross-sectional area of leg 2 larger than that of leg 1.

$$A_1 < A_2$$

$$A_S = A_2 - A_1$$

This will cause a shadow region in leg 2 having a cross-sectional area of A_S. Any input pulse on leg 1 will first switch into the shadow region of leg 2 and only after all the flux in leg 2 is in the upward direction will the flux switch round the larger aperture. This will artificially increase the threshold m.m.f. required to switch the device.

When the device is in its blocked state the flux round the larger aperture must be in a clockwise direction and the local flux closures must be limited to leg 2. Hence, the cross-sectional area of the material round the larger aperture must be smaller than the sum of the cross-sectional areas of legs 1 and 2.

$$A_L < A_1 + A_2$$

The shadow region in leg 2 must be limited so as to ensure that only a small portion of the flux available in leg 1 will switch into leg 2, while the remainder of the flux in leg 1 will switch round the larger aperture.

$$A_1 \gg A_S$$

Furthermore, the requirement stated in the previous section, that leg 3 must be fully switched by an input pulse must still be fulfilled. Thus the previous equation given as $A_1 \geqslant A_3$ will have to be modified to a new form.

$$A_1 - A_S \geqslant A_3$$

$$\therefore 2A_1 - A_2 \geqslant A_3$$

Another method of clipping the noise input pulses and preventing them from switching into leg 3, is to have the effective path length along leg 1 longer than that of leg 2. This will cause the effective m.m.f. required to switch leg 2 to be relatively smaller than that of leg 1 and consequently spurious input noise pulses in leg 1 will create elastic switching in leg 2. From this it can be seen that the desired shape of the device would be that shown in Fig. 9 rather than that shown in Fig. 10. The effective path length of leg 1 could be further increased by enlarging the input aperture as shown in Fig. 13. The cross-sectional area of the legs and the relation between them must still remain as before.

The Output Aperture

The read-out pulses are usually taken from leg 4 of the output minor aperture, whether the read-out is in the form of destructive or non-destructive pulses. During the input phase, the flux in leg 3 is first switched from the downward direction to the upward direction. Then during the second phase, the interrogate or the prime m.m.f. switches the upward direction flux from leg 3 to leg 4. In order to enable maximum interchange of flux between legs 3 and 4, the cross-sectional areas of these two legs should be made approximately equal.

$$A_3 \simeq A_4$$

However, care must be taken to ensure that the noise output pulses from leg 4 will be as small as possible. Hence, when the device is in its blocked state, the flux in leg 4 must be fully saturated in the downward direction. This may be achieved by making the cross-sectional area of leg 4 equal or slightly smaller than that of leg 3.

$$A_3 \geqslant A_4$$

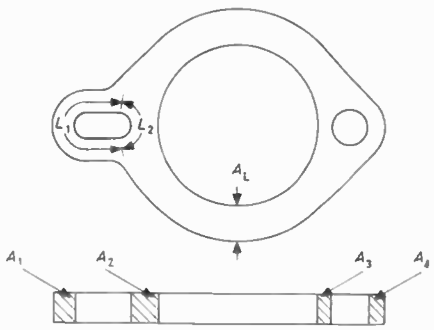


Fig. 13. Dimensions of the input aperture
$A_1 < A_2$
$A_1 + A_2 > A_L$
$L_1 > L_2$

As the ratio of diameters of the larger and smaller apertures is increased, the path lengths round the larger aperture and along leg 3 or leg 4 will become approximately equal, $L_{131} \simeq L_{141}$. Although, theoretically no switching should take place in leg 4, when the device is set, but only in leg 3, it is inevitable that some switching will take place in leg 4, since the path lengths are approximately equal. This effect may cause some complications when using the device for non-destructive read-out stores. The danger of switching into leg 4 may be diminished by reducing the diameter ratio of the apertures, but as shown before when discussing Fig. 11, this will consequently also reduce the operation range of the prime and the interrogate m.m.f.'s. Another method of increasing the difference between the path lengths along leg 3 and along leg 4, is by enlarging the output aperture as shown in Fig. 14. In this configuration the effective path length of leg 4 is increased in relation to the effective path length of leg 3.

$$L_{131} < L_{141}$$

$$L_3 < L_4$$

In this configuration of the device, the ratio of diameters of the large and the input aperture is not affected, only the ratio of the large and output apertures is changed. This will cause a change in the relation between the two curves which were shown in Fig. 11. No variation will take place regarding the 'setting' curve, but the signal-to-noise ratio curve would be shifted further to the right. This is due to the increase of the interrogate switching path round the output aperture.

The configuration shown in Fig. 14 is important when the device is used for non-destructive stores, since a high signal-to-noise ratio is required from the output in leg 4. However, in all-magnetic circuits there is no danger that there will be any switching in leg 4. Although, in the device shown in Fig. 12, the path lengths along leg 3 and along leg 4 are approximately equal, $L_{131} \simeq L_{141}$, the reluctance along leg 4 will be much higher than along leg 3. In all-magnetic circuits, the extremely low impedance, imposed on leg 4 by the coupling loop, increases the reluctance along that leg but does not affect the reluctance along leg 3. Thus, when setting the device in all-magnetic circuits, only the flux in leg 3 is switched while the flux in leg 4 will remain undisturbed. The shape of the output aperture which should be used for all-magnetic circuits is that shown in Fig. 12 or in Fig. 15. In the shape shown in Fig. 15 the shadowed regions are minimized and the effective path length of leg 3 is equal to that of leg 4. Furthermore the elastic switching caused in leg 3 and leg 4, when the device is in its blocked state, will be equal.

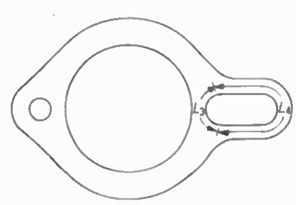


Fig. 14. Extended output aperture for non-destructive stores
$L_3 < L_4$

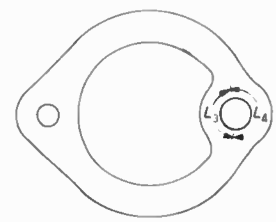


Fig. 15. Shaped output aperture for all-magnetic circuits
$L_3 = L_4$

In one half cycle, leg 3 is driven further into saturation by the drive m.m.f., and in the second half-cycle, leg 4 is driven further into saturation by the hold m.m.f. (which prevents back-flow-of-information).

Thickness of the Material

It has been shown before that the ratio of the large to the small apertures must be as large as possible. This requirement of the device was suggested in order to achieve a wide operation range of the prime and interrogate m.m.f's. On the other hand, the amplitude of the read-out pulses is required to be as large as possible. This could be achieved by increasing the diameter of the output aperture which consequently increases the amount of material switched. However, due to the limitation stated above, by increasing the diameter of the output aperture, the whole structure of the device must be increased by the same proportion.

A more practical method for increasing the amplitude of the read-out pulses is to increase the thickness of the device. By using this method the restrictions laid down before, regarding the device geometry, are not affected. The ratio limitation of the cross-sectional areas of the legs will remain as laid down before, since all the cross-sectional areas are increased in the same proportion. Nevertheless, the flux which is switched will increase and hence the read-out pulses will also increase.

$$\Phi = 2BA = 2Bh\,(r_2 - r_1)$$

$$V_{out} = Nd\Phi/dt = 2Nh\,(r_2 - r_1)\,dB/dt$$

Where h is the thickness of the material, r_1 is the inner radius of the output aperture and r_2 is the outer radius of the minimum width of the output aperture.

The magnetomotive-force required to switch the device will not be affected by the increase of the thickness, since the flux switching path lengths are not varied. However, the device would require more switching power with the increase in the thickness, since there will be an increase in the back e.m.f. across the windings.

Conclusions

Although multi-aperture devices have been known for several years, their applications have been rather limited. However, these devices are starting to arouse some interest in this country mainly in all-magnetic circuits where the connexion between the devices is of wire only with no extra components necessary. With further improvements of the geometry more applications are possible.

This article has shown the complex nature of the shape of the multi-aperture device and the geometry of its apertures and legs. There is still a wide scope for more research to be carried out in future on different shapes.

Acknowledgments

The author wishes to thank English Electric-Leo Computers Ltd for permission and facilities to publish this article. In preparation of this article the author made use of some information in his paper entitled "Performance of an All-Magnetic Shift Register" which was published by the Institution of Electrical Engineers as Paper No. E4362, and published in the Proceeding of that Institution, February, 1964.

REFERENCES

1. Rajchman, J. A., Lo, A. W. The Transfluxor. *Proc. Instn. Radio Engrs. 44*. 321 (1956).
2. Crane, H. D. High Speed Logical System Using Magnetic Elements and Connecting Wire Only. *Proc. Instn. Radio Engrs. 47*, 63 (1959).
3. Rowe, I. H., Slemon, G. R. On the Process of Flux Reversal in Multi-Aperture Ferrite Cores. *Communication and Electronics* Paper No. 61 (1961).
4. Morgan, W. I. Transfluxor Design Considerations. *IRE Trans. Electronic Devices. ED-8*, No. 2. 155 (1961).
5. Cohler, E. V., Baker, T. E. Geometric Factors in Multi-Aperture Ferrite Devices. Sp. Tech. Conference on Non-Linear Magnetics, sponsored by AIEE and IRE (November 1961).
6. Nitzan, D. Flux Switching in Multipath Cores. Rep. I, Standford Research Inst., U.S. Government Research Report No. AD 275 533 (November 1961).
7. Jensen, H. J. Multi-aperture Ferrite Cores for Infinite Time Constant Memory Devices. U.S. Government Technical Memorandum, Case No. 13.804.90 (March 1962).

Automatic Control for a Shoe Warehouse

Lancashire Dynamo Electronic Products (M.I. Group) of Rugeley, Staffordshire, has recently installed electronic control equipment at The British Shoe Corporation warehouse at Braunstone, Leicester.

The warehouse, used for the storage of shoes for a number of the retail shops of the British Shoe Corporation, covers an area of approximately 500 000ft^2 including storage area for goods.

Goods arrive at the warehouse by road and are conveyed to a high level temporary holding area in the roof of the building. The goods are then routed to the required destination on the warehouse floor where they are manually unloaded from the conveyor, unpacked and placed on the storage racks.

Orders are picked manually from the storage racks, packed into cartons, and conveyed to another high level storage area, to await transportation by lorry to the various retail shops.

Two control rooms are provided in the warehouse, both at high level, from where four operators control the movement of goods into and out of the warehouse. A standard sized carton which accommodates a number of pairs of shoes, depending upon size and type, is used. The warehouse is designed to handle 15 000 cartons per day, both into and out of the warehouse.

The control equipment provides remote control of the conveyor drive motors and divertor mechanisms. The large distances involved in the warehouse and the high rate of handling of cartons require the use of high speed, highly reliable, automatic control techniques.

For example, the transport of cartons from the inwards holding area to the floor of the warehouse may take some twelve minutes. The selection of a divertor mechanism may not be possible until the goods are near to their final destination.

Minimum demand is made of the operator by the use of automatic control techniques whereby when he despatches the goods from the holding area, he indicates the destination on a desk mounted push-button. The leading carton of the train of goods is automatically identified by means of a reflective label and the label is used to initiate the diversion of the goods to the correct destination automatically.

Indication of all running conveyors is given to the operators by means of a mimic diagram incorporated in the control desk. Television closed-circuit monitors are grouped around the desk and telephone and public address equipment are mounted on the desk top together with the instrument type push-buttons used for the control of manually started conveyors.

Use is made of illuminated push-buttons on the desk top to indicate the state of automatically selected functions.

Static control techniques are used throughout the installation to ensure the reliability of the equipment and allow high speed operation. The electronic control equipment is built exclusively from the standard LDEP range of 'Digital System' transistorized logic industrial control units.

Goods routed to a lane of the inwards holding area are automatically re-routed to an empty holding lane should the lane to which they were initially routed become full.

A programme of four of the 36 holding lane can be set at the control desk to discharge in any order. The lanes will automatically be discharged in a sequence which will ensure maximum usage of conveyor space for transport of the goods to the warehouse floor. The initiation of discharge of a selected lane when a previously selected lane becomes empty is automatic. Indication is given to the operator when the automatic programme is complete by release of a desk push-button.

Routing of goods from the inwards holding area to the warehouse floor is automatic in that the operator selects a destination by means of a control desk push-button. When the programme of four trains of goods is released from the holding area, the routing to the warehouse floor is automatic and a new programme of four more trains may be discharged.

Control is achieved by automatic marking of the leading carton of a train of goods with a reflective label. The selection of the leading carton is made by means of carton counting. The reflective label is detected by photo-electric equipment at each junction on the conveyor route to the warehouse floor.

On detection of a marked carton, an electronic store is interrogated and a decision made whether the junction is set to the correct destination. If a change in direction of the junction is required, this is made automatically. As the carton approaches the floor level conveyor at its destination, the conveyor is automatically started.

In the outwards goods system the assembly of lorry loads of goods from individual orders is achieved semi-automatically by use of manually applied reflective labels and photo-electric equipment. The equipment is used to sequence the orders correctly into lorry loads.

The conveyors forming the outwards holding area are grouped in ten banks of six conveyors each. Each conveyor being approximately 120ft long. The automatic control equipment allows the banks of six conveyors to be used as though it were one long conveyor. Thus lorry loads of goods assembled in the correct sequence on the warehouse floor are maintained in that sequence for lorry loading, during storage at the holding area and when being despatched to the lorry loading dock.

During their passage into and out of the warehouse, the cartons are counted four times and the count displayed at the control desk. The cartons are counted immediately after they are unloaded from the lorry, as they enter the holding lanes of the inwards holding area, as they enter the holding lanes of the outwards holding area, and as they are loaded on to lorries at the despatch bay.

Display of the total quantity of goods transferred both into and out of the warehouse is also displayed to the warehouse manager in his office. Displays are also given to the lorry loading supervisors of the number of cartons loaded and unloaded from the lorries at the loading dock.

Systematic Development of Cohn Structure for H.F. Band-Pass Filters

By K. E. Brown*, B.Sc.

A modification of an existing narrow band transformation technique is presented yielding a lumped component band-pass filter circuit with many practical advantages. Filters designed on this basis may have responses centred on frequencies of up to 200Mc/s with bandwidth ratios ranging from 1 per cent to 20 per cent. The inductance values may be specified throughout and design formulae are presented for the case of equal inductances. The inductances occur only in the shunt arms of the ladder network and each inductance has a bridging capacitance; further, specified input and output capacitances are provided, thus yielding a capacitive path to earth at each node of the ladder. This method can readily be applied when the termination resistances are not equal (including open-circuit load conditions). Use of the standard technique of predistortion to compensate for the dissipation in the inductances is shown to hold good for this transformation.

(Voir page 503 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 509)

S. B. COHN[1] introduced a new low-pass to band-pass transformation design method for direct coupled band-pass filters resulting in a much better response than had previously been possible. This method, applied to lumped component filters, is here presented in a slightly modified form giving yet a further degree of freedom: namely specified input and output capacitances. Networks designed on this basis are capable of operation over a wide frequency range up to 200Mc/s and with a relative bandwidth ranging from 1 per cent to 20 per cent. The transformation involves extensive use of impedance inverting devices which have not up to now been used to any great extent in connexion with lumped component circuits.

A severe disadvantage of conventional narrow bandwidth filters is the large ratio of inductance values in the series branches to those in the shunt branches which in general is of the order of Q_B^2 where Q_B is the inverse bandwidth ratio. As a consequence, at very high frequencies exceedingly small inductance values occur in the shunt branches while at low frequencies the dual situation arises when series arm inductances are often impracticably large. Use of Norton transformation can alleviate the situation in some cases but this seldom applies for relative bandwidths less than 20 per cent. This disadvantage is overcome by the method presented here and design formulae are obtained for a band-pass filter in which the inductances all have the same specified value. A further advantage of this network is that every inductance has a bridging capacitance (a necessary practical condition at high frequencies), and because the inductances only occur in the shunt branches of the resulting ladder, capacitance to earth from every node is provided.

The disadvantages of this realization are:

(1) The approximation breaks down for relative bandwidths greater than about 20 per cent.

(2) This article is concerned only with filters of constant k configuration; this class of filters includes filters with Chebyshev equal ripple and Butterworth maximally flat pass-band characteristics.

(3) Since this method does not allow finite frequency attenuation poles, a large number of sections may be necessary to meet steep skirt attenuation requirements.

(4) The transformation does not yield a symmetric response; the loss in the lower stop-band is increased while that in the upper stop band is decreased with respect to the rejection that can be obtained with a conventional realization.

(5) The number of capacitors in the circuit is more than doubled by this transformation (compared with the standard realization).

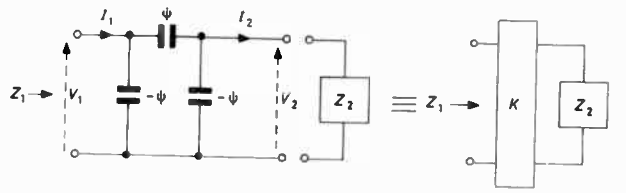


Fig. 1. Capacitance invertor

Quarter Wavelength Invertors and Impedance Inverting Networks

An ideal invertor is a two-port device with the property that if an impedance Z is connected to one port the impedance seen at the other port is K^2/Z where in the ideal case K is independent of frequency. It is found that transmission lines and waveguides which are a quarter wavelength long at the operating frequency exhibit this property over a narrow band of frequencies. Practical impedance inverting networks for use at lower frequencies contain negative components, which in general can be absorbed into neighbouring positive components. The network used in the following transformation is a Π of capacitors as shown in Fig. 1.

The chain matrix of this network is

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ -p\psi & 1 \end{bmatrix}\begin{bmatrix} 1 & 1/p\psi \\ 0 & 1 \end{bmatrix}\begin{bmatrix} 1 & 0 \\ -p\psi & 1 \end{bmatrix}$$

$$= \begin{bmatrix} 0 & 1/p\psi \\ -p\psi & 0 \end{bmatrix}$$

$$\therefore Z_1 = V_1/I_1 = (I_2/V_2).(-1/p^2\psi^2) = -1/Z_2p^2\psi^2 = K^2/Z_2$$

where

$$K = 1/\omega\psi \text{ and } p = j\omega. \quad \ldots\ldots\ldots\ldots (1)$$

Since the above circuit representing the impedance invertor is composed entirely of capacitors the phase change introduced is independent of frequency and equal to $-\pi/2$; this is in general of no consequence. Several other circuits having the same inverting property exist, but are not needed for this analysis.

The frequency dependence of the above invertors when

* The Marconi Co. Ltd.

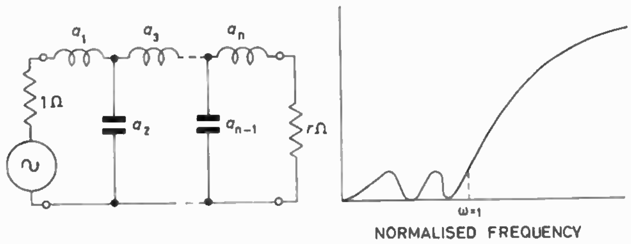


Fig. 2. Low-pass prototype

used in the final band-pass circuit is partially corrected by the alternative formulae for mid-band and bandwidth frequencies derived by Cohn from waveguide considerations extending the useful bandwidth to about 20 per cent of mid-band or operating frequency. These formulae are given in the following section.

Direct Coupled Band-Pass Filters

Consider the basic normalized low-pass prototype filter configuration of order n, either TT for an odd number of branches or TII for an even number of branches, having ladder coefficients $a_1 \ldots a_n$ as shown in Fig. 2.

A typical response of such a configuration is also shown in Fig. 2 for a Chebyshev filter containing five branches.

Conventional frequency transformation yields the band-pass filter of Fig. 3 and the corresponding band-pass response, again for a Chebyshev filter of five branches, is also shown schematically in Fig. 3.

The constants E, F, G, H in Fig. 3 are given by

$$E = R/\omega_B$$
$$F = \omega_B/R\omega_0^2$$
$$G = 1/R\omega_B$$
$$H = R\omega_B/\omega_0^2 \qquad \ldots\ldots\ldots (2)$$

where the conventional transformation frequencies are given by

$$\omega_0 = 2\pi\sqrt{(f_1 \cdot f_2)}$$

and

$$\omega_B = 2\pi(f_2 - f_1).$$

The alternative transformation frequencies suggested by Cohn which partially correct for the frequency dependence of the inverting networks are

$$\omega_0 = 2\pi(f_1 + f_2 - \sqrt{[(f_2-f_1)^2 + f_1 f_2]}) \qquad \ldots\ldots\ldots (3)$$

and

$$\omega_B = \rho \cdot \omega_0 = \frac{f_0(f_2-f_1)}{f_1 \cdot f_2} \cdot \omega_0 \qquad \ldots\ldots\ldots (4)$$

For very narrow bandwidths these frequency definitions tend to the conventional definitions given above.

The network of Fig. 4 can be readily obtained by introducing capacitive inverting networks between each branch and at the input and output of the circuit in Fig. 3 in the positions indicated by the dotted lines (see Appendix).

By suitable choice of the invertor transformer ratios K (equation (1)) it is possible to arrange for all the inductances to be the same specified value. The input and output capacitances C_1 and C_{2n+3} are obtained by a further narrow band approximation replacing the original negative capacitances by positive capacitances. The component values for this case are given below (these formulae are derived in the Appendix).

Given the ladder coefficients a_s of the low-pass prototype the following quantities must be specified:

L; the common inductance value, this should be of the order R_{01}/ω_0.

R_{01}, R_{02}; the required termination impedances.

C_1, C_{2n+3}; the required input and output capacitors.

The series arm capacitances are then given by the formulae

$$C_2 = \frac{\psi_1(1+\omega_0^2C_1^2R_{01}^2)(1+\omega_0^2\psi_1^2R^2)}{(1+\omega_0^2C_1^2R_{01}^2)-\omega_0^2C_1\psi_1R_{01}^2(1+\omega_0^2\psi_1^2R^2)} \qquad \ldots\ldots (5)$$

where

$$R = \frac{R_{01}}{(1+\omega_0^2C_1^2R_{01}^2) - \dfrac{\rho R_{01}}{a_1\omega_0 L}} \qquad \ldots\ldots\ldots (6)$$

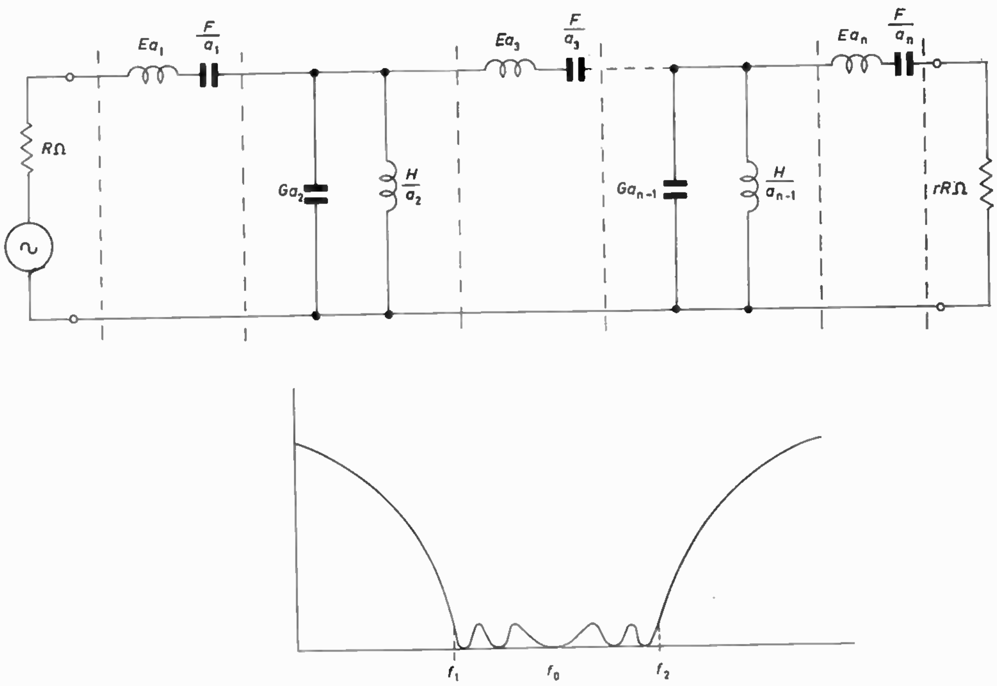
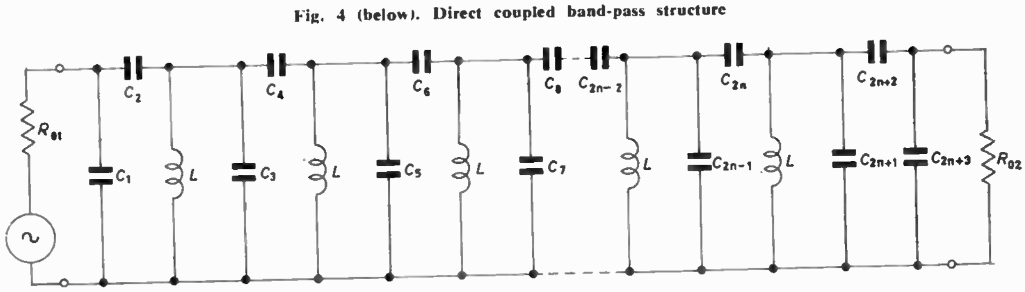


Fig. 3 (above). Standard band-pass realization

Fig. 4 (below). Direct coupled band-pass structure



and

$$\psi_1 = \sqrt{\left(\frac{\rho C_0}{a_1\omega_0 R}\right)} \text{ where } C_0 = 1/\omega_0^2 L \quad \ldots\ldots(7)$$

$$C_{2s} = \frac{\rho C_0}{\sqrt{(a_{s-1}a_s)}} \; (s = 2, \ldots, n). \quad \ldots\ldots\ldots\ (8)$$

$$C_{2n+2} = \frac{\psi_{n+1}(1+\omega_0^2 C_{2n+3}^2 R_{02}^2)(1+\omega_0^2\psi_{n+1}^2 R'^2)}{(1+\omega_0^2 C_{2n+3}^2 R_{02}^2) - \omega_0^2 C_{2n+3}\psi_{n+1} R_{02}^2 (1+\omega_0^2\psi_{n+1}^2 R'^2)} \quad \ldots\ldots\ldots\ (9)$$

where for n odd

$$R' = \frac{R_{02}}{(1+\omega_0^2 C_{2n+3}^2 R_{02}^2) - \dfrac{r\rho R_{02}}{a_n\omega_0 L}} \quad \ldots\ldots\ldots\ (10)$$

and

$$\psi_{n+1} = \sqrt{(r\rho C_0/a_n\omega_0 R')}, \quad \ldots\ldots\ldots\ (11)$$

and for n even

$$R' = \frac{R_{02}}{(1+\omega_0^2 C_{2n+3}^2 R_{02}^2) - \dfrac{\rho R_{02}}{r a_n\omega_0 L}} \quad \ldots\ldots\ldots\ (12)$$

and

$$\psi_{n+1} = \sqrt{(\rho C_0 . / r a_n\omega_0 R')}. \quad \ldots\ldots\ldots\ (13)$$

The shunt arm capacitors are then given in terms of the above equations

$$C_3 = C_0 - \psi_1 - C_4,$$
$$C_{2s+1} = C_0 - C_{2s} - C_{2s+2} \quad (s=2, \ldots, (n-1)),$$
$$C_{2n+1} = C_0 - \psi_{n+1} - C_{2n}. \quad \ldots\ldots\ldots\ (14)$$

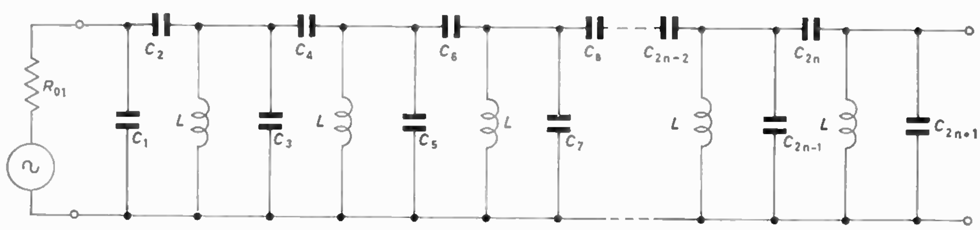
There is not complete freedom of choice of the common value of the inductances L; too high a value may render some of the shunt capacitors negative or some of the series capacitors too small to be practical. This latter condition applies particularly to very narrow bandwidth filters. There is also an upper limit to the value of the input and output capacitors but this is unlikely to be approached in practice since it is in general sufficiently high virtually to constitute a short-circuit between the leads. It should be noted that there is also a restriction on the termination ratio, but again this limitation should not be met in practice. Thus non-equally terminated filters can readily be designed from an equally terminated prototype and, conversely, equally terminated filters may be designed from a non-equally terminated prototype.

Open-Circuit Realization

Filters required to operate into very high impedance loads or in which the ratio of load impedance to generator impedance is more than about 20:1 should be designed from a short-circuit T-section prototype for n odd and from an open circuit TΠ-section prototype for n even. In both cases only n invertors are required. The final circuit is shown in Fig. 5. All the component values except C_{2n+1} remain the same as for the terminated case already quoted. The last capacitance is given by

$$C_{2n+1} = C_0 - C_{2n}. \quad \ldots\ldots\ldots\ (15)$$

Fig. 5. Open-circuit realization



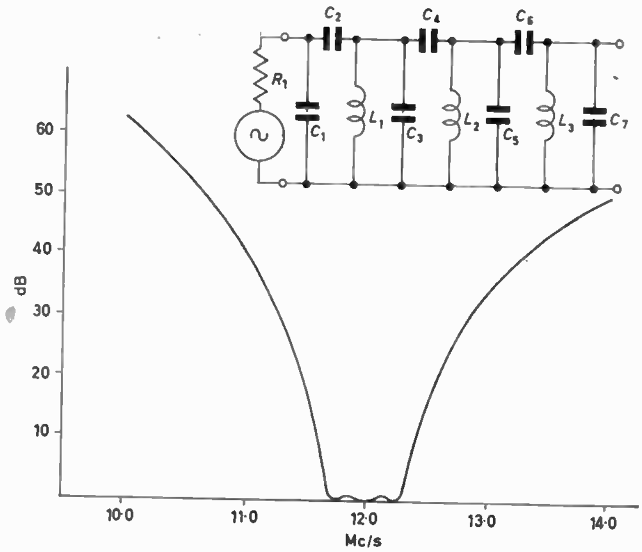


Fig. 6. Example of open-circuit realization
R_1 75Ω
C_1 20pF, C_2 41·66pF, C_3 130pF C_4 7·58pF, C_5 162pF, C_6 6·21pF, C_7 170pF
L_1, L_2, L_3 1μH

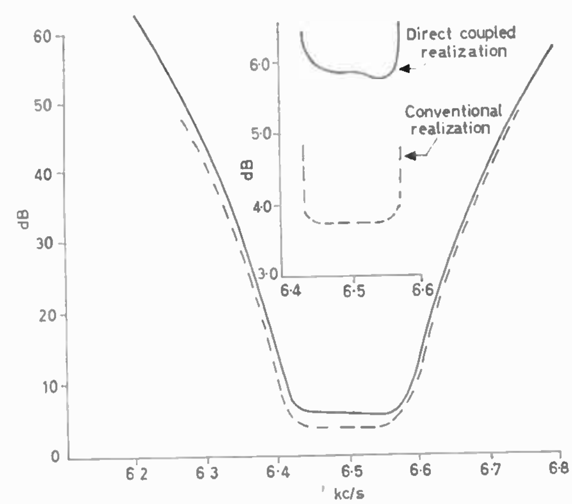
An example of such a filter and its response are shown in Fig. 6 for $n = 3$ giving 5 per cent bandwidth at 12Mc/s.

Predistortion

For narrow bandwidth filters the effects of inductive dissipation can be disastrous in the pass-band. This can be overcome by predistorting the prototype low-pass filter by the standard method[2]. The reactive elements of the final filter are realized by the method given using the predistorted ladder coefficients. The effective inductive resistances are then inserted to obtain the final response. This response, however, is not as good as the predistorted lossy response obtained by standard realization. The example (Fig. 7) shows a filter of bandwidth less than 2 per cent based on a five branch maximally flat prototype predistorted for an inductive Q of 280 at mid-band frequency of 6·5kc/s. The corresponding low-pass prototype component Q is 35·13. It can be seen that the basic loss introduced by predistortion is 6dB and the pass-band is no longer maximally flat. The basic loss introduced by a predistorted standard band-pass filter is 3·75dB. The difference between the basic losses is due to the mismatching action of the direct coupled filter in bringing the termination ratio down from the original 4·25 on the predistorted prototype to unity. In spite of this discrepancy the obvious advantages in component sizes can be readily appreciated. (The input and output capacitances in this case are set to zero.)

Stop Band Response

The out-of-band response given by this realization is not symmetrical. This can be seen by considering the limiting forms of the circuit as the frequency tends to zero and infinity. At zero frequency the response has $2n+1$ poles of attenuation and at infinite frequency 1 pole of attenuation contrasted to



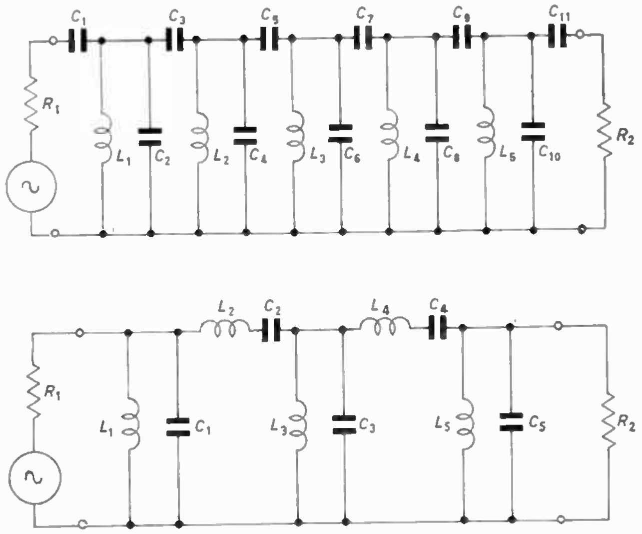


Fig. 7. Example of predistorted realization

Upper filter; direct coupled realization

R_1, R_2 600Ω

C_1 8856pF, C_2 20435pF, C_3 1089pF, C_4 28365pF, C_5 525pF, C_6 29062pF, C_7 382pF, C_8 28993pF, C_9 594pF, C_{10} 26338pF, C_{11} 3072pF

L_1, L_2, L_3, L_4, L_5 20mH

Lower filter; standard realization

R_1 600Ω, R_2 2·55kΩ

C_1 0·665μF, C_2 879pF, C_3 2·866μF, C_4 491pF, C_5 1·25μF

L_1 0·901mH, L_2 682mH, L_3 0·209mH, L_4 1221mH, L_5 0·480mH

n poles at both zero and infinite frequency obtained on standard realization. Thus the transfer function of the approximate transformation is:

$$\Lambda = \frac{\sum_{s=0}^{2n+2} d_s p^s}{p^{2n+1}} \qquad \ldots\ldots\ldots (16)$$

and of the exact realization

$$\Lambda = \frac{\sum_{s=0}^{2n} d_s p^s}{p^n} \qquad \ldots\ldots\ldots (17)$$

Finite Frequency Poles

Sharper cut-off rates can be obtained at the expense of poorer pass-band responses if some or all of the invertors are represented by a Π of tuned circuits. Maintaining equal inductances throughout imposes a condition on the pole frequencies, but since these frequencies are close to the cut-off frequencies this inflexibility is not in general disadvantageous.

This procedure, however, has very detrimental effects to the pass-band characteristic and is not recommended.

Conclusions

An approximate low-pass to band-pass transformation is described which yields a practical circuit suitable for narrow band operation and is particularly convenient at high frequencies. It is shown that it is possible to prescribe a constant inductance value throughout the filter; moreover, the inductances occur only in the shunt branches and each has a bridging capacitance. In addition, specified input and output capacitances are provided, thus yielding a capacitance to earth at each node. The nodes are coupled by further capacitances. Any termination ratio including open-circuit realization can be accommodated. Although predistortion does not yield quite the expected response when inductive losses are included, it does, however, give a flat pass-band response for narrow band filters.

A reasonably good match to the original response is obtained for relative bandwidths up to 20 per cent. As an example of a wide-band filter designed on this basis a five-branch Chebyshev filter with bandwidth of 20 per cent at 100Mc/s was used as a basis for a Cohn structure filter. The pass-band, i.e. the band whose response is within the pass-band distortion of the original Chebyshev filter, was reduced to 17·5 per cent. The rejection in the upper stop-band, measured in decibels, was reduced by a factor of two-thirds. It is possible that some improvement may be obtained by choosing different values of the shunt inductance and the terminating capacitances, but this has not been investigated.

It will be seen from the example responses that the bandwidth is slightly reduced by this transformation and allowance should be made for this in designing filters by this method.

Acknowledgment

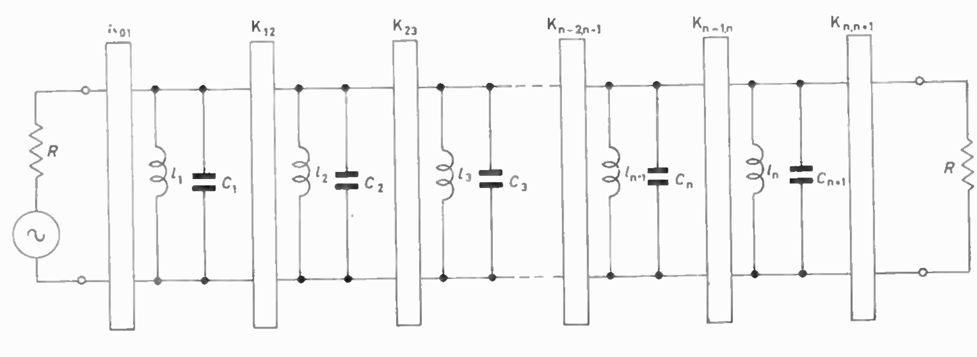
The author wishes to acknowledge the help and encouragement of Mr. J. K. Skwirzynski at whose suggestion this work was carried out.

Appendix

Derivation of Component Values

With reference to Fig. 3 the TT or TΠ network is equivalent to an ideal invertor followed by a ΠΠ or ΠT network respectively. The network to the right of the resulting first tuned circuit again begins as a T type

Fig. 8. Band-pass filter with ideal invertors



impedance and can also be replaced by an invertor followed by a Π type impedance. Thus the original network can be replaced by a sequence of invertors and parallel tuned circuits in shunt as shown in Fig. 8, where:

$$l_1 = K_{01}^2F/a_1 \qquad c_1 = Ea_1/K_{01}^2$$
$$l_2 = (K_{12}^2/K_{01}^2)(H/a_2) \qquad c_2 = K_{01}^2Ga_2/K_{12}^2$$
$$l_3 = K_{01}^2K_{23}^2F/K_{12}^2a_3 \qquad c_3 = K_{12}^2Ea_3/K_{01}^2K_{23}^2$$
$$\text{etc.} \qquad \ldots\ldots\ldots\ldots (18)$$

and

$$R' = \frac{K_{12}^2K_{34}^2\ldots K_{n,n+1}^2}{K_{01}^2K_{23}^2\ldots K_{n-1,n}^2}\,.\,Rr \text{ for } n \text{ odd}$$
$$= \frac{K_{01}^2K_{23}^2\ldots K_{n,n+1}^2}{K_{12}^2K_{34}^2\ldots K_{n,n-1}^2}\,.\,1/Rr \text{ for } n \text{ even}$$
$$\ldots\ldots\ldots\ldots (19)$$

Equating all the inductances to the specified value L and replacing the invertors by the Π of capacitors of Fig. 1, all the capacitors of the circuit except C_1, C_2, C_{2n+2} and C_{2n+3} are given in terms of L and R. The last invertor $K_{n,n+1}$ determines the final termination ratio.

It now remains to relate the design resistances R and R' and the capacitors C_2 and C_{2n+2} to the required termination resistance R_{01} *and* R_{02} and the specified termination capacitors C_1 and C_{2n+3}.

Consider the generator end of the ladder. When the first invertor K_{01} is replaced by the corresponding Π of capacitors, one of the negative capacitors cannot be absorbed into a positive capacitor (see Fig. 9(a)). This situation is overcome by replacing the circuit of Fig. 9(a) by that of Fig. 9(b) (in which R_{01} and C_1 are specified), equating the real and imaginary parts of the two impedances at mid-band frequency thus:

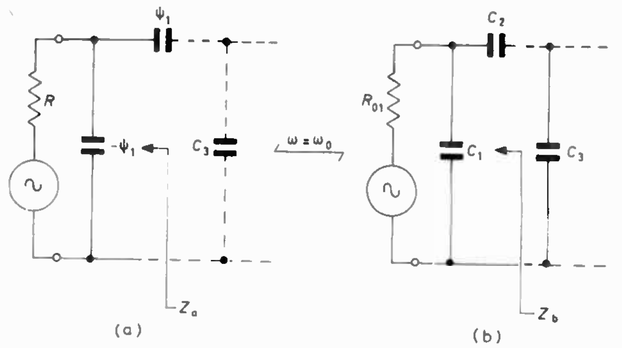


Fig. 9. Termination approximation

$$\frac{R}{1+\omega_0^2\psi_1^2R^2} = \frac{R_{01}}{1+\omega_0^2C_1^2R_{01}^2} \qquad \ldots\ldots (20)$$

and

$$\frac{1}{\omega_0\psi_1(1+\omega_0^2\psi_1^2R^2)} = 1/\omega_0C_2 + \frac{\omega_0C_1R_{01}^2}{1+\omega_0^2C_1^2R_{01}^2}$$
$$\ldots\ldots\ldots\ldots (21)$$

Also from equation (18):

$$L = L_1 = K_{01}^2F/a_1 = \frac{\rho}{a_1\omega_0^3\psi_1^2R} \qquad \ldots\ldots (22)$$

These three equations enable the three quantities C_2, ψ_1 and R to be evaluated. The quantities C_{2n+2}, ψ_{n+1} and R' may be evaluated in like manner.

REFERENCES

1. Cohn, S. B. Direct Coupled Resonator Filters. *Proc. Inst. Radio Engrs.* (Feb. 1957).
2. Darlington, S. Synthesis of Reactance Four Poles. Bell Syst. Monograph B.1186, p. 79.
3. Skwirzynski, J. K., Zdunek, S. Design Data for Symmetrical Darlington Filters. Instn. Elect. Engrs. Monograph 227R, p. 14.

A New Radio Telescope at Jodrell Bank

A NEW Mark II Radio Telescope has now been constructed at Jodrell Bank under the control of the Ministry of Public Building and Works acting as agent for the Department of Scientific and Industrial Research, which made a grant of £300 000 to Manchester University to cover the construction cost.

The new telescope which will shortly be brought into service has been designed to operate on wavelengths of 21cm or less and to have a still more accurate beam width and guidance system than the Mark I. Its immediate research programme is the measurement of the intensity of the more distant radio sources up to 4 to 5000 million light years away, at wavelengths of 15, 10 and 6cm.

The instrument has an accurately balanced paraboloid steel bowl with elliptical aperture as distinct from the 250ft Mark I Jodrell Bank telescope, which has a circular aperture with the focus in the aperture plane. This elliptical aperture measures 125ft on the major axis and 83ft 4in on the minor.

The Mark II telescope has a maximum speed on both motions of 90°/min with smooth speed control down to at least 1/8°min. It can continue operating in winds of up to 40 mile/h, which are rarely experienced at Jodrell Bank, and when used for radio astronomy purposes it will have a following accuracy better than 2 minutes of arc.

The revolving superstructure which carries the bowl is of concrete; in this respect and also in the form of the azimuth and elevation drives the design is similar to that of the 85ft diameter satellite communications aerial at Goonhilly Downs. Bowl tilting (elevation) motion is by screw, nut and connecting rod, the lower end of the rod being attached to the nut and the upper end to the bowl structure, which is able to pivot through bearings mounted on the top horizontal beam of the pre-stressed concrete structure. The screw, which is over 30ft long and 8½in in diameter, is driven through precision reduction gearing by an electric motor, causing the nut and hence the connecting rod, to move up or down according to the direction of the rotation of the screw. Superstructure rotation (azimuth) motion is by means of a 5in pitch Renold chain, engaging with teeth set round the periphery of the superstructure, and a horizontal sprocket. The sprocket is driven through precision gearing by an electric motor, and special precautions are taken to avoid backlash and give a smooth drive under all operational conditions. The concrete rotating superstructure, weighing about 1000 tons, is carried on 54 tapered rollers running between manganese steel tracks and located about a heavy vertical centre pivot roller bearing by a roller cage and 27 spokes. The diameter of the roller tracks is over 40ft. The tracks were machined as complete rings in the works and erected to extremely close tolerances of level circularity.

A unique feature of the instrument is the use of a Ferranti ARGUS 100 digital computer.

This is the first occasion on which a direct control system of this kind has been installed, and it is expected to provide considerable improvements over previous control methods in the accuracy and flexibility.

The drive and control equipment which was supplied by the Brush Electrical Engineering Co. Ltd consists of 75 h.p. and 50 h.p. Ward Leonard Drives for the azimuth and elevation motions respectively, together with all the contactor and sequence equipment for the drives and ancillary units.

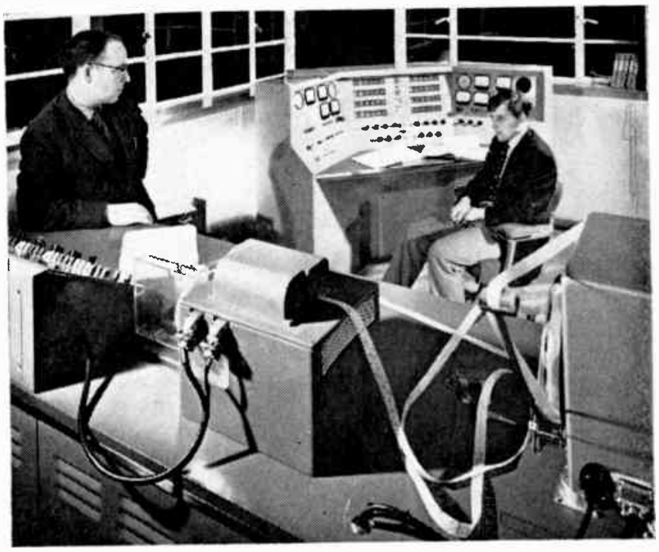
Control of the drive motors is by a closed loop velocity servo from an input analogue velocity demand signal. The speed range of the azimuth and elevation drives is 720: 1, armature voltage control being used. The drives are reversible by the use of twin field exciters and push-pull amplifiers. Maximum speed of the motors is 1000rev/min which corresponds to an aerial speed of 90°/min.

The servo equipment includes all the necessary stabilizing networks and incorporates current and speed limiting features. Fully transistorized plug-in units are employed.

Since the telescope is on an altitude-azimuth mount, it is necessary to compute the angles to which it is set continuously during the time it is used even in order to direct it to a fixed star, since the steady rotation of the earth does not bear a simple relation to the rotations of the two telescope axes. If search patterns are to be carried out, this increases the severity of the computational problems.

Similar calculations are performed in reverse order to display or print galactic or terrestial co-ordinates for the astronomers from electrical information on the telescope position provided by optical shaft digitizers made by S. G. Brown Ltd to the same design as those used on the Goonhilly aerial. The digitizers provide 16-bit information, which corresponds to one part in 64 000 or about 20sec of arc.

The new telescope



Ferranti ARGUS 100 computer and control desk in the control room

The intensity of radiation found by the telescope when pointing to any particular spot in the sky is converted to a digital value, and recorded via the computer. Sequences of readings are either passed directly to the Manchester University ATLAS computer or are pre-processed by ARGUS to reduce the quantity of data before final analysis.

Several miles of drive, control and research apparatus cabling have been installed on the instrument and to connect it to the control desk and computer housed in a building approximately 100 yards away from the telescope. No sliprings have been used, cable looping devices above the centre pivot and on the bowl pivot bearings ensure that every cable has a free loop of sufficient length to accommodate the arcs of travel, which are 420° in azimuth and 95° in elevation, giving full sky coverage.

The main (remote) control desk displays all the essential operational information and includes all necessary safety and warning devices. A focusing device to adjust the position of the aerial array in two planes is remotely controlled from the desk. Research apparatus mounting space is provided in a cabin at the focus, a second cabin in the bowl structure, and a large concrete laboratory on the turntable. The first two cabins tilt with the bowl and the third rotates in azimuth; the focus cabin is fitted with an air conditioning system for use with high gain amplifiers.

The total of the grants made or promised by D.S.I.R. to Jodrell Bank, since 1952 is now over £1M of which between 75 per cent and 80 per cent has been for radio telescopes and the remainder to finance individual research programmes as follows:

£360 000 Mark I Radio-telescope completed 1957

£300 000 Mark II Radio-telescope

£140 000 Mark III Transportable Radio-telescope (125ft) to be completed Summer 1965

£4000 Design study for improvement of the Mark I

£212 000 Research programmes using the Marks I and II.

Further extensions to Jodrell Bank may be expected when the Fleck Committee on Radio Astronomy publishes its report in the autumn of this year.

Some Precision Direct Coupled Transistor Amplifiers and Their Approximate Design

(Part 1)

By C. W. B. Grigson*

Detailed mathematical analyses are of little practical use in actual design work unless the parameter values of the individual transistors are known. In this article design methods are given which enable precision circuits to be developed with only an approximate knowledge of transistor β, V_{be} and V_{knee}. The circuits dealt with are (1) the ring-of-three voltage amplifier, (2) its application as a voltage amplifier with a drift of $20\mu V/°C$, (3) a gain-stabilized amplifier with full-scale sensitivity of $3 \times 10^{-11}A$, (4) a 0 to 20A current stabilized source with an output resistance of 300Ω, (5) a class-B current amplifier with a capacity of 20A peak-to-peak.

(Voir page 503 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 510)

MANY sophisticated studies of transistors in circuits have been made and they have led to the present extensive knowledge of transistor behaviour. Unfortunately, valuable though the detailed mathematical analyses are, they may not be much help in actual design if one is ignorant of the parameter values of the individual transistors. The aim of this article is to give design methods which enable some precision d.c. circuits to be developed. The design methods will assume only a rough knowledge of transistor β, V_{be} and V_{knee}; that in some cases input pairs will be matched for β; and that high frequency stabilization will be done experimentally. The advantage of this realistic assessment of the actual data likely to be available is simplicity: nothing but a first-order design is justified.

The article shows that for a wide range of special purpose equipment useful in instrumentation, this primitive approach is sufficient, reducing design and development effort. Detailed methods are given for the design of class-A ring-of-three d.c. amplifiers, which are then applied to three problems of precision instrumentation: measurement of small voltages, of small currents, and the provision of large constant currents. Similar design methods are also given for the class-B ring-of-three d.c. amplifiers originally developed for use as slow sweep drives in electron beam instruments[1]. These are of high precision, do not require matched power transistors, and may be of high power.

The circuits are (1) the ring-of-three voltage amplifier, (2) its application as a voltage amplifier with a drift of $20\mu V/°C$, (3) a gain-stabilized amplifier with full-scale sensitivity of $3 \times 10^{-11}A$, (4) a 0 to 20A current stabilized source with an output resistance of 300Ω, (5) a class-B current amplifier with a capacity of 20A peak-to-peak.

The design procedure is adequate to predict the values of nearly all components, and to design sufficient uncritical adjustments to cover adequately the wide parameter variation of the transistors. It takes perhaps two hours of paper work (as will become evident) to design a typical circuit, and this must be supplemented by a similar amount of laboratory work. The latter consists of checking levels without the d.c. feedback, and in reducing the high-frequency loop gain enough to prevent high-frequency oscillations. In all these circuits a high degree of feedback is essential for precision and for thermal stabilization.

For laboratory equipment an elaborate paper design is seldom meaningful, since sufficient data about the transistors, for example for a gain-phase analysis, is not available. In all the circuits described it is easy experimentally to prevent self oscillation. It is worth emphasizing that for d.c. applications a precision circuit can be developed with as little design effort as a valve-circuit for similar functions, if indeed it is possible to have a valve-circuit as an alternative. The transistorized circuits are much simpler to construct, and much cheaper to energize, especially as precision circuits should in general be fed by stabilized d.c. supplies. For these reasons such circuits should be much cheaper to produce than valved ones for comparable functions.

The success of design methods as primitive and simple as the ones to be given stems from a principle, two facts, and a matter of ignorance. The principle is that good circuits should use much feedback; the first fact is that such circuits are quite easy to stabilize against self-oscillation; the second that if the transistors are to work at all, V_{be} is confined to a narrow range, about 650mV for silicon, 300mV for germanium. The matter of ignorance is the transistor β, unknown to a factor of three: therefore the predictions of a more refined design method are really no more accurate than those of a first-order one.

Class-A Ring-of-Three Voltage Amplifier

It consists of two long-tailed pairs followed by a common-collector stage, and the design depends on the fact that alternate pnp and npn pairs, and emitter followers, set their own d.c. levels automatically. The design procedure entails (1) designing for consistent d.c. levels *starting at the output end.* (2) Checking that the d.c. circuit will give an adequate variational performance. (3) Checking that the impedance levels are suitable for source and load and for the correct operation of the feedback.

The gain may be positive or negative according to the way the pairs are interconnected, or which phase the last stage is driven by. The feedback may be operational; what might, from valve practice, be called 'classical ring of three'; or via the difference characteristics of the input long-tailed pair.

Suppose that the circuit of Fig. 1 is required to supply a load G or R with $\pm I_p$ peak current at $\pm V_p$ peak voltage. The d.c. lines will be supposed low impedance, value B_1 and B_2; the transistors all silicon, of current amplification factor* β.

The circuit has pnp transistors for the first stage; this is to permit the use of cheap alloy transistors in the first stage with improved drift properties compared with dif-

* *Engineering Department, University of Cambridge.*

* *The verbose terms usual for β are deplored: one does not refer to the μ of a valve as the 'open-circuit voltage amplification factor'.*

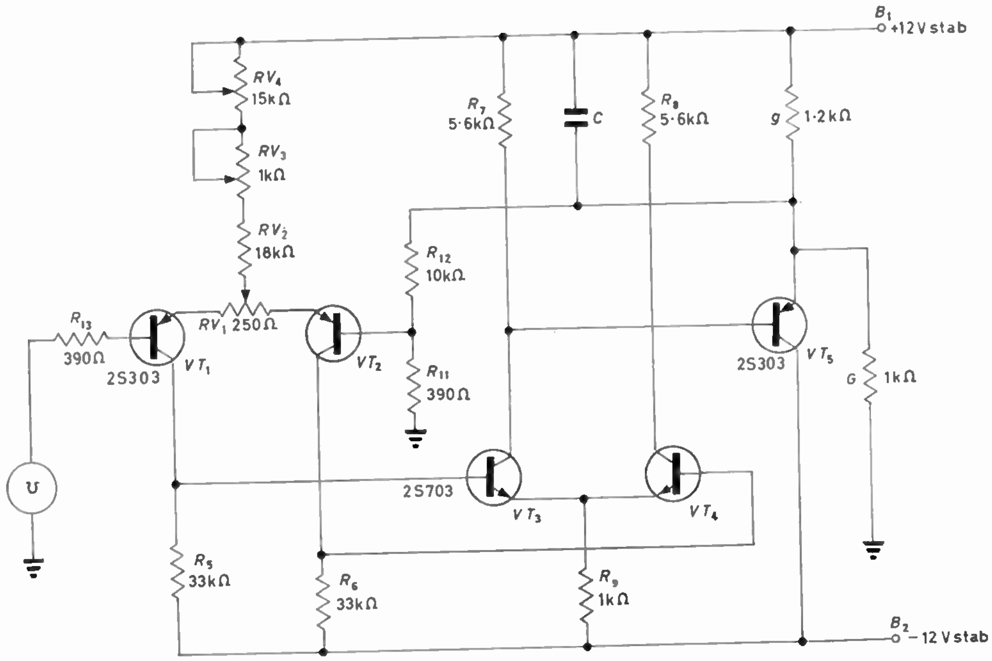


Fig. 1. Class-A ring of three d.c. amplifier

fused junction ones; ideally one would prefer planar expitaxials.

The Output Stage

Let the smallest value of the transistor emitter current be I_1 and the largest I_2, these values correspond with the maximum and minimum values of the load voltage, V_2, V_1 respectively. Since it is required to solve for the specified

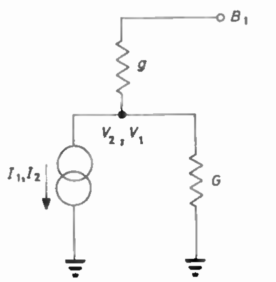


Fig. 2. Current soorce equivalent to the emitter of VT_5 and its loads g and G

current I_1, or I_2, the equivalent circuit of Fig. 2 is permissible. Nodal equations are then:

$$-I_1 = V_2(g + G) - B_1 g \qquad (1)$$

$$-I_2 = V_1(g + G) - B_1 g \qquad (2)$$

I_1 is chosen by the designer so that* $I_1 \gg \beta I_{co}$.

Having decided a suitable value for minimum current, equation (1) fixes g. Equation (2) then gives maximum current. If quiescent and peak values are denoted by suffices q and p

$I_{qe} = 1/2(I_2 + I_1)$, $I_{pe} = 1/2(I_2 - I_1)$ and conditions at the base of the emitter-follower are then obtained:

$$I_{qb} = I_{qe}/\beta - I_{co} \qquad (3)$$

$$I_{pb} = I_{pe}/\beta \qquad (4)$$

$$V_{qb} = -0{\cdot}65\text{V} \qquad (5)$$

$$V_{pb} = V_p \qquad (6)$$

The input impedance looking into the base of VT_5 is

$$R_{in} = \beta/(g + G) \qquad (7)$$

* *No distinction is made between β and $\beta + 1$.*

The Second Stage

The collector of VT_3 must provide the drive demanded by equations (3) to (6). The collector should not be loaded appreciably by the d.c. base current to VT_5, or there is an imbalance between VT_3 and VT_4. A standing collector current of

$$I_{qc} = 10\, I_{qb} \qquad (8)$$

will ensure this; the factor 10 may seem over-safe, but it must allow for a variation of 3 : 1 in I_{qb} because of the β variation possible in VT_5. The precise factor is arbitrary, is the choice of the designer and a smaller one will often work.

Equations (5) and (8) now give $R_{7,8}$ (Fig. 3).

$$R_{7,8} = \frac{B_1 - V_{qb}}{I_{qc}} \qquad (9)$$

When the load signal is at minimum level V_1 the transistor VT_3 must not bottom, its collector-emitter voltage must be at least equal to V_{knee}.

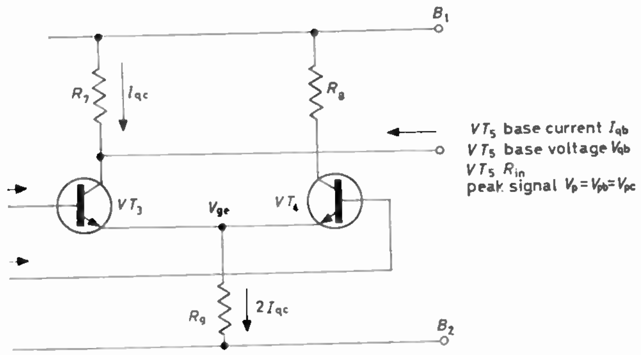


Fig. 3. Corrents aod voltages in the second stage

$$V_{qe} = V_{qc} + V_1 - V_{knee} \qquad (10)$$

V_{qc} is the same as equation (5), V_{knee} is typically 0·5V, hence V_{qe}, the d.c. level of the emitters of the second pair. Therefore

$$R_9 = \frac{V_{qe} - B_2}{2I_{qc}} \qquad (11)$$

The quiescent levels of VT_3 and VT_4 bases must be

$$V_{qb} = V_{qe} + 0{\cdot}65 \qquad (12)$$

and d.c. base current must be

$$I_{qb} = (I_{qe}/\beta) - I_{co} \qquad (13)$$

Notice that equation (13) can vary by a factor of 3:1 because of the variation of β: for silicon transistors in 'normal' laboratory use, the leakage is by comparison negligible.

So far the second stage design has been based entirely on quiescent states. However it must be ensured that VT_3 collector can supply enough signal current to VT_5 base, and since $R_7 \ll R_{in}$ this may not always be possible. For the peak base current of VT_5 is given by

$$I_{pb} = V_{pc}/R_{in} \qquad \ldots\ldots\ldots\ldots\ldots\ldots (14)$$

and also

$$I_{pb} \leqslant I_{qb} = (I_{qc}/10) = \frac{B_1 - V_{qc}}{10\,R_7} \qquad \ldots\ldots\ldots\ldots (15)$$

Thus equation (8), which ensures adequate balance in stage 2, and equation (14) together imply a relation between V_{pc} ($= V_p$, the load voltage) and the d.c. supply voltage. Since $B_1 \gg V_{qc}$, equation (14) and (15) show that

$$V_p/R_{in} \leqslant B_1/10R_7$$

or

$$V_p/B_1 \leqslant R_{in}/10R_7 \qquad \ldots\ldots (16)$$

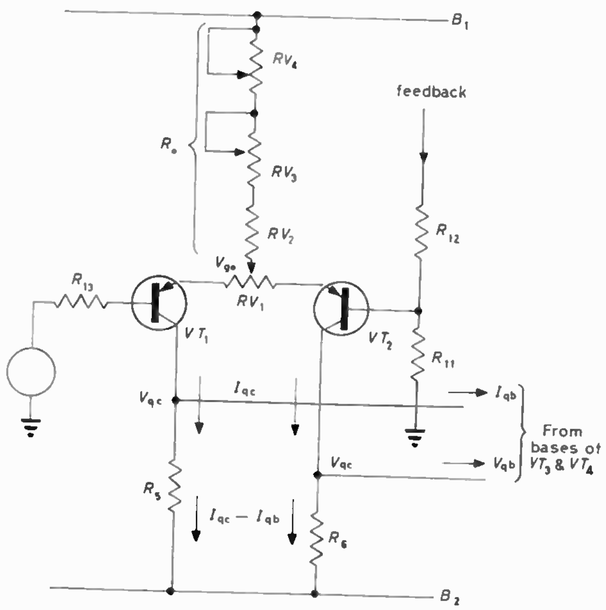


Fig. 4. Currents and voltages in the input stage

Note that because of R_{in} the inequality (16) is very sensitive to variation of β.

The Input Stage

Continuing as before, the collectors of VT_1 and VT_2 (Fig. 4) must supply the current, equation (13), to the bases of VT_3 and VT_4, which are to be at the same level as in equation (12). The collectors see the variational input impedance to VT_3 and VT_4, which is small and balanced on each side. It is possible to drive VT_3 and VT_4 with *currents*, since $R_{5,6}$ are much greater than the input resistance to VT_3 and VT_4. Quiescent collector currents of VT_1 and VT_2 could therefore be only slightly larger than quiescent base drives of VT_3 and VT_4, but the latter may be so small, in this part of the amplifier, that VT_1 and VT_2 might then be working in the region where β falls off. On the other hand, small values of I_{qc} enable large collector load resistors to be used, tending to increase the gains. Suitable compromise values of I_c must be decided by the designer. But

$$I_{qc} \mid \geqslant I_{qc}/\beta_{min} \mid \qquad \ldots\ldots\ldots\ldots\ldots (17)$$

for VT_1 or VT_2 for VT_3 or VT_4

When I_{qc} has been decided, the values of R_5 and R_6 follow:

$$R_{5,6} = \frac{V_{qc} - B_2}{I_{qc} - I_{qb}} \qquad \ldots\ldots\ldots\ldots\ldots (18)$$

where V_{qc}, I_{qb} are given by equations (12) and (13), and I_{qc} must satisfy equation (14).

If I_{qc} was close to the equality in (17) then $R_{5,6}$ would be very large, conversely if β were at its maximum, $R_{5,6}$ would be small. Variation of β in transistors VT_3 and VT_4 demands either a variation of R_5 and R_6 to satisfy equation (18) or a variation in I_{qc} to do so. The latter course is preferable, and is discussed later.

The bases of VT_1 and VT_2 have to be quite closely at zero voltage, so that the emitters are at

$$V_{qe} = +0{\cdot}65\text{V} \qquad \ldots\ldots\ldots\ldots\ldots (19)$$

$$R_e = \frac{B_1 - V_{qe}}{2I_{qc}} \qquad \ldots\ldots\ldots\ldots\ldots (20)$$

In fact the bases of the input transistors set themselves at

$$(-I_{qc}/\beta) \times R_{11,13} \qquad \ldots\ldots\ldots\ldots\ldots (21)$$

This is typically a few millivolts. Both bases should see the same d.c. resistance to earth, otherwise there is a voltage imbalance which is amplified by the subsequent d.c. gain, and this is often big enough to cut off part of the amplifier.

Adjustments

There are two, or possibly three. RV_1 compensates for differences between the base-emitter voltage characteristics of the input pair. RV_4 and RV_3, coarse and fine, enable the emitter currents to be adjusted after they have been equalized by R_1. If a precise gain figure is wanted then R_{12} might be adjustable to set the feedback ratio precisely.

V_{be} may vary by $\pm 0{\cdot}05$V for a fixed emitter current, R_1 must be large enough so that

$$R_1 I_{qc} \geqslant 0{\cdot}05 \qquad \ldots\ldots\ldots\ldots\ldots\ldots (22)$$

Rearranging equation (18) in terms of I_{qc}, the collector current of VT_1 or VT_2,

$$I_{qc} = \frac{V_{qc} - B_2}{R_{5,6}} + I_{qb}$$

where I_{qb} is the base current of VT_3 or VT_4.

TABLE 1

QUANTITY	ESTIMATED VALUE	VALUE USED	ACTUAL OR FINAL VALUE
RV_1	250	250	250 (kΩ)
$R_2+RV_3+RV_4$	26	18+1+10	32 (kΩ)
RV_3 & RV_4	6	1+10	1+15 (kΩ)
$R_{5,6}$	35·3	33	33 (kΩ)
$R_{7,8}$	5·75	5·6	5·6 (kΩ)
R_9	1·09	1·0	1·0 (kΩ)
$(1/g)$	1·2	1·2	1·2 (kΩ)
A	33	—	32 (V/V)
Ratio V_p/B_1 from equator (16)	0·475	—	0·375
Limit (16) using actual β of VT_5	0·36	—	0·375
D.C. Levels			
Base VT_5 = collector VT_3	−0·65		−0·55 (V)
Collector VT_4	—		+0·25 (V)
Bases VT_4 & VT_3 = collectors VT_2 & VT_1	−6·5		−6·8 (V)
Emitters VT_4 & VT_3	−7·2		−7·6 (V)
Emitters VT_2 & VT_1	+0·65		+0·6 (V)
β values at appropriate current VT_1	50		36
VT_2	50		36
VT_3	25–75		70
VT_4	50		100
VT_5	50		37

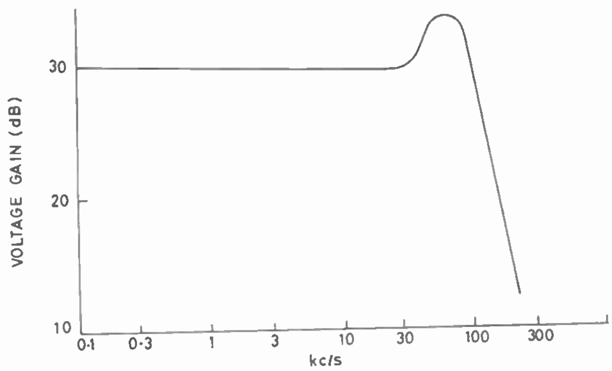


Fig. 5. Gain-frequency characteristic of d.c. amplifier of Fig. 1

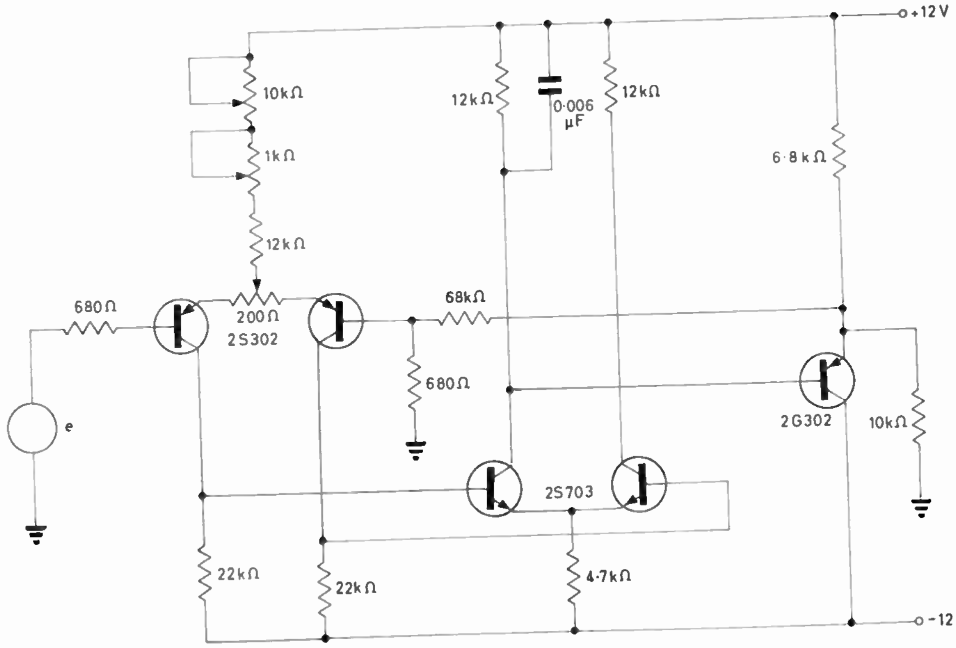


Fig. 6. Part of low drift 'straight' voltage amplifier

Therefore

$$\delta I_{qc} = \delta I_{qb} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (23)$$

where the change in base current on the right-hand side is to be obtained from the possible scatter of β values in equation (13). Inserting the least and greatest values $I_{qc} - \delta I_{qc}$ and $I_{qc} + \delta I_{qe}$ in equation (20) one obtains the necessary variation for R_e.

$$\delta R_e / R_e = \frac{(\delta I_{qc}/I_{qe})}{1 - (\delta I_{qc}/I_{qc})^2} \quad \ldots\ldots\ldots \quad (24)$$

This can be considerable, if $\delta I_{qc}/I_{qc} = \frac{1}{2}$ then $\delta R_e/R_e = \frac{2}{3}$.

Check of the Variation Performance

The input impedance between a base and earth of the long-tailed pairs is:

$$R_{in} = 2(r_b + \beta r_e) \quad \ldots\ldots\ldots\ldots \quad (25)$$

r_b must include a resistor like R_{11} in the base lead, r_e must include $\frac{1}{2}R_1$.

The current gain from base-to-collector load is:

$$|A_i| = \frac{\beta}{1 + R_c/r_d} \quad \ldots\ldots\ldots\ldots\ldots \quad (26)$$

(R_c/r_d) is rarely greater than 0·3, in view of the ignorance of β one might just as well take the denominator as unity.

The voltage gain from base to collector is

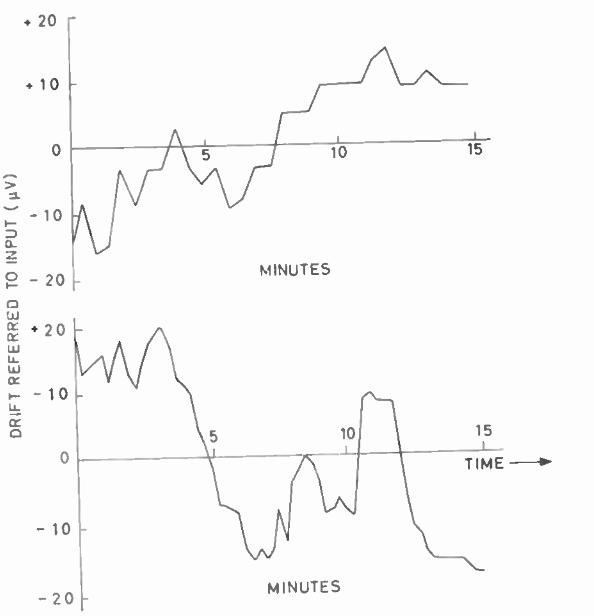
$$|A_v| = \frac{R_c A_i}{2(r_b + \beta r_e)} \quad \ldots\ldots\ldots\ldots\ldots \quad (27)$$

In these expressions r_b, r_e, r_d, β are the common-emitter T parameters, R_c the effective collector load resistance. R_{in} in equation (25) can easily be 10kΩ or more; if the feedback factor B is to be governed by the ratio $R_{11}/(R_{11} + R_{12})$ then $R_{11} \ll R_{in}$. Stage voltage gains can be high, especially if high β transistors are used. In contrast with d.c. valve circuits there is no gain-loss in transfer networks from one stage to the next.

Performance of a Typical Amplifier and Conclusion

This procedure was used for the class-A designs given in the following sections, and enabled voltage amplifiers quickly to be tailored to suit different impedances and supply voltages. As a detailed example, Table 1 shows the quantities predicted for the amplifier of Fig. 1, which was designed to give 6mA into 1kΩ load, with ±12V stabilized d.c. lines. Transistors VT_1 and VT_2 were matched for β, the β values used in design were assumed to be 50, with scatters between 25 and 75. The feedback was designed to give a voltage gain of +33. The open-loop voltage gain was × 1140, × 13·9 in stage 1, × 82 in stage 2. The gain-frequency characteristic is shown in Fig. 5 for the amplifier with feedback. Oscillations occurred at 500kc/s, but were prevented by the addition of C (Fig. 1) = 0·01μF. The laboratory work of setting up and

Fig. 7. Drift runs over 15min for two different amplifiers according to Fig. 6



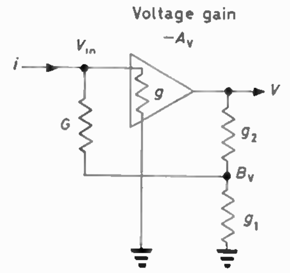


Fig. 8. Schematic diagram of transresistance amplifier

The voltage gain is $-A_v$, input conductance g

A Current Amplifier to Measure 3×10^{-11}A Full Scale[6]

A d.c. amplifier with the correct type of feedback for converting a small current to a useful voltage is shown schematically in Fig. 8. If the output impedance is negligibly small, then a nodal analysis shows immediately that

$$v/i = 1/BG \left[1 - \frac{G + g}{A_v BG} \right] \quad \text{(28)}$$

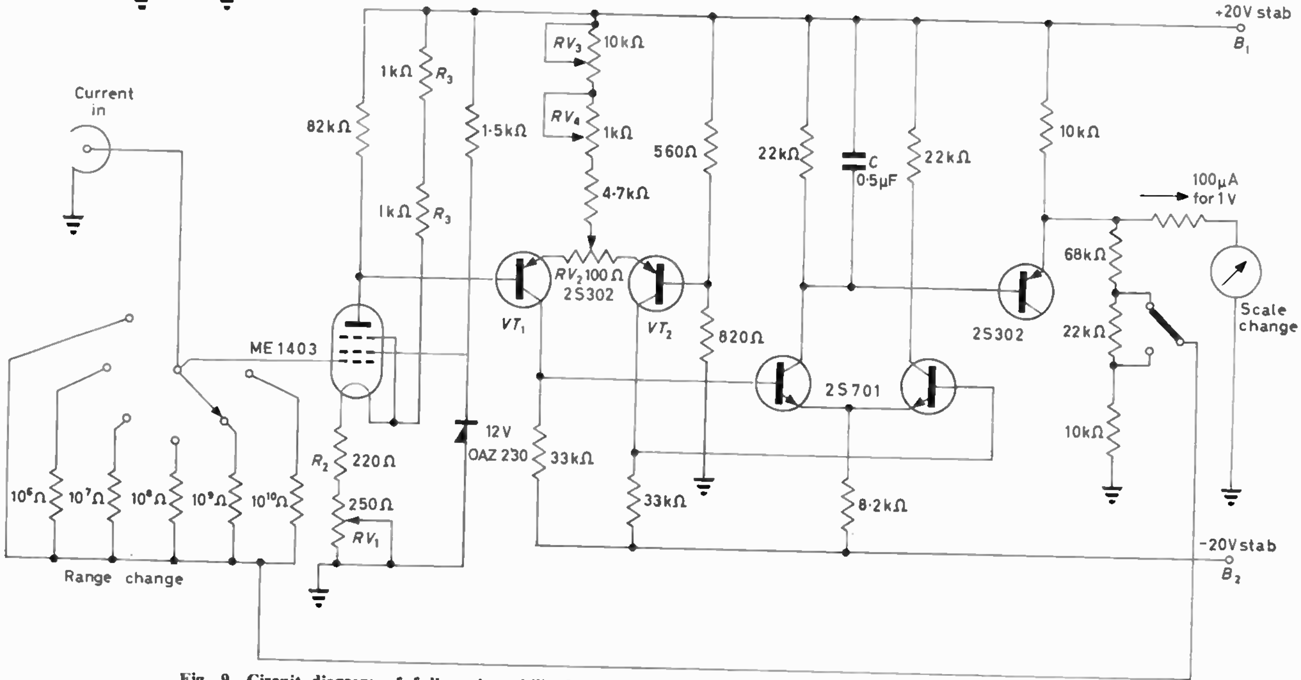


Fig. 9. Circuit diagram of fully gain-stabilized transresistance, amplifier for measuring 3×10^{-11}A full scale

stabilizing took an hour, and the entire process of design, construction and testing 5½ hours.

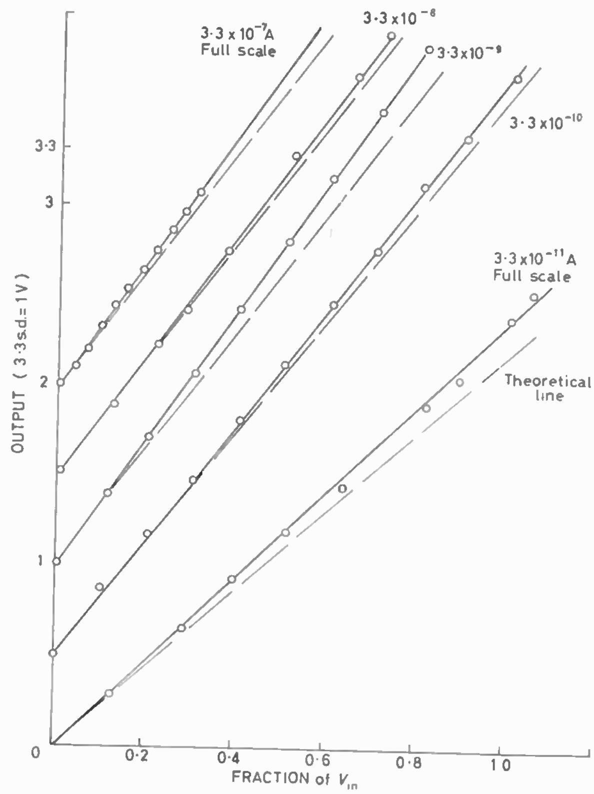
The inadequacies of the paper design were a wrong prediction for $R_2 + RV_3 + RV_4$, and of RV_4; and the limit for V_p/B_1 estimated from equation (16). In designing the second stage the lower limit of VT_5 β ought to have been used. A more elaborate procedure does not seem worth while.

A Low Drift 'Straight' Voltage Amplifier[2,3,4]

Two voltage amplifiers whose circuit is given in Fig. 6 were cascaded to raise a signal of 1mV to 8V. The open loop gain of each of these was greater than 2000, gain with feedback was 90, and no stabilizing capacitor was necessary in the single amplifier. When connected in series and energized from the same stabilized supplies, a 6000pF capacitor was required across R_7. Fig. 7 shows that the drift referred to input over 15min is better than 20μV°C. To achieve this figure, which is low for a non-chopper type amplifier, the input pairs are thermally coupled by mounting the first four transistors in an aluminium block which is in good thermal contact with the transistor cases. In applications where resetting the d.c. zero is undesirable, it is a simple matter to maintain the block at a uniform temperature[5]. Thus for applications such as thermocouple measurements, where drift stabilities of 100μV is all that is required, straight amplifiers of this kind are more than adequate, without going to the high performance and extra complication of chopper amplifiers.

Fig. 10. Response to a small voltage V_{in} in series with 100± 1MΩ resistor

Curves for the different ranges are displaced upwards for clarity. Broken lines are the theoretical curves. Tolerance on feed-in resistors was 5 per cent; tolerance on feedback resistors was also 5 per cent



and the impedance presented to the incoming current is

$$R_{in} = 1/A_vBG\left[1 - \frac{G+g}{A_vBG}\right] \quad \ldots\ldots\ldots\ldots (29)$$

The correction factors in the brackets can easily be made very small, in which case

$$i = -BG_v \quad \ldots\ldots\ldots\ldots (30)$$

$$G_{in} = A_vBG \quad \ldots\ldots\ldots\ldots (31)$$

In these expressions $-A_v$ is the voltage gain of the amplifier, B quite accurately equals $g_2/(g_1+g_2)$, since the path in parallel with g_2 is of small admittance (typically $10^{-14}\Omega^{-1}$ compared with $3 \times 10^{-5}\Omega^{-1}$). The output voltage is governed entirely by the feed-in conductance G and the feedback ratio B. Moreover the input impedance of the amplifier itself does not have to be especially high if the loop gain is large. Suppose $A_vB = 100$, and $g = G$. Then the correction factor in the brackets is 0·98, comparable with the uncertainty in the value of G.

Suppose currents of 10^{-8}A full scale are to be measured (ion currents in run-of-the-mill high vacuum systems for example) then the amplifier could be wholly transistorized. If $v = 0{\cdot}25$V, $B = 0{\cdot}1$, $G = 4 \times 10^{-7}$, then equation (30) gives $i = 10^{-8}$A. For a 2 per cent error $A_vB = 100$ and the input resistance of the amplifier must exceed 2·5MΩ, this is quite feasible[7]. A drift stability referred to input of the voltage amplifier is $0{\cdot}25/10^3$ or 250μV. This presents no problem.

However, if currents of 10^{-10} are to be measured, then, for 1V output $BG = 10^{-10}$. *If* $B = 1/10$, G equals 10^{-9} mho. Therefore the open amplifier input impedance must be thousands of megohms and this calls for an electrometer valve as first stage. Fortunately the d.c. supplies for operating transistors are very suitable for operating an electrometer. Fig. 9 shows the circuit of such a current meter, with full scale sensitivity of 3×10^{-11}A, using an ME1403 pentode as input valve. The latter can have the unnecessarily high input impedance of $10^{15}\Omega$, hence the effect of g in equation (28) is negligible with feed-in resistors of $10^{10}\Omega$. $A_v = 600$, $B = 0{\cdot}33$ or 0·1, giving a loop gain of 200 or 60. The chief inaccuracy inherent in such a circuit is the value of the feed-in conductance, G.

The electrometer is designed to work at an anode and screen voltage of 12V; and the value of anode voltage is set equal to the voltage of VT_2 base via resistance RV_1. The filament is fed essentially with constant current $(B_1/[R_2 + R_3 + RV_1])$. The transistor amplifier is a ring-of-three tailored to suit $\pm$20V lines. RV_2 balances VT_1 and VT_2 as usual, RV_3 and RV_4 set the emitter currents of VT_1 and VT_2 correctly. In operation RV_4 acts as amplifier set zero. Current ranges are changed by switching feed-in conductances from 10^{-6} to 10^{-10} mhos, and a 1:3·3 scale change is available by switching the value of B.

Fig. 10 shows the output voltage in response to a varying input voltage in series with 100MΩ; on the most sensitive range the input impedance is $(10^{10}/200)\Omega = 5 \times 10^7\Omega$. Therefore the abscissae are scaled down by 2/3. Drift stability is good, except on the range giving 10^{-11}A full scale, $\beta = 1/10$, when drifts of 10 per cent of full scale, equivalent to 10^{-12}A, can occur in 10sec. The origin of these is a little uncertain since the voltage level at the input is typically $10^{-11}\text{A} \times 10^{10}\Omega = 100$mV, but the measurements of Fig. 10 were carried out in conditions of poor screening. If the insulation resistance of the input connexions, and the screening were improved, there appears to be no reason why feed-in resistors of $10^{12}\Omega$ might not be used, giving fully gain-stabilized conversion gains of 10^{-13}A/V.

REFERENCES

1. Grigson, C. W. B. On Scanning Electron Diffraction. *J. Electronics and Control. 12,* 209 (1962).
2. Beneteau, P. J. The Design of High Stability D.C. Amplifiers. *Semiconductor Prod. 4,* 35 (1961).
3. Hilbibier, D. F. A D.C. Transistor Differential Amplifier. *I.R.E. Trans. PGCT-8,* 434 (1961).
4. Beneteau, P. J., Murari, E. D.C. Amplifiers using Transistors. *Electronic Engng. 35,* 257 (1963).
5. Rignall, M. W. Constant Temperature Transistor Enclosure. *Electronic Engng. 33,* 452 (1961).
6. Leck, J. H., Austin, W. E. A Multirange Electrometer Amplifier using Variable Feedback. *Electronic Engng. 32,* 1006 (1960).
7. MacNichol, E. F. Negative Impedance Electrometer Amplifiers—Introduction. *Proc. Inst. Radio Engrs. 50,* 1909 (1962).

(To be continued)

A River Radar

The Kelvin Hughes Division of S. Smith & Sons (England) Ltd has introduced a first 3cm transistorized radar designed specifically for vessels navigating the Rhine. Designated the type 17R, this new equipment has already received the approval of the Rhine River Commission, and is now being demonstrated at main centres along the Rhine River. It offers distinct advantages over other radar available at this time.

Consisting of four compact and lightweight units (aerial, transmitter/receiver, display unit and motor alternator) the equipment presents a high definition p.p.i. display on a 9in c.r.t. An optical magnifier can be provided to increase the size of the display to an equivalent of 12in. Range and bearing discrimination and minimum range performance are well above the specification prescribed by the Rhine River Commission. To assist the Master in short range navigation when moving in limited dock spaces it has been found possible to incorporate for the first time a minimum range scale of 300 metres. This provides a particularly open scale for short range navigation. There are in all 8 range scales up to a maximum range of 10km with off-centring facilities on each. Three pulse lengths and two pulse repetition frequencies are included, while picture alignment is automatic.

Except for the cathode-ray tube the display unit is completely transistorized. In the transmitter, too, valves have been kept to a practical minimum. By using transistors, in this manner instead of conventional valves both the size and weight of the radar units are reduced. This is of real benefit in the small wheel houses necessary in river craft. The demand made by the radar on the ship's electrical supply is 350W only; an important factor where the power available is limited.

As with all current Kelvin Hughes radar the aerial supplied is of the slotted waveguide type, 6ft in length. With this highly efficient design disturbing side lobe echoes are eliminated resulting in a picture of exceptional clarity and definition.

A 'rate of turn indicator' can be provided at an extra charge with a repeater which is mounted directly at the radar tube face.

The display unit

A Low Speed Single Supply NOR Logic System

By G. Flanagan* and L. Molyneux*

The NOR logic system described uses one value of resistor, the cheapest transistor of its range and only one power supply. It is slow, maximum pulse rate 5 000/sec, but is suitable for conditions where speed of the operations is governed by electromechanical components, as in data processing.

(Voir page 503 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 510)

A NOR logic system[1-4] has the important virtue of only requiring one type of element to perform all logical functions. The system described is of this kind and has the additional attractions of using one value of resistor, the cheapest silicon transistor of its range and only one power supply. The system is slow, with pulse rates up to 5 000 pulses per second, but it is very suitable for a system containing electromechanical components. For instance it has been used for the punched tape equipment of a computer type setting project[5]. This system comprises a type-

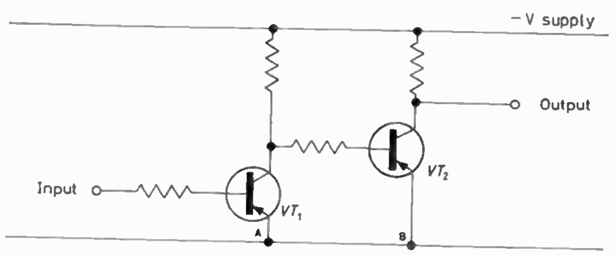


Fig. 1. Two invertors

writer coupled to a tape punch, a tape reader coupled to a typewriter and a tape reader coupled to a 'Monotype' 31-hole tape punch.

Silicon pnp alloy transistors have the important feature, which they share with the more expensive surface barrier transistors, of having a collector saturation voltage considerably less than their working base-emitter voltage. This means that they may be coupled together without need for bias supplies. For example consider Fig. 1.

When transistor VT_1 is 'on' (i.e. saturated, $\beta I_b > I_c$) its collector voltage will be about 100mV negative with respect to the neutral line. As silicon transistors have negligible leakage current, the base of transistor VT_2 will be at this value, and, as no current will flow till the base emitter voltage of a transistor reaches about 0·5V, transistor VT_2 will not draw any collector current, and is said to be 'off'. On the other hand, when VT_1 is off, no current is taken by the collector of VT_1 and the current is therefore available to saturate VT_2. Logically speaking, the two transistors are invertors. For other logical functions it is necessary that a transistor can be maintained 'on' from more than one source, (fan-in greater than 1) and it is preferable that its load resistor should provide enough current to saturate more than one other transistor (fan-out greater than 1). In short, it is desirable that the fan capacity both in and out should be as great as possible, though low fan capacity is acceptable if, as is the case, there is a simple method of increasing it where required.

* *University of Newcastle upon Tyne.*

Design Considerations

Fig. 2 shows typical I_c/V_{ce} and I_c/V_{be} curves for a pnp silicon alloy transistor. It shows the collector current is largely independent of collector emitter voltage for values of V_{ce} greater than about 200mV. Therefore, provided V_{ce}

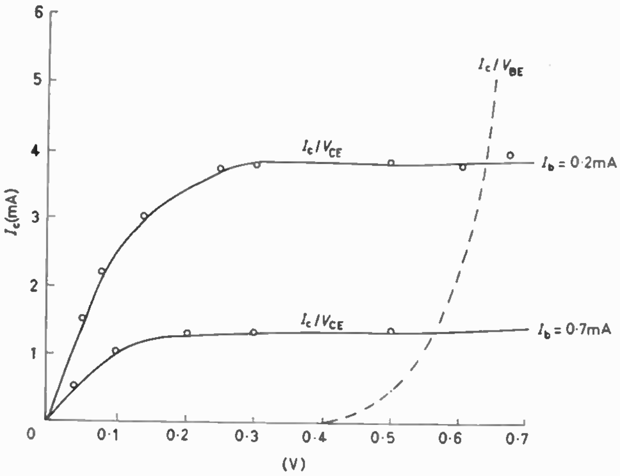


Fig. 2. Input and output characteristics for pnp silicon transistors

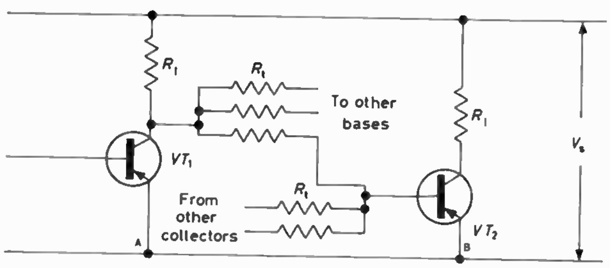


Fig. 3. Circuit used to evaluate worst case design

is below the knee of the I_c/V_{ec} curve, which will be the case if $\beta I_b > I_c$ then the collector will be less negative than the critical base voltage and a tranistor connected to it will not draw current, and will be 'off'.

The essential feature of NOR logic is that when transistor VT_1 (Fig. 3) is 'off', its load resistor (R_1) must be able to provide enough base current to saturate all the transistors connected to the transfer resistors (R_t). In the worst case, all the other inputs to these transistors are connected by other transfer resistors to the collectors of transistors in the saturated condition. It being assumed that this saturated condition occurs at $V_{ce} = 0$ volt. For transistor VT_2 to just reach saturation, $\beta I_b = V_s/R_1$, the current in the collector load which supplies this is $\dfrac{V_s - V_{be}}{R_1 + R_t/M}$ so that the current available at the base connexion of each transistor is $\frac{V_s - V_{be}}{MR_l + R_t}$. However, not all of this current flows in the base emitter circuit since a fraction given by $(N-1)V_{be}/R_t$ is wasted in the other input connexions.

Therefore:

$$I_b = \frac{V_s - V_{be}}{MR_l + R_t} - \frac{(N-1)\,V_{be}}{R_t}$$

where V_s = supply voltage

V_{be} = critical base emitter voltage at which transistor conducts

M = fan-out

N = fan-in

β = current gain

However $I_b = V_s/\beta R_l$

Therefore:

$$V_s/BR_l = \frac{V_s - V_{be}}{MR_l + R_t} - \frac{(N-1)\,V_{be}}{R_t}$$

Or, if x is put equal to R_t/R and V_{be} is neglected with respect to V_s then:

$$M = \frac{1}{(1/\beta) + ((N-1)/x)\,V_{be}/V_s} - x$$

It can be shown that, although values of x of slightly greater than unity increase the fan capacity, the increase is only small so that the simplification of circuit design and construction obtained by making R_t equal to R_l is worthwhile.

The expression then becomes:

$$M = \frac{1}{(1/\beta) + (N-1)\,V_{be}/V_s} - 1$$

The expression suggests an inverse relationship between fan-in and fan-out, i.e. as fan-in is increased fan-out decreases and vice versa. Further, the fan-out for any value of fan-in may be increased by increasing β and/or V_s, which is tantamount to saying that the fan capacity may be increased by increasing either of these quantities.

V_s is limited by the breakdown voltage of the transistor and a figure incorporating a safety factor is usually given by the manufacturer, though it may be convenient to use a lower value.

For a given range of transistor β usually varies with price, transistors with higher β's commanding higher prices.

In a logic system, and particularly in a NOR logic system, it is not uncommon to find a considerable proportion of the elements using only a small fan capacity. As fan capacity can easily be increased where required, the most economic system will be that which uses the least costly transistor, for the basic unit even though the concomitant low β will restrict its fan capacity.

The design problem therefore reduces to working out a safe value for fan capacity for a given transistor. Consider for example the OC200 (Mullard Ltd). The main difficulty is to find the appropriate value of β for design purposes. The relevant β is the d.c. gain when the transistor is just on the verge of saturation, say, $V_{ce} = 200\text{mV}$. Curves for this are not given, but a value can be found by combining the common emitter transfer characteristics with the common emitter output characteristics. It is important that the design should work with all transistors so that the transfer characteristic marked minimum is selected. This gives a minimum low voltage β of 8. Taking this value, Fig. 4 shows a plot of fan-in/fan-out for various supply voltages. V_{be} is taken as 0·5V.

It is usually found that fan-in capacity is more useful than fan-out, so that an asymmetrical system is preferred. For a minimum supply voltage of 8V a fan-in of 3 and fan-out of 2 is the chosen design. This gives considerable margin of safety in the worst case, which would be an unlikely combination of low β transistors with fully committed fan capacity. The margin can be improved by increasing the minimum supply voltage, 10V is a suitable value.

In a 'worst case' condition at this voltage, (i.e. two logic units each with two inputs connected to saturated elements which must be maintained 'on' by the output current from the same 'off' unit) each of the transistors will have twice the minimum current required for saturation. At 0°C where the β is reduced to 7·5 and V_{be} rises to 0·54V, each transistor will then have 1·8 times the minimum current.

At higher temperatures, say, 80°C, β increases and the safety factor on base current increases to 2·5. This increase in safety factor on base current does not, however, completely describe the situation. As temperature increases the

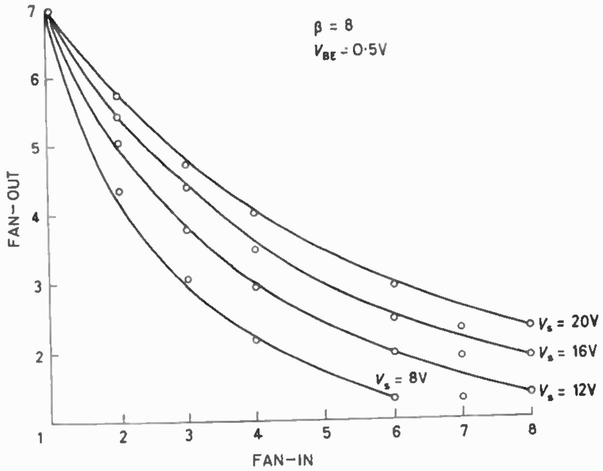


Fig. 4. Fan-in/fan-oot relationship based on worst case conditions

working base-emitter voltage decreases at a rate of 2mV/°C and the collector saturation voltage increases slightly, thus narrowing the gap between them. Eventually at high temperatures, some way above 100°C, the two voltages become equal and a bias network becomes necessary. However, in all ordinary situations there is an adequate margin.

The design is completed by choosing a suitable value for load and transfer resistors, which have the same value. The actual value does not come into the design equation and is chosen mainly to give an adequate β. A value of 10kΩ is a suitable choice for both the OC200 (Mullard Ltd) and the 2S322 (Texas Instruments Ltd).

Lower values than 10kΩ would give higher currents, and in the case of the 2S322 higher β. But on the other hand 10kΩ makes any calculations concerned with coupling the logic to other circuits easy, and mainly for this reason is recommended.

Increasing the Fan Capacity

Fan-in may be increased by connecting in parallel the collectors of a number of transistors and providing only one supply resistor, the fan capacity is then increased by 3 for each transistor added.

Since the leakage current of silicon transistors is low, a large number of transistors may be connected in this way. The maximum leakage current of an OC200 transistor is stated to be $3\mu A$ at 80°C, 10 transistors connected in parallel, all with this leakage current would degrade the safety factor from 2·5 to 2·3 at 80°C.

The added collector capacitance would, of course, reduce the ultimate pulse rate, but the figure of 5000/sec is a very cautious estimate and is not affected. More than 10 transistors could be connected in parallel and would imply a fan-in of more than 30, a limit of 10 parallel transistors therefore does not impose serious limitations, and is suggested. Fan-out may be increased by connecting a transistor as an emitter-follower to the output from a NOR logic unit. The fan-out capacity is then multiplied by approximately the β of the emitter follower. However, the voltage difference between the base and emitter restricts the increase in fan capacity to something less than this and for this reason a factor of 5 is suggested. Thus a NOR element connected directly to an emitter-follower will have a fan-out of 10, while an element with one ordinary load and an emitter-follower will have a total fan-out of 6.

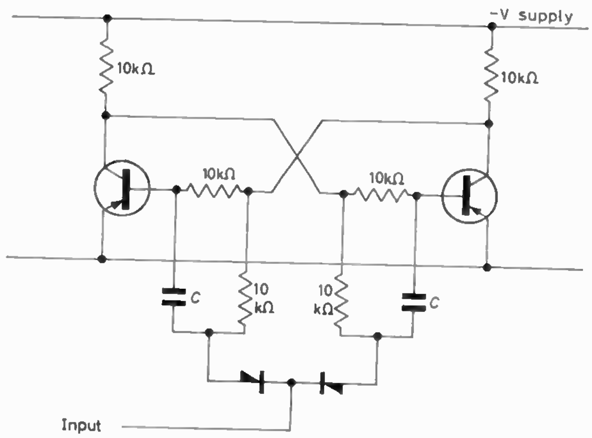


Fig. 5. Scale of two circuit

Neither of these methods for increased fan capacity increases the variety of components required, and greater fan capacity can be created by logical connexions of groups of NOR elements.

Additional Circuits

So far only the logic elements have been described as the scheme is intended for use by those with a knowledge of electronic, as well as logical design. There are, however, two circuits of wide application that unlike input and output circuits, do not depend on external requirements and can reasonably be considered as part of a logic system. Both of these use additional components. A scale of two circuit (Fig. 5) can be formed from a pair of NOR units and a diode pulse routing network.

The minimum value of C is governed by the need to feed sufficient charge into the off transistor to turn it on. Experiments show $0{\cdot}002\mu F$ to be a minimum value. This value allows most units to work at rates in excess of 10kc/s. The penalty for increasing it is loss of speed due to the time-constants involved, though large values are desirable as they ensure that the circuit will work on slowly rising waveforms.

A value of $0{\cdot}01\mu F$ is suggested for general use and this value allows the counter to work at rates in excess of 1kp/s.

As one load is already connected to each collector, the usable fan-out is only 1, but this can easily be increased by the use of an emitter-follower. The circuit is capable of driving other similar stages, and complete counter chains may be constructed. The circuit is not sensitive to pulse length, provided that it exceeds about $20\mu sec$; it must, however, be fed with pulses of about 0·5V supply amplitude. This is arranged automatically if the circuits feed each other as in a counter chain, or the circuit may be fed from a NOR element which has one other load.

In the time delay circuit (Fig. 6) base current must flow through R_t to saturate VT_1. VT_2 is therefore off. If VT_3 is turned on, the base of VT_1 is taken positive and VT_1 turns off, thus maintaining VT_2 on. The base of VT_1 however returns to the critical value in a time given by $t \simeq 0{\cdot}7CR_t$. Therefore a pulse of $0{\cdot}7CR_t$sec is available at the collector of VT_1 which can drive one NOR unit. The trailing edge of the pulse is thus delayed with respect to the leading edge. The pulse width at VT_2 may have any value in excess of about $20\mu sec$.

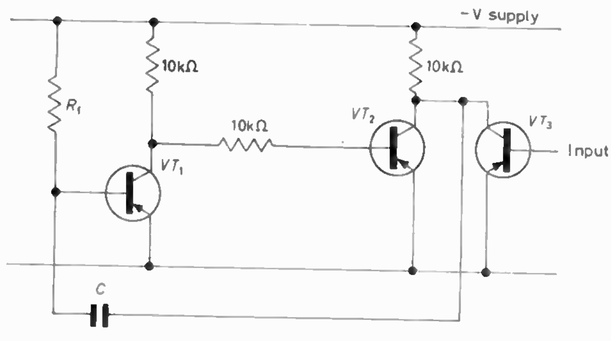


Fig. 6. Time delay circuit

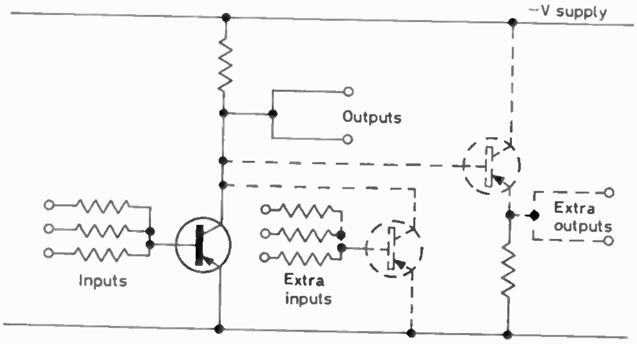


Fig. 7. Basic NOR logic unit (full lines) with methods of increasing fan capacity (dotted lines)

Complete Specification

Fig. 7 shows the basic unit together with the methods of increasing fan capacity.

Transistor	OC200 (Mullard Ltd) 2S322, 2S302 (Texas Instrument Ltd)
Resistor	10kΩ ½W 10% tolerance
Fan-in	3 (up to 30 with parallel transistors)
Fan-out	2 (up to 10 with emitter-follower output)

Supply voltage: min. 10V, max. 20V

Max. pulse rate 5000/sec

Temperature range 0° to 80°C.

Input: Logical '1' $\geqslant \frac{1}{3}$V supply.
Logical '0' $\leqslant$ 200mV.

Conclusion

The system is inexpensive but limited to low speed applications. It is primarily intended for the designer who needs some logical functions as part of a larger circuit design.

Acknowledgment

The work was done as part of a D.S.I.R. project on 'Investigation of Computer Program Writing and Proving for Automatic Type Composition, in English' and the authors gratefully acknowledge financial assistance from this source.

REFERENCES

1. Rowe, W. D. Transistor NOR Circuit Design. *I.R.E. Wescon Conv. Rec.*, p. 231, pt. 4 (1957).
2. Marcovitz, M. W., Seif, E. Analytical Design of Resistor-Coupled Transistor Logic Circuits. *I.R.E. Trans. EC7*, 109 (1958).
3. Kellett, P. The Elliott-Sheffer Stroke Static Switching System. *Electronic Engng. 32*, 534 (1960).
4. Nichols, K. G. Pulse counting and Scaling by Sheffer Stroke Logic. *Electronic Engng. 35*, 296 (1963).
5. Duncan, C. J. *et al.* Computer Type Setting: an Evaluation of the Problem. *Printing Technol.* (in press).

University of London Computing Service

The University of London's £2M Atlas computer, which was formally inaugurated on 26 May, will be used by two newly created University organizations.

The new Institute of Computer Science, promoted on 25 March from the seven-year-old University Computer Unit, will carry out a programme of research covering all aspects of the design and use of computers, will provide courses of lectures on the theory and use of computers for internal and external students, and will give a service to all University departments needing the use of a large computer in their own research. The creation of the Institute represents a full recognition of the importance of computer research and of the use of computers in many departments of University work.

The University of London Atlas Computing Service is a completely new type of enterprise. Operated by a private company set up and owned by the University, with its own staff of experienced consultants, analysts and programmers, a full range of computing services will be provided on a commercial basis for industry, commerce and government. For the user, new techniques developed in the University departments will be made more readily available; for the University, a substantial contribution will be provided towards the cost of the Atlas. For both, there will be the opportunity to benefit from an interchange of knowledge and experience.

One very large industrial user of Atlas will be the British Petroleum Company, which gave the University essential support at the start of the Atlas project by agreeing to buy in advance a substantial amount of computer time over the first five years of operation.

The I.C.T. Atlas is one of the most powerful computer systems in the world; it represents in many ways a completely new approach to computer design and use. Apart from the speed and power of the central computer, the most significant feature is the internal arrangement by which the system schedules and organizes its own work; a number of quite separate jobs can all be in process of operation at the same time, each using only those parts of the whole system which it needs. By this means even the smallest problems can be solved economically by this very large installation, while the whole system can be turned over when needed to an individual, large and intractable computation.

The Atlas will be connected by high-speed data links, due to be installed during this summer, with several of the London Colleges and possibly also with some of the larger users of the service.

Brief details of the installation are as follows:

Data word

48 bits or 8 characters of 6 bits.

Storage

Main store, cores 32 768 words, magnetic drums 98 304 words; fixed store 8 192 words; subsidiary working store 1 024 words.

Magnetic tape

This installation will have 13 to 15 Ampex tape units.

Speed

Atlas I can execute on the average 400 000 basic instructions per second; about 60 times more than the Mercury computer. (The last five years' work on the Mercury could be repeated in 3 weeks on Atlas.)

Input and output

Input may be by 5, 7 or 8 channel paper tape (4 tape readers provided) or 80 column punched cards (2 card readers). Output may be on line printer (2 Anelex printers), paper tape (3 tape punches) or punched cards (1 card punch). Input and output will also be possible via data links to other locations, e.g. a fast link to Imperial College, slower links to Queen Mary College, King's College, University College and probably other centres.

Other peripheral units

Three teleprinters provide information to engineers and operators.

Parallel operation

Under the action of the 'Supervisor' (a program stored permanently in the machine) all input and output devices can be in action simultaneously, and up to 8 magnetic tape units. The Supervisor assembles programs, controls the order of their execution, provides monitoring information to the operators and all essential logging data.

Programming languages

The following languages may be accepted: Mercury Autocode (CHLF 3); Extended Mercury Autocode (EMA); Algol 60; Fortran; Combined Programming Language (CPL); Lunacode; Atlas Commercial Language (ACL); Atlas Basic Language (ABL).

Power and cooling

The Atlas installation has a loading of up to 220kVA. and the cooling system up to 320kVA. Units in the basement are water-cooled, the total capacity of the cooling system being 90 tons of refrigeration. Excess heat is dispersed through the cooling tower.

The operating area on the upper floor showing on the left paper tape readers and punches, magnetic tape units (at rear), main control position, high-speed printers, and on the right card readers and punches

Extraction of Luminance from the SECAM Composite Signal

By G. Melchior* and J. P. Doury*

For certain purposes it is necessary to extract the luminance signal from the composite SECAM signal. In this article a method is described by means of which this can be accomplished without serious loss of definition.

(Voir page 503 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 510)

IN receivers equipped with single-gun tubes, studio equipment (mixers—special effects) and for conversion of colour signals to black-and-white signals, it is necessary to extract the luminous signal from the composite SECAM signal.

Known Solutions

The common solution consists in extracting luminance by filtering the composite signal by inserting a trap for the chromatic channel. But this trap has to be fairly wide in order to be effective in spite of sub-carrier modulation, and in practice a low-pass filter has to be used, whose cut-off frequency cannot exceed 3·5Mc/s or so if colour signals are to be effectively suppressed. This causes a marked loss of definition.

The resulting blur can be avoided by 'crispening', but this cleaning up of definition is only partial and is incapable for instance, of reproducing a high definition test pattern.

The method described below avoids this criticism.

Principle

The principle consists in removing the colour signal from the composite signal, the colour signal constituting frequency modulation and not amplitude modulation of the sub-carrier.

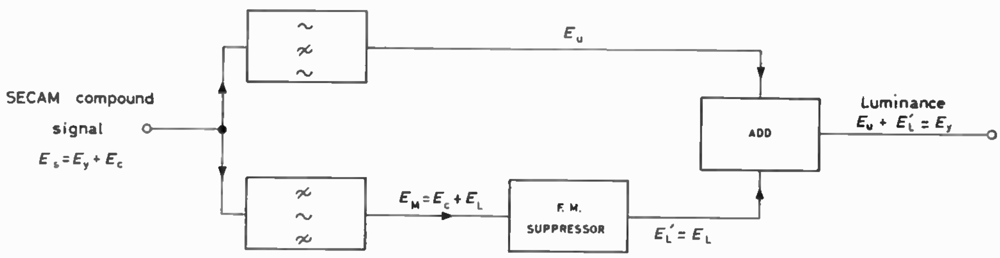


Fig. 1. Luminance extraction

Removal of Frequency-Modulated Signals

The principal element of the arrangement is a frequency-modulated signal suppressor whose mode of action can be defined as follows.

The 'compound signal' which occupies the frequency band of the chromatic channel consists of the sum of the modulated sub-carrier and of the luminance components included in this channel.

The modulated sub-carrier is expressed by:

$$E_C = A_0 \cos\left[\omega_0 t + \theta(t)\right]$$

where ω_0 is the undeviated frequency, and $\theta(t)$ is the phase modulation resulting from the frequency modulation by chrominance.

The compound signal can be expressed in the form of an oscillation modulated both in amplitude and in phase:

$$E_M = \left[A_0 + A(t)\right] \cos\left[\omega_0 t = \theta(t) + P(t)\right]$$

where: $A(t)$ is the amplitude modulation, and $P(t)$ is the phase modulation impressed on it by the luminance components of the compound signal.

The luminance components E_L contained in the compound signal could be obtained by subtracting the colour signal E_c from it.

This rigorous extraction cannot be carried out in practice. Although it is impossible to remove precisely E_c it is possible to extract a somewhat similar signal constituted by the sub-carrier of amplitude A_0, subjected to the spurious phase modulation $P(t)$ due to luminance, in addition to its phase modulation due to chrominance $\theta(t)$.

Then, removing

$$E_C' = A_0 \cos\left[\omega_0 t + \theta(t) + P(t)\right]$$

the remaining signal

$$E_L' = E_M = E_C'$$

is a satisfactory approximation of the luminance components.

The calculations for this approximation are given in the appendix.

Description of the Luminance Extractor

Fig. 1 shows the circuit arrangements for complete luminance extraction from a SECAM composite video signal.

The video composite signal E_s, equal to the sum of the luminance signal E_Y and the colour signal E_c, is separated by means of two complementary filters into:

(1) A 'univocal signal' E_u containing no chromatic components.

(2) A 'compound signal' E_M containing the colour signal as well as the luminance components in the chromatic channel.

The colour signal is removed from the compound signal by a 'frequency-modulated signal suppressor' whose action is as described below.

This gives a signal E_L' which is added to E_u to produce the final signal.

N.B. Strictly speaking, it should be noted that in addition it is necessary:

(a) To cover either the complete system, or the frequency-modulated signal suppressor, by a 'bell-shaped' filter (or r.f. de-emphasis) and an anti-bellshape' filter (or r.f. pre-emphasis).

(b) To equalize transit times between the signal E_L' and the signal E_u by a delay inserted in the channel of the E_u signal.

* *Compagnie Française de Television.*

The Frequency Modulation Suppressor

A suppressor of frequency-modulated signals operating on the principle described above can be produced in different ways: The two solutions which appear to be the simplest are given below: suppression by subtraction, and suppression by transposition.

Suppression by Subtraction

The signal E_c' which has to be subtracted from the compound signal E_M can be obtained by limiting the compound signal. So the principle of an extractor operating by subtraction (Fig. 2) is simple.

The signal E_c' obtained from a limiter is subtracted from the signal E_M, this limiter acting on the signal E_M. The output level of this limiter is automatically adapted to the level of the colour sub-carrier through an automatic control operating on the limitation threshold and fed from an amplitude comparator. The amplitude comparator is intentionally slow-acting in order not to respond to modulation caused by luminance.

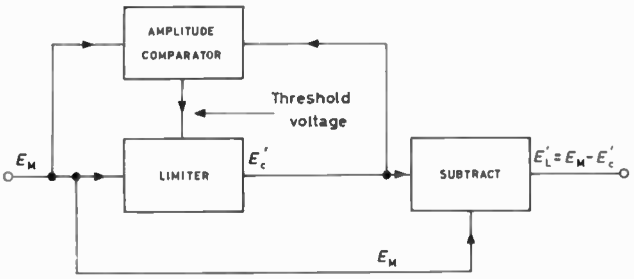


Fig. 2. F.M. suppression by subtraction

Suppression by Transposition

By this method the signal E_c' is suppressed by a highly selective notch filter. This is made possible by the use of 'auto-transposition', i.e. transposition of the signal E_M by its own frequency so that it is converted to a fixed frequency carrier modulated in amplitude. Fig. 3 shows the principle of this arrangement.

The signal passes through a 'linear channel' with two successive frequency transpositions M_1 and M_2. These two transpositions are effected by means of a common heterodyne signal E_H and are of opposite sense so that the output signal would be in accord with the input signal if no filter were inserted between them.

The compound signal is applied to a limiter which delivers a signal proportional to E_c', i.e. of constant amplitude and of the same frequency as E_M. This limited signal is then transposed to M_3 by a fixed frequency F_1. This produces the signal E_H which follows exactly the same frequency modulation as the signal E_M.

The transposition to M_1 of the compound signal by the signal E_H thus provides a fixed frequency signal F_1 which, however, has retained the amplitude modulation of E_M.

The notch filter adjusted to F_1 suppresses the carrier of frequency F_1 and the final transposition M_2 resets the useful retained components to their original frequency.

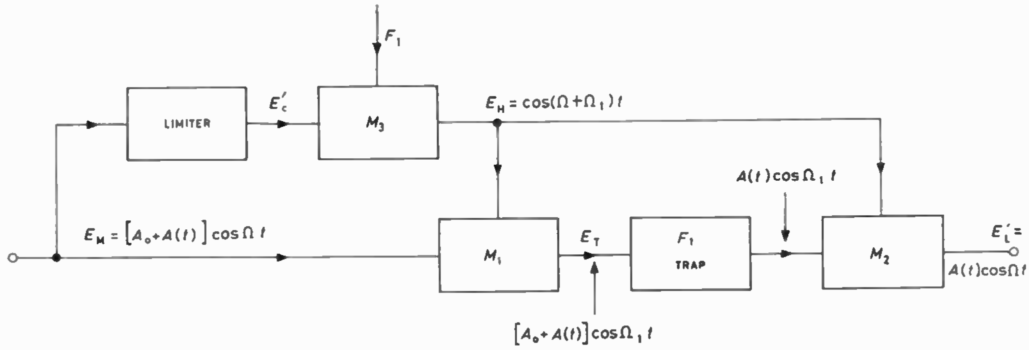


Fig. 3. Suppression f.m. by transposition

Appendix

Calculation of the Difference $E_L' - E_L$

The compound signal E_M is composed of the sum of the colour signal E_c and of the luminance components E_L. Their rigorous extraction would require subtracting the colour signal from the compound signal.

$$E_L = E_M - E_c$$

The operation actually carried out by the suppressor is:

$$E_L' = E_M - E_c'$$

that is to say that instead of subtracting E_c, an oscillation E_c' of the same frequency as E_M and of constant amplitude is subtracted.

This approximation will be calculated.

Writing for simplification:

$$Z = \omega_0 t + \theta(t)$$

the expression for the colour signal is:

$$E_c = A_0 \cos Z$$

The components of the luminance signal included in the compound signal can be expressed by the integral:

$$E_L = (1/\pi) \int_{\omega_1}^{\omega_0} A(\omega) \cos \left[\omega t + \phi(\omega)\right] d\omega$$

writing:

$$\omega t + \phi(\omega) = Z - z$$

and noting that $\cos[\omega t + \phi(\omega)]$ is then equal to:

$$\cos z \cos Z + \sin z \sin Z$$

The compound signal is expressed by:

$$E_M = E_L + E_c = A_0 \cos Z +$$
$$(1/\pi) \cos Z \int A \cos z \, dz + (1/\pi) \sin Z \int A \sin z \, dz$$

Assuming $1/\pi \int A \sin z \, dz$ is small compared to A_0, development of a constant amplitude oscillation with the same phase modulation as E_M would give:

$$E_c' = A_0 \cos Z + (1/\pi) \sin Z \int A \sin z \, dz + \ldots.$$

Considering only terms of the first order, then:

$$E_L' = E_M - E_c' = (1/\pi) \cos Z \int A \cos z \, dz =$$
$$(1/2\pi) \int A \cos (Z - z) \, dz + (1/2\pi) \int A \cos (Z + z) \, dz$$

The first term is equal to E_L except for the 1/2 factor, but this is only a matter of adjustment of the relative gain between the two channels when this signal is finally added to the 'univocal signal' E_u.

The second term shows that spurious components are generated whose frequencies are symmetrical with the useful components about the frequency of the colour signal.

Experience has shown that these spurious components are not troublesome if the pass-bands of the compound signal are limited to the intermediate frequency range of the colour signal (4·43 ± 0·5Mc/s).

An 85Mc/s Transistorized Transmitter

By A. E. Hilling*, B.Sc.

An 85Mc/s transmitter using the AFY19 alloy-diffused germanium transistor is described. This transmitter produces a peak envelope power of 1W when 100 per cent amplitude modulated. This circuit would be suitable for low power portable transmitters in the 71·5 to 88Mc/s band.

(Voir page 503 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 510)

As the third article on the design of transistorized transmitters[1,2] an example working in the 80Mc/s band is chosen. The advantages of transistorizing portable equipment are obvious and there appears to be interest in this band in low power a.m. telephony transmitters.

A transmitter is designed giving 1W r.m.s. output at 85Mc/s and is capable of being amplitude modulated up to 100 per cent. The transmitter (Fig. 1) consists of five stages, a 21·25Mc/s oscillator, two doubler stages to bring the frequency up to 85Mc/s, a driver stage and an output stage.

The AFY19 alloy-diffused germanium transistor is used in the last three stages, the output stage employing two in parallel. Small signal y parameters are used to predict the input admittances of the amplifier stages. The use of these small signal parameters cannot be justified theoretically but owing to the lack of a universal, easily-used high frequency equivalent circuit, they are employed to give an approximate idea of the circuit components needed.

Interstage π coupling networks are used throughout, inductive coupling being avoided due to increased losses and difficulty of specification and adjustment encountered at this frequency.

The supply voltage for this transmitter is 12V, usual in portable applications.

General Details

Driver and Output Stages

Two AFY19 transistors are used in the output stage in parallel to avoid the phase changing networks necessary in push-pull working. One AFY19 transistor is used in the driver stage. The common emitter configuration is used for both stages, the operating conditions being nominally class-B. Collector modulation is applied to both stages; further details are given in the modulation section. The full design of these stages is considered later.

Multiplier Stages

These consist of two doubler stages, the first using an OC170 transistor, the second an AFY19 transistor. The two stages are similar in outline, the transistors being self-biased into class-C and the output tuned to the second harmonic of the input frequency. Each output network contains a series tuned circuit (Fig. 2) which is resonant at the input frequency. If the first shunt arm of the π network has a reactance X_A at the second harmonic frequency, then for the arm to be series resonant at the fundamental frequency, it must be composed of an inductance and a capacitance in series.

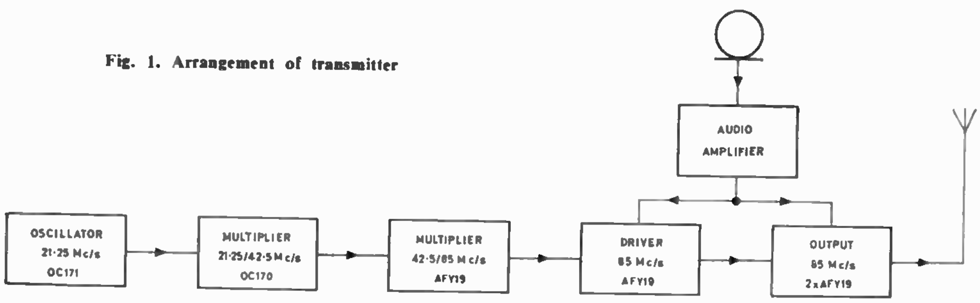
Where:

$$L_A = \frac{X_A}{(1-n^2)\omega_1}$$

$$C_A = \frac{(1-n^2)}{n^2\omega_1 X_A}$$

ω_1 is the angular velocity at which the arm is to have a reactance X_A.

Fig. 1. Arrangement of transmitter



$$n = \frac{\text{Frequency at which arm is series resonant}}{\text{Frequency at which arm has reactance } X_A}$$

i.e. $n = 0·5$ for a doubler stage.

A more detailed examination of this type of circuit is given by Everitt[3].

The theoretical design of transistor multiplier stages is difficult; the best operating conditions in these multipliers are determined experimentally. The two stages have a total gain of 15dB, the current taken by the first and second stages being 4mA and 30mA respectively. The efficiency of fundamental frequency suppression by the coupling networks can be appreciated by examining the transmitter output waveform. All frequencies below 85Mc/s. i.e. 21·25Mc/s, 42·5Mc/s, etc. are at least 50dB down on the carrier frequency.

Oscillator Stage

The oscillator is of the Colpitts type with a third overtone 21·25Mc/s crystal in the collector emitter feedback loop. An output power of about 0·5mW is obtained from the OC171 transistor, the oscillator being loosely coupled to the first multiplier stage. The transistor and bias networks together take a current of 2mA.

Design of Driver and Output Stages

Load Requirements

The r.f. output power required from the output stage is 1W when working from a 12V supply. Allowing for knee voltage of 2V, a peak collector voltage swing (V_{CM}) of 10V is available. The knee voltage at r.f. is usually higher than the d.c. value.

* *Formerly Mullard Ltd.*

The output power from two transistors in parallel is given by:

$$P_{out} = 2\,.\,(V_{CM}/\sqrt{2})\,.\,\frac{(0{\cdot}5\,I_{CM})}{\sqrt{2}}$$

where I_{CM} is the peak collector current per transistor.

i.e.:

$$I_{CM} = 200\text{mA}$$

The d.c. taken by the output stage is, therefore, $\dfrac{2 \times 200\text{mA}}{\pi}$

$$I_{dc} = 128\text{mA}$$

The load which must be presented to the collectors of the paralleled transistors is then:

$$R_T = \tfrac{1}{2}\,.\,\frac{V_{CM}}{(0{\cdot}5\,I_{CM})}$$

$$R_T = 50\Omega$$

The load per transistor (R_L) is, therefore, 100Ω

i.e. the load admittance/transistor (G_L) is 10mmho.

A gain of 9dB is typical for the AFY19 at this frequency and power. The input power required by the output stage is, therefore, 120mW. This figure is approximate owing to the losses incurred by the coupling networks.

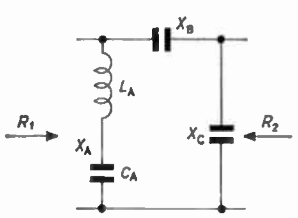


Fig. 2. Multiplier stage output network

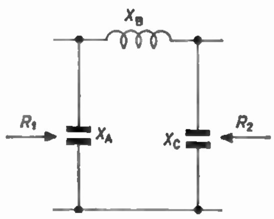


Fig. 3. π matching network

The driver stage is required to supply approximately 120mW of r.f. power with a supply voltage of 12V.

Allowing for a knee voltage of 2V a peak collector voltage swing (V_{CM}) of 10V is available.

$$\therefore\ P_{out} = (V_{CM}/\sqrt{2})\,.\,\frac{(0{\cdot}5\,I_{CM})}{\sqrt{2}}$$

where I_{CM} is the peak collector current

i.e. $I_{CM} = 48\text{mA}$

The d.c. taken by the driver stage is, therefore, $48/\pi$ mA

$$I_{dc} = 15\text{mA}$$

The load which must be presented to the collector of the driver transistor is then:

$$R_L = \frac{V_{CM}}{(0{\cdot}5\,I_{CM})}$$

$$R_L = 420\Omega$$

The load admittance G_L is then 2·4mmho.

A gain of 9dB is typical for the AFY19 under these operating conditions, the driver stage, therefore, requiring approximately 15mW from the second multiplier stage.

Transistor *y* Parameters

The measured values of *y* parameters of the AFY19 at 85Mc/s are:

Output Stage: ($V_{CE} = -6$V; $I_E = 100$mA)

$$y_{ie} = 80\text{mmho}$$
$$y_{fe} = -240\text{j mmho}$$
$$y_{oe} = 1{\cdot}2 + 8\text{j mmho}$$
$$y_{re} = -6{\cdot}3\text{j mmho}$$

Driver Stage: ($V_{CE} = -6$V; $I_E = 20$mA)

$$y_{ie} = 80\text{mmho}$$
$$y_{fe} = -236\text{j mmho}$$
$$y_{oe} = 4 + 12\text{j mmho}$$
$$y_{re} = -5{\cdot}7\text{j mmho}$$

It may be shown that the input admittance of a four-pole network is given by:

$$Y_{in} \simeq y_{ie} - \frac{y_{fe}\,.\,y_{re}}{g_{oe} + G_L} \quad (\text{If } b_{oe} \simeq b_{out} \simeq -B_L)$$

The input admittance of each transistor in the output stage is then:

$$Y_{in} \simeq 80 - \frac{(-240\text{j})\,(-6{\cdot}3\text{j})}{1{\cdot}2 + 10}\ \text{mmho}$$

where $G_L = 10$ mmho

$$\therefore\ Y_{in} \simeq 215\ \text{mmho}$$

The input admittance of the output stage (i.e. two paralleled transistors) is, therefore, 430 mmho.

i.e. the input resistance $R_{in} \simeq 2{\cdot}3\Omega$

The input admittance of the driver stage is then:

$$Y_{in} \simeq 80 - \frac{(-236\text{j})\,(-5{\cdot}7\text{j})}{4 + 2{\cdot}4}\ \text{mmho}$$

where $G_L = 2{\cdot}4$ mmho

$$Y_{in} \simeq 290\ \text{mmho}$$
$$R_{in} \simeq 4\Omega$$

Output Network

The output stage is required to deliver the 1W of r.f. power into a 50Ω load via a frequency selective coupling network. A π network, which also acts as a low-pass filter is used. The relevant formulae are quoted below; a detailed analysis of this type of network is given by Everitt[3].

Referring to Fig. 3:

$$X_A = \frac{-R_1X_B}{R_1 \pm \sqrt{(R_1R_2 + X_B^2)}}$$

$$|X_B| \leqslant \sqrt{(R_1R_2)}$$

$$X_C = \frac{-R_2X_B}{R_2 \pm \sqrt{(R_1R_2 - X_B^2)}}$$

In this case $R_2 = 50\Omega$ and $R_1 = R_T = 50\Omega$

$$\therefore\ |X_{B(max)}| = 50\Omega$$

Choosing $X_B = 32\Omega$, i.e. an inductance of 0·06μH. A coil with an unloaded Q, Q_o, of 200 was constructed.

$$X_A = X_C = \frac{-50 \times 32}{50 \pm 38{\cdot}5}\ \Omega$$

Taking the positive sign in the denominator gives a convenient value.

$$X_A = X_C = -18\Omega$$

i.e $C_A = C_C = 105$pF

But C_A is shunted by the output capacitance of the two paralleled transistors:

where C_{out} per transistor $\simeq b_{oe}/\omega$

$$\therefore\ \text{Total output capacitance} \simeq \frac{2 \times 8 \times 10^{-3} \times 10^{12}}{535 \times 10^6}\ \text{pF}$$

$$\simeq 30\text{pF}$$

$$\because\ C_A\ (\text{effective}) = 105 - 30\text{pF}$$
$$= 75\text{pF}$$

The theoretical values of the output network are then:

$$L_B = 0{\cdot}06\mu\text{H},\ C_A = 75\text{pF},\ C_C = 105\text{pF}$$

The values found by experimental adjustment were:

$$L_B = 0{\cdot}06\mu H, \; C_A = 80pF, \; C_C = 90pF$$

Interstage Network

The input resistance of the output stage ($R_{in} = 2{\cdot}3\Omega$) must be transformed to become the load resistance required by the driver stage ($R_L = 420\Omega$).

A π network is again used here but an additional series matching capacitor (C_M) must be used, the transformation ratio being too large to be handled easily by the π network alone.

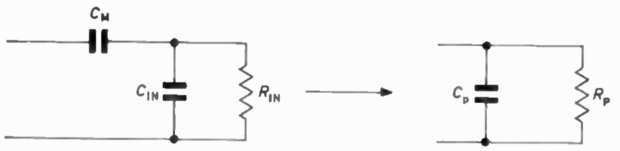


Fig. 4. Series to parallel network conversion

But C_A is shunted by the output capacitance of the driver stage, where

$$C_{out} \simeq b_{oe}/\omega$$

$$C_{out} \simeq \frac{12 \times 10^{-3} \times 10^{12}}{535 \times 10^6} \text{ pF}$$

Output capacitance $\simeq$ 23pF

$$\therefore \; C_A \text{ (effective)} = 90 - 23pF = 67pF$$

Also C_C is shunted by C_P

$$\therefore \; C_C \text{ (effective)} = 107 - 68pF = 39pF$$

The theoretical values of the interstage network are then:

$$C_M = 68pF; \; L_B = 0{\cdot}07\mu H; \; C_A = 67pF; \; C_C = 39pF$$

The values found by experimental adjustment were:

$$C_M = 68pF; \; L_B = 0{\cdot}07\mu H; \; C_A = 50pF; \; C_C = 40pF$$

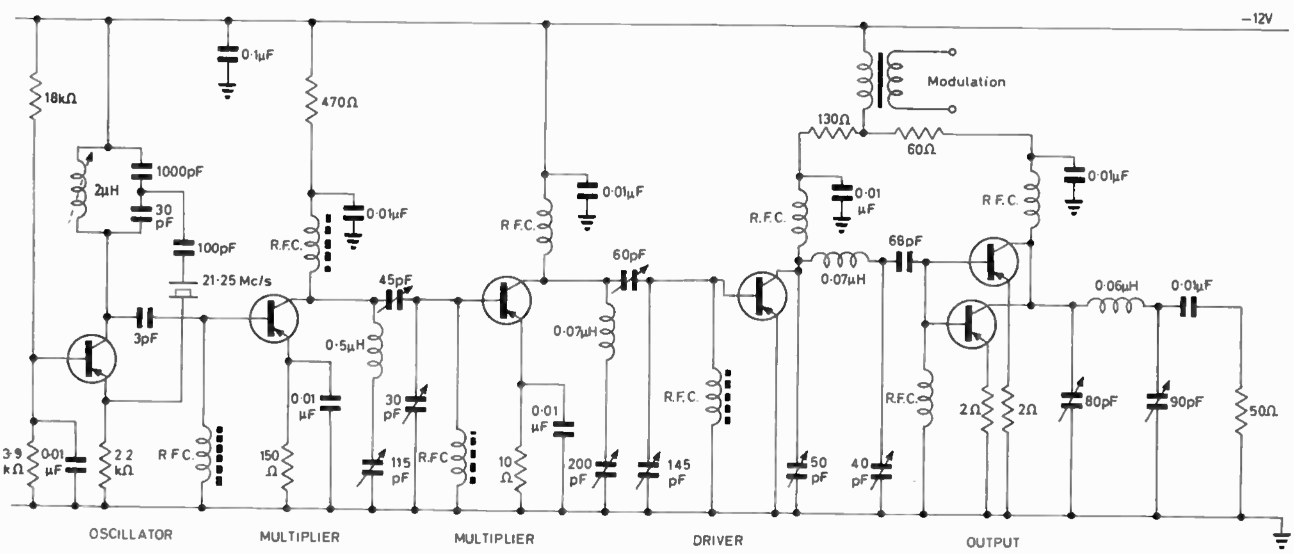


Fig. 5. The complete transmitter

Referring to Fig. 4, it may be shown that:

$$R_P = \frac{1}{\omega^2 C_M{}^2 R_{in}} + \frac{R_{in}(C_M + C_{in})^2}{C_M{}^2}$$

$$C_P = \frac{C_M + \omega^2 C_{in} \,.\, C_M \,.\, R_{in}{}^2 (C_M + C_{in})}{1 + \omega^2 R_{in}{}^2 (C_M + C_{in})^2}$$

Now $R_{in} = 2{\cdot}3\Omega$ and $C_{in} = 0$

Choosing $C_M = 68pF$

then $R_P = 330\Omega$

$C_P = 68pF$

Using the formulae given in the output network design, where:

$$R_1 = R_L = 420\Omega, \; R_2 = R_P = 330\Omega$$

$$|X_{B(max)}| = 370\Omega$$

Choosing $X_B = 37{\cdot}5\Omega$, i.e. an inductance of $0{\cdot}07\mu H$. A coil with a Q_o of 200 was constructed.

$$X_A = \frac{-420 \times 37{\cdot}5}{420 \pm 370}\,\Omega \quad X_C = \frac{-330 \times 37{\cdot}5}{330 \pm 370}\,\Omega$$

Taking the positive sign in the denominator

$$X_A = -21\Omega \qquad X_C = -17{\cdot}5\Omega$$

$$C_A = 90pF \qquad C_C = 107pF$$

Thermal Design

The AFY19 has a maximum collector junction temperature ($T_{j(max)}$) of 90°C and a thermal resistance junction to case (θ_m) of 35°C/W.

Assuming an efficiency of 50 per cent in the output stage, each transistor must be capable of dissipating 500mW(P_C).

The thermal resistance (θ_h) of the heatsink required when operating the transmitter in an ambient temperature (T_{amb}) of 45°C is then:

$$\theta_h \simeq \frac{T_{j(max)} - T_{amb}}{P_C} - \theta_m$$

$$\theta_h \simeq \frac{90 - 45}{0{\cdot}5} - 35^\circ C/W$$

$$\theta_h \simeq 55^\circ C/W$$

A Redpoint heatsink type 5C with a thermal resistance of 30°C/W is used.

The driver stage is working well below its maximum thermal rating, when a Redpoint 5C heatsink of thermal resistance 30°C/W is used.

Performance

The circuit gave its required output power with all the AFY19 transistors tried. The average currents of the driver and output stages were 30mA and 160mA respectively, giving efficiencies of 34 per cent and 52 per cent. Losses due to the unbypassed emitter resistors and coupling networks have not been considered.

The second and third harmonics of the carrier frequency were 40dB and 50dB down respectively. If it is necessary to reduce these harmonics and the lower multiplier frequencies still further, a simple *m*-derived band-pass aerial filter can be used.

Modulation

Collector modulation of both the driver and output stages is employed in this transmitter. The modulation of the driver stage is found to be necessary to obtain modulation depths of 100 per cent overall without appreciable distortion. The transistors in the output stage have 2Ω unbypassed emitter resistors giving a small amount of negative feedback. This is found to be necessary to obtain a clean modulation envelope. The d.c. supply voltage to the driver and output stages is reduced by resistors in the supply lines thereby reducing the unmodulated output. The drive conditions are unchanged as each stage must be capable of supplying the power necessary at the peak of the modulation envelope, the r.m.s. output power at this point being 1W. Each stage is modulated approximately 50 per cent giving an overall modulation depth of 100 per cent. The audio power required when 100 per cent modulated is approximately 330mW; the impedance seen by the secondary of the modulation transformer is 93Ω. It must be remembered that the secondary winding carries appreciable direct current.

General Design Notes

The tuning of this transmitter (Fig. 5) follows the normal practice. Starting at low power the interstage networks are successively adjusted from the output and the oscillator power increased until maximum output power is obtained. The two series tuned circuits in the multiplier stages are adjusted each time to give minimum outputs at 21·25Mc/s and 42·5Mc/s respectively.

Care should be taken when building the transmitter to ensure that the layout is as compact as possible. Leads should be short and preferably of copper or silver foil. Screening should be introduced around components liable to radiate and cause feedback problems. Care must be taken in choosing r.f. chokes and decoupling capacitors, otherwise instability problems are likely. The decoupling capacitors used in this transmitter are Erie PAZ and PAX types. The chokes are, where ferrite, wound on Neosid choke cores. Where air cored, the chokes are 20 turns of 30 s.w.g. TNA wire wound on a 1cm diameter former.

The values of capacitance shown as variable in Fig. 5 are composed of Philips' trimmers type E7879 supplemented where necessary by miniature silver mica capacitors.

Conclusion

The outline of the design procedure used in the amplifier stages of this transmitter has been given together with a description of the complete system.

When adjusted to give a c.w. output of 1W the d.c. power taken by the complete transmitter is 2·7W giving an overall efficiency of 37 per cent.

The design and construction of transmitters in this power and frequency range begin to pose new problems, due mainly to the low impedance nature of the circuits. These problems would appear to increase with power and frequency and may require a reappraisal of circuit ideas.

Acknowledgments

The author wishes to thank the General Manager of Mullard Southampton Works and the Directors of Associated Semiconductor Manufacturers Ltd for permission to publish this article.

REFERENCES

1. Nowicki, J. R. An Introduction to Transistor Transmitters. *Electronic Engng. 34,* 744 (1962).
2. Hilling, A. E., Nowicki, J. R. The Design of a 4W 8·31Mc/s Transmitter. *Electronic Engng. 35,* 292 (1963).
3. Everitt, W. L. Communication Engineering, pp. 263-268 (McGraw Hill Book Co., 1937).

An Image Amplifier For Quality-Control

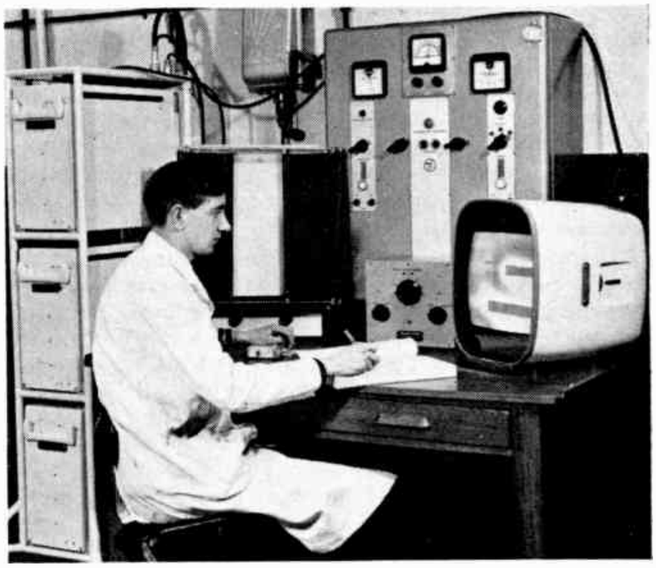
Already well-known in medical circles for the high degree of accuracy and convenience it has brought to both diagnosis and treatment, the Marconi Instruments twelve-inch image amplifier is now available in an industrial version, known as type OE 1280B. The first production model has been installed in the Willans Works of The English Electric Co. Ltd. at Rugby, where it is materially assisting the production of steam turbine blades of consistently high quality.

The image amplifier is a high-definition, closed-circuit television system designed for direct viewing of a fluorescent image produced by the passage of X-rays through the specimen under examination. The image is focused by a mirror and lens system on to the photo-cathode of a 4½in image orthicon tube, and is finally displayed on one or more 17in television monitor screens which, by means of electronic magnification, can have an 800-line definition in the central eight inches of the screen. Outlets are also available for recording on video tape, 16mm ciné film, 100mm negative film or 4in by 5in Polaroid positive film.

At the English Electric Works, the installation is being used to examine the adhesion of the Stellite erosion shields which are brazed, by induction heating, on to the leading edges of low-pressure reaction blades ranging in length from 3in to 36in. Each blade is placed on a variable-speed, remotely-controlled trolley which passes between the X-ray source and fluorescent screen, a complete detailed examination taking less than five minutes per blade. Maximum thickness of steel that can be examined visually, using a 250kV source at 10mA, is 1½in, giving a sensitivity of 2 per cent.

No special room or building is required as the inspection unit is of radiation-proof construction. If a thicker lead lining is fitted, a 300kV source can be employed. An interlocking mechanism ensures that the X-ray tube cannot be energized with the cover open.

The viewing position



Ferrite Core Parameter Measurements at Very High Switching Currents

By M. C. Stevens*, Dip. Tech. (Eng.), M.Sc. (Eng.)

Apparatus is described for producing current pulses of up to 10A amplitude with a rise time of a third of a nanosecond. The apparatus has been used in the measurement of the switching parameters of rectangular hysteresis loop ferrite cores for the three regions where the reversal mechanism is by domain wall movement, non-uniform rotation and uniform rotation respectively. The measured parameters of eight types of core are given.

(Voir page 503 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 510)

APART from its use in coincident current matrix computer stores, the rectangular hysteresis loop ferrite core is finding increasing application as a circuit element in industrial and military electronics. Manufacturers publish data for only the low-level, coincident current applications. However, for other uses information is often required concerning the core's performance at high drive currents.

A very convenient form for this information is a description of the core's response to a current step. Its response to any current waveform can then be determined by considering the new waveform to be made up of a large number of small steps[1]. The current step should be as nearly ideal as possible, for if its rise time were greater than about 1 per cent of the total switching time of the core, the core response would be modified.

This article describes apparatus for producing a variable amplitude, near ideal, current step and gives details of the response of eight types of ferrite core at high switching currents.

The Current Step Generator

Requirements of the Step Generator

Switching m.m.f.'s up to at least 6At are desirable, so as to be able to switch well into the region of uniform rotation reported by Lee Sheval[2,3]. To avoid increasing the rise time of the current step unnecessarily, stray inductance, and hence primary turns, must be kept to a minimum. This condition is met by a single turn consisting of a short straight length of wire passing through the centre of the core.

Step generators built around transistors, thermionic valves, s.c.r.'s, hydrogen thyratrons, or avalanche transistors, all have the disadvantages of providing an output at too low a current, at too low a voltage or with too slow a rise time. The device adopted was the mercury-wetted relay[5], although it had its own disadvantage of a low p.r.f. (50 to 100p/s). The relay is used to discharge a pulse forming cable into a delay line at the end of which is a mount holding the ferrite core.

The mercury-wetted relay used is an Elliott type RP5441. It is soldered directly to the coaxial delay line and pulse forming cables and screened with similar braiding to that of the cable. The assembly is clamped in a closely fitting brass tube for mechanical protection. The reed is sprung so that it is normally touching contact *b* (see Fig. 1). When the external electromagnet is energized the reed is attracted to contact *a* and the pulse forming cable is charged to the h.t. potential. On releasing, the reed returns to *b* discharging the cable into the delay line. The pulse forming cable is 600ft of 50Ω Uniradio 67 and is open-circuited at the far end. It produces a pulse of 1·8μsec duration. The delay line is of similar cable, 80ft in length providing a delay of 120nsec which allows a measuring oscilloscope to be pre-triggered. The electromagnet is driven by a 6·3V 50c/s half-wave rectified supply, to produce a p.r.f. of 50p/s.

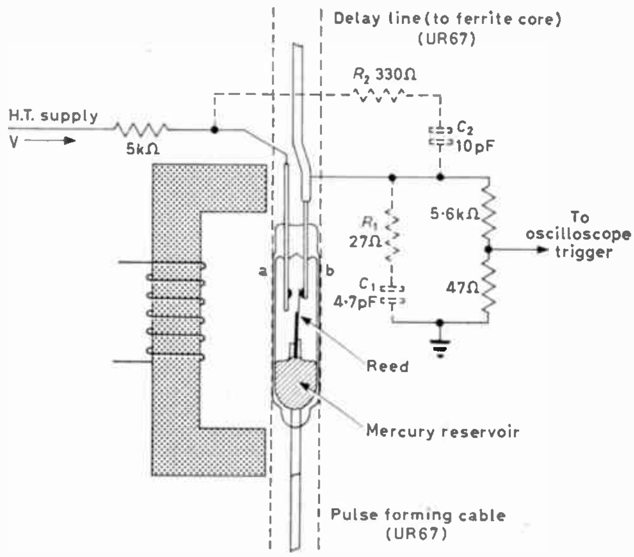


Fig. 1. Mercury wetted relay mounted in a coaxial line

Core Test Rig

The core is mounted on a ¼in length of copper wire held at each end in silver-plated spring slips (Fig. 2). This wire carries the current step. Also threaded through the core, and at right-angles to the first wire, is the output turn.

The 100Ω resistor is necessary to damp out 500Mc/s ringing in the output circuit. It has negligible effect on the core output waveform. The output turn is also used to pass a 1A pulse through the core between the main current steps. This pulse is produced by a blocking oscillator and reverses the flux in the core to its original direction in readiness for the next current step. A 1S916 diode is included in the circuit to isolate the blocking oscillator from the test circuit during the period in which the blocking oscillator is inoperative. At zero bias the junction capacitance of the 1S916 is 2pF. This is reduced by taking the anode to −9V at the blocking oscillator, which also ensures that the diode does not become forward biased when the output turn has a voltage induced in it by the core.

* *Cossor Electronics Ltd. on leave of absence at Northampton College, London.*

Performance

The manufacturer's maximum voltage rating for the mercury wetted relay is 500V. In this apparatus it is being used up to 1kV. Although at the higher voltages the hydrogen-mercury atmosphere surrounding the contacts glows fairly violently, no injurious effects have been noticed over the past two years. The pulse-forming cable has fully discharged long before the relay contacts open.

The pulse forming cable when charged to a voltage V can be considered as a voltage generator of amplitude V with a generator impedance of 50Ω—the characteristic impedance of the cable. When the relay contacts close, the voltage appearing across the load, also at 50Ω, should be $1/2V$; but this is not realized in practice.

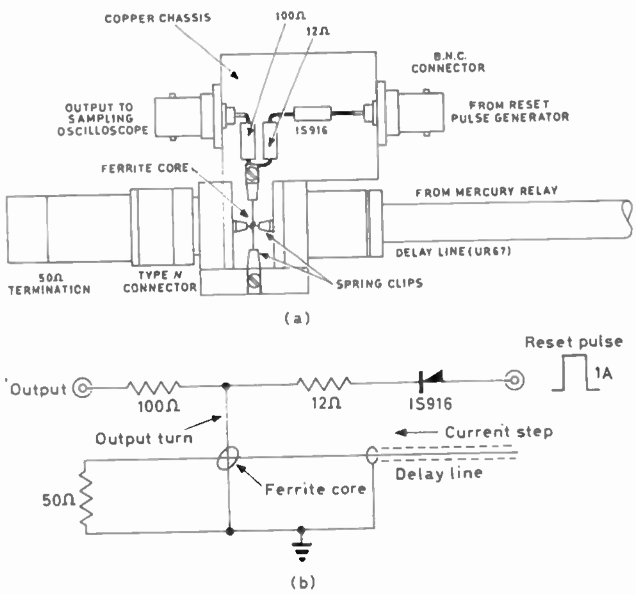


Fig. 2(a). Core test rig
(b). Core test rig—schematic diagram

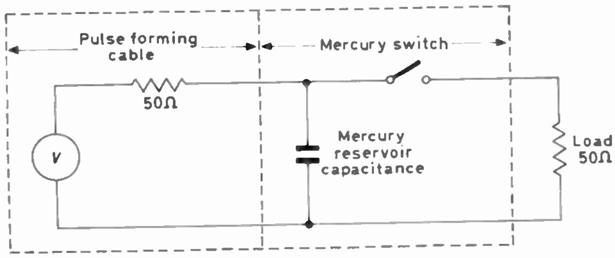


Fig. 3. Mercury relay equivalent circuit

Referring to Fig. 1 it can be seen that the relay is not a perfect match into the coaxial line, the principal discontinuity being the capacitance of the mercury reservoir to the outer conductor. The equivalent circuit is shown in Fig. 3. The capacitance is charged to the same potential as the pulse forming cable with the result that when the relay closes, the initial voltage sent down the delay line to the load is V and produces the overshoot in Fig. 4(a).

The overshoot observed is not of full amplitude V due to the greater attenuation of higher frequencies by the delay cable and to the insufficiently fine resolution of the sampling oscilloscope. A compensating capacitor C_1 (shown dotted in Fig. 1) is connected across the output of the relay so that the surplus charge transmitted in discharging from V to $1/2V$ is used to charge C_1 from zero to $1/2V$. The small resistor R_1 prevents ringing.

The slow rise immediately after the overshoot, shown in Fig. 4(a), is partly due to the greater attenuation of the higher frequency components in the step by the 80ft delay cable. This effect is compensated by connecting a series resistor-capacitor circuit R_2 C_2 (shown dotted in Fig. 1) across the pulse forming cable at the relay end. Its effect is somewhat similar to the capacitance between the mercury reservoir and the outer braid, but C_2 discharges with a long time-constant. The resulting current

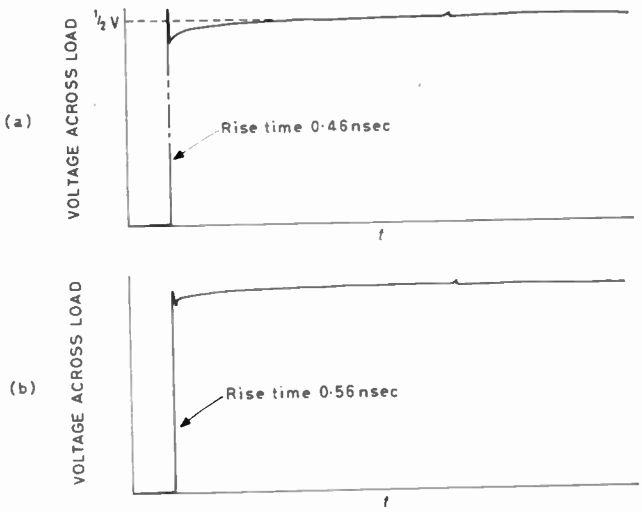


Fig. 4. Current waveform
(a) uncompensated, (b) compensated

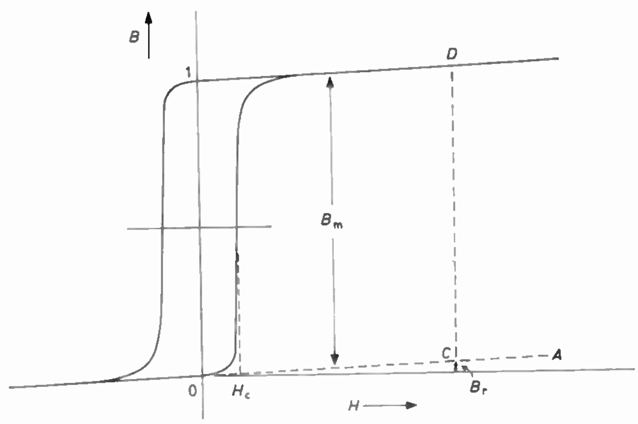


Fig. 5. Typical low-frequency hysteresis loop

waveform measured as a voltage across the 50Ω termination is shown in Fig. 4(b).

The rise time has increased from 0·46nsec to 0·56nsec. These figures include the rise time of the oscilloscope, which is not known but has a theoretical value of 0·44nsec. This would give actual rise times of the current step of 0·13nsec and 0·34nsec respectively.

The voltage appearing across the terminating resistor is actually a little less than the theoretical $1/2V$ due to small mismatches and attenuation in the system. The amplitude of the current pulse in amperes is equal to the h.t. voltage V in hundreds of volts multiplied by 0·92.

By means of a 1 000Mc/s sampling oscilloscope, an x-y plotter, and a slow ramp generator, which completed the equipment, the nano-second output waveforms of the switched ferrite core were recorded directly on paper.

Ferrite Core Switching Characteristics

General

The low frequency hysteresis loop of a typical ferrite core is shown in Fig. 5. The flux is normally remanent at the position marked 0, and is switched to 1 by the application of an m.m.f. (the step current). It is reset to 0 by the reset pulse. There is no significant change if the applied field is less than the coercive field H_c.

Fig. 6 shows a typical output voltage waveform from the arrangement shown in Fig. 2 for an input step current. The initial undershoot is due to capacitive coupling between the input and output wires while the first positive going peak is a result of the small instantaneous reversible flux change in the core material. This appears either as a slight kink or not at all if the rise time of the current is comparable to the switching time of the core itself. The great majority of the flux change is due to domain

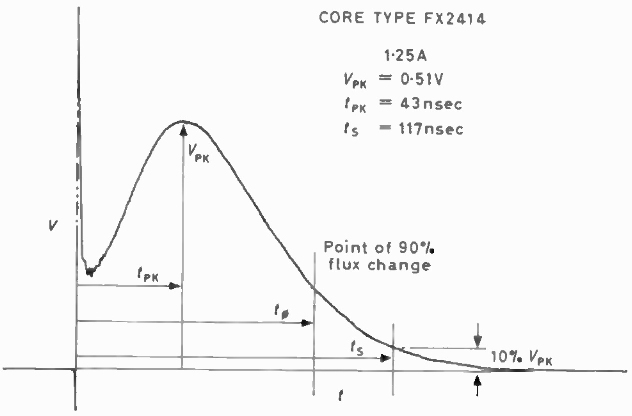


Fig. 6. Typical core output voltage waveform

wall movement and is irreversible; this is a time dependent mechanism, and gives rise to the wide main peak in Fig. 6.

The rise of the current step, and hence the field, has been shown to be very fast—of the order of ¼nsec; the irreversible flux component cannot respond to this and so the point on the *B-H* curve in Fig. 5 moves along *OA* to some point *C*, say, corresponding to the amplitude of the current step. Since the line *OA* has a small slope there is a correspondingly small flux density change B_r which is the reversible one mentioned earlier. While the field remains constant, the flux density changes from *C* to *D* as the main flux changes. Some time later when the current step ceases the field falls to zero, and the point on the *B-H* curve follows the path *D*1, giving rise to a flux density change of $-B_r$.

Each core was tested over a wide range of switching currents and measurements were made of the voltage V_{PK} and the times t_{PK} and t_S. These are defined in Fig. 6. The variation of V_{PK}, t_{PK} and t_S with switching field is shown in Fig. 7.

H_o is the pseudo-coercive force and differs only slightly from H_c, the coercive force for the major hysteresis loop. It is approximately 30 per cent greater than that quoted for coincident current applications when an inner loop is used.

Flux reversal is by domain wall movement when the applied field has a value between H_o and H_1, but this mechanism is superseded by non-uniform rotation for fields between H_1 and H_2, and by uniform rotation for fields above H_2. For the first region the following relationship[6] have been written:

$$V_{PK} = R_o(H - H_o),\ t_s = \frac{S_o}{H - H_o},$$

where H is the applied field, R_o is the slope of the V_{PK}/H curve and S_o is a switching constant equal to the inverse of the slope of the $1/t_s$ against H curve.

Instead of measuring t_s, the time at which the output voltage has fallen to 10 per cent of its peak value, some workers measure t_ϕ, the time at which 90 per cent of the reversible flux has switched. Determination of the latter involves the integration of the output voltage, and allows measurements to be made when the switching time of the core is comparable with the rise time of the switching pulse being used. t_ϕ, however, is less immediately useful in circuit analysis.

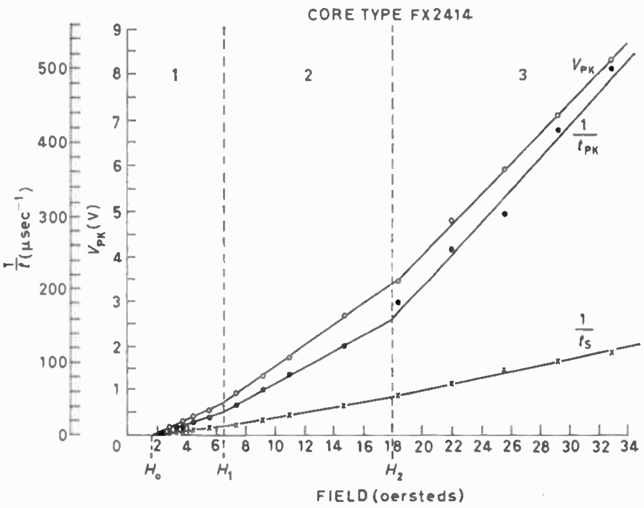


Fig. 7. Variation of V_{PK}, t_{PK} and t_s with increasing switching field

Since the curves in the second and third regions of Fig. 7 are also straight lines, the above empirical relationships can be modified for use at the higher fields. For applied fields between H_1 and H_2 in value they become

$$V_{PK} = R_o(H_1 - H_o) + R_1(H - H_1)$$

$$\text{and } t_s = \frac{S_o S_1}{S_1(H_1 - H_o) + S_o(H - H_1)}$$

and for applied fields above H_2

$$V_{PK} = R_o(H_1 - H_o) + R_1(H_2 - H_1) + R_2(H - H_2)$$

$$\text{and } t_s = \frac{S_o S_1 S_2}{S_1 S_2(H_1 - H_o) + S_2 S_o(H_2 - H_1) + S_o S_1(H - H_2)}$$

where R_1 and R_2 and S_1 and S_2 are defined in a similar way to R_o and S_o for applied fields of greater than H_1 and H_2 respectively.

Results

The parameters of ten cores from a single batch for each of eight different core types were measured at various switching currents from 0 to 9·2A and for some core types to 17A. The results are tabulated in Table 1 and are given as functions of switching current instead of switching field since the former is of more immediate interest to the circuit designer.

The parameters can be directly substituted into the expressions given in the previous section with current *I* replacing field *H*.

The spread in measurements between individual cores

TABLE 1

CORE TYPE	Region 1 R	Region 1 S_{PK}	Region 1 S_S	Region 1 I	Region 2 R	Region 2 S_{PK}	Region 2 S_S	Region 2 I	Region 3 R	Region 3 S_{PK}	Region 3 S_S	Region 3 I	SIZE (IN. O.D.)
FX1897	2·8	0·13	0·35	0·46	3·4	0·081	0·26	1·6	5·3	0·044	0·22	4·2	0·150
FX1508	0·68	0·085	0·22	0·49	1·0	0·049	0·14	2·1	1·3	0·042	0·13	5·0	0·080
FX1948	0·63	0·10	0·26	0·42	0·92	0·066	0·18	1·8	1·1	0·050	0·16	5·5	,,
FX2423	0·93	0·060	0·16	0·37	1·2	0·036	0·12	1·5	2·1	0·017	0·09	3·2	,,
WP3884/SF11/SC500	0·49	0·042	0·11	0·49	0·72	0·023	0·077	2·2	0·97	0·017	0·075	5·0	,,
FX1899	0·86	0·030	0·091	0·42	1·0	0·022	0·070	1·8	1·6	0·014	0·055	3·0	0·050
FX2414	0·60	0·039	0·11	0·40	0·95	0·023	0·071	1·6	1·3	0·012	0·062	4·5	,,
FX2551	0·30	0·014	0·043	0·44	0·44	0·007	0·033	1·6	—	—	—	>7	0·032

I is the critical switching current in amperes. In region 1 it is the current necessary to produce a field equal to the pseudo-coercive force Ho, in region 2 equal to the force H_1 and in region 3 equal to the force H_2.

R is the characteristic resistance in ohms and is equal to the slope of the peak voltage against switching current curve.

Spk and S_s are the switching constants (μsec.A) and are the inverse slopes of the $1/t_{pk}$ against switching current and $1/t_s$ against switching current curves respectively. It was not practically possible to measure the critical current for region 3 for core type FX 2551. It would appear to be greater than 7A.

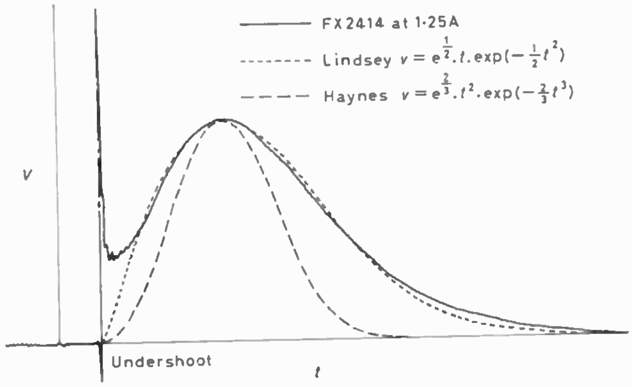


Fig. 8. Comparison of actual and theoretical voltage waveforms

of the same type was less than ±10 per cent (including experimental error) with a coefficient of variation of less than 5 per cent. For some core types it was considerably better than this, and for the FX2423 these figures were ±5 and 3 per cent respectively.

There have been several attempts[7-10] to derive an expression for the shape of output voltage waveform for the region where flux reversal is by domain wall movement. Two of these expressions[7,8] have been normalized, and are given in Fig. 8 with the response of an actual core, switching by domain wall movement. Lindsey's curve appears to be in good agreement, although the degree of agreement varies a little with switching current amplitude, and his comment on Haynes' theory seems fully justified.

As the switching current is increased the shape of the response is modified slightly and the tail becomes more extended. In region 1 the ratio of t_s to t_{PK} is typically around 2·6. It increases to between 2·9 and 3·4 in region 2, and more rapidly to between 3·6 and 4·5 in region 3. This is almost certainly due to a fraction of the flux still being reversed by a slower mechanism. This is reflected also in the fact that the switching constant S_s, relating to t_s, has a value substantially the same in region 3 to that in region 2. Determination of the ratio of t_s to t_{PK} for a particular switching current would appear to give a reliable indication of the mechanism of the majority flux reversal involved. Fig. 9 shows typical output voltage curves for each of the three regions.

The results obtained have been used to compute the performance of a transistor regenerative amplifier with a ferrite core as the feedback element operating at p.r.f.'s of 10Mc/s and higher.

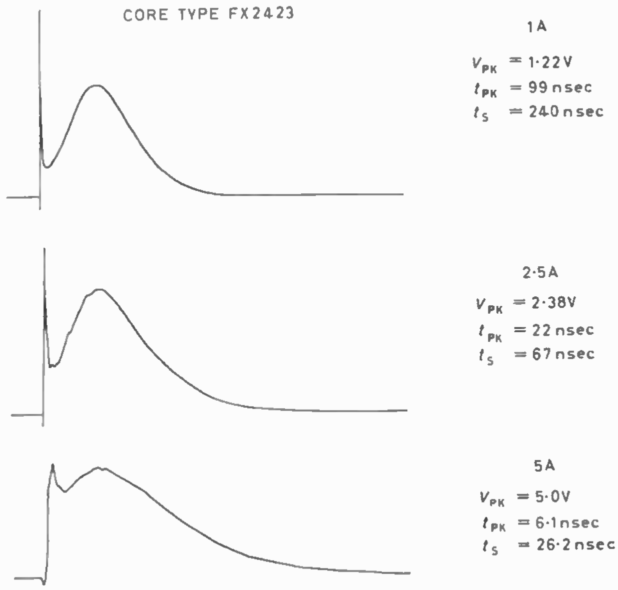


Fig. 9. Voltage waveforms for core switching in Regions 1, 2 and 3

Acknowledgments

The author wishes to thank Dr. F. C. Widdis for his encouragement and the Head of the Electrical Engineering Department, Northampton College, London, for providing research facilities. He particularly acknowledges the assistance of Mr. D. Webb who carefully made many of the routine measurements on the cores. He is also very grateful to Cossor Electronics Ltd for the award of a Research Scholarship.

REFERENCES

1. Betts, R., Bishop, G. Ferrite Toroid Core Circuit Analysis. *IRE Trans. Electronic Computers. EG10*, 51 (1961).
2. Shevel, L. W. Observations of Rotational Switching in Ferrites. *IBM. J.* 93 (1959).
3. Shevel, L. W. Millimicrosecond Switching Properties of Ferrite Computer Elements. *J. Appl. Phys. 30*, No. 4, 47s. (1959).
4. Gyorgy, E. M., Hagedorn, F. B. Uniform Rotational Flux Reversal of Ferrite Toroids. *J. Appl. Phys. 30s*, No. 4, 308s. (1959).
5. Garwin, R. L. Pulse Generator for the Millimicrosecond Range. *Rev. Sci. Instrum. 21*, 903 (1950).
6. Menyuk, N., Goodenough, J. B. Magnetic Materials for Digital Computer Components. 1—A Theory of Flux Reversal in Polycrystalline Ferromagnetics. *J. Appl. Phys. 26*, 8 (1955).
7. Lindsey, C. H. The Square-Loop Ferrite Core as a Circuit Element. *Proc. Instn. Elect. Engrs. 106C*, 117 (1959).
8. Haynes, M. K. Model for Nonlinear Flux Reversals of Square-loop Polycrystalline Magnetic Cores. *J. Appl. Phys. 29*, 472 (1958).
9. Hilberg, W. Toroids of Square Loop Ferrite as Switching Elements. *Arch elekt. Ubertragung. 15*, No. 3, 145 (1961).
10. Chen, T. C., Papoulis, A. Terminal Properties of Magnetic Cores. *Proc. Inst. Radio Engrs. 46*, 839 (1958).

Improvements in the Design of Precision Wirewound Resistors

By J. R. Price*, A.M.I.E.E.

One of the most important parameters of the precision wirewound resistor is its long-term stability. A new type of construction has been developed in which the effects of winding tension and bobbin material have been obviated: in these resistors the winding is completely loose and free to move on its bobbin. Stability measurements on typical resistors are given.

(Voir page 503 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 510)

ONE of the most important parameters of the precision wirewound resistor is its long-term stability. The factors influencing this are discussed and some results are shown for resistors embodying the various desirable design features.

The long-term stability can be affected by the following:

(1) Type of wire.

(2) Winding tension.

(3) Heat treatment.

(4) Bobbin material.

(5) Sealing.

(6) Termination.

Type of Wire

The type of wire used is dependent upon the temperature coefficient of resistance (t.c.r.), the thermal e.m.f. requirement and the most suitable resistivity for the particular application. 'Manganin' or copper nickel is used for low value resistors and 'Karma', 'Evanohm' or 'Nikrothal' for high values.

Where resistors of low ohmic value have to work over a wide temperature range, it is more usual to use copper nickel, as the t.c.r. of manganin is non-linear (Fig. 1). The disadvantage with copper nickel is the high thermal e.m.f. against copper, see Table 1.

The high resistivity low t.c.r. wires have very similar characteristics and the selection of wires becomes the choice of the resistor manufacturer. Low t.c.r. wires can be selected, they are reasonably linear and the thermal e.m.f. against copper is low (Table 1).

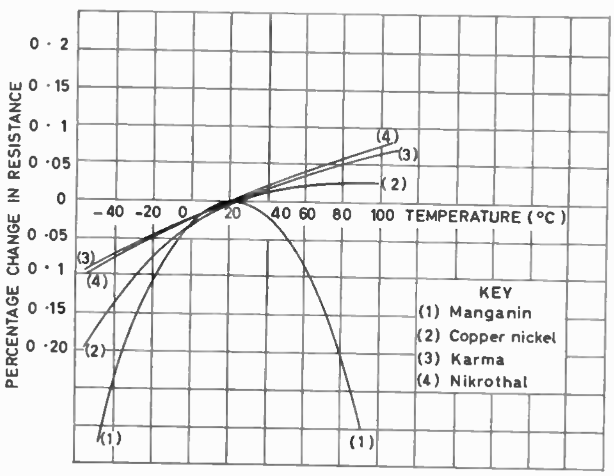


Fig. 1. Graph showing the temperature coefficient of resistance characteristics from −60°C to +100°C for 'Manganin', copper nickel, 'Nikrothal' and 'Karma' resistance wire

* *Alma Components Ltd.*

Winding Tension

The winding tension can play a very large part in causing bad long-term stability. It is therefore essential to keep the winding tension to a minimum. The wire manufacturer usually gives safe figures for a given size of wire.

Fig. 2 shows the resistive movement due to various winding tensions during heat treatment at 190°C. Fig. 3 shows the resistance change after 160 hours at 190°C related to wire tension.

Heat Treatment

All resistors change with time. This change is much greater initially and can be accelerated by heating. The initial change of resistance would take place over a much longer period without being artificially aged.

It is also essential that the heat treatment temperature should be well above the highest temperature which the resistor is ever likely to reach. Fig. 4 illustrates the movement caused when the temperature is raised.

Bobbin Material

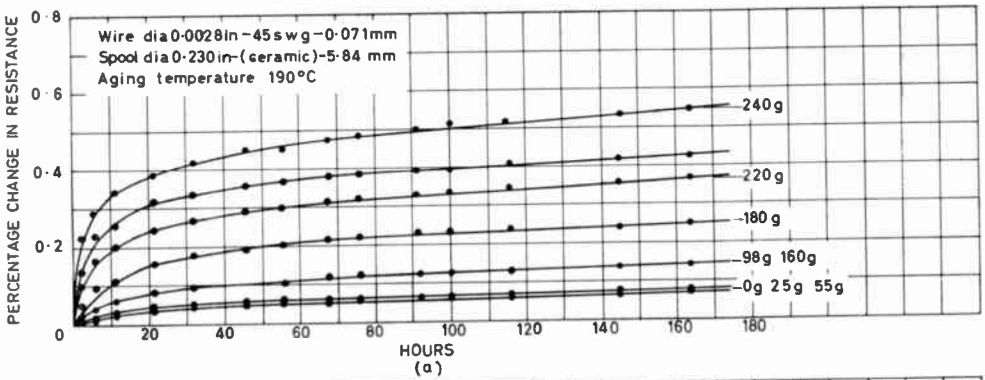
Choice of bobbin material on which to wind precision wirewound resistors is of the utmost importance. If material is chosen with a mechanical temperature coefficient (m.t.c.) of expansion greater than than of the wire, the t.c.r. characteristics of the wire will be modified by the bobbin m.t.c. (strain gauge effect). This also applies if the resistor is impregnated with a material. The effect has even been noticed when turns of wire have adhered to each other due to the enamel insulation softening during the heat treatment. Fig. 5 shows the modification of the t.c.r. characteristics due to complete encapsulation in resin.

The bobbin material must not be hygroscopic, otherwise a low resistance path will form and adversely affect the stability, particularly with the high ohmic values.

The diameter of the bobbin also plays a part contributing to the instability. This is most noticeable when a large

TABLE 1
The Thermal E.M.F. for several resistance wires in common use

Wire	Thermal E.M.F. against Copper $\mu V/°C$
'Karma'	2
'Nikrothal'	2
Copper Nickel	43
'Manganin'	0·5



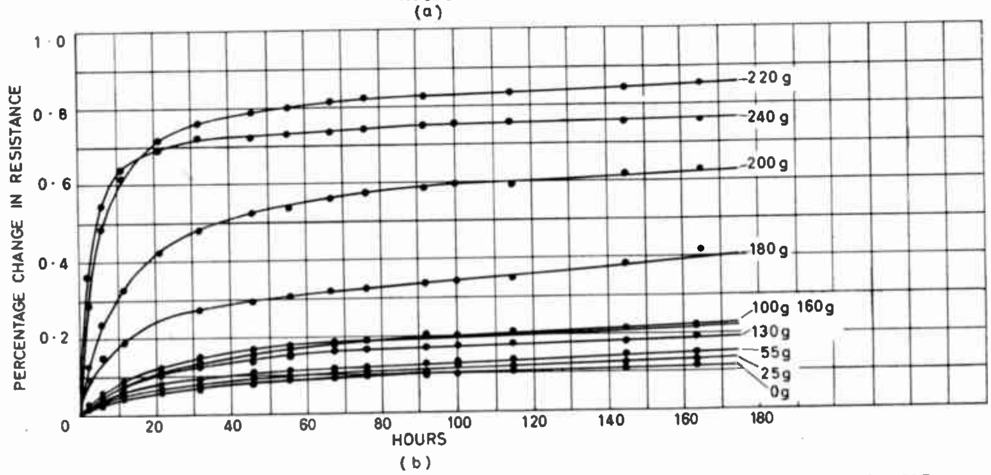


Fig. 2 (a) The graph illustrates the percentage change in resistance at an aging temperature of 190°C for various winding tensions, using 'Nikrothal' wire

(b) The graph illustrates the percentage change in resistance at an aging temperature of 190°C for various winding tensions, using 'Karma' wire

diameter wire is wound round a small diameter former. This can be seen from the graph, Fig. 6. The loop of wire is also moving a small amount and this is thought to be due to the stress and strain put into the wire by the manufacturer when spooling. In winding a wire of diameter d on a spool with diameter D, the outer fibre of the wire is elongated by a factor $d/(D + d) \times 100$ per cent, while the inner fibre of the wire is compressed by the same amount.

Sealing

Sealing is important, as even a small amount of moisture within the component can set up electrolytic action at the termination which is accentuated if the component is working on a light d.c. load. Shunt paths can form and also cause breakdown trouble.

Termination

Without a good contact between the resistance material and the termination, the component is quite useless. It is difficult to soft solder some of the high resistivity wires and welding is usually the only answer. In the case of manganin and copper nickel, this can be soft soldered without difficulty.

The Loose Wound Resistor

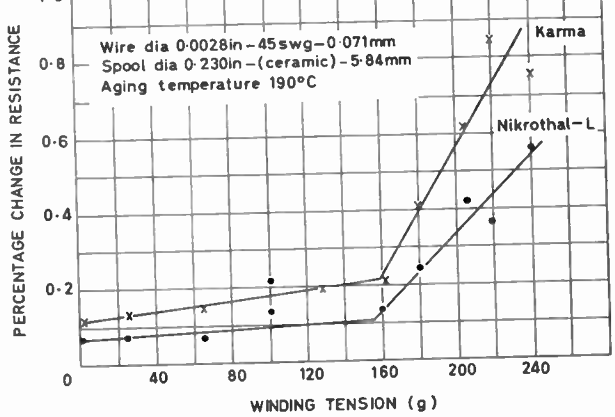
The various factors have now been considered which can influence the performance and in particular the long-term stability.

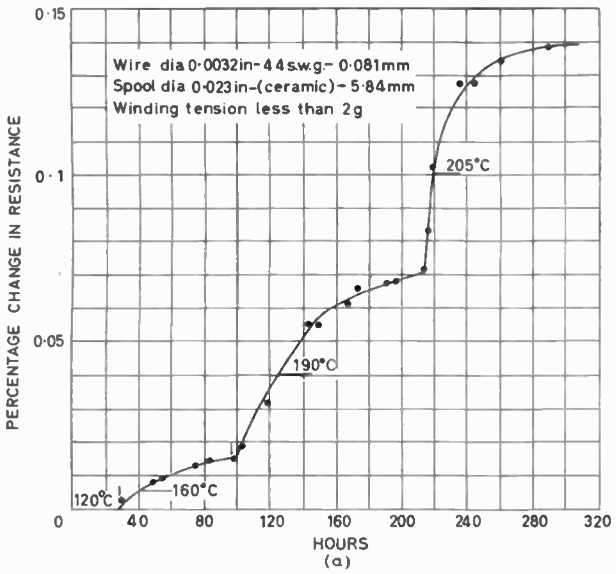
A new type of resistor attempting to embody all the desirable features listed has been designed and is now commercially available.

Ideally the wire should have zero temperature coefficient; in practice better than $\pm 5 \times 10^{-6}$ is obtainable by selection. The resistor should be wound at very low speeds and the tension must be kept to a minimum.

After considerable development, a new technique has now been devised so that the finished resistor is virtually without winding tension. The winding is completely loose and free to move on its bobbin; this not only ensures freedom from winding strain, but also means the bobbin itself plays no part in modifying the t.c.r. and also completely removes the strain gauge effect. Since the wire has to be bent round a finite diameter, it is still necessary to heat treat which assists in relieving the small stresses and strains.

Fig. 3. Shows the percentage resistance change of 'Karma' and 'Nikrothal' wire after 160 hours at 190°C related to the winding tension





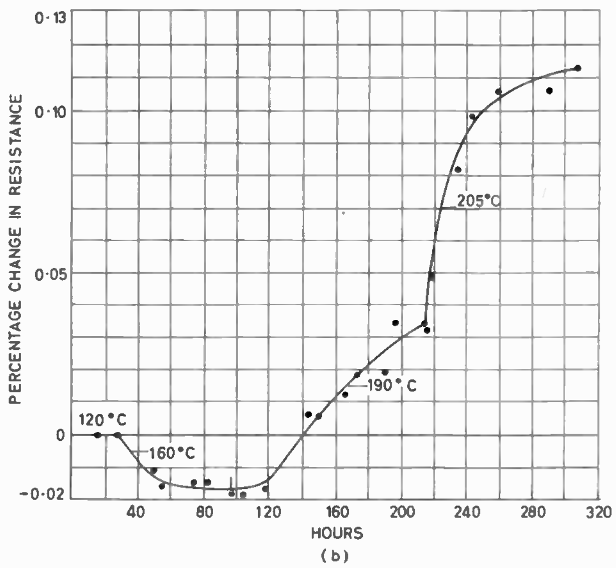
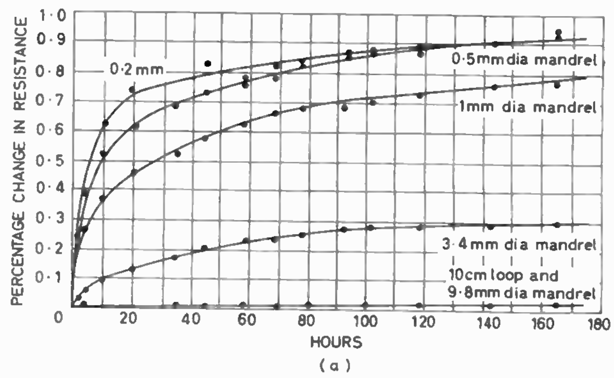


Fig. 4 (a) Shows the change in resistance as the aging temperature is increased for 'Nikrothal' wire

(b) Shows the change in resistance as the aging temperature is increased for 'Karma' wire



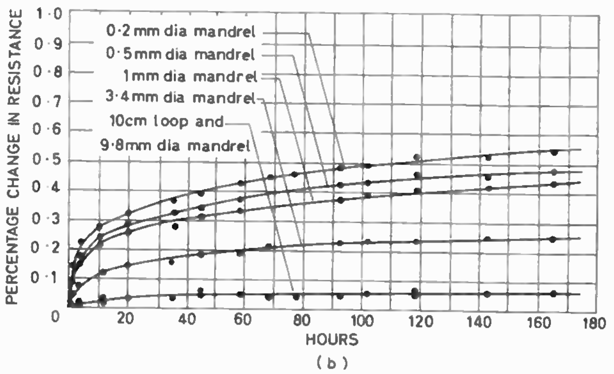


Fig. 6 (a) The graph shows the percentage change in resistance for 44 s.w.g. wire wound round various diameter bobbins for 'Nikrothal' wire

(b) The graph shows the percentage change in resistance for 44 s.w.g. wire wound round various diameter bobbins for 'Karma' wire

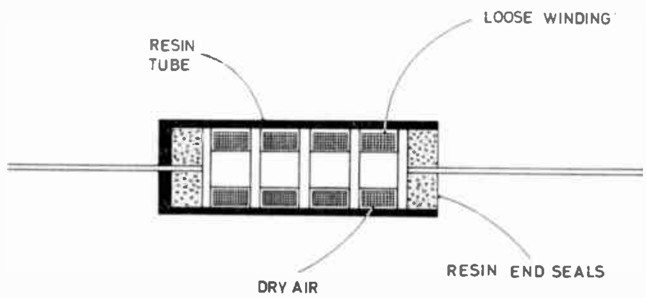


Fig. 7. The drawing shows the cross-section of loose wound resistors, showing the end seal and the winding completely free to move

Fig. 5. The graph illustrates the modification of the temperature coefficient characteristic, due to the complete encapsulation in resin and also shows that the t.c.r. characteristic of the loose wound resistor is similar to the wire characteristic

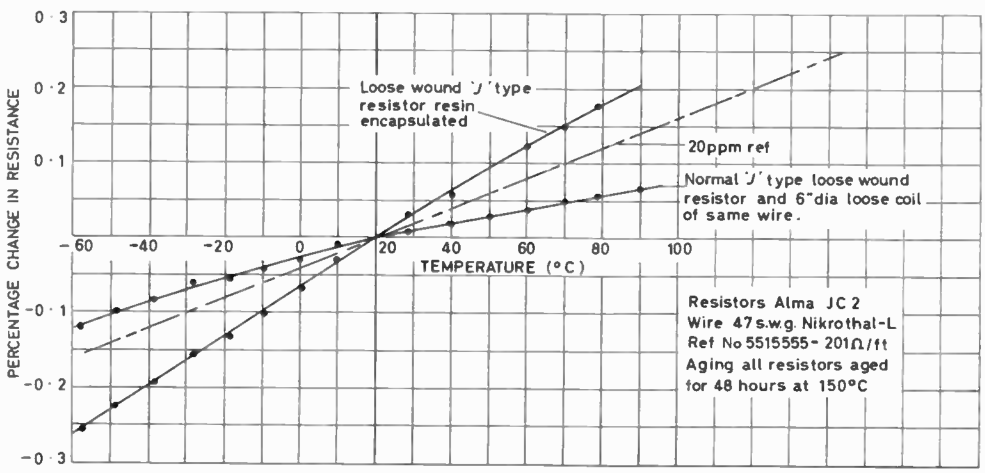


TABLE 2

Typical stability figures for the loose wound range of resistors when loaded to rated power in an ambient of 20°C

TYPE	RESISTANCE VALUE	POWER RATING	1000H	2000H	3000H
JA8	10kΩ	$\frac{1}{8}$W	·005	·004	·004
JA8	10kΩ	,,	−·002	0	−·002
JA8	10kΩ	,,	−·001	0	−·001
JA8	10kΩ	,,	−·001	−·001	−·001
JA8	300kΩ	,,	0	0	+·003
JA8	300kΩ	,,	0	+·003	−·002
JB4	100kΩ	$\frac{1}{4}$W	−·003	*	*
JC2	100kΩ	$\frac{1}{2}$W	·004	*	*
JC2	1kΩ	,,	−·002	*	*
JC2	100kΩ	,,	−·001	*	*
JC2	700kΩ	,,	0	*	*
JC2	700kΩ	,,	−·003	*	*
JC2	700kΩ	,,	−·001	*	*
JC2	700kΩ	,,	−·004	*	*
JC2	700kΩ	,,	+·001	*	*
JD$\frac{3}{4}$	100Ω	$\frac{3}{4}$W	+·003	*	*
JD$\frac{3}{4}$	100Ω	,,	−·003	*	*
JD$\frac{3}{4}$	1kΩ	,,	+·001	*	*
JD$\frac{3}{4}$	1kΩ	,,	+·002	*	*
JE1	100kΩ	1W	·004	*	*
JE1	100kΩ	1W	·004	*	*
JG2	100kΩ	2W	−·001	*	*
JG2	100kΩ	2W	+·002	*	*

All the above figures were obtained by measurement against a ·001 per cent standard.

* *Figures not yet available.*

TYPE	DIMENSIONS LENGTH (IN)	DIAMETER (IN)
JA8	$\frac{9}{16}$	$\frac{1}{4}$
JB4	1	$\frac{1}{4}$
JC2	1	$\frac{3}{8}$
JD$\frac{3}{4}$	$1\frac{1}{4}$	$\frac{3}{8}$
JE1	1	$\frac{1}{2}$
JG2	2	$\frac{1}{2}$

The loose wound resistor can be wound over a resin bobbin which makes the sealing in a resin tube extremely easy. The m.t.c. of bobbin and tube can be made to match exactly and providing the sealing resin does not touch the winding, the stability is not affected.

The production resistors (Fig. 7) are sealed in an oven after the component has been heated to 150°C for 16 hours. The air inside is then free from moisture, and the winding is still free to move slightly.

Due to the close match of the m.t.c., the component is able to work over a wide temperature range (−55°C to +125C°), without any fear of cracking. In the more conventional ceramic bobbin covered by resin, due to the difference in m.t.c. (about 30×10^{-6} dependent on type of resin) crazing almost invariably occurs below −10°C or if it does not, the pressure put on the winding can have an extremely adverse affect.

The case for the loose wound resistors can be summarized as follows:

DISADVANTAGES

(1) The loose wound component is slightly larger than its conventionally wound equivalent
(2) The possible adverse behaviour of the winding under high vibration conditions (still under investigation).

ADVANTAGES

(1) Considerable improvement in the long-term stability.
(2) No modification of the t.c.r. characteristic by the bobbin.
(3) Use of a resin bobbin enables the lead out wires to be embedded in the bobbin and reduces the risk of open-circuited components when handled.
(4) Ease of sealing in a resin tube enables the component to work over a wide temperature range.
(5) The inductive coupling is reduced between turns resulting in a lower overall inductance.
(6) The capacitance between turns is reduced.

A quantity of resistors has been on test and typical results are shown in Table 2.

Acknowledgments

The author acknowledges with grateful thanks, the assistance given by various members of his resistor laboratory staff and in particular Mr. K. Rackham.

An Ultrasonic Level Gauge

An ultrasonic level gauge, designed to measure the level of liquid in a sealed tank or which can be used to monitor the rise and fall of tidal waters, has been developed by Electronic Applications (Commercial) Ltd of London.

The instrument operates on an ultrasonic echo sounding principle through air. Direct measurement of distance is obtained by counting the number of pulses from a clock frequency source during the period between the transmission pulse and echo. These clock pulses are counted by a three decade transistor counter and the result displayed on a three digit in-line read-out of the edge lit Perspex type. A relay may be set to operate through a range of any ten readings characterized by the first two digits of the read-out.

The instrument is intended primarily to measure the level of liquids in tanks and is particularly suitable where problems of hygiene prevent the use of a probe into the liquid.

The transducer is installed with its face exactly parallel with the surface of the liquid and at a height of about one foot above the liquid surface when this is at its maximum. The instrument will give a reading directly proportional to the distance between the liquid surface and the transducer face.

A range of 8ft is possible with a smooth liquid surface. Excessive foam or turbulence reduces this. If the foam is not dense the measurement will be to the liquid surface., The calibration may be adjusted and a reading of one count for 0·1in is typical.

A sounding rate of about two per second is used, a figure resulting from each sounding being displayed on the read-out. At short range variation between soundings is plus or minus one digit, at maximum range plus or minus two digits through air turbulence. An accuracy of 0·2 per cent may thus be obtained.

When the levels in a number of tanks are to be monitored, a single display instrument can be used to indicate or register the condition of each tank in sequence, showing considering economy in the installation.

A Simple Null Filter with Variable Notch Frequency

By J. L. Douce*, Ph.D., M.Sc., and
K. H. Edwards*, B.Sc. (Eng.)

This article describes how a modification of the parallel-T network enables the frequency of minimum transmission to be adjusted by a single potentiometer. An analysis of the filter is given in an appendix

(Voir page 503 pour le résumé en français:
Zusammenfassung in deutscher Sprache auf Seite 510)

THE basic parallel-T network[1] shown in Fig. 1 is readily arranged to give zero transmission at a particular

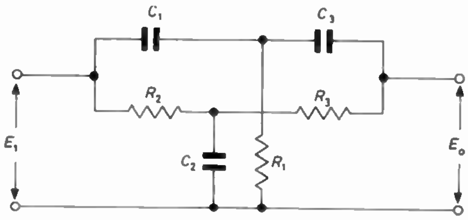


Fig. 1. The basic parallel-T network

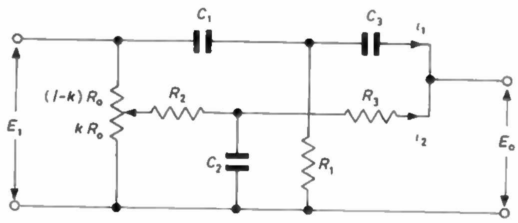


Fig. 2. The modified parallel-T network

frequency. For a null at the angular frequency ω_o the following relationships must be satisfied simultaneously.

$$\omega_o^2 = \frac{C_1 + C_3}{C_1C_2C_3R_2R_3} \quad \ldots\ldots\ldots\ldots\ldots\ldots \text{(1a)}$$

$$\omega_o^2 = \frac{1}{C_1C_3R_1(R_2 + R_3)} \quad \ldots\ldots\ldots\ldots \text{(1b)}$$

For a null

$$R_1(R_2 + R_3)(C_1 + C_3) = C_2R_2R_3 \quad \ldots\ldots\ldots \text{(1c)}$$

By inspection of these equations it is seen that ω_o cannot be varied by adjustment of any one component.

To vary the frequency of null transmission two components, R_1 and C_2 for example, must be simultaneously adjusted. Accurate ganging of the adjustable components is required to maintain a high rejection for a wide range of ω_o.

The Modified Filter

The null frequency of the parallel-T can much more easily be adjusted if one of the parallel paths from the input signal is supplied via a low resistance potentiometer. One suitable arrangement is shown in Fig. 2. As k, the gain of the potential divider, is reduced from unity to zero, the frequency of the null falls from that given by equations (1) towards zero.

At a null the output current into any load must be zero. Consider the output current $(i_1 + i_2)$ which flows when the output is short-circuited to earth. If this current is zero, the output voltage is zero for any load impedance.

For a sinusoidal input signal of angular frequency ω, neglecting the resistance of the potentiometer:

$$i_1 = \frac{-\omega^2C_1C_3R_1e_i}{1 + j\omega R_1(C_1 + C_3)}$$

$$i_2 = \frac{ke_i}{(R_2 + R_3) + j\omega C_2R_2R_3}$$

For a null $i_1 + i_2 = 0$ so, equating real and imaginary parts to zero:

$$(C_1 + C_3)k = \omega^2C_1C_2C_3R_2R_3 \quad \ldots\ldots\ldots\ldots \text{(2a)}$$

$$k = \omega^2C_1C_3R_1(R_2 + R_3) \quad \ldots\ldots \text{(2b)}$$

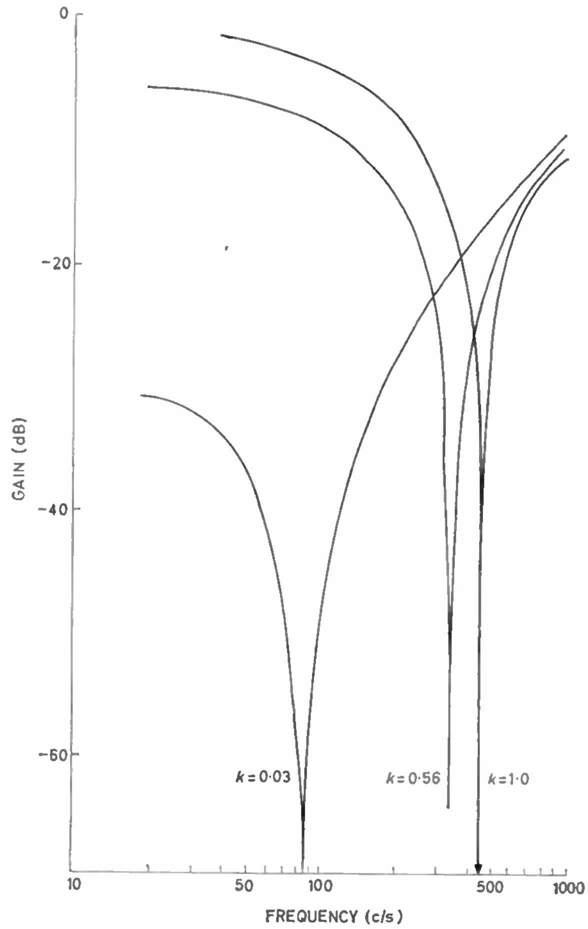


Fig. 3. Frequency response of modified network

Therefore the condition for a null is:

$$R_1(R_2 + R_3)(C_1 + C_3) = C_2R_2R_3 \quad \ldots\ldots\ldots \text{(2c)}$$

Equation (2c), which is independent of k is the necessary and sufficient condition for a null. This is identical to the condition previously stated for a null of the basic parallel-T, equation (1c). The frequency of null transmission is proportional to $k^{\frac{1}{2}}$. The appendix gives an analysis of the network employing a potentiometer whose impedance is not zero.

Experimental Results

A filter has been constructed with $R_2 = R_3 = 2R_1 = 100\text{k}\Omega$, $C_1 = C_3 = \frac{1}{2}C_2 = 3\,500\text{pF}$, and a potentiometer of overall resistance $10\text{k}\Omega$. Fig. 3 shows the open-circuit

* *The Queens University of Belfast.*

frequency response for three settings of the potentiometer. With the potentiometer in the resistive path of the parallel-T the gain is unity at high frequencies and tends to k as the frequency approaches zero. The reverse holds if the potentiometer is in the capacitive path. Fig. 4 shows the variation of maximum attenuation and the frequency at which it occurs with different values of k. A rejection of 1500:1 can be maintained over a range of 100c/s for a reduction in gain at low frequencies of about 0·6:1.

Conclusions

The filter was originally designed for the stabilization of an a.c. control system, with facility for continuous adjustment of the null frequency to follow fluctuations in supply frequency. In this application the potentiometer is replaced by an electronic multiplier[2]. With the multiplier placed in the low-pass branch of the filter, a simple mark space type may be used, and no additional smoothing is required to remove the high frequency ripple introduced by switching. A good notch is maintained for a carrier frequency variation of ± 25 per cent from the nominal value (400c/s).

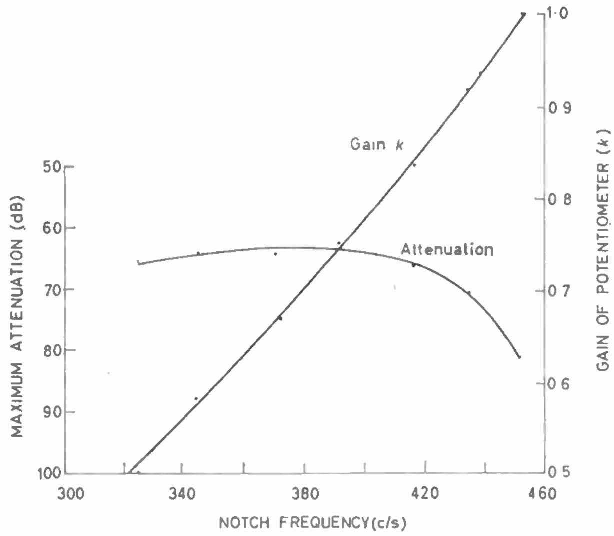


Fig. 4. Characteristics of adjustable notch

Appendix

An exact analysis of the modified filter is obtained from a star delta transformation at ABC in Fig. 5(a).

In Fig. 5(b) let Z_2 be the impedance of R_b and C_2 in parallel.

Then:

$$R_a = 1/k\,\{R_2 + k(1-k)R_o\} \quad \ldots\ldots\ldots\ldots \quad (3a)$$

$$R_b = \frac{1}{1-k}\,\{R_2 + k(1-k)R_o\} \quad \ldots\ldots\ldots \quad (3b)$$

$$R_c = R_o/R_2\,\{R_2 + k(1-k)R_o\} \quad \ldots\ldots\ldots \quad (3c)$$

$$Z_2 = R_b\,\frac{1}{1 + pC_2R_b} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3d)$$

The network of Fig. 5(b) is a parallel-T filter having a null at a frequency such that $i_1 + i_2 = 0$.

Let the input voltage be sinusoidal and of angular frequency ω.

$$i_1 = \frac{e_i(-\omega^2)C_1C_3R_1}{1 + j\omega R_1(C_1 + C_3)}$$

$$i_2 = \frac{e_iZ_2}{R_aR_3 + Z_2(R_a + R_3)}$$

from equation (3d) for frequency ω:

$$Z_2 = R_b\,\frac{1}{1 + j\omega C_2R_b}$$

so:

$$i_2 = \frac{e_iR_b}{R_aR_3(1 + j\omega C_2R_b) + R_b(R_a + R_3)}$$

At a null $i_1 + i_2 = 0$, so equating real and imaginary parts to zero.

$$\omega_o^2 = (1/R_a)\,.\,\frac{C_1 + C_3}{C_1C_2C_3R_3} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4a)$$

$$\omega_o^2 = (1/R_a)\,.\,\frac{1}{C_1C_3R_1\,\{1 + R_3((1/R_a) + (1/R_b))\}} \quad \ldots\ldots\ldots\ldots \quad (4b)$$

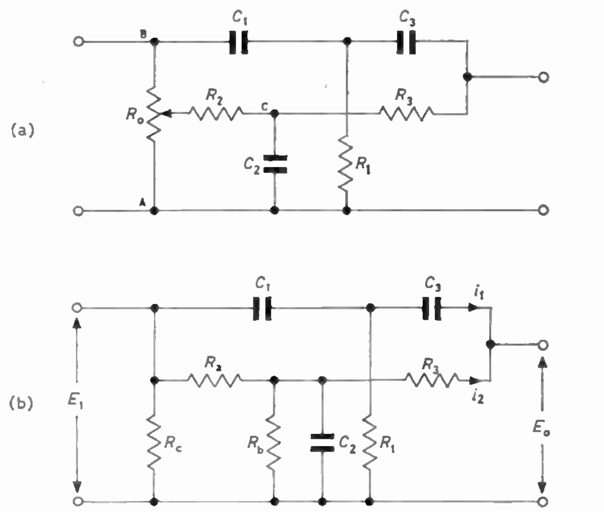


Fig. 5. Analysis of modified parallel-T network

Equations (4a) and 4(b) are consistent over a range of values of R_a if the term $R_3((1/R_a) + (1/R_b))$ is either negligible or constant.

$$R_3((1/R_a)+(1/R_b)) = R_3\,.\,\frac{1}{R_2 + k(1-k)R_o}$$

$$= (R_3/R_2)\,.\,\frac{1}{1 + (R_o/R_2)\,k(1-k)}$$

The maximum value of $k(1-k)$ is $\frac{1}{4}$ at $k = \frac{1}{2}\,(k \leqslant 1)$ so for $R_2 \geqslant 3R_o$ the term $R_3((1/R_a) + (1/R_b))$ is almost constant.

From equation (4a):

$$\omega_o^2 \propto 1/R_a$$

$$\propto k\,\frac{1}{R_2 + k(1-k)R_o}$$

$$\propto k/R_2\,\frac{1}{1 + k(1-k)\,R_o/R_2}$$

$\propto k$ under the same conditions as before.

The inconsistency between equations (4a) and (4b), however small, indicates that the filter cannot be a true null filter at more than one frequency. This is shown clearly in Fig. 4 where the attenuation is about 65dB except for one frequency. At this frequency the attenuation is 80dB. An attenuation of about 1500:1 over a range of 100c/s has been found quite satisfactory in practice.

References

1. Sobczyk, A. Parallel-T Stabilizing Networks for A.C. Servos. MIT Radiation Laboratory Report No. 811 (1946).
2. Barber, D. L. A. A High Speed Analogue Multiplier. *Electronic Engng.* 35, 242 (1963).

An Improved Control System for A.C. Voltage Stabilizers

By R. G. Ackland*, B.Sc., A.Inst.P., and N. M. Buckland*, A.R.M.T.C.

This article describes a simple a.c. voltage sensing and servomotor drive system which, when used to drive the Variac in motor-driven a.c. line voltage stabilizers, can hold the output voltage constant to within ±0·1 per cent. The recovery rate for step changes is 18V/sec.

(Voir page 503 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 510)

SINCE the publication, early in the last decade, of papers describing early motor-operated Variac type a.c. line voltage stabilizers[1-3], there has been an increasing demand for stabilizers of this type. They are, of course, pre-eminent for those applications where freedom from distortion of the voltage waveform is necessary, and especially where outputs of 1 000VA or more are required. Moreover, for the same power handling capacities, motor-operated types are generally smaller and lighter than are other types.

As a consequence there are now several stabilizers of this kind being marketed by a number of manufacturers. Many of these still incorporate driving motors controlled by electromechanical relays, with discontinuous or step-wise correction characteristics, and generally limited sensitivity (c. ±1 per cent) and low speed of response (1 to 10V/sec). However, there are now also available a few stabilizers incorporating continuous drives of the electronically controlled type; these stabilizers generally employ multi-valve sensing and control systems, with their attendant d.c. anode supplies.

The sensing and servomotor drive system to be described in this article is a very simple and reliable one, incorporating as its sensing device a very much under-run lamp bridge, followed by a twin triode which drives the servomotor. No. d.c. anode supply voltages are required.

Despite the simplicity of this control system, it is capable of driving 500VA and 2kVA Variacs, which, with appropriate buck-boost transformers, go to make up 2kVA and 9kVA stabilizers. With the r.m.s. output voltage held constant to within ±0·1 per cent, and a correction rate of 18V/sec (with a buck-boost transformer ratio of 4·5:1) overall performance of these stabilizers is superior to most, if not all, of the commercially available units.

Principle of Operation (Fig. 1)

The sensing unit is a low voltage lamp bridge, energized from a step-down transformer across the stabilizer output terminals. If the line voltage departs from the value at which the bridge is balanced, there appears at the output terminals of the bridge an a.c. signal of magnitude dependent on the line voltage excursion, and in phase or 180° out of phase depending on whether the line voltage is above or below the balance value.

This out-of-balance signal is fed through a matching

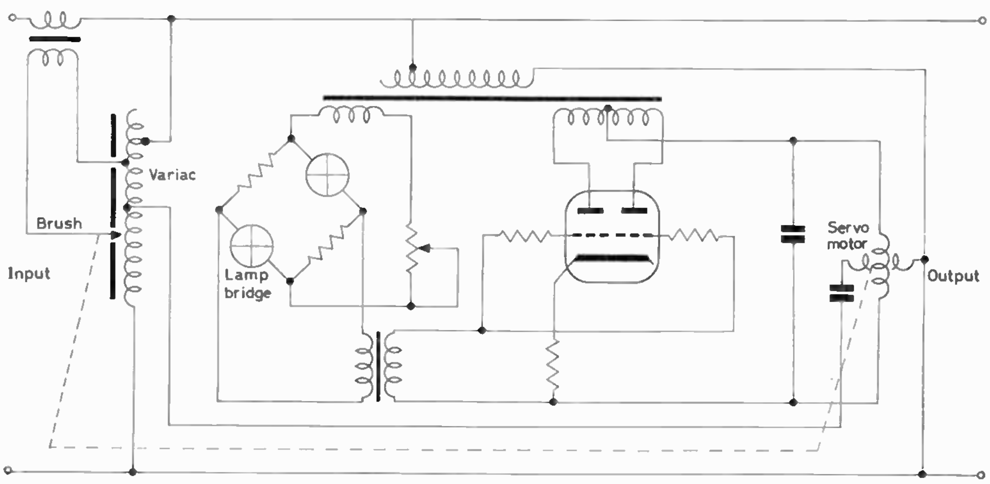


Fig. 1. The stabilizer circuit

transformer to both grids of the twin triode phase-sensitive amplifier. Anode current flows through that side of the valve whose anode is positive when its grid is positive, and does not flow in the next half-cycle when the anode is negative. Very little current flows in the other side of the valve as in the first half cycle its anode is negative and in the second half cycle its grid is negative. The resultant half-wave current flows through the servomotor control winding, in phase or 180° out of phase with the supply voltage, according to which half of the triode is conducting, i.e. whether the bridge voltage is higher, or lower, than it should be.

A rather unsatisfactory feature of most a.c. voltage sensing bridges is that they produce an 'unwanted' output signal when the bridge is balanced. This signal comprises two components, one of supply frequency but 90° out of phase, and one of three times supply frequency[4]. Its normal effect is to obscure the balance point of the bridge, but in this case the amplifier valve and servomotor are virtually only sensitive to 50c/s signals in phase or 180° out of phase, so the unwanted output is of no consequence. With lamps in opposite arms of a symmetrical bridge, sensitivity of the control system is such that the lamps can be run at half their rated voltage,

*Department of Supply, Australian Defence Scientific Service, Defence Standards Laboratories, Australia.

thereby greatly enhancing their stability and greatly extending their useful lives.

Lamp Bridge

Of the possible non-linear devices which may be used in a.c. voltage sensing bridges, viz., lamps, barretters, and thermistors, lamps were chosen as having the best all round properties, despite the somewhat lower figures of merit, or sensitivities, of bridges incorporating them. A significant advantage of lamps is that they are not very sensitive to ambient temperature changes. The measured value of the temperature coefficient of the type of lamp used in the bridge, was found to be 0·005 per cent/°C with the lamps operating at half normal voltage. This is easily and satisfactorily compensated for by using nichrome resistors in the other two arms of the bridge.

Factors taken into account when surveying the lamp field for suitable types were that

(a) The filament should be supported at its two ends only, or alternatively, if fitted with a centre support, this should be welded to the filament. Changes in thermal loss rates at loop supports, due to movement of the filament in the loop can result in local changes in temperature and hence resistance, changing the bridge balance voltage by up to 1 per cent.

(b) If of the coiled type, the filament should be thick enough to remain free of distortion, at operating temperature, irrespective of orientation.

(c) Power consumption and physical size should be reasonably small.

The above considerations narrowed the field down very considerably; maximum compatibility of (b) and (c) indicated a low voltage lamp. After a fairly extensive amount of preliminary testing of various available types, a 24V, 6W lamp, Osram type 150, was selected. This lamp has a bulb diameter of about 18mm, and has two helically coiled filaments in parallel, supported at their ends only. (Lamps have also been seen of the same type number but with a single filament with a loop support.) To avoid spurious resistances, leads are soldered directly to the lamp terminals.

Lamp Stability

Several of these lamps were assembled into opposite arms of approximately equal arm bridge units which were submitted to short term and long term stability tests. The bridges were operated on a.c., but were switched over to d.c. for measurement of their balance voltages. This was necessary as a.c. measurements of the desired precision (c. ± 0·05 per cent) would not have been possible even if the 'unwanted' output were not present to complicate matters. As each bridge was switched from a.c. to d.c. a galvanometer was connected to its detector terminals, and a potentional divider and potentiometer were used to measure its d.c. balance voltage.

Initial short terms tests on as-received lamps operated at their normal voltage showed them to have fairly high drift rates. This could not be attributed to annealing, since the drifts were in the direction of increasing filament resistance[5]. As shown by the upper curve in Fig. 2, continued operation of the lamps at 26V resulted in progressive reduction of the drift rate over the first 10 days, after which it remained more or less constant to the end of filament life. This is consistent with the proposition, based on Larrabee's work[6], that after high initial 'impurity' evaporation had taken place there would be lower, more or less constant, rate of increase in resistance (i.e. reduction in lamp bridge balance voltage) due to evaporation of tungsten.

Since the rate of tungsten evaporation is very dependent on filament temperature, tests were then carried out to determine drift rates of the lamps at various reduced operating voltages, after an initial 'conditioning' period of 10 days at 26V during which impurity evaporation was virtually completed. Results of these tests are shown in Fig. 2, from which it will be seen that, expressed in terms of change in bridge balance voltage, and for a bridge containing two lamps, the drift was approximately 10 per cent for 50 days with the lamps at 26V, less than 1 per cent for 100 days at 20V, and less than 0·2 per cent for 100 days at 16V. At 12V no regular drift was measured, the readings obtained varying in a random manner over a range of about 0·1 per cent.

Whereas at 26V end-of-life of the lamps tested generally occurred at about 60 days, and was due to fracture of the filament, probably because of embrittlement following recrystallization[5], no life period has been established for the lamps at 12V. None of the lamps which have been operating in several stabilizers for about three years has failed so far.

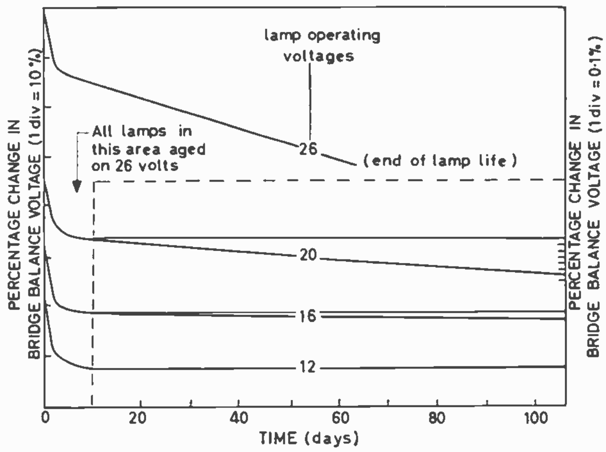


Fig. 2. Percentage change in bridge balance voltage against ageing time.

Each curve is representative of the several bridges tested in each ageing voltage group. (The curves are displaced vertically for clarity. Note also the expanded scale in boxed area)

Amplifier, Servomotor and Variac

The lamp bridge has an internal impedance of about 80Ω, and its output is coupled to the twin triode through a 50Ω to 100kΩ matching transformer. The twin triode used is the Brimar type 6060T (a ruggedized version of the 12AT7) which has the most suitable combination of transconductance and amplification factor. In the interests of enhanced reliability and sensivity, two parallel connected valves are used.

The servomotor is a Brown (Honeywell) type 362497-3, which has an output shaft speed of 20rev/min at no load on 50c/s. It is direct coupled through a simple universal coupling to the Variac shaft. The latter is fitted with ballraces, and in the 500VA model, the contact arm is counter-balanced in order that the operating voltage should not be dependent on orientation of the complete unit. In the case of 2kVA Variacs the rather 'springy' lead to the moving contact is replaced by a more flexible one so as not to impose any restoring torque on the shaft whatever its position.

Performance

Some initial difficulty was encountered in checking the sensitivity and stability of the output voltage of the stabilizers. A narrow span d.c. electronic potentiometer recorder with suppressed zero was used in conjunction with a rectifier and filter for short run tests on sensitivity, but on runs of a day or more, more or less systematic variations of up to 0·3 per cent occurred in recorder reading. Checks with a reflecting dynamometer voltmeter and split photocell amplifier and recorder[7] confirmed that the r.m.s. value of the output voltage was being kept within ±0·1 per cent. The discrepancy was found to be due to small variations in the voltage waveform of the supply mains. Subsequent r.m.s. voltage stability tests were made with a temperature controlled vacuum thermocouple as the a.c.-d.c. convertor.

The ±0·1 per cent sensitivity achieved is very close to the limit attainable with this type of stabilizer, since it represents a range of only plus or minus two single turns of 500VA Variac winding, or plus or minus one turn on a 2kVA Variac.

The rate of response, as measured with a high speed recorder and oscilloscope, is 18V/sec for large input variations, with a minimum response time of about 0·25sec.

Acknowledgment

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

REFERENCES

1. Collinge, B., Marsham, T. N. An A.C. Stabilizer. *J. Sci. Instrum. 28*, 374 (1951).
2. Long, J. V. P. A Simple Electro-Mechanical Stabilizer. *Electronic Engng. 24*, 26 (1952).
3. Ackland, R. G. An A.C. Voltage Stabilizer. *Electronic Engng. 26*, 143 (1954); also *Aust. J. Instrum. Tech. 10*, 3 (1954).
4. Patchett, G. N. Automatic Voltage Regulators and Stabilizers, pp. 156-157 (Pitman, 1954).
5. Smithells, C. J. Tungsten (Chapman and Hall, 1952).
6. Larrabee, R. D. The Spectral Emissivity and Optical Properties of Tungsten. M.I.T. Technical Report 328 (May 1957).
7. Ackland, R. G. Simple Technique for Recording Variations of a Few Centimeters in Galvanometer Reading. *J. Sci. Instrum. 37*, 39 (1960).

A Laser Rangefinder

Barr & Stroud Ltd have developed what is believed to be the first commercially available laser rangefinder. It is a compact instrument capable of being transported and operated by one man to give fast, extremely accurate ranging of objects up to 10 000 metres distant.

Since the demonstration by Dr. Maiman in California, of the laser principle some four years ago, Barr & Stroud have been applying this principle to rangefinding. Recently an agreement was signed between Barr & Stroud and the Hughes Aircraft Company, U.S.A., which allows the research teams of both firms to pool their resources. This, coupled with the experience of Barr & Stroud in optical rangefinding, has resulted in the present instrument.

The laser rangefinder consists of three units, the laser head unit, the power unit and the tripod.

The head unit embodies the laser transmitter, the receiver, the ranging electronics and read-out, the telescopic sight and all associated optical equipment. The head unit is fully sealed and desiccated with all the components in modular form to allow ease of replacement.

The transmitter uses a 'Q'-switched ruby laser with a peak output power of more than 1MW and a beam divergence of 0·5 milliradians.

A repetition rate of 2 shots/min is usual but 6 shots/min can be achieved over short periods.

The receiver has an aperture of 2in diameter and uses a tri-alkali photomultiplier as the detector incorporating a narrow band filter.

The range read-out is in digital form with illumination provided. Should there be no range return or a return in excess of 10 000 metres, the reading clears to 0000.

The sighting telescope magnification is ×6 with an exit pupil of 7mm and an angular field of view of 5·3° in the object space. A cross-wire graticule or other form is provided and eyepiece focusing is included.

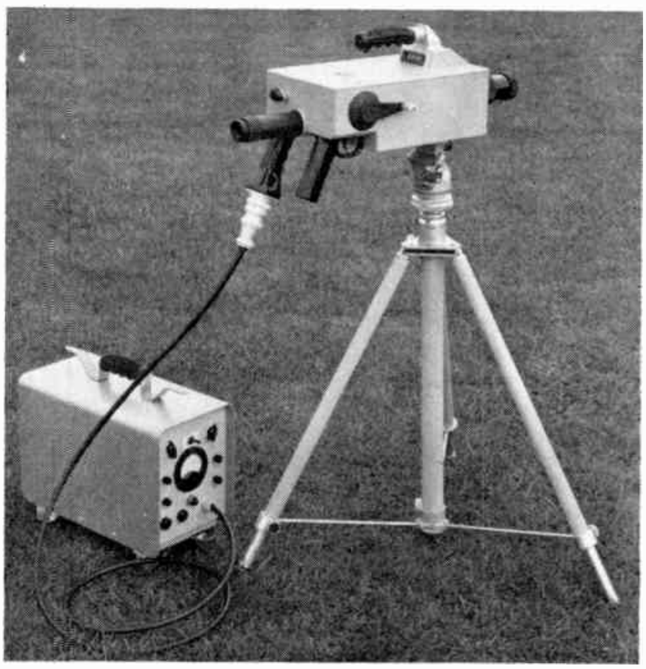
The power unit contains the batteries, power circuits and a capacitor bank. A meter is provided to show the state of charge of the batteries. Not less than 50 shots can be obtained before recharging or replacement of the batteries is required.

The laser rangefinder's minimum range is 300m and the maximum range is 10 000m. Maximum range is dependent on weather conditions and is reduced by poor meteorological visibility.

Effective ranging can be achieved with target areas equivalent to 0·5 milliradians beam-width at maximum range. Target areas can be either normal to both the sun direction and line of sight or up to 45° to these directions and can have a diffuse reflectivity of 0·1.

The tripod allows all round movement in azimuth and ±25° (with respect to horizontal) in elevation. Suitable scales and level indication are provided. The range accuracy is ±10m and the range resolution is 5m.

The laser rangefinder



Wide Band Current Transformers for the Examination of Current Waveforms

By F. J. U. Ritson*, M.Sc., and J. Wood†, B.Sc.

This article describes a current transformer and its associated amplifier for use in the examination of current waveforms. The unit has a very wide bandwidth, and introduces negligible impedance into the circuit being observed.

(Voir page 503 pour la traduction en français; Deutsche Übersetzung Seite 510)

IN circuit development work it is often necessary to examine current waveforms. This is usually done by inserting a resistor in the circuit with a value large enough to produce a voltage waveform sufficient to operate an oscilloscope. However, the maximum sensitivity of most modern oscilloscopes is 50mV/cm, and a voltage of at least 50mV, peak-to-peak, is therefore required to obtain a satisfactory picture. This voltage is not negligible when compared with the voltage levels existing in transistor circuits, and it follows that the insertion of a current-measuring resistance may disturb the operation of the circuit, giving a misleading current waveform. The stray inductance associated with the resistor may also introduce difficulties.

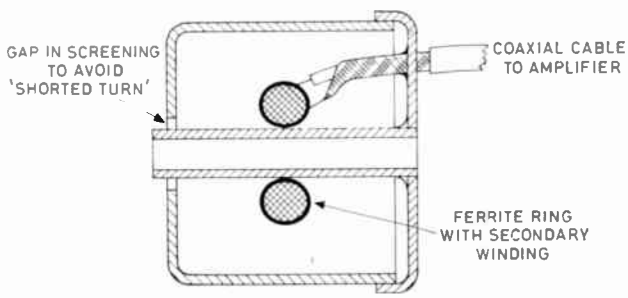


Fig. 1. Construction of current transformer and screening assembly

This article describes a current probe designed to overcome these difficulties. It consists of a wide-band current transformer, of which the primary is the conductor carrying the current to be observed, followed by a transistor amplifier whose output voltage is made proportional to the secondary current supplied by this transformer.

This arrangement has several advantages. A bandwidth of several hundred megacycles per second may easily be achieved in the transformer; the absence of a conducting path between probe and circuit is a convenience and the reflected impedance introduced into the primary circuit is very small. The instrument has been found particularly useful when investigating circuits using transistors in conjunction with square-loop ferrite cores.

The Wide Band Current Transformer

The transformer is wound on a ¼in o.d. ferrite ring core (Mullard FX1593), the secondary being a single-layer winding evenly distributed round the circumference of the core, while the primary, which carries the current to be measured, is a single conductor passed through the centre of the core. Screening between the windings is necessary to avoid unwanted electrostatic coupling and the construction is as shown in Fig. 1. A coaxial cable is used to connect the secondary winding to the amplifier, the cable being terminated at the amplifier end to avoid reflection.

The equivalent circuit of the transformer may be derived from the complete 'lumped constant' equivalent circuit shown in Fig. 2(a). Considering, first, the primary side; the leakage inductance, l_p, is due to flux paths which link with the primary and not with the secondary. These paths are entirely in air, and their presence is independent of the core and secondary winding. It follows that l_p is a property of the circuit in which the measurement is taking place rather than that of the transformer, and may be considered as part of the source impedance. Similarly C_p and r_p may be lumped together with l_p and R_s to form a source impedance Z_s.

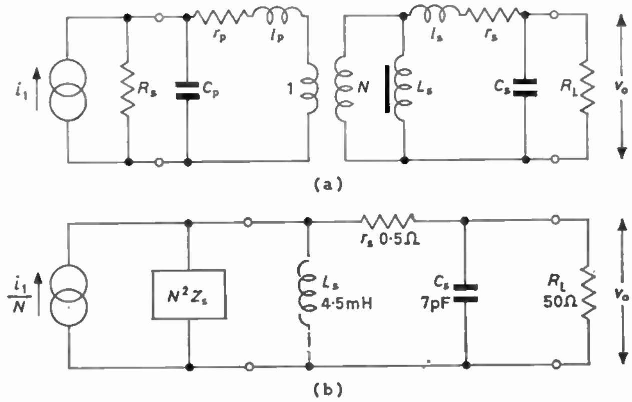


Fig. 2(a). Complete equivalent circuit of a transformer
(b). Equivalent circuit of current transformer, with typical component values

Consider now the secondary side of the transformer. If an alternating current is applied to the secondary, with the primary open-circuited, then by symmetry, all the flux paths must be confined to the inside of the secondary, and must, therefore, all link with the primary. It follows that l_s must be zero. This is the important feature of this form of construction, and is the reason for the excellent h.f. performance which is possible.

The equivalent circuit of the current transformer is therefore as shown in Fig. 2(b), drawn with all components referred to the secondary. The component values shown are for a 73 turn secondary of 36 s.w.g. wire.

The impedance reflected into the primary by the transformer never exceeds $\frac{r_s + R_L}{N^2}$ (for the values given in Fig. 2(b), about 0·01Ω) and this is so small compared with the practical values of the parallel impedance Z_s that the latter may be neglected. The mid-band transfer resistance of the transformer is therefore R_L/N volts per ampere.

* *The University of Newcastle upon Tyne.*

† *Formerly The University of Newcastle upon Tyne, now The Royal Radar Establishment, Malvern.*

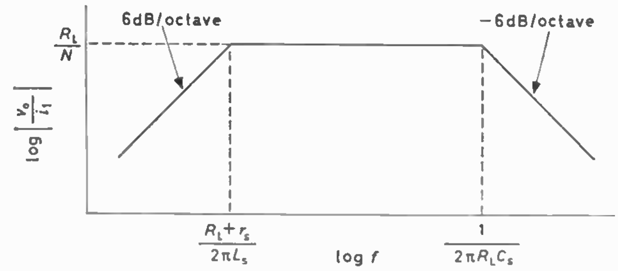


Fig. 3. Theoretical frequency response of transformer

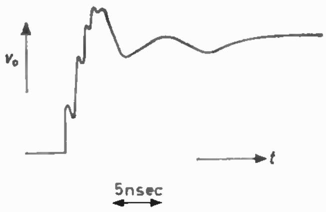


Fig. 4 (left). Transformer transient response to a step change of primary current

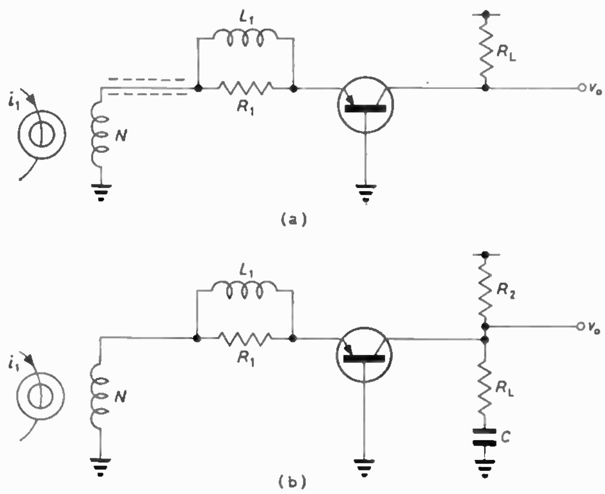


Fig. 5. Amplifier input stages

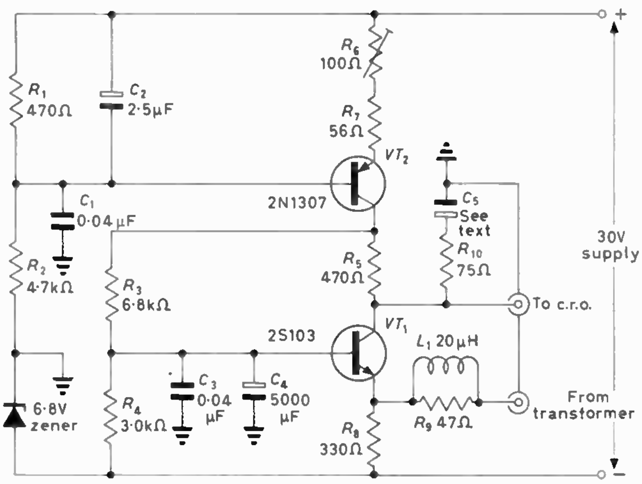


Fig. 6. Complete circuit of amplifier

Fig. 3 shows the theoretical frequency response of the transformer, with the l.f. corner-frequency occurring at $(R_L + r_s)/2\pi L_s$ c/s and the h.f. corner at $1/2\pi R_L C_s$ c/s. For the values given on Fig. 2(b), the predicted h.f. corner frequency is at 450Mc/s, corresponding to a rise time of 0·8nsec.

The transformer transient response was tested by passing through it a step change of current with a rise time of 0·7nsec, the output waveform being recorded on a sampling oscilloscope with a 1 000Mc/s bandwidth: this waveform is shown in Fig. 4. It will be seen that the rise time is of the right order, but ringing occurs, which cannot be predicted from the equivalent circuit. This happens because the 'lumped constant' approximation for the secondary winding is not valid at very high frequencies.

Some experimental work was done on attempting to improve the transient response by using different forms of winding, and it was found that the best arrangement was a single-layer secondary, distributed evenly round the whole circumference of the core to form a toroid. At high frequencies the permeability of the core material falls to a low value and unless the winding is closed, flux leakage occurs with the result that the secondary leakage inductance is no longer zero.

The l.f. response of the transformer of Fig. 2(b) is −3dB at 1·8kc/s, and this will be inadequate in most applications. Some reduction in the l.f. turn-over frequency is possible by reducing R_L, but at the expense of reducing the output. One solution would be to replace R_L by the input resistance of a virtual earth type of amplifier, which gives voltage gain. However, it is difficult to achieve a wide bandwidth in the amplifier while retaining stability, and a better solution is to use a common base amplifier.

Amplifiers for use with the Current Transformer

The transfer resistance of the transformer using the common base amplifier of Fig. 5(a) is $\alpha R_L/N$, almost as before, but the transformer load resistance at low frequencies is the input resistance of the common base amplifier, $r_e + r_b/\beta$, which may be made as low as 5Ω by running the transistor at a high current. At high frequencies, the resistance is increased by R_1, to provide a match for the cable, although, since the input impedance for the common base stage rises at high frequencies, a perfect match is not obtained. The bandwidth of the

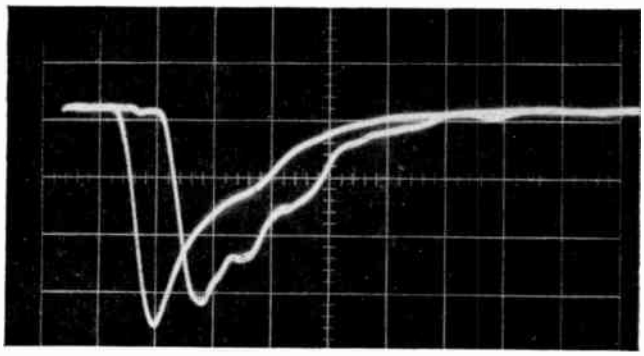
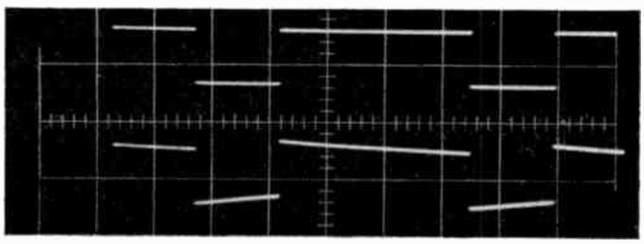


Fig. 7(a) (above). H.F. transient response of transformer and amplifier
Time scale: 10nsec/cm
(b) (below). L.F. transient response of transformer and amplifier
Upper trace: input current, 100mA/cm
Lower trace: output voltage, 100mV/cm
Time scale: 2msec/cm



common base amplifier extends to the α cut-off frequency, which may be several hundred megacycles per second with modern transistors.

The l.f. response may be further improved by using a collector load as in Fig. 5(b). If R_LC is made equal to the transformer time-constant $L_s/(r_s+r_e+r_b/\beta)$, the low frequency response is improved by a factor $(R_2 + R_L)/R_L$.

A practical circuit using the principles described above is shown in Fig. 6. With a 75Ω collector load, R_{10}, and a 73 turn secondary winding to allow for loss in VT_1 and R_3, the overall transfer resistance is 1V/A. The d.c. working point of VT_1 is controlled by feedback through R_3 and by the constant current circuit using VT_2. This current is adjusted to give no d.c. in the transformer secondary winding. The low frequency response of the transformer and VT_1 alone is -3dB at about 250c/s, but the overall response is improved by the method shown in Fig. 5(b). To achieve a large measure of improvement, the d.c. supply is provided by the high impedance (constant current) source VT_2, and R_5 is included to allow the value of R_3

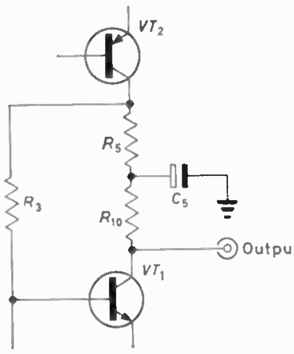


Fig. 8. Modification to Fig. 6 to avoid effect of VT_2 collector capacitance

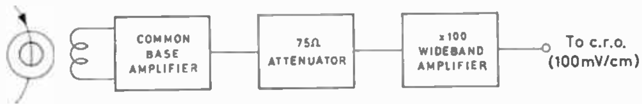


Fig. 9. Arrangement of high sensitivity current transformer and amplifier

to be increased. By choosing the value of C_5 correctly, the l.f. transient response may be made as shown in Fig. 7(a) showing a droop of 3 per cent per millisecond.

The h.f. response is shown in Fig. 7(b); the 5nsec rise time of the test pulse is not increased, but there is a slight reduction in amplitude, probably because at high frequencies the 470Ω resistor R_5 is connected across the 75Ω load R_{10} by the collector-base capacitance of VT_2. This may be avoided by rearranging the circuit as in Fig. 8.

The slight ringing on the trailing edge of the pulse is probably due to a mismatch of the 50Ω coaxial cable at high frequencies: the effect is not present when the transformer has a 50Ω load directly connected and it could be eliminated by building VT_1 into the transformer assembly, omitting R_9 and L_1, and connecting the transistor to the rest of the circuit by 75Ω coaxial cable, which would be correctly terminated at high frequencies by R_{10}.

The sensitivity of this unit (1V/A) may be rather low for use with some oscilloscopes and a wide-band amplifier may be required between the common base stage and the oscilloscope. Fig. 9 gives the block diagram of an instrument designed to provide a maximum sensitivity of 1mA/cm using an oscilloscope with a sensitivity of 100mV/cm. This requires an amplifier with a gain of 100, and some form of range switching must be used to prevent the amplifier from overloading on large inputs. This is conveniently done by an attenuator in the collector circuit of the common base stage, but this modification prevents the method of improving the l.f. response shown in Fig. 5(b) from being used. However, the same result may be achieved by shaping the l.f. response of the amplifier.

Conclusions

It has been shown that wide-band current transformers may easily be constructed, using small ferrite ring cores. A current probe may be made by using such a transformer in conjunction with a suitable amplifier. Transformer bandwidths of up to 200Mc/s have been measured, so the overall bandwidth of the probe is likely to be determined in the amplifier.

The probe could be made easier to use if the transformer core were split, enabling it to be clipped round a conductor instead of having to break the circuit in order to thread the conductor through the core. The effect of the small air-gap in the split core is to reduce the l.f. response slightly. It is, however, difficult to distribute the secondary winding evenly round a split core and some flux leakage takes place, resulting in a reduction in transformer bandwidth. In this context, it is noted that commercially available instruments using clip-on probes have bandwidths of about 20Mc/s.

Apart from its use as a test instrument, the current transformer could also be incorporated in production equipments as a current monitoring point. If the secondary were connected to a shorting-type jack socket, a single amplifier could be used in conjunction with many transformer test points.

Acknowledgment

The authors wish to thank Professor R. L. Russell of the Department of Electrical Engineering, University of Newcastle upon Tyne, for the use of the facilities of the Department. Thanks are also due to Mr. S. W. Noble, of the Physics Division, R.R.E., for making available facilities for measurement of the current transformer transient response.

One of the authors (J.W.) was in receipt of a D.S.I.R. Studentship at the time the work described was carried out.

A Miniature Photo-Electric Edge Guide

Hird-Brown Limited have devised a comprehensive system of photocell controls for edges of materials, utilizing a range of miniature fork-type scanning heads. These scanning heads, which contain a light source and a photo-transistor, operate reliably in all ambient lighting conditions and are even insensitive to changes in daylight. The distance between the lamp and cell can be varied between 1/16in. and 1½in by varying the dimension of the centre distance piece.

Both the photocell and lamp units are fitted with lenses, which results in the relay(s) being consistently operated at a definite cut-off point. Normally two heads are mounted side by side, but staggered, at one edge of the material, each operating independently a separate relay to give a signal 'right' or 'left'.

The control relay unit, which is housed in a sheet steel case size 10in by 6in by 4in, contains all necessary electronic circuits for the independent operation of both relays, and the control units will operate the relay immediately at each change in light caused by the material obscuring the beam.

Alternatively, time delays are incorporated so that rough edges of material causing temporary shadows are ignored. Additionally, a time delay governing the relay action time can be incorporated either on its own or together with the time delay previously mentioned. The relays have two sets of change-over contacts each rated at 5A at 240V a.c.

The fully transistorized amplifier unit can be mounted at a distance from the viewing position, which is often an advantage where centralized control of a process is required.

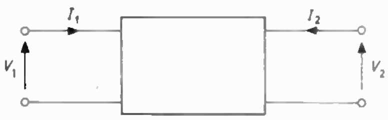
LETTERS TO THE EDITOR

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

Signal Flow Graphs

DEAR SIR,—When dealing with transistor circuits as defined by a four-terminal linear matrix, it was decided to attempt to apply 'Signal Flow Graph' techniques to various problems, to see whether a short solution to the problem of parameter transformation could be obtained.

Hybrid, h, parameters are the most common set used and are defined as below:



$$\begin{bmatrix} V_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix} \begin{bmatrix} I_1 \\ V_2 \end{bmatrix}$$

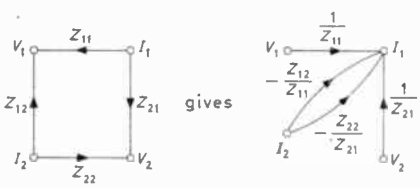
other commonly used parameters are the Z and Y sets:

$$\begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} Z_{11} & Z_{12} \\ Z_{21} & Z_{22} \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix}$$

and

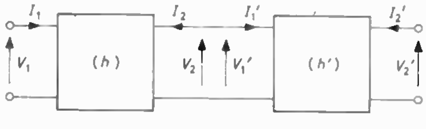
$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} Y_{11} & Y_{12} \\ Y_{21} & Y_{22} \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \end{bmatrix}$$

the Y parameters can be obtained from the Z set by straightforward matrix inversion. However, attempting to use the s.f.g. technique to perform this inversion breaks down as shown in the following s.f.g.'s.



Since for the Y set of parameters V_1 and V_2 are required to be 'sources', and I_1 and I_2 'sinks', and as can be seen by investing Z_{11} and Z_{21}, to make I_1 into a 'sink', one can only obtain I_1 in terms of V_1, V_2 and I_2.

However, where inversion is not required some useful results can be obtained. For example, to find the combined h parameter for two cascaded networks of h parameters, as shown in the diagram.

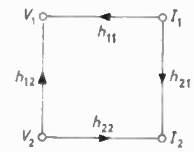


The conventional approach is tedious and time consuming, the s.f.g. technique gives a fast solution.

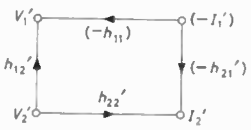
Letting the combined parameters be h''; then h'' will be defined as below:

$$\begin{bmatrix} V_1 \\ I_2' \end{bmatrix} = \begin{bmatrix} h_{11}'' & h_{12}'' \\ h_{21}'' & h_{22}'' \end{bmatrix} \begin{bmatrix} I_1 \\ V_2' \end{bmatrix}$$

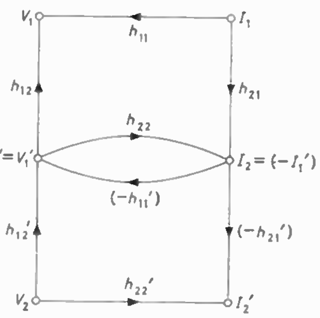
For the first network the s.f.g. is:



and for the second network is similar, h being replaced by h'. However, since $I_2 = -I_1'$, a slight modification to the s.f.g. for h' is required as shown below:



Now, $V_2 = V_1'$ and $I_2 = (-I_1')$, also V_2 and $(-I_1')$ are 'sources' and I_2 and V_1' are 'sinks', thus one can combine the two s.f.g.'s.



Now since I_1 and V_2' are the only two 'sources' present and V_1 and I_2' are the only two 'sinks' present, one can write down by inspection V_1 and I_2' in terms of I_1 and V_2'.

Hence:

$$V_1 = I_1\left[h_{11} + h_{21} \cdot h_{12}\left(\frac{h_{11}'}{1-(-h_{11}')h_{22}}\right)\right] + V_2'\left[\frac{h_{12}'h_{12}}{1-(-h_{11}')h_{22}}\right]$$

and

$$I_2' = I_1\left[\frac{h_{21}(-h_{21}')}{1+h_{22}h_{11}'}\right] + V_2'\left[h_{22}' + \frac{h_{12}'h_{22}(-h_{21}')}{1+h_{11}'h_{22}}\right]$$

Thus

$$h_{11}'' = h_{11} - \frac{h_{12}h_{21}h_{11}'}{1+h_{22}h_{11}'}$$

$$h_{12}'' = \frac{h_{12}h_{11}'}{1+h_{22}h_{11}'}$$

$$h_{21}'' = \frac{-h_{21}h_{21}'}{1+h_{22}h_{11}'}$$

$$h_{22}'' = h_{22}' - \frac{h_{12}'h_{21}'h_{22}}{1+h_{22}h_{11}'}$$

Now that the combined h parameters have been found, the cascade network can be considered as a single element h'', and system gain can be calculated in terms of h'', for which a simpler expression exists.

Also if CR coupling exists between the two cascaded elements, the h parameters can be written down for the coupling network from simple circuit theory. Then the combined parameters for the first network and coupling network can be found from the previous equations. Using these parameters in conjunction with those from the second network, and the previous set of equations, the overall expression can be found. Thus any number of cascaded networks can be dealt with, and equated to a single network.

Yours faithfully,

O. J. WHEATON,
Loughborough College
of Technology.

REFERENCES

1. MASON, S. J. Feedback Theory—Some Properties of Signal Flow Graphs. *Proc. Inst. Radio Engrs.* *41*, 1144 (1953).
2. MASON, S. J. Feedback Theory—Further Properties of Signal Flow Graphs. *Proc. Inst. Radio Engrs.* *44*, 920 (1956).

Transistor Low Drift D.C. Amplifiers

DEAR SIR,—Mr. Pinto's article in your issue of May 1964 has interested me very much. May I be permitted to add to his contribution by raising a simple point which is in danger of being overlooked.

Low drift amplification is a common requirement in d.c. stabilizers, in which there is one great advantage: gain stabilization is relatively unimportant. Furthermore, the sampling and reference networks are usually of low resistance so that only drifts arising from the *so called* V_{be} changes are important. I say 'so called' because, as most writers fail to point out, this source of drift actually results from a change in I_{eo}, the emitter saturation current, and the net drift only manifests itself truly as a change in V_{be} when the drift is physically referred back to the input by negative feedback as in the case of power supply stabilization.

In this particular application, then, very good results are obtained using a simple balanced amplifier but it is vital that the emitters be directly coupled,

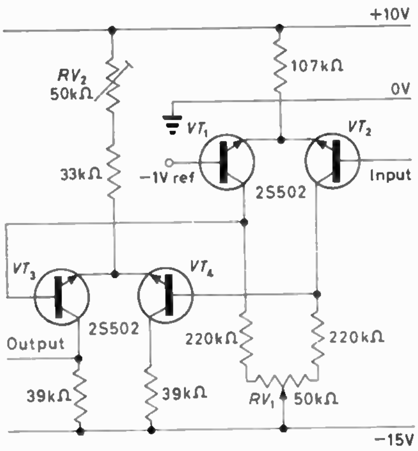


Fig. 1. Stabilized amplifier

zero adjustment being made in the collector circuit. Having chosen a suitable working current, the V_{be}'s are matched at that current (planar transistors having very close tolerances in this respect as the author points out), whence simple theory shows that the net drift is

$$\frac{V_{be1} - V_{be2}}{T} \text{ V/°C.}$$

Thus if the ambient temperature can be held within 0·1°C and a maximum drift of 1μV were sought the V_{be}'s must be matched to better than 3mV—not a difficult task. I have assumed, of course, that ohmic effects are negligible. This is justified for a typical base current of 1μA since r_{bb}', which represents the lumped ohmic resistance of the junction referred to the base, is of the order of 300Ω, giving a value of 0·3mV for V_{bb}', the neglected portion.

The practical stabilizer amplifier requires at least two stages in balanced mode after one can utilize the single-ended signal. The diagram (Fig. 1) shows a typical design with which I have obtained a drift rate of less than 3μV/h without the use of a temperature controlled oven. The choice of balancing controls is very important and they are set up as follows: Short the bases of VT_1 and VT_2 together. Adjust RV_1 for equal collector voltages. Unshort bases and adjust RV_2 until the stabilizing loop equalizes the potentials of the bases of VT_1 and VT_2.

Yours faithfully,

N. Bett,
University of Cambridge.

Strain Rate-of-Change Switch

Dear Sir,—In order to obtain high speed photographs of crack propagation in selected specimens, a request was received for a crack detector based on the change of strain rate in the specimen. The very simple device described below was put together and tested within several hours and performed satisfactorily.

A strain gauge (or gauges) was attached to the specimen at right-angles to the direction of crack propagation and in the predicted path of the crack. This active gauge was connected in a bridge circuit which was energized from a battery. The bridge output being fed to a large capacitor, output from the capacitor appeared only when strain *changes* occurred or when the rate-of-change of strain was altered. Balance of the gauge bridge was unnecessary.

In order that the sensitivity to 'roughness' in the (say) testing machine be prevented from triggering the device, gain control is essential. This is provided from the decade d.c. wide band amplifier and the resistance R—which is also useful for reducing mains ripple pick-up. The circuit shown in Fig. 1 includes a relay drive unit which incorporates a pair of diodes for output limiting together with Zener diode protection against induced voltages arising from relay operation.

As the device is unidirectional it is necessary to watch the polarity of the

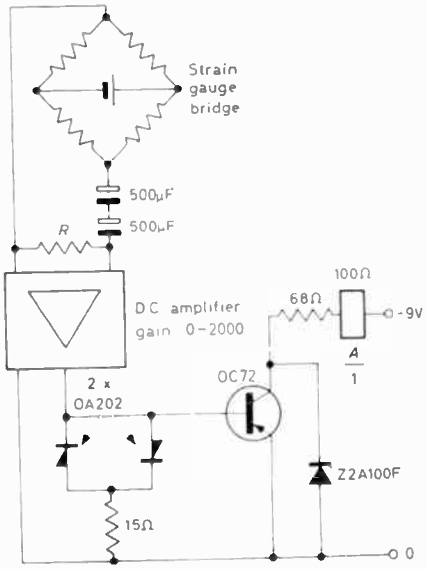


Fig. 1. Rate of change-of-strain measurement

battery together with the direction of strain change.

Permission to publish this letter has been granted by the Chief Scientist, Australian Defence Scientific Service, Department of Supply, Melbourne, Australia.

Yours faithfully,

I. G. Scott
H. A. Ramsbottom,
Grad., I.E. Aust.
Australian Defence Scientific Service.

Transistors as Rectifiers

Dear Sir,—As Mr. Linsley Hood points out (January 1964) a transistor may be used as a diode with low forward voltage drop if its base is connected to its collector.

I have used the OC74 in this configuration but, where space is not a problem, prefer to use the diode OA31 which

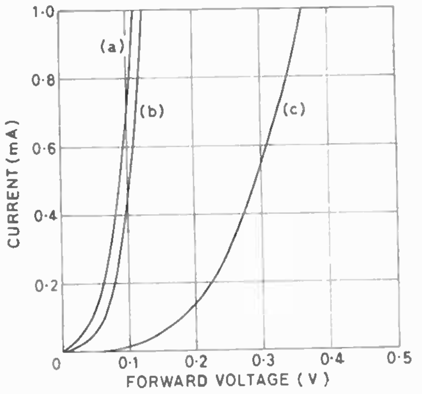


Fig. 1. Forward characteristics of (a) OA31, (b) OC74 (as diode), (c) OA95

has a slightly lower voltage drop. See Fig. 1 comparing OA31, OC74 (diode connected) and OA95 curves.

Furthermore, reference to the published data would show the OA31 to be generally superior to the 'diodes' in Fig. 2 of Mr. Linsley Hood's letter.

Perhaps this diode junction would be suitable for manufacturing full-wave units for meter rectifiers.

Yours faithfully,

F. J. Maher
Commonwealth Scientific and Industrial Research Organization, Australia

In Reply:

Dear Sir,—While it is undoubtedly practicable, as Mr. Maher suggests, to use high current diodes, of the OA31 type, as low current rectifiers having a low forward voltage, the disadvantage usually encountered in this usage is that these diodes have a relatively high reverse leakage current. This difficulty is, in fact, suggested by the makers' published characteristics for the OA31, which is quoted as having a reverse leakage current, at −1·0V, of 25μA at 25°C, increasing to 1·5mA at 75°C.

Although CB/E connected transistors ('super-diodes') have a leakage current which is higher than that characteristic of a normal diode junction, the magnitude of this leakage current in a small dissipation transistor will normally be much lower than that of a power rectifier, such as the OA31. The performance of such a device would, therefore, be expected to be better as an instrument rectifier than a power rectifier having an identical forward characteristic. The comparisons I have made in my letter, in your January issue, were between diodes which were considered to be of similar junction area and degree of doping, but as Mr. J. W. McPherson subsequently pointed out (March 1964), much better performances than those quoted could be obtained with other transistor types in this configuration.

Yours faithfully,

J. L. Linsley Hood
British Cellophane Ltd.

BOOK REVIEWS

Basic Electric Circuits

By A. M. P. Brookes. 131 pp. Crown 8vo. Pergamon Press. 1963. Price 10s.

THIS book is number 123 in a new series of soft-cover technical books published by Pergamon Press under the general title of "The Commonwealth and International Library of Science, Technology, Engineering and Liberal Studies". The editor for the Electrical Engineering Division of the series is Mr. P. Hammond of the Engineering Department of Cambridge University, and the author of the present volume is also at Cambridge. In the preface it is stated that the book is based on lectures given to "undergraduates in their first or second year of a university engineering course". The content may well be suitable for courses of the Cambridge type where engineering is taught in the wider sense; however, most universities now have separate departments of electrical engineering, and for such departments the present book is insufficient for the requirements of the first year.

The first two chapters deal with the nature of resistance, capacitance and inductance. Chapter three covers resistive circuits and includes Kirchhoff's and Thévenin's theorems, together with the star-delta transformation. This accounts for nearly half of the book and subsequent chapters consider the current and voltage relations in capacitance inductance and in *RCL* circuits. The book concludes with a chapter on graphical methods and one on simple transients. The operator j is not mentioned and presumably the standard methods of solving a.c. circuits are intended to be taught later. The book is well produced and not unduly expensive, but it seems to have little to recommend it in preference to existing texts.

V. H. ATTREE.

Boolean Algebra and its Application

H. G. Flegg. 262 pp. Med. 8vo. Blackie & Son. 1964. Price 50s.

MORE elaborate electronic apparatus, miniaturization, automation and computers have all led to an increasing interest in switching problems. This interest is by no means spent. Greater use of printed circuits and other similar developments which need plane circuits are likely to produce even more interest.

This book is an admirable account of the logical analysis of these problems, using Boolean algebra. The book is very pleasing and it should be popular. The author, who is at Cranfield, assumes no more than a background knowledge of the electrical problems and, having developed from scratch the algebra he requires, applies it skilfully to a number of examples. A variety of techniques are described which simplify the solution of the more elaborate problems including the use of tables, graphs and other devices, all discussed in detail in the middle chapter.

The final chapters on the matrix formulations were found fascinating. There does seem scope here for extensions of the method which will make it possible to handle very large numbers of variables. Two brief comments on the structure of the book. Each chapter contains a number of exercises which are carefully chosen and especially useful in a 'course' book. Secondly, the author has deliberately avoided mention of the electrical 'hardware' to which the theory is applied. Although this is almost certainly right for engineers, it is doubtful whether it is so for the mathematics students and the teachers who he hopes will use his book. The students ought to be kept down to earth and the teachers might have to explain the hardware to their intolerant and knowledgeable pupils.

G. J. KYNCH.

Properties and Applications of Transistors

By J. P. Vasseur. 434. Med. 8vo. Pergamon Press. 1964. Price £5

WITH a field as large as that implied by "Properties and Applications of Transistors", it is to be expected that an author will select, for emphasis, those parts of the subject with which he is most familiar and give briefer coverage to other parts. In this book chapters 5 and 6, which deal with "Bias Circuits" and "Maximum Ratings of a Transistor" respectively, stand out as those which have received the author's main attention; chapter 7 on "Transistor Noise" is also reasonably good.

The treatment of the majority of the remainder of the material leaves much to be desired. The text as a whole is devoted almost exclusively to the 'uniform base resistivity' or 'diffusion' transistor—the drift transistor is dismissed in less than two pages, whereas several pages are devoted to the long-obsolete point contact transistor. There are too many categorical mis-statements of the kind: "All transistors are in practice manufactured from single crystals of germanium or silicon", p. 2; "The current gain α, equal to unity to a first approximation, is always slightly less than 1", p. 32; "Phototransistors . . . are simply transistors in which the base is not connected and can be illuminated", p. 54—and many more. There is an unfortunate choice of terminology in section 1.2, where the dynamic depletion-layer or transition-region capacitance is referred to as a static capacitance. The significance of this capacitance is not really brought out.

It is somewhat surprising to find no mention of non-linear applications at all; there is nothing on oscillators, waveform generation or pulse amplifiers. Taken overall this is a disappointing book.

F. J. HYDE.

Assembly of Radio and Allied Electronic Equipment

By V. O. Bukler and Yu. I. Rabinovich. 250 pp. Med. 8vo. Oldbourne Press. 1964. Price £6

THIS is a translation of a Russian book published in 1960, and translated by The Israel Program for Scientific Translation in 1963.

The first comment is that the type is extremely small, as also are the illustrations, and the photographs are poor in definition.

The book is concerned with radio, radar and electronic production techniques and was written primarily for radio equipment assemblers. It is concerned solely with valve equipment; transistors are mentioned only as a new technique just being introduced. The first few chapters are elementary and cover mechanical tools and their use; for example, pictures are shown of hacksaws, files, drills, hammers, etc.

Considerable detail is given on soldering processes and the making of soldered and welded joints, but no mention is made of wrapped joints. Surface treatment of metals for chassis, etc., are covered in an elementary but thorough fashion. Inductor and transformer winding and construction is fully covered, and also the assembly of variable capacitors. Potting and sealing of components and assemblies is covered for protection against climatic effects, and a quite large chapter is devoted to the actual assembly of components and the manufacture of chassis. Printed circuits are covered in a detailed and practical manner, with particular reference to radio and television production. It is interesting to note the use of printed wiring transformer windings, which was published in this country many years ago. The assembly of radar and u.h.f. equipment is dealt with separately, again in an elementary but thorough manner. The penultimate chapter covers the assembly of electrical measuring instruments and shows many devices and jigs for assembly and adjustment. The final chapter briefly deals with inspection and quality control.

The book is undoubtedly a good record of assembly practice of valve electronic equipment and as such is a useful record of practice in the U.S.S.R. The value of the book is somewhat doubtful as most people engaged in the electronic industry are already aware of these techniques and indeed are more concerned with the assembly of transistorized equipment. The price is somewhat high for the amount of information contained.

G. W. A. DUMMER.

Stereophony

N. V. Franssen. 85 pp. Med. 8vo. Philips Technical Library. 1964. Price 21s.

PERHAPS the most noteworthy thing about stereophony is the length of time it has taken to become generally available. All the equipment has been at hand since Harvey Fletcher started his experiments in 1933.

It is difficult to describe methods which are not stabilized and about which there is no recognized technique. Thus it is possibly as well that Mr. Franssen makes no attempt to specify system designs for stereophonic sound, but confines himself to the mechanism of auditory perspective. The Philips laboratories have made many significant contributions to electroacoustics, and this book summarizes most of the information currently available. In dealing with effects which can only be assessed subjectively, the author is faced with extreme difficulties. For example, a musician does not interpret the mixed sound spectrum in the same way as a non-musical listener. A technician may worry far too much about pass-bands and tends to listen for things to satisfy his knowledge. There is as yet no finality in stereophonic balance and the many systems all perform differently. Often multi-channel reproduction does not enhance the material at all.

Mr. Franssen summarizes the whole subject in one sentence when he says: "it is not so much the interpretations to be placed on the reading of one type of voltmeter or another, but the subjective aural assessment that is the ultimate measure." Bearing this in mind, it is not surprising to find a good deal of qualification of some of the experiments described in the book. The author has made a very good job of a controversial subject, the book is full of authentic detail and for those interested in the theories of binaural hearing it is excellent value. As is usual with Philips publications, the type and illustrations are good and the author must indeed be pleased that production took less than six months. Altogether a very welcome addition to the bookshelf.

A. DOUGLAS.

Electronics for Students of Mechanical Engineering

By H. Sutcliffe. 353 pp. Demy 8vo. Longmans Green & Co. 1964. Price 37s. 6d. (soft cover) 50s. (cased edition)

ELECTRONIC methods are widely used by mechanical engineers and many courses for students of mechanical engineering now have some electronic content. Relatively few electronic textbooks are suitable for mechanical engineers and the present book is intended to make good this deficiency. A mechanical course would be expected to include material in the field of heavy electrical engineering so that the time available for purely electronic studies is strictly limited. In the preface the author states that "the text could be read by most professional mechanical engineers and second-year students of engineering at universities and colleges of technology". Mechanical engineers may well disagree with this statement as the book assumes a fair knowledge of theory and the design of circuits is considered in some detail.

The introduction discusses the specification of linear detecting systems and chapter two gives a summary of the required electric circuit theory. This theory includes the circuit theorems, j notation and transients. Chapter three deals with various electronic devices including semiconductors. The pn junction, transistor, silicon controlled rectifier and Zener diode are all covered in the space of seventeen pages. The next three chapters deal with the application of these devices in actual circuits. Extensive use is made of the load-line construction but the treatment is probably too detailed for mechanical engineers. Chapter seven covers transducers and chapter eight gives a full discussion of the rather specialist topic of modulated systems. Feedack is then considered and the book is completed by a description, with detailed circuit diagrams, of a number of electronic instruments. This book contains much that is of interest to a mechanical engineer but does not appear to be wholly suitable for use by students.

V. H. ATTREE.

Automat und Mensch (Automatic Machines and Man)

By Dr.-Ing. Karl Steinbuch. 392 pp. Demy 8vo. 2nd edition. Springer Verlag. 1963. Price DM 36

A SECOND, enlarged edition two years after the first one shows an unusual interest by the wider public. Some indication of the encouragement received by the author may be seen in the change of the sub-title from "About human and mechanical intelligence" to "Cybernetic Facts and Hypotheses". The first issue brought a great deal of correspondence, little of which was negative. The criticism has been grouped under four headings which give the key to the author's thinking. To those pointing out factual deficiencies he bluntly answers that even the second edition will not be free of mistakes and errors. Others claim that he goes beyond what is permissible, and those critics are asked who fixes what is permissible. He shows little patience with those telling him that what he says runs counter to what they have learned in their lifetime. The last point raised, "The consequences of his philosophy are unbearable", engages his mind very much. In the last chapter of the book he deals with this question by stating why writing it seemed essential to him. Whether one agrees with him or not, the book is still fascinating reading for those with a fluent command of the German language.

E. R. FRIEDLAENDER.

Electronic Circuits

By E. J. Angelo, Jr. 652 pp. 2nd Edition. Med. 8vo. McGraw-Hill Book Co. 1964. Price 89s.

This book was first published in 1958 with the object of serving as a basis for one-year courses in electronic engineering.

Among the changes in the second edition is the addition of two chapters, one on power amplifiers and the other on controlled rectifiers.

The use of the Laplace Transform for circuit analysis studies has also been included.

Diodes and Transistors General Principles

By G. Fontaine. 470 pp. Demy 8vo. Philips Technical Library. 1963. Price 55s.

This book provides a detailed, non-mathematical treatment of the fundamental section of semiconductor devices, their characteristics and equivalent circuits.

Considerable use is made of diagrams, many of which are in colour.

Abnormal Loads on Power Systems

By J. E. L. Robinson. 186 pp. Demy 4to. Institution of Electrical Engineers. 1964. Price £4 10s.

This book is concerned with three types of Abnormal Load—those which are intermittent in character and cause variation in voltage, and those which distort the voltage waveshape or unbalance the phase voltages.

The subject is of interest to engineers working in the fields of electricity supply, telecommunication, broadcasting, computers, steel production and electric transport.

The Radio Amateur's Handbook

Edited by B. Goodman. 720 pp. 41st Edition. Med. 8vo. American Radio Relay League. 1964. Price $3.50

The Forty-First Edition of The Radio Amateur's Handbook has been published on the occasion of the fiftieth anniversary of the American Radio Relay League, 1964, by the Headquarters Staff of the American Radio Relay League (Byron Goodman, WIDX, Editor).

Short Wave Amateur Radio

By J. Schaap. 160 pp. Demy 8vo. Philips Technical Library. 1963. Price 21s.

This book is intended primarily as a guide for those radio amateurs who wish to specialize in short wave transmissions. It tells the beginner what these amateur transmissions entail and what he has to know to be able to practise this hobby.

The author's aim has been to present the newcomer with a concise reference book which will introduce him to amateur radio; he has not attempted to give a complete treatment of the subject.

Computer Organization

By A. A. Barnum and M. A. Knapp. 242 pp. Med. 8vo. Cleaver-Hume Press. 1963. Price 60s.

This volume comprises the eight papers presented and the subsequent discussions at the Workshop on Computer Organizations held on October 2 and 3, 1962 at Westinghouse Air Arm Division in Baltimore, Md. The conference, which was sponsored by the Information Processing Laboratory of Rome Air Development Center and the Westinghouse Air Arm Division was attended by approximately 80 people who were at the time concerned directly with computer organizations.

Microwaves (Proceedings of the 4th International Congress)

813 pp. Demy 4to. Centrex Publishing Co. 1963. Price £13

This volume records in full the 138 papers presented by some 200 authors at the Fourth International Congress held in the Netherlands in September 1962.

ELECTRONIC EQUIPMENT AT HANOVER FAIR

A description, compiled from information supplied by the manufacturers, of some of the German exhibits shown at the Hanover Fair from 26 April to 5 May 1964.

(Voir page 497 pour la traduction en français; Deutsche Übersetzung Seite 504)

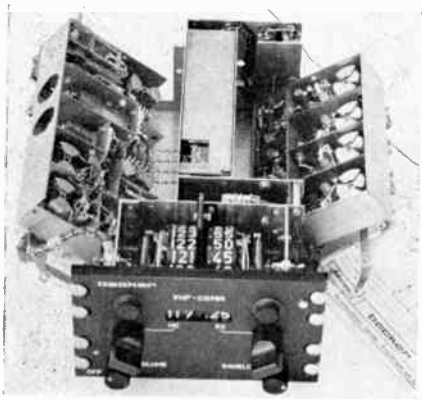
Becker Flugfunkwerk GmbH

Flugplatz, 757 Baden-Baden-Oos, Germany

AIRBORNE TRANSCEIVER

(Illustrated below)

The AR380 is a compact, crystal-controlled transceiver offering 380 transmit and receive channels at 50kc/s spacing within the frequency range 117·0 to 135·95Mc/s. Receiver and power/modulator unit are fully transistorized, but the transmitter employs three valves to obtain an output of 10W. Low



equipment weight as well as ruggedness and low power consumption are achieved by the use of transistors.

With dimensions of 3·75in × 5·75in the front panel corresponds to ARINC standards; it is illuminated from inside and large figures as well as suitably shaped controls facilitate accurate frequency setting even in the dark. A monitor tone, generated in a supplementary wide-band amplifier during the transmission and fed into the headphones after amplification, safeguards against unobserved failure of the equipment or aerial system.

It is an important feature of the AR 380 that it is designed in unit construction. The installation fulfils all requirements of standards and regulations in force; the transceiver weighs approximately 5½ lb, the power pack 4½ lb.

EE 71 751 *for further details*

Bölkow-Apparatebau GmbH

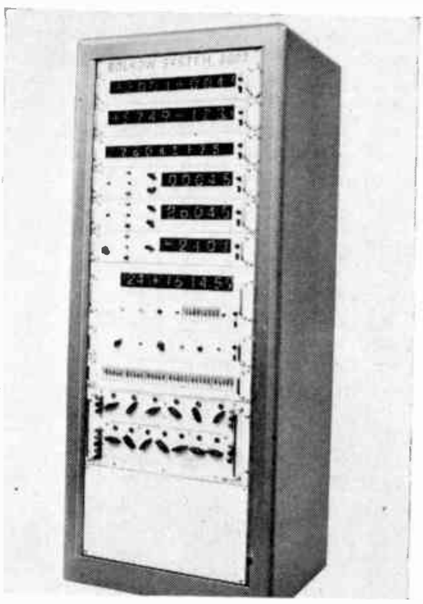
7311 Nabern-Teck, Germany

DATA LOGGING SYSTEM

(Illustrated on right)

The Bölkow System 2000 is a data logging system in which the very divers values of temperature, pressure, chemical composition of gases, acceleration, vibration, flow, etc., to be monitored are converted into d.c. and a.c. voltages by means of transducers. The accuracy of the transducers varies between 0·1 and 10 per cent according to type and version. Basic units of the system are the analogue-digital convertor and the digital frequency meter which measure voltages with an accuracy of approximately 0·01 to 0·2 per cent, frequency and time with one of 1×10^{-5} to 1×10^{-7}, and feed them to other units of the system for processing and registration. Available for this purpose are digital clocks, code convertors, digital read-outs, printers, digital preselector units and programmed pulse generators.

Complicated and extensive studies of development and research problems have to be broken down into investigations of their different aspects and it has proved very advantageous that instruments may be used for these part studies which can afterwards be integrated into a large programmed system for the examination of the project as a whole. Special control and computer systems for the solution of small or part problems can be designed using standardized digital units of the Bölkow System 2000. Large installations can be assembled using digital computers of Compagnie des Compteur (SETI) which may be connected into the System 2000 direct. All units of the Bölkow System 2000 are supplied for 19in rack mounting.



EE 71 752 *for further details*

braun electronic gmbh

109 Lange Str., 7808 Waldkirch/Breisgau, Germany

CONTACTLESS TEMPERATURE MEASUREMENT

Individual units of an electronic measuring system for temperature radiation were introduced by braun electronic gmbh under the designation 'TASTOTHERM-INFRA'. Equipment for the contactless measurement in standard ranges between +100°C and +3 000°C consists in every case of a sensor connected to a basic unit comprising amplifier, power pack and—where required—limit switches.

The signal is available as an electrical quantity and the method of measurement lends itself readily to remote indication and registration of temperature as a function of time. The incident heat radiation acting on the sensor is converted into an electrical signal and pre-amplified. The size of the measured spot and its distance from the sensor may be chosen freely between certain limits. In addition, the emission factor of the object to be measured has to be considered, i.e. the ratio between the radiation power of the material in question and that of a black body, as well as the diameter of the aperture through which the measurement is taken. Equipment is either supplied for special-purpose installation for a given material or as a multi-purpose installation calibrated for several emission factors.

EE 71 753 *for further details*

Brown, Boveri & Cie AG

68 Mannheim, Germany

MICROWAVE EQUIPMENT

(Illustrated on page 491)

BBC introduced its new fully transistorized equipment FM12/7000 for microwave links operating at 7Gc/s. The exclusive use of semiconductor techniques has created equipment with a new concept, employing silicon transistors and diodes exclusively. Thus, complicated microwave tubes are eliminated, consumption is reduced to a fraction of what has hitherto been usual, and crystal control improves the fre-

quency to such a degree that channel spacings of 1Mc/s can be maintained.

The equipment is designed for use in microwave links and covers the range 6 400 to 8 400Mc/s, accommodating either 12 or 24 telephony channels. The unit is also suitable for the transmission of a high quality music channel; it has a consumption of 10W. In the standard version, transmitter, receiver and aerial reflector are assembled into one unit; it is, however, possible to insert waveguides of up to 165ft length where remote installation of the reflector is desired.

EE 71 754 *for further details*

DeTeWe
Deutsche Telephonwerke und Kabelindustrie AG

100 Wrangelstrasse, 1 Berlin 36, Germany

MULTI-PURPOSE MEASURING EQUIPMENT

Various different measurements and tests required in the operation of telephone exchanges and control engineering can be carried out with this new combination equipment comprising a universally applicable pulse generator and a timer/frequency meter.

Five decade switches are provided for the selection of the pulse width and a further five for the adjustment of the interval between pulses. The number of pulses generated may be preset between 1 and 12, or they may be continuous and pulse trains may be repeated automatically.

Sine wave or pulse inputs are counted over periods of 1 or 10sec and the result is shown by illuminated figures with decimal point indication. The range from 10 to 99 999c/s is covered for frequency measurements.

During the measurement of time high stability oscillations of the time-base are fed to the electronic counter via a gate, i.e. the gate is controlled by the input signal and the digital read-out corresponds to the time during which the gate is open. A time range from 1 to 99 999msec is covered by the equipment.

EE 71 755 *for further details*

Friesecke & Hoepfner GmbH

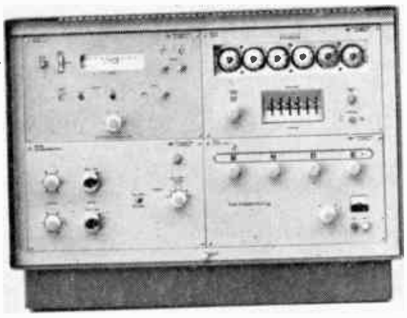
P.O. Box 72, 8520 Erlangen-Bruck, Germany

NUCLEONIC INSTRUMENTATION

(Illustrated below)

As the result of further developments of its range of nucleonic instruments, Friesecke & Hoepfner showed for the first time the transistorized building blocks of their transistor equipment programme TGP. These building blocks are housed in standardized plug-in units UC/2, two of which fit next to each other into the standard 19in rack-mounting unit. At the same time bench cabinets have been designed for the various combinations.

The equipment illustrated consists of the high tension generator FHT 101A1, the linear rate meter FHT 110B, the pulse counter FHT 120A2 and the timer FHT 120A3. The mains-operated high tension generator provides for the detectors voltages between 200 and 500V at a maximum of 5mA, a stability of better



than 0·2 per cent over 24 hours, and a switching facility for negative and positive polarity. The absolute accuracy of adjustment is ±2 per cent, the smallest step being 1V.

The linear rate meter indicates the mean time value of a pulse train in eleven ranges between 10^2 and 10^7p/min. The two pulse rate inputs can also be connected against each other to provide a differential circuit. The pulse output from the radiation detectors is amplified and fed into the pulse counter which is fitted with a six digit decade counting tube read-out. A count control is incorporated, but terminals marked 'multi-purpose control' permit connexion of associated control systems, e.g. of the single channel pulse height analyser FHT 300A. The built-in automatic stop can be preset in tens. The crystal-controlled timer provides the time control for the measurement with a stability of better than 1×10^{-3} and a resolution of 1µsec. The digital read-out of the values may be used to drive a printer.

The range also covers all the necessary amplifier units and nearly all building blocks can be supplied for either mains or battery operation.

EE 71 756 *for further details*

fuba, Werk elektronischer Bauteile u. Geräte
Hans Kolbe & Co.

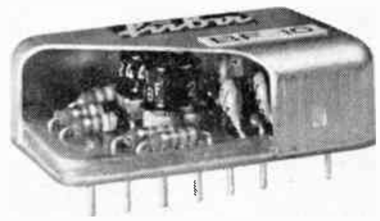
3371 Gittelde üb. Seesen, Germany

CURRENT-STABILIZED 400A POWER UNIT

The basic development of this unit has been carried out by Deutsche Elektronen Synchrotron (DESY) at Hamburg. It provides the power supply, for instance, for deflexion magnets of particle accelerators and other inductive loads requiring a high and stable direct current. The output current can be selected in steps of 40mA by means of a 4-digit numerical switch, e.g. it may be set to 384·8A or 38·48A. Once the current is set it is substantially independent of input voltage or load fluctuations. The maximum nominal output is 16kW and the output voltage may be selected between 0 and 40V.

The 20kVA mains transformer and the 400A choke are permanently installed in a cabinet with overall dimensions of 22·8 by 21·8 by 64·6in. The functional units are designed for 19in rack mounting. Additional provision has been made for external operational control and monitoring facilities on request of the customer. The equipment weighs approximately 880 lb.

EE 71 757 *for further details*



DIGITAL PLUG-IN UNITS

(Illustrated above)

The digital building blocks of the system Darmstadt, first shown in 1963, are now available in three speed series for clock frequencies of 250kc/s, 5Mc/s, and above 10Mc/s. The important advantage of the system is the strict standardization of the matching requirements of the individual units which considerably simplifies the design of equipment. Standardized connexions facilitate the use of combinations of different series within one circuit.

Flip-flops, invertors, crystal oscillators, pulse shapers, output amplifiers and drivers for Nixie tubes are available.

EE 71 758 *for further details*

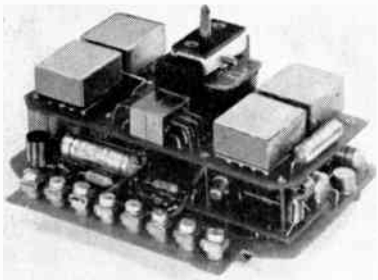
P. Gossen & Co. GmbH

25 Nägelsbachstr., 8520 Erlangen, Germany

D.C. AMPLIFIER

(Illustrated on page 492)

The output voltages of thermocouples are too low at low temperatures to drive even highly sensitive moving-coil instruments reliably. The measurement of even the lowest output voltages of thermocouples with robust standard moving-coil movements has, however, now been achieved by means of magnetic amplifiers. A demonstration model showed how the voltage of 3·19mV generated by an iron-constantan thermocouple at +60°C (140°F) is converted into a current of 1A by the d.c. pre-amplifier PMV10, permitting reliable indication of the temperature range 0° to +60°C on



an insensitive moving-coil ammeter with 250° pointer deflexion.

A square wave oscillator supplies the magnetic amplifier with an auxiliary voltage at 1kc/s. Small dimensions, a low time-constant (<0·1sec) and low output current ripple are achieved by the choice of this frequency. The pre-amplifier may be operated with 24V d.c. or 220V, 50c/s a.c. and can be mounted on the base plate of panel instruments with flanges of 5·7in × 5·7in from the rear.

EE 71 759 *for further details*

Gebr. Grieshaber

P.O. Box 30, 762 Wolfach, Germany
Distribution: VEGA Gmbh, P.O. Box 1649, 78 Freiburg/Breisgau, Germany

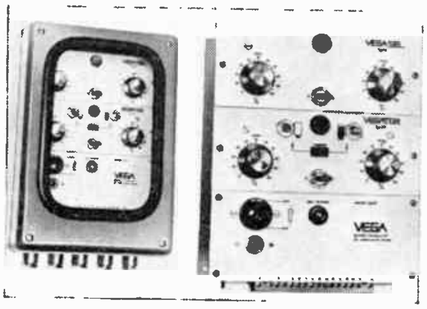
Level Controller

(Illustrated below)

The new transistorized level measuring equipment exhibited by Grieshaber employs silicon planar transistors to meet the demands for the greatest possible reliability. The design uses the capacitive measuring method, but determines the current of the capacitor instead of its voltage; in this way an electrical value is obtained automatically which maintains a linear proportional relation to the variable capacitance.

Long unscreened leads have been made possible by transferring the oscillator to the head of the electrodes. The electrode head insert as well as the electronic unit proper may be used in applications covering the range 10 to 10 000pF without changing any of the electronic components, and without affecting the response of the limit indicators to capacitance changes of less than 0·5pF. Particular attention has been paid to the thermal stability of the unit.

The equipment is designed in unit construction. The illustration shows the power supply common to all models, together with the new double limit switch Vegator/Vegasel as wall mounting model and as supplied for incorporation into other equipment.



EE 71 760 *for further details*

Grundig Werke

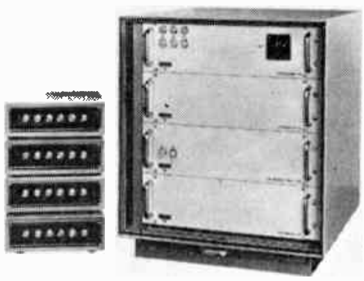
8510 Fürth, Germany

Numerical Machine Tool Control

(Illustrated below)

This numerical positioning system for machine tools—first exhibited in Milan six months ago—was introduced as an improved system and with 8-channel punched tape input. It is a positioning control with 'control information' in building block construction which can be assembled according to requirements and for a wide range of machines, and allows for expansion to suit additional requirements at a later date. In this context 'control information' refers to the control of additional functions of the machine, e.g. different feeds and speeds, change of tools or supply of cooling agents. The numerical control can be supplied for one to three co-ordinates, and in the version for two co-ordinates up to 20 control information types may now be programmed in four independent groups.

The control rack is assembled from plug-in units and provision is made for a readily exchangeable punched tape cassette where tape loops are used. The serial number of the instruction being



processed at any moment is presented digitally at the control desk. The system is not only suitable for drilling but also for plain milling.

The datum point can be chosen freely within the range of travel, which covers nearly 33ft, and the result of the photo-electric measurement of the travel is presented digitally by means of a bi-directional counter.

A dummy panel takes the place of the punched tape reader in the control rack of the simplified version for manual input of two co-ordinates, thus facilitating expansion at a later date.

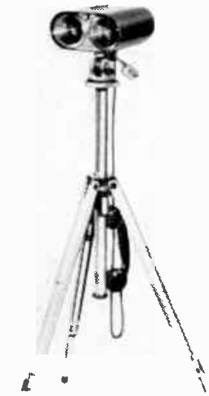
EE 71 761 *for further details*

Optical Telephone Equipment

(Illustrated above right)

Lightweight, portable optical telephone equipment comprising two units LiG3, each consisting of a transmitting and receiving device employing modulated light, allows two-way communication over distances up to 2200 yards. These units may be mounted on commercial camera tripods with tilting heads and are lined up by means of sighting telescopes.

A conventional handset with press-to-speak key is used for speaking and listening, but ancillary units as, for example, amplifiers and loudspeakers or



tape recorders may be connected. An available remote control also provides a complete call facility including buzzer and call key. It is also possible to indicate the intention to speak to the other station by a buzzer tone transmitted via the light beam.

The light waves are modulated directly by means of the operating current of the incandescent lamp employed as light source. The visible part of the spectrum may be excluded without noticeable deterioration of the transmission characteristic, for which purpose a hinged optical filter is provided. A photo-diode serves as photo-electric convertor at the receiving end. The unit, which operates on four batteries supplying 6V, has a frequency range from 300 to 2500c/s and employs, including remote control, ten transistors and three diodes. The overall dimensions are 12·4in by 9·45in by 4·5in and the weight approximately 5¼ lb.

EE 71 762 *for further details*

Hartmann & Braun AG

97 Gräfstr., 6 Frankfurt-West, Germany

D.C. Multimeter

(Illustrated below)

Hartmann & Braun exhibited the ELAVITRON, a d.c. multimeter satisfying the highest demands, from the range of instruments manufactured by ELIMA GmbH. High sensitivity is ensured by

Leadership in Semiconductors

TEXAS INSTRUMENTS

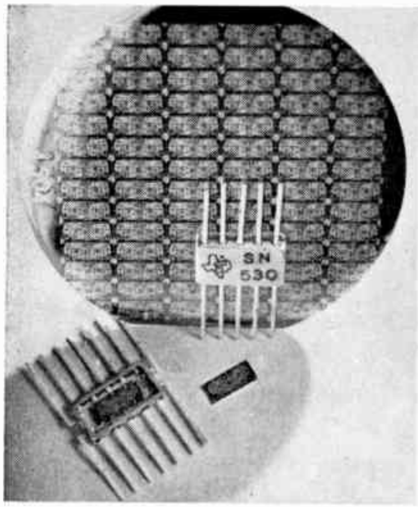
Looking logically at integrated circuits

New high-speed digital semiconductor networks

Series 53 SOLID CIRCUIT* semiconductor networks provide the designer with maximum flexibility through the use of AND/OR INVERT logic with the minimum number of different units.

Features include operation to 5 Mc/s, propogation delay times as low as 5 nanoseconds per AND gate, excellent loading capabilities and the ability to cascade non-inverting logic gates. Series 53 can fulfill most digital requirements in present day computer system applications.

Here is the Series 53 range:
SN530 Single phase, J-K flip-flop
SN531 5-input NAND gate
SN532 5-input AND gate
SN533 Dual 3-input AND gate
SN534 2 and 3-input AND gate
SN535 Clock driver/buffer



Series 53 network, Master Slice wafer*

Master Slice* wafer allows flexibility in mass production

Series 53 as well as TI's established Series 51 and Series 52 networks are made using the Master Slice* fabrication technique. All circuits in a series begin with a slice of silicon containing 50 or more sets of equivalent components.

For Series 53, each component set contains the equivalent of 28 NPN transistors, 10 PNP transistors, 5 capacitors and 26 resistors. Late in the production process, interconnections are made on the Master Slice* to form standard networks, or special networks built to customer specifications. This technique produces maximum network flexibility without sacrificing the economy and consistent reliability of the TI mass production system.

* *Texas Instruments Trademark*

NOW is the time to design semiconductor networks into new equipment

Semiconductor networks may be the answer to your design requirements for reliable, compact and economical circuitry... sooner than you think.

Why now?

Semiconductor networks are penetrating the electronics industry, in Europe as well as in the U.S., even faster than the transistor. Why? Partly because the industry now has the overall experience needed to avoid the sharp transition which we remember when vacuum tube equipment was "transistorized".

Reliability

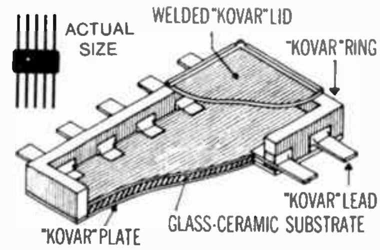
But primarily because experience in transistor reliability evaluation has quickly revealed the remarkable reliability improvements which semiconductor networks will bring. Early tests showed how an integrated circuit containing the equivalent of 20 parts could have the same failure rate as a single conventional transistor. Manufacturers now believe that in a few years a failure rate of around .0001 percent per 1000 hours will be achievable.

Industrial, consumer applications

In addition to the growing military market, industrial and consumer product manufacturers in the U. S. and Europe are moving rapidly to utilize reliable, compact integrated circuits.
For example, Zenith Radio Corporation in the U. S. recently announced a semiconductor hearing aid, developed with TI.

Production, prices

The industry stands just at the beginning of true mass production. Production at TI in 1963 increased eight-fold and more networks were shipped by TI in the fourth quarter of 1963 than the entire industry shipped in the second. And TI is confident that network prices will follow the pattern of steady downward adjustments established by transistors.
In short, the time is *now* for forward looking logic designers to plan for new standards of system reliability and economy with integrated circuits.

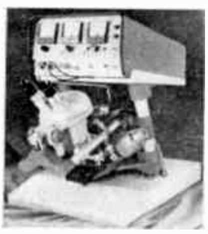


Flat package concept becomes standard

In 1959, while most manufacturers were packing more conventional components into transistor cans, TI took a bold short cut and announced the first fully integrated networks in flat packages. It is estimated that 60 to 70% of all integrated circuits sold today are in the flat package configuration.

The flat package provides the most efficient form factor for systems packaging. Shorter bonding leads permit a more rugged construction. Interconnections are easier. And the package is readily welded to printed circuit boards. The package shown above provides a 10 to 1 reduction in size over TO-5 cans when mounted. Total weight is less than 0.1 gram. Package dimensions are 6.4 x 3.2 x 0.9 cm.

Total capability... materials to systems assembly



As semiconductor network applications have developed, there has been a need for related test and assembly equipment. Such equipment is now available from TI for use with networks from TI and other manufacturers. A unique parallel gap welder has been developed to weld networks to printed circuit board. For large users, an in-line tester is available. And special carriers and test jigs assist handling and testing.

Data is now available on this equipment and on the complete line of TI Series 51, 52 and 53 networks. Write for it today

TEXAS INSTRUMENTS LIMITED
MANTON LANE · BEDFORD · ENGLAND
TELEPHONE: BEDFORD 67466 · CABLES: TEXINLIM BEDFORD · TELEX: 82178

the use of a reliable transistorized amplifier (transistor push-pull chopper). A total of 24 ranges allows the measurement of voltages and currents between 0·1mV and 1kV and 0·1μA to 1A. 100mV plug-in shunts permit the current ranges to be extended to 100A. The quality grade 1·5 ensures high indicating accuracy, the large and clear mirror scale with only two calibrations give high reading accuracy. The measuring movement as well as the amplifier are protected against gross overloads by fuses.

EE 71 763 *for further details*

Herfurth GmbH

6-8 Beerenweg, 2000 Hamburg-Altona, Germany

RADIOACTIVE AIR MONITOR

The radioactive air monitor H 1346 (System KFA Jülich) has been designed for the determination of the effective half-life time by means of the automatic measurement of the decay characteristics. The first of the detectors employed measures the activity direct during the enrichment period, the second one immediately afterwards, and the third one after a considerable delay which may be extended to as much as several days. The following detection limits apply to artificially radioactive aerosols after a three-hour enrichment cycle and a natural activity of 300pc/m³:

after 18 minutes	500pc/m³
after 6 hours	5pc/m³
after 48 hours	0·3pc/m³

The electronic equipment comprises three fully transistorized logarithmic ratemeters covering a range from 10 to 10 000 counts per second. The equipment has been designed for use with all commercially available detectors to ensure universal application. The equipment is constructed for mounting in 19in racks.

EE 71 764 *for further details*

Hewlett-Packard Vertriebs GmbH

95 Kurhessenstr., 6 Frankfurt/Main, Germany

20Mc/s OSCILLOSCOPE

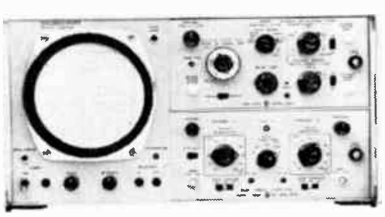
(Illustrated below)

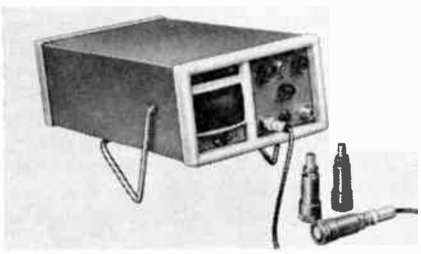
Versatile plug-in units facilitate universal laboratory application of a new oscilloscope manufactured by Hewlett-Packard in Germany; it serves alternatively as a 20Mc/s oscilloscope or as sensitive 10μV instrument for the presentation of slower events. A transit time reflectometer with 90psec rise time and using sampling techniques has been developed for microwave work; it can measure and locate the smallest discontinuities in coaxial transmission systems.

In the model 140A the two plug-in units are arranged one above the other. They have the same overall dimensions and are directly connected to the deflexion system of the c.r.t. An *X-Y* oscilloscope requires two identical plug-in units for vertical deflexion. An internal 10cm by 10cm graticule eliminates errors due to parallax on the non-glare, mat c.r.t. screen. Nearly all plug-in units are designed as differential amplifiers on which d.c. or a.c. coupled inputs may be selected.

The following plug-in units are available: Type 1400A: adjustable bandwidth 0 to 400kc/s, 100μV/cm, one channel; type 1401A: 0 to 450kc/s, 1mV/cm, two channels; type 1402A: 0 to 2Mc/s, 5mV/cm, two channels; type 1403A, 0·1c/s to 400kc/s, adjustable bandwidth, 10μV/cm, one channel; type 1415A: sampling unit, 90psec rise time, 1mV/cm; type 1420A: time-base, 0·5μsec/cm to 5sec/cm, automatic triggering; type 1421A: time-base with delay generator, 0·1μsec/cm to 1sec/cm.

EE 71 765 *for further details*





Dr. Lehfeldt & Co. GmbH

P.O. Box 449, 6148 Heppenheim/Bergstr., Germany

ULTRASONIC TEST EQUIPMENT

(Illustrated above)

The "Echoskop MPT1" is a transistorized, battery-operated ultrasonic tester employing the pulse echo method. However, it can also be used with two separate probes operating in accordance with the transmission or reflection system. A small c.r.t. has been chosen to keep consumption low, but the presentation is enlarged by means of a lens mounted in front of the screen.

The receiver is a wide-band amplifier with a range from 1 to 6Mc/s and probes for any frequency within this range can be connected; however, operating frequencies between 2 and 4Mc/s are preferred.

Thickness measurements are possible above 2mm (0·08in) and maximum penetration in steel is about 6ft 6in for longitudinal waves. The capacity of the battery is sufficient for eight hours' operation and it can be used as buffer battery in combination with a trickle charger. The equipment weighs only 10½ lb including battery.

A plug-in signal accessory can be supplied from the same battery. With this accessory, echo pulses encountered in a freely adjustable depth range and of a minimum height which is also adjustable are indicated by a signal lamp.

EE 71 766 *for further details*

Metrawatt AG

50-54 Schopperhofstr., 8500 Nuremberg, Germany

THERMO-VOLTAGE COMPENSATOR

(Illustrated above)

In addition to the usual indication of thermo-voltages between 0 and 50mV, the thermo-voltage compensator 'Thermokomp' gives a direct reading of temperatures on three scales: 0 to 1112°F (iron-constantan), 0 to 2192°F (chromel-alumel), and 0 to 2912°F (PtRh-Pt). The cold junction temperature can be continuously adjusted between 32°F and 122°F.

An electronically stabilized voltage supplied by dry batteries is applied to a slide wire from which the compensating voltage is obtained. This compensating voltage is compared with the voltage to be measured by means of a galvanometer and adjusted until the galvanometer reads zero. The indicator of the scale is mechanically coupled to the slider. No load is applied to the source of the compensating voltage once balance has been achieved and the result is, therefore, substantially independent of the internal resistance of the source and the resistance of the leads.

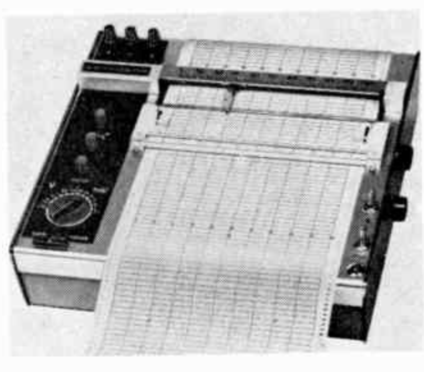
The scale is 7·87in long, the accuracy ±0·3 per cent f.s.d., and the life of the batteries approximately 100 hours. The overall dimensions of the instrument are 11·8in by 8·3in by 5·9in, its weight about 14 lb inclusive of batteries. The instrument is not only suitable for the measurement of thermo-voltages and d.c. voltages, but can also be applied to the calibration of instruments used in connexion with thermocouples on site.

EE 71 767 *for further details*

POTENTIOMETRIC RECORDER

(Illustrated below)

The 'Servogor' is a portable self-



compensating potentiometric recorder for bench work and for laboratory applications. The measurement of d.c. voltages is carried out in accordance with the potentiometer method, and the slider of a potentiometer in a bridge circuit is therefore moved simultaneously with the writing stylus by a servo motor. The equipment has eleven switchable measuring ranges from 0 to 2mV up to 0 to 20V. The effective full deflexion may be adjusted to any intermediate value between 40 and 100 per cent of the range chosen, i.e. when the measuring range 10mV has been selected it may be adjusted to cover any range between 0 to 4mV and 0 to 10mV. The maximum deflexion of the stylus is 7·87in on 9in wide paper of an 82ft long roll, or 7in on a sheet of paper size A4 when the usable length will be 9·84in. Switchable paper feeds are 1·2in—4·7in—23·6in/h and 1·2in—4·7in—23·6in/min. A time marking system provides a time scale for the events registered. The limit of error is ±0·5 per cent f.s.d., but not better than ±20μV (±0·2 where intermediate values are set); the response time is 1sec. The dimensions of the unit are 14·2in by 13·8in by 3·9in, its weight about 17½ lb.

EE 71 768 *for further details*

Olympia Werke AG

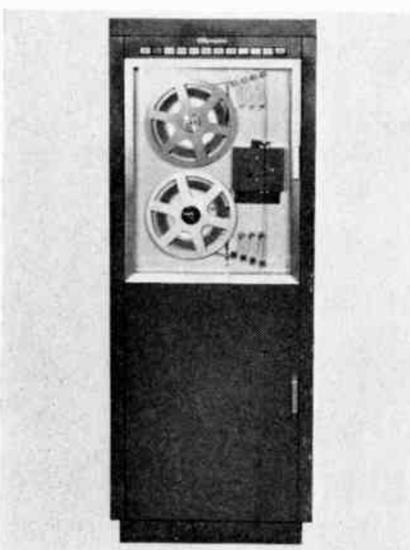
P.O. Box 960, 2940 Wilhelmshaven, Germany

PUNCHED TAPE READER

(Illustrated below)

The prototype of a photo-electric punched tape reader with run-off and re-wind facilities was exhibited, in which punched tapes are read out at a speed of 9·8ft/sec, i.e. 1200 characters per second. The deck has a spring-loaded compensating device and can be stopped within one character.

Forward transport and re-winding take place at the same speed and only requires one instruction or the pressing on a push-button. The prototype was demonstrated with the eight-channel tape code, but readers can also be supplied for other codes. Each character is read out twice, i.e. the first time into a store from which it is only transmitted on parity with the second reading.



The two tape reels may readily be exchanged against a spacious cassette when tape loops are to be analysed, or for similar work.

EE 71 769 *for further details*

P-E-K Electronic

Dr.-Ing. Paul E. Klein

P.O. Box 80, 7992 Tettnang, Germany

ELECTRONIC MULTIPLE INDICATOR

The electronic 'Multiscop 3122' provides a large screen c.r.t. monitor on which a large number of measured values are made visible in quick succession, so that up to 400 measured values appear to the viewer as if presented simultaneuosly. Non-electric values are converted into electrical values by means of transducers and pre-amplified where necessary. In the standard model applied voltages are fed into the monitor via a transistor-controlled relay switch at intervals of 20msec.

Usually, the indication takes the form of light columns arranged next to each other and having deflexions in the directions corresponding to the negative or positive amplitudes. Where amplitudes occur in one direction, the trace may be twice as long. Calibration lines are written simultaneously with the measured values whose deflexion can be compared with the calibration value and adjusted accordingly. The size of the display is 11in × 7·9in on a screen size of 12·6in × 8·7in. All measured values can be fed to an additional rack which can be supplied for conversion into digital values suitable for high speed storage or print-out.

EE 71 770 *for further details*

Rohde & Schwarz

15 Mühldorfstr., 8000 Monich 8, Germany

STANDARD FREQUENCY GENERATOR

(Illustrated above right)

The new Rohde & Schwarz standard frequency generator XUC provides frequencies for measurements with high accuracy. The frequency drift of the built-in frequency standard is less than 2×10^{-9}.

The output frequency can be set on two scales between 470 and 1 000Mc/s. It is formed by a component based on the crystal-stabilized frequency standard which is adjustable in steps of 10Mc/s and the continuously variable frequency of an interpolation oscillator. The error of the continuously variable setting remains below 5kc/s. The output voltage of the standard frequency generator is adjustable in steps or continuously between 0·1mV and 1·5V (into 50Ω).

Where still higher requirements have to be met the interpolation oscillator may be replaced by the standard frequency generator XUA which employs the same basic operating principle. In this way the setting error is reduced to 0·5c/s. It is also possible to use an external frequency standard of higher stability to

drive the generator. By employing, e.g., the frequency standard XSD in this way, a frequency of 1 000Mc/s can be set up with an absolute accuracy of 0·2c/s.

The equipment weighs 154 lb and has overall dimensions of 21in × 17·5in × 24·2in.

EE 71 771 *for further details*

RMS Regelungs- und Messtechnik

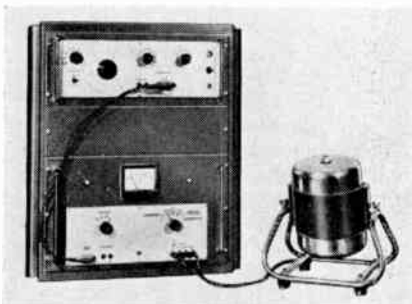
Dipl.-Ing. Schmidt KG

Gutenbergstr., 2057 Reinbeck bei Hamburg, Germany

VIBRATION TEST EQUIPMENT

(Illustrated below)

German and international committees have agreed appropriate standards for



vibration tests which form part of the specifications for numerous units and components on an increasing scale. RMS vibration equipment meets the requirements of these test specifications.

The illustration shows the equipment type SW100 comprising an *RC* sine wave generator, an l.f. power amplifier, and the electro-dynamic vibrator. At the top of the cylinder a socket is visible for the mounting of the test specimen.

The SW100 is designed for test specimen weighing about 2¼ lb and can produce vibrations between 5 and 2 000c/s with maximum amplitudes of ±0·12in. The corresponding values for the equipment SW1000 are 11 lb for the weight of the test specimen and 5 to 800c/s for the frequency range.

EE 71 772 *for further details*

Sadowski & Co.

7301 Esslingen-Berkheim, Germany

DOUBLE PULSE GENERATOR

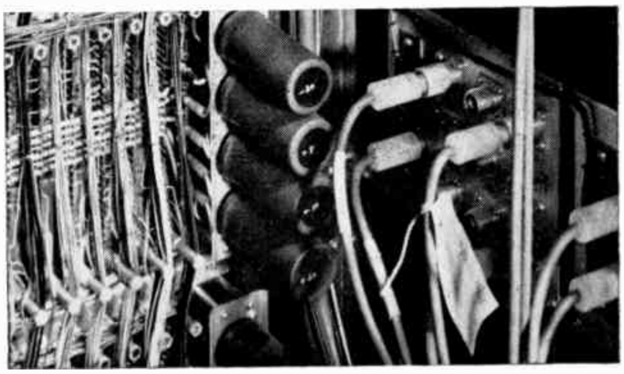
The double pulse generator type DIG 121 uses silicon transistors throughout and has been designed for universal application; its small size and low weight are combined with an unusually high reliability.

EE 71 140 *for further details*

MORE THAN 2 MILLION FEET

OF WANDLESIDE P.T.F.E. CONNECTING WIRES AND FLEXIBLES ARE USED ANNUALLY IN WIRING ENGLISH ELECTRIC LEO COMPUTERS

The delicate and sensitive "brain" of a computer requires the most intricate of wiring systems. Wiring is the life-blood of a computer, so it is essential that it is of the highest quality. Wandleside P.T.F.E. insulated wires and flexibles guarantee absolute accuracy and reliability under the most stringent conditions. We can manufacture to both Ministry and Commercial specifications.



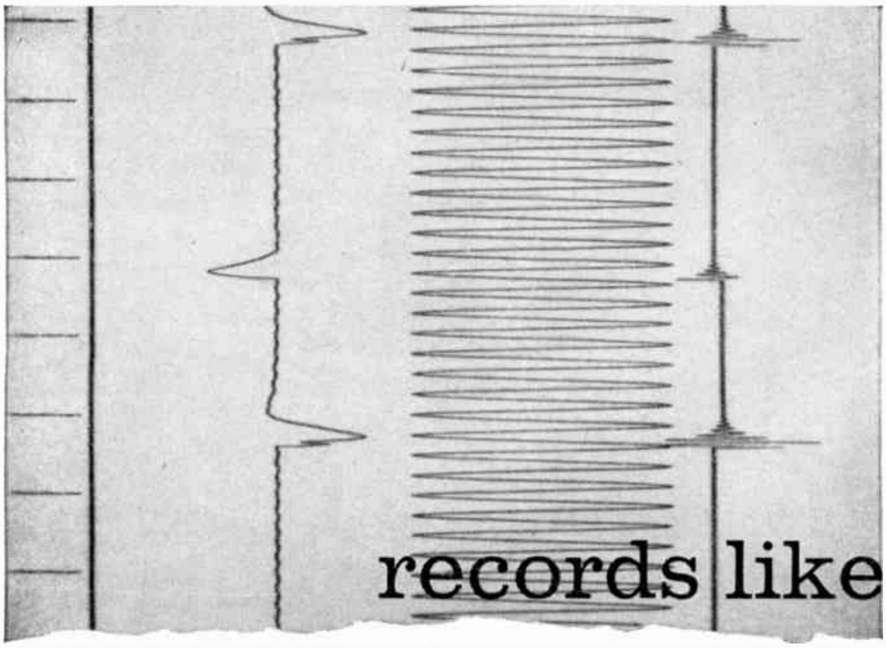
If you have problems with

● MINIATURIZATION ● HIGH AMBIENT TEMPERATURE CONDITIONS ● SOLDERING IN CONFINED AREAS ● SOLDERING TO MINIATURE CONNECTORS OR COMPONENTS

Consult **WANDLESIDE WARREN WIRE COMPANY** *In Association with* WANDLESIDE CABLE WORKS LIMITED *(Falks Group)*

Our Technical Advisory Service is at your disposal. Samples and prices on application.
P.T.F.E. insulated wires and flexibles are available in eleven standard colours and a full range of bi-colours and tri-colours. Stocks are carried at both our London and Dunmurry works.
Our inspection has full A.I.D. and A.R.B. approval. We hold M.O.A. design approval for P.T.F.E. aircraft cables.

WANDLESIDE WARREN WIRE CO. LTD. DUNMURRY, NORTHERN IRELAND. TELEPHONE: DUNMURRY 2536. TELEX: 74339. TELEGRAMS: DERRIAGHY, DUNMURRY, N. IRELAND. London Office: 106 Garratt Lane, S.W.18. TELEPHONE: VANDYKE 7544. TELEX: 261355. TELEGRAMS: WANDLESIDE, LONDON, S.W.18.

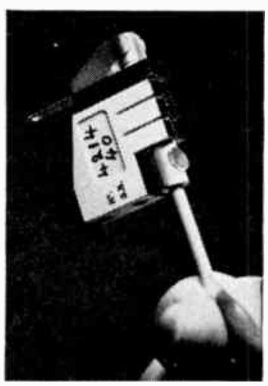


Un-retouched
12 cm wide trace

records like this cost less than you think with the Honeywell 1706 Visicorder— designed simply for accuracy!

A galvanometer recorder with only five controls—yet the Honeywell 1706 *Visicorder* oscillograph embodies all the features necessary for high resolution recording of recurrent and transient phenomena; and by using an ultra-violet system, images are rendered visible for analysis, rapidly and without processing.
Outstandingly compact and easy to operate, the 1706 offers 4 independent channels . . . 8 paper speeds from 6 to 800 mm/sec . . . 0.1 and 1 sec timing intervals . . . two datum lines . . . drop in paper loading . . . terminal inputs . . . *all at a price well within the budget of even the smallest laboratory.*
The 1706 is one of the range of Honeywell high speed recorders which includes types 1185: 36 channels on 12-inch wide paper—and 2500: 12 channels on 12-centimetre wide paper.

1706 *Visicorder* oscillograph uses robust Honeywell Series BB mirror galvanometers with responses from 0-30c/s to 0-5kc/s



Honeywell

SCIENTIFIC AND MEDICAL INSTRUMENTS

Honeywell Controls Limited,
Greenford, Middlesex. Waxlow 2333

HONEYWELL INTERNATIONAL Sales and Service Offices in all principal cities of the world. Manufacturing in United Kingdom, U.S.A., Canada, Netherlands, Germany, France, Japan.

The instrument comprises the following units: spacing generator, pulse generator 1, delay stage, and pulse generator 2. The spacing generator produces trigger pulses with a spacing adjustable between 1μsec and 1sec and may be synchronized with an external sine wave or pulse voltage. The output of the spacing generator triggers the pulse generator 1; its square wave pulse output has a width which can be varied between 0·1μsec and 0·1sec. The pulse generator 2 is of similar design and is triggered via a delay circuit giving adjustable delay times of between 0·1μsec and 0·1sec. The pulse generator 2 is of similar design and is triggered via a delay circuit giving adjustable delay times of between 0·1μsec and 0·1sec. The rise time of the pulses is 0·02μsec, the delay time 0·03μsec.

The amplitude of the output pulses is continuously variable between 0 and 10V, the polarity is reversible. The outputs of the two pulse generators are available either separate or, via a switch, as double pulse. An external voltage source (sine wave or pulse) may be used in place of the spacing generator to trigger the pulse generators. Triggered externally the DIG 121 may also be used as a delay time generator.

EE 71 773 *for further details*

Sodeco
Société des Compteurs de Genève
70, Grand-Pré, Geneva, Switzerland

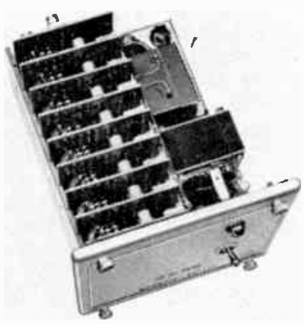
TRANSISTORIZED SUMMING UNIT

(Illustrated below)

The transistorized summing unit type SC58 is constructed as a plug-in unit and can be applied wherever electric pulses originating from different sources are to be registered by a common counter or printer without losses.

The summing unit, designed on printed circuits, automatically stores free from error pulses which may arrive independently of each other in time and from different sources. The store is immediately scanned with a correspondingly high frequency. The output of the summing unit can either be connected direct to the high speed pulse counter SC60 (up to 250 counts per sec) or via the pulse convertor type 3Tu1 to all Sodeco pulse counter types including those equipped with a printer.

Up to 24 input channels may be employed and the input signals may be derived from photo-electric devices or electromechanical contacts and can consist of pulses of different shape and width.



EE 71 774 *for further details*

PRINTING PULSE COUNTER

The basic equipment consists of three units, viz. a counting magnet which drives simultaneously the 6-digit indicator and the 6-digit printer, an electric printing mechanism, and an electric zeroing device. Two of the basic equipments can be arranged next to each other and for printing on the same paper roll.

Three different counters may be incorporated:

(1) Counter with carry-over of tens.
(2) Time counter with indication in hours - minutes - seconds - tenth of seconds, or hours-minutes-seconds.
(3) Date counter without electric zeroing, driven by timed pulses.

The pulse counters are available for d.c. supply (mains or batteries) or for a.c. supply and have a maximum counting rate of 10 counts/sec.

EE 71 775 *for further details*

Telefunken AG
Division High Frequency Installations
3 Elisabethenstr., 79 Ulm, Germany

DIGITAL COMPUTER

Telefunken demonstrated for the first time the digital computer type TR10, which is not only a complementary machine to the fast and powerful larger computer TR4 but also a versatile digital computer in its own right. It is suitable for application over the whole field of technical and scientific studies, to the solution of commercial data processing problems, and in the control of industrial manufacturing processes.

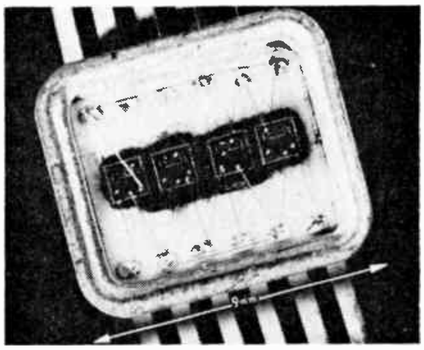
The TR10, a decimal digital computer with stored programme control requiring very little space, is characterized by its excellent performance and particularly by its capacity for simultaneous operation between the processor itself and its peripheral equipments. Facilities are available for the processing of punched cards and tapes; larger quantities of data may be wrtiten out quickly by a high speed printer. The internal core stores may be augmented by external digital magnetic tape stores and magnetic disk stores. The processor may be connected to Telex subscribers viâ a Telex exchange, and a traffic control may be assembled from building blocks to organize the transfer from points of measurement as well as to points of control in industrial processes. The computer programme may also be interrupted by external means to react to priority signals.

The TR10 is a decimal computer and the decimal figures, 26 letters and 28 symbols are read out of the computing store serially and assembled to be processed as words of any length. The capacity of the computing stores may be selected in steps of 10 000 between 10 000 and 80 000 digits. The cycle time of the ferrite core store is 8μsec.

Planar transistors are chiefly employed in the TR10 and the peripheral equipment. The central processor is housed in a modern cabinet with overall dimensions of 70·8in × 23·6in × 15·7in and weighs about 440 lb.

Besides the technical performance of its equipment, the TR10 offers also very flexible programming facilities for the most varied applications. Addresses can be modified as often as required and independent of each other. Instructions for the handling of the individual peripheral units by the computer may differ from unit to unit to suit special conditions. The instructions are transmitted from the computer to the peripheral units, which carry them out independently and simultaneously with the computer.

EE 71 776 *for further details*



Telefunken AG
Valve Division, Distribution Dept.
100 Söflingerstr., 79 Ulm, Germany

SOLID STATE CIRCUITS

(Illustrated above)

The known solid state circuits consist in general of silicon into which the active and passive components are diffused in accordance with the planar techniques. The isolation of the components in one wafer of silicon against each other is produced by diffusing barrier layers into the silicon. The silicon surface is covered with a silicon oxide layer on to which metal conductors are deposited by evaporation to connect the components.

Resistances produced by diffusion have a temperature coefficient of $3 \times 10^{-3}/°C$. An applied barrier voltage of 1V will result in a capacitance of about 150pF/mm² between such a resistance and the base material, or for a typical resistor of 10kΩ in a capacitance of 5pF distributed over the whole of the resistor.

These disadvantages are avoided by employing the new hybrid manufacturing techniques for solid state circuits. Active elements are produced by using planar techniques as before, passive elements are deposited by evaporation in a high vacuum. This reduces the temperature coefficient to the substantially lower value of $2 \times 10^{-4}/°C$. The capacitance to the base material is now only 1/10 of that of diffused resistors.

The illustration shows half a shift register consisting of 4 NOR gates, each side being 0·35in long.

EE 71 777 *for further details*

Short News Items

The Scientific Instrument Manufacturers' Association announces that approximately £½M worth of business has been done as a result of the Peking exhibition of British scientific instruments organized by S.I.M.A. on 15 to 25 April this year.

At the exhibition 28 British instrument manufacturers showed their equipment to members of the Chinese Government, commercial, industrial and research organizations.

The Electronics Industry Council which was formed by the British Valve Manufacturers' Association, the Electronic Engineering Association and Radio and Electronic Component Manufacturers' Federation in March 1960 to speak on behalf of the capital goods sector of the electronic industry in negotiations with Government departments has been dissolved.

The functions of the Council have now been taken over by the Conference of the Electronics Industry which was formed in July 1963.

The I.E.E. Science and General Division Group on Education and Training is to hold a residential conference on audio-visual teaching aids at Loughborough College of Technology on 10 to 12 September 1964.

The aim of the conference is to bring together tutorial staff, industrial training officers, and those interested in education and training to discuss the progress that is being made in the development of aids to teaching, and the possibility of arranging for co-operation in the production of more.

A scientific exhibition will run concurrently with the discussion sessions. Offers of exhibits are invited within the following categories:

(a) Audio visual aids, including teaching machines, language laboratories, closed-circuit television, video tapes and loop films.

(b) Electrical engineering laboratory demonstration equipment.

(c) Apparatus developed at teaching establishments, including laboratory and demonstration equipment for electrical engineering courses and general and special courses for technicians, and college-built audio-visual aids.

Those wishing to offer exhibits for consideration for inclusion in the exhibition should apply to the Secretary of the Institution for application forms as soon as possible.

Further information and registration forms are available from the Secretary, the Institution of Electrical Engineers, Savoy Place, London, W.C.2.

English Electric-Leo Computers Ltd has received an order from Messrs. Common Brothers of Newcastle upon Tyne for marine data logging and monitoring systems to be installed in two ships.

The first ship, due to be completed at the end of this year, is being built by the Furness Shipbuilding Company of Haverton-on-Tees. It will be powered by a six-cylinder Fairfield-Sulzer diesel engine and the data logging systems installed in the engine room will monitor all important pressures, levels and temperatures and the running or alarm condition of some 40 motors. An alarm lamp system mounted on the front of the logger cubicle will provide indication of the location of all alarm conditions. An electric typewriter will be included in the system for producing the log sheets. A manual select feature is incorporated to permit the immediate display, in illuminated decimal digits, of the value of any selected variable.

The second system, with similar facilities to those described, will be installed in the engine room of a turbine vessel, being built by Fairfield-Rowans of Glasgow.

The Ministry of Defence (Navy Department) has placed an order with Standard Telephones and Cables Ltd for thirty-five new automatic radio transmitters.

These 7kW high frequency transmitters are to improve long-distance communications between Admiralty stations at home and overseas. They are of the latest design, specifically for unattended automatic operation.

These transmitters are similar to the company's QT-3A units which have been supplied in quantity to the British Post Office and to a number of countries including the U.S.A., Kenya, Kuwait, Greece, Argentina, Brazil and Spain.

The QT-3A transmitter, which has a frequency range of 2·3 to 28Mc/s, is for use in medium and long distance radio links for independent sideband and double sideband telephony and for single and multi-channel telegraphy.

The BBC announces that this year's Reith lectures will be given by Sir Leon Bagrit, chairman of Elliott-Automation Ltd.

His subject will be The Age of Automation—the nature and possibilities of systems of computerized automation and the implications he foresees in industrial, economic, political, educational and social aspects of society.

Sir Leon will develop his ideas in the course of six lectures beginning on Sunday, 8 November, in the Home Service. Each lecture will be repeated in the Third Programme later in the week.

English Electric Valve Co. Ltd products are now to be manufactured in the U.S.A. as a result of arrangements made with Compagnie Generale de Telegraphie Sans Fils (C.S.F.) in Paris. The two companies have exchanged manufacturing information, so that each now makes certain products of the other, and this facility is now available to Warnecke Electron Tubes Inc. in Chicago, U.S.A., which is a subsidiary of C.S.F.

An Automatic Control Convention is to be held at the University of Nottingham in April 1965.

The Convention, under the aegis of the United Kingdom Automation Council, is being organized by the Institution of Mechanical Engineers on behalf of the Institutions of Mechanical, Electrical, Chemical and Production Engineers, The Royal Aeronautical Society, The Institution of Electronic and Radio Engineers and the Society of Instrument Technology.

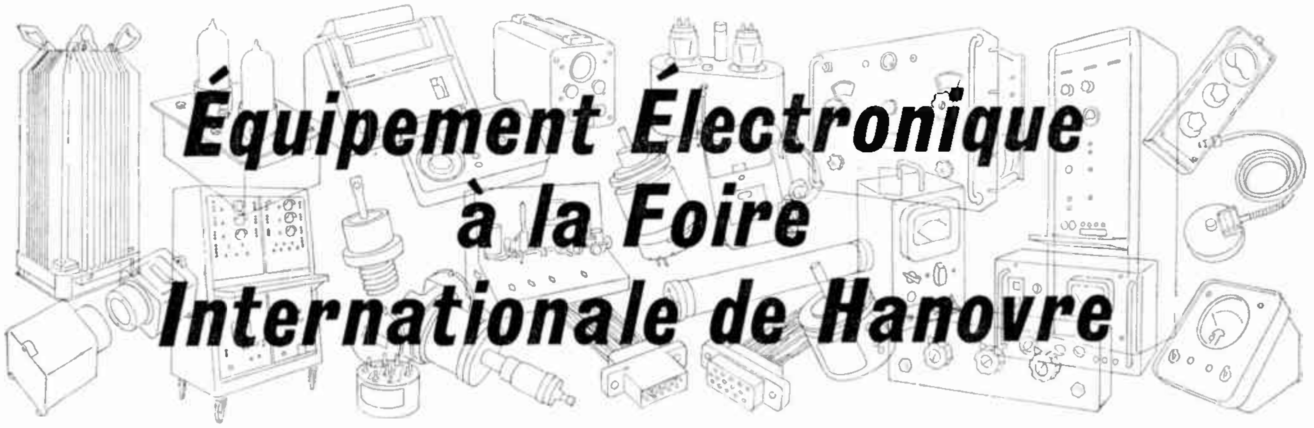
Papers offered should be sent, together with brief synopses, to: R. J. Millson, Editor of Proceedings, The Institution of Mechanical Engineers, 1 Birdcage Walk, Westminster, London, S.W.1.

The Avo Valve Data Manual (16th edition) is now published. It provides information on approximately 6 000 obsolete, obsolescent, current and prototype valves.

The publication is available, ex stock, from Avo Limited (M.I. Group), 92/96 Vauxhall Bridge Road, London, S.W.1, price 35s. post free.

The Central Research Laboratory of Associated Electrical Industries at Rugby has developed single-junction silicon power diodes capable of withstanding more than 5kV. This reverse voltage capability could previously only be achieved with a stack of carefully matched individual units each limited to 1-2kV, resulting in a high voltage drop in the conducting direction. The very high breakdown voltage of the new units is partly the result of using an epitaxial deposition process in forming the rectifying p-n junctions.

A unit rated at 5-10A is at an advanced stage of development and experiments are in progress on units capable of 100A or more. The 5-10A units can withstand surge currents of 200A for 5msec.



Équipement Électronique à la Foire Internationale de Hanovre

Description, basée sur des renseignements fournis par les fabricants de certains des appareils et composants allemands exposés à la Foire de Hanovre du 26 avril au 5 mai 1964.

Traduction des pages 490 à 495

Becker Flugfunkwerk GmbH

Flugplatz, 757 Baden-Baden-Oos, Allemagne

ÉMETTEUR-RÉCEPTEUR DE BORD

(Illustration à la page 490)

L'émetteur-récepteur piloté au quartz AR380 est un appareil très compact offrant 380 canaux d'émission et de réception à intervalles de 50 kHz dans la gamme de fréquence de 117 à 135,95 MHz. Le récepteur et le modulator de puissance sont entièrement transistorisés, mais l'émetteur emploie trois tubes pour obtenir une sortie de 10 W. Le poids réduit de l'appareil ainsi que sa robustesse et sa consommation électrique insignifiante sont le fruit de l'emploi de transistors.

Le panneau frontal mesurant 9,52 cm × 14,6 cm est conforme aux normes de l'ARINC. Il est à éclairage interne et les grands chiffres ainsi que les commandes de forme appropriée facilitent la précision du réglage de fréquence, même dans l'obscurité. La tonalité de contrôle, produite par un amplificateur supplémentaire à large bande au cours de l'émission et transmise aux casques après amplification, assure la protection contre toute panne non observée de l'appareil ou de l'antenne.

L'appareil se caractérise par sa construction à éléments. L'installation répond à toutes les conditions de normes et règlements actuellement en vigueur. L'émetteur-récepteur ne pèse que 3 kg environ et le bloc d'alimentation en pèse 2.

EE 71 751 *pour plus amples renseignements*

Bölkow-Apparategebau GmbH

7311 Nabern-Teck, Allemagne

APPAREIL À CONCENTRATION DE DONNÉES

(Illustration à la page 490)

L'appareil à concentration de données Bölkow 2000 est un système dans lequel les valeurs—très diverses en soi—de température, de pression, de composition chimique des gaz, de vibration, de débit, etc. devant être contrôlées, sont converties en tensions continues et alternatives au moyen de transducteurs. La précision de ces transducteurs varie entre 0,1 et 10%, selon le type et la version. Les éléments de base du système sont le convertisseur analogique-numérique et le fréquencemètre numérique qui mesure les tensions avec une précision d'environ 0,01 à 0,2%, la fréquence et la durée avec une précision de 1×10^{-5} à 1×10^{-7}, et les injecte à d'autres éléments du système aux fins de traitement et d'enregistrement. Ce processus s'effectue à l'aide de minuteries numériques, de convertisseurs de codes, de lecteurs numériques, d'imprimeurs, de présélecteurs numériques et de générateurs d'impulsions de programmation.

Des études compliquées et étendues sur les problèmes de recherche et de réalisation doivent être subdivisées en recherches sur leurs différents aspects et il s'est avéré fort avantageux de pouvoir utiliser pour ces études partielles des instruments qui peuvent être intégrés par la suite dans un important système de programmation pour l'examen du projet dans son ensemble. Des systèmes spéciaux de contrôle et de calcul pour la solution de petits problèmes ou de problèmes partiels peuvent être conçus à l'aide d'éléments numériques standardisés du système Bölkow 2000. De grandes installations peuvent être assemblées au moyen des calculateurs numériques de la Compagnie des Compteurs (SETI) qui peuvent être reliés directement au système 2000. Tous les éléments du système Bölkow 2000 sont fournis pour montage sur bâti de 48,2 cm. Le système est déjà utilisé pour le projet ELDO.

EE 71 752 *pour plus amples renseignements*

Braun Electronic GmbH

109 Lange Str., 7808 Waldkirch/Breisgau, Allemagne

MESURE DE LA TEMPÉRATURE SANS CONTACT

Des éléments individuels d'un système électronique pour mesurer le rayonnement de la température ont été créés par Braun Electronic GmbH sous la désignation de TASTOTHERM-INFRA. L'équipement pour la mesure sans contact en gammes standard de +100° C à +3000° C consiste toujours en un élément sensible relié à un élément de base comprenant des commutateurs d'amplificateur, de bloc d'alimentation et, si nécessaire, de limite.

Le signal est fourni sous forme de quantité électrique et la méthode de mesure se prête aisément à l'indication à distance et à l'enregistrement de la température en fonction du temps. Le rayonnement de chaleur incidente touchant l'élément sensible est converti en un signal électrique et préamplifié. La grandeur du point mesuré et sa distance de l'élément sensible peuvent être choisies librement entre certaines limites. De plus, le facteur d'émission de l'objet devant être mesuré doit être considéré, c'est à dire le rapport entre la puissance de radiation du matériau en question et celle d'un corps noir, de même que le diamètre de l'ouverture à travers laquelle s'effectue la mesure. L'équipement est fourni soit pour installation particulière en vue d'un matériau donné soit pour installation universelle étalonnée pour plusieurs facteurs d'émission.

EE 71 753 *pour plus amples renseignements*

Brown, Boveri & Cie AG

68, Mannheim, Allemagne

EQUIPEMENT MICROONDES

(Illustration à la page 491)

La société Brown, Boveri & Cie a présenté son nouvel équipement entièrement transistorisé FM12/7000 pour les liaisons par microondes à 7 GHz. L'emploi exclusif de méthodes à semi-conducteurs a créé du matériel d'une conception nouvelle n'utilisant que des transistors au silicium et des diodes. La consommation a été réduite à une fraction de ce qu'elle était auparavant et le pilotage au quartz améliore la frèquence à un tel degré qu'on peut maintenir des espacements de canaux de 1 MHz.

L'équipement a été conçu pour les liaisons microondes et il couvre la gamme de 6400 à 8400 MHz; il peut recevoir 12 ou 24 canaux de téléphonie. Il peut être employé également pour l'émission d'un canal musical de haute qualité. Sa consommation électrique est de 10 W. Dans le modèle standard, l'émetteur, le récepteur et le réflecteur d'antenne sont assemblés en un seul élément. On peut, cependant, insérer des guides d'ondes mesurant jusqu'à 50 m de long lorsqu'on veut installer le réflecteur à distance.

EE 71 754 *pour plus amples renseignements*

DeTeWe
Deutsche Telephonwerke und Kabelindustrie AG
100 Wrangelstrasse, 1 Berlin 36, Allemagne

GÉNÉRATEUR D'IMPULSIONS

Ce nouvel équipement combiné comprenant un générateur d'impulsions et un fréquencemètre/minuterie universal permet d'effecteur les divers essais et mesures qu'exige l'utilisation de centrales téléphoniques et les applications techniques de contrôle.

Cinq commutateurs de décades sont prévus pour la sélection de la largeur d'impulsion et cinq autres pour le réglage de l'intervalle entre les impulsions. Le nombre d'impulsions produites peut être préréglé entre 1 et 12, ou elles peuvent être continues et les trains d'impulsions peuvent être répétés automatiquement.

Les ondes sinusoïdales ou les entrées d'impulsions peuvent être comptées pendant des périodes de 1 à 10 secondes. le résultat est indiqué par chiffres illuminés et décimales. La gamme de 10 à 99 999 Hz est couverte pour les mesures de fréquence.

Durant la mesure de temps, des oscillations à haute stabilité de la base de temps sont injectées au compteur électronique au moyen d'un circuit de porte, c'est à dire que le circuit de porte est contrôlé par le signal d'entrée et que la lecture numérique correspond au temps pendant lequel le circuit de porte est ouvert. Une gamme de temps de 1 à 99 999 msec est couverte par l'équipement.

EE 71 755 *pour plus amples renseignements*

Friesecke & Hoepfner GmbH
B.P. 72, 8520 Erlangen-Bruck, Allemagne

INSTRUMENTS NUCLÉONIQUES

(Illustration à la page 491)

A la suite de perfectionnements apportés à sa gamme d'instruments nucléoniques, la société Friesecke & Hoepfner a pu exposer pour la première fois les éléments transistorisés de son matériel électronique TGP. Ces éléments transistorisés sont logés dans des blocs standardisés interchangeables UC/2, dont deux s'insèrent côte à côte dans le bâti de montage standard de 48 cm. En même temps, des coffrets de banc d'essai ont été réalisés pour les diverses combinaisons.

Le matériel que l'on voit dans nos gravures comprend le générateur de haute tension FHT 101A1, l'intensimètre linéaire FHT 110B, le compteur d'impulsions FHT 120A2 et la minuterie FHT 120 A3. Le générateur à haute tension fonctionnant sur courant secteur fournit les tensions de détecteurs entre 200 et 500 V à un maximum de 5 mA, une stabilité supérieure à 0,2% pendant 24 heures et la possibilité de commutation pour obtenir une polarité négative ou positive. La précision absolue du réglage est de ±2%, le plot le plus court étant de 1 V.

L'intensimètre linéaire indique la valeur de temps moyenne d'un train d'impulsions en onze gammes entre 10^2 et 10^7 impulsions par minute. Les deux entrées de taux d'impulsions peuvent être reliées entre elles afin de fournir un courant différentiel. La sortie d'impulsions des détecteurs de radiations est amplifiée et injectée au compteur d'impulsions qui est muni d'un dispositif de lecture à tube de comptage de décades à six chiffres. L'appareil comprend une commande de comptage, tandis que des bornes marquées "commande universelle" permettent de relier des systèmes de commande connexes, tel que l'analyseur de hauteur FHT 300 A à une voie. L'arrêt automatique incorporé peut être préréglé en dixièmes. La minuterie pilotée au quartz assure le contrôle horaire pour la mesure avec une stabilité supérieure à 1×10^{-3} et une résolution de 1 μsec. Le lecteur numérique des valeurs peut être utilisé pour actionner un imprimeur.

La gamme couvre également tous les amplificateurs et presque tous les éléments peuvent être founis pour fonctionnement sur secteur ou batterie.

EE 71 756 *pour plus amples renseignements*

Fuba, Werk elektronischer Bauteile u. Geräte
Hans Kolbe & Co.
3371, Gittelde üb. Seesen, Allemagne

BLOC D'ALIMENTATION DE 400A À COURANT STABILISÉ

La conception de base de cet appareil s'est effectuée à Hambourg par la Deutsche Elektronen Synchrotron (DESY). Il fournit le courant, par exemple, aux aimants de déviation d'accélérateurs de particules ainsi que d'autres charges inductives exigeant un courant continu stable et élevé. Les utilisateurs sont essentiellement les instituts de recherches physiques, les centres de recherches nucléaires et les laboratoires de physique nucléaire à haute énergie. Le courant de sortie peut être pris en plots de 40 mA au moyen d'un commutateur numérique à 4 chiffres. Il peut, par exemple, être fixé à 384,8 A ou à 38,48 A. Une fois le courant réglé, l'appareil est largement indépendant des fluctuations de tension d'entrée ou de charge. La sortie nominale maxima de 16 kW et la tension de sortie peuvent être choisies entre 0 et 40 V.

Le transformateur de courant secteur de 20 kVA et la bobine d'arrêt sont fixés de manière permanente dans un coffret mesurant 580 × 555 × 1640 mm. Les éléments fonctionnels sont conçus pour le montage sur bâti de 48,2 cm. On a prévu, en outre, des dispositifs de contrôle extérieur pouvant être fournis sur demande. L'appareil pèse environ 400 kg.

EE 71 757 *pour plus amples renseignements*

ÉLÉMENTS NUMÉRIQUES INTERCHANGEABLES

(Illustration à la page 491)

Les éléments numériques du système Darmstadt, qu'on a pu voir pour la première fois en 1963, sont maintenant livrables en trois séries de vitesses pour des fréquences de 250 kHz, 5 MHz et au-dessus de 10 MHz. Un des avantages les plus importants de ce système est la standardisation rigoureuse des conditions d'adaptation des éléments individuels, ce qui simplifie considérablement la réalisation de l'installation. Des connexions normalisées facilitent l'emploi de combinaisons de différentes séries à l'intérieur d'un seul circuit.

On a prévu comme accessoires des univibrateurs, des inverseurs, des oscillateurs piézoélectriques, des correcteurs d'impulsions, des amplificateurs de sortie et des dispositifs d'entraînement pour tubes Nixie.

EE 71 758 *pour plus amples renseignements*

P. Gossen & Co. GmbH
25 Nägelbachstr., 8520 Erlangen, Allemagne

AMPLIFICATEUR DE COURANT CONTINU

(Illustration à la page 492)

Les tensions de sortie des thermocouples sont beaucoup trop faibles aux températures réduites pour pouvoir actionner de manière sûre des instruments à cadre mobile, même s'ils sont d'une grande sensibilité. Grâce aux amplificateurs magnétiques on peut maintenant mesurer les tensions de sortie les plus faibles des thermocouples à cadres mobiles standards. Un modèle de démonstration a montré comment la tension de 3,19 mV produite par un thermocouple ferconstantan à +60° C (140° F) est convertie en un courant de 1 A par le préamplificateur de courant continu PMV10, permettant une indication sûre de la gamme de température de 0° à +60° C sur un ampèremètre à cadre mobile insensible avec déviation de l'aiguille de 250°.

Un oscillateur à ondes carrées fournit

à l'amplificateur magnétique une tension auxiliaire à 1 kHz. Grâce au choix de cette fréquence, on a pu assurer à l'appareil des dimensions réduites, une faible constante de temps (<0,1 sec) et un faible facteur d'ondulation de courant de sortie. Le préamplificateur peut être actoinné sur 24 V c.c. ou 220 V, 50 Hz c.a. et il peut être monté sur la plaque de base d'instruments de panneau avec des ailettes de 144 × 144 mm de l'arrière.

EE 71 759 *pour plus amples renseignements*

Gebr. Grieshaber

B.P. 30, 762 Wolfach, Allemagne
Distributeurs: Vega GmbH, P.O.B. 1649, 78 Freiburg/Breisgau, Allemagne

CONTRÔLEUR DE NIVEAU

(Illustration à la page 492)

Le nouveau contrôleur de niveau transistorisé exposé par la société Grieshaber emploie exclusivement ou de préférence des transistors planaires au silicium afin d'assurer la plus grande fiabilité possible. Ils sont basés sur la méthode de mesure capacitive mais déterminent le courant du condensateur au lieu de sa tension. De cette manière une valeur électrique est obtenue automatiquement. Elle maintient un rapport proportionnel linéaire avec la capacité variable.

De longs conducteurs sans blindage ont pu être réalisés en transférant l'oscillateur à la tête des électrodes. La tête de l'électrode de même que l'élément électronique proprement dit peuvent être utilisés dans des applications couvrant la gamme de 10 à 10 000 pF sans aucun changements des composants électroniques et sans affecter la réponse des indicateurs de limite de cette série à des changements de capacité inférieurs à 0,5 pF. Un soin particulier a été apporté à la stabilité thermique de l'appareil.

L'appareil est à éléments. Notre gravure montre le bloc d'alimentation commun à tous les modèles, ainsi que le nouveau commutateur à double limite Vegator/Vegasel, en tant que modèle de montage mural, tel que fourni pour être incorporé à d'autres appareils.

EE 71 760 *pour plus amples renseignements*

Grundig Werke

8510 Fürth, Allemagne

CONTRÔLE NUMÉRIQUE DE MACHINES-OUTILS

(Illustration à la page 492)

Le système de positionnement numérique pour machines-outils, exposé pour la première fois à Milan il y a six mois, a été introduit comme un système perfectionné et avec une entrée à carte perforée de 8 canaux. Il s'agit d'une commande de positionnement avec "information de commande" de construction à éléments pouvant être assemblés selon les besoins et pour une gamme étendue de machines. Il peut être étendu par la suite pour des usages supplémentaires. Dans ce contexte, "l'information de commande" se rapporte à la commande des fonctions supplémentaires de la machine, comme par exemple différentes vitesses et différentes alimentations, changement d'outils ou fourniture de réfrigerants. La commande numérique peut être fournie pour une à trois coordonnées, et dans la version pour deux co-ordonnées on peut mettre en programme actuellement, en quatre groupes indépendants, jusqu'à 20 types d'informations de contrôle.

Le châssis de commande est assemblé à partir d'éléments à fiches. Il comporte, en outre, une cassette facilement interchangeable contenant les bandes perforées. Le numéro de série de l'instruction étant traité à n'importe quel moment donné est présenté sous forme numérique au pupitre de commande. Le système n'est pas seulement utile pour le forage mais également pour le fraisage ordinaire.

Le point de référence peut être choisi librement à l'intérieur de la gamme de propagation qui s'étend sur près de 10 m, et le résultat de la mesure photoélectrique de la propagation est présenté numériquement au moyen d'un compteur bidirectionnel.

Un panneau factice remplace le lecteur de bande perforée dans le châssis de commande de la version simplifiée pour l'entrée manuelle de deux coordonnées, facilitant ainsi l'expansion par la suite.

EE 71 761 *pour plus amples renseignements*

EQUIPEMENT TÉLÉPHONIQUE OPTIQUE

(Illustration à la page 492)

Il s'agit ici d'un matériel téléphonique optique, portatif et de poids léger, comprenant deux éléments LiG3, dont chacun consiste en un dispositif d'émission et de réception à lumière modulée. Ce dispositif permet la communication dans les deux sens sur des distances allant jusqu'à 2 km. Ces éléments peuvent être montés sur trépieds de caméra commerciaux avec têtes basculantes et ils sont alignés au moyen de télescopes de mire.

Un casque classique avec clé de fonctionnement est utilisé pour l'écoute et la parole, mais des éléments auxiliaires comme, par exemple, des amplificateurs et des haut-parleurs ou des magnétophones peuvent être reliés. Une télécommande met en jeu un dispositif d'appel complet, comprenant vibreur et clé d'appel. On peut également indiquer son intention de parler à une autre station au moyen d'une tonnalité transmise par le faisceau lumineux.

Les ondes lumineuses sont modulées directement par le courant de fonctionnement de l'ampoule incandescente utilisée comme source lumineuse. La partie visible du spectre peut être exclue sans dégradation visible de la caractéristique d'émission. Un filtre optique à charnière est d'ailleurs prévu à cette fin. Une diode photoélectrique sert de convertisseur photoélectrique à l'extrémité de réception. Cet élément, qui fonctionne sur quatre batteries fournissant 6 V, a une gamme de fréquence de 300 à 2500 Hz et emploie, en plus de la télécommande, 10 transistors et 3 diodes. Ses dimensions hors tout sont 315 × 240 × 115 mm et il pèse 2,5 kg.

EE 71 762 *pour plus amples renseignements*

Hartmann & Braun AG

97 Gräfstr., 6, Frankfurt-West, Allemagne

MULTIMÈTRE À COURANT CONTINU

(Illustration à la page 492)

La société Hartmann & Braun a exposé l'ELAVITRON, un multimètre à courant continu répondant aux conditions les plus rigoureuses, appartenant à la gamme d'instruments construits par ELIMA GmbH. Une haute sensibilité est assurée par un amplificateur transistorisé des plus sûrs (relais modulateur transistorisé à push-pull). Un total de 24 gammes permet la mesure de tensions et de courants entre 0,1 mV et 1000 V et entre 0,1 μA et 1 A respectivement. Des dispositifs de dérivation à fiches de 100 mV permettent d'étendre les gammes de courant à 100 A. Le grade de qualité 1,5 assure une grande précision d'indication, tandis que la grande et claire échelle à miroir avec deux étalonnages assure une grande précision de lecture. Le mouvement de mesure ainsi que l'amplificateur sont protégés par des fusibles contre les surtensions. L'ELAVITRON peut également être relié à des enregistreurs.

EE 71 763 *pour plus amples renseignements*

Herfurth GmbH

6-8, Beerenweg, 2000 Hambourg-Altona, Allemagne

CONTRÔLEUR D'AIR RADIOACTIF

Le contrôleur d'air radioactif H1346 (système KFA Jülich) a été conçu pour déterminer la mi-durée de vie effective en mesurant automatiquement les caractéristiques de persistance. Le premier des détecteurs employés mesure l'activité directement durant la période dite d'enrichissement, le second immédiatement après, et le troisième après un retard considérable pouvant s'étendre sur plusieurs jours. Les limites de détection suivantes s'appliquent à des aérosols artificiellement radioactifs après un cycle d'enrichissement de 3 heures et une activité naturelle de 300 pc/m^3:

après 18 minutes	500 pc/m^3
après 6 heures	5 pc/m^3
après 48 heures	0,3 pc/m^3

L'équipement électronique comprend trois intensimètres logarithmiques entièrement transistorisés couvrant une gamme de 10 à 10 000 coups de comptage par seconde. L'équipement a été conçu pour l'emploi avec tous les détecteurs obtenus dans le commerce afin d'assurer une application universelle. L'équipement a été construit pour le montage sur bâtis de 48 cm.

EE 71 764 *pour plus amples renseignements*

Hewlett-Packard Vertriebs GmbH

95 Kurhessenstr., 6 Francfort sur le Main, Allemagne

OSCILLOSCOPE DE 20 MHz

(Illustration à la page 493)

Des éléments à fiches d'une grande souplesse d'emploi facilitent les applications universelles en laboratoire d'un nouvel oscilloscope construit par Hewlett-Packard en Allemagne. Cet appareil peut servir soit comme oscilloscope de 20 MHz soit comme instrument très sensible de 10 μV pour la présentation d'événements plus lents. Un réflectomètre à temps transitoire avec temps de montée de 90 psec et utilisant des techniques d'échantillonnage a été réalisé pour les travaux microondes. Il peut mesurer et déceler les plus légères discontinuités dans les systèmes de transmissions coaxiales.

Dans le modèle 140 A, les deux éléments à fiches sont disposés l'un sur l'autre. Ils ont les mêmes dimensions hors-tout et sont directement reliés au système de déviation du tube cathodique. Un oscilloscope X-Y exige deux éléments à fiches identiques pour la déviation verticale. Un micromètre intérieur de 10 cm × 10 cm élimine les erreurs dues aux parallaxe de l'écran cathodique mat et non éblouissant Presque tous les éléments à fiches sont conçus comme amplificateurs différentiels sur lesquels des entrées couplées sur courant continu ou alternatif peuvent être choisies.

Les types suivants d'éléments à fiches peuvent être obtenus: Type 1400 A: largeur de bande réglable de 0 à 400 kHz, 100 μV/cm, une voie; type 1401 A: 0 à 450 kHz, 1 mV/cm, deux voies; type 1402A: 0 à 2 MHz, 5 mV/cm, deux voies; type 1403A: 0,1 Hz à 400 kHz, largeur de bande réglable, 10 μV/cm, une voie; type 1415A: élément d'échantillonnage, temps de montée de 90 psec, 1 mV/cm; type 1420A: base de temps de 0.5 μsec/cm à 5 sec/cm, déclenchement automatique; type 1421A, base de temps avec générateur de retard, 0,1 μsec/cm à 1 sec/cm.

EE 71 765 *pour plus amples renseignements*

Dr. Lehfeldt & Co. GmbH

B.P. 449, 6148 Heppenheim/Bergstr., Allemagne

MATÉRIEL DE CONTRÔLE À ULTRA-SONS

(Illustration à la page 493)

Le contrôleur à ultra-sons "Echoskop MPT1" est un appareil transistorisé fonctionnant sur batterie et employant la méthode à échos d'impulsions. Il peut également être employé avec deux sondes séparées fonctionnant suivant le système de transmission ou de réflexion. Un petit tube cathodique a été choisi pour réduire la consommation électrique mais la présentation est élargie au moyen d'une lentille montée devant l'écran. L'ensemble de l'appareil est transistorisé à l'exception du tube cathodique.

Le récepteur consiste en un amplificateur à large bande avec une gamme de 1 à 6 MHz et des sondes de n'importe quelle fréquence à l'intérieur de cette gamme peuvent être reliées. Toutefois, des fréquences de fonctionnement de 2 et de 4 MHz sont préférables.

On peut aussi effectuer des mesures d'épaisseur au dessus de de 2 mm et la pénétration maxima dans l'acier est d'environ 2 m pour les ondes longitudinales. La capacité de la batterie est suffisante pour un fonctionnement de 8 heures et elle peut être utilisée comme batterie tampon en liaison avec un chargeur d'entretien. L'appareil ne pèse que 4,75 kg, batterie comprise.

Un accessoire à signaux interchangeable peut être fourni pour la même batterie. Grâce à cet accessoire, des impulsions d'échos rencontrées dans une gamme de profondeur librement réglable et d'une hauteur minima également réglable sont indiquées par une lampe de signalisation.

EE 71 766 *pour plus amples renseignements*

Metrawatt AG

50-54 Schopperhofstr., 8500 Nuremberg, Allemagne

COMPENSATEUR DE THERMO-TENSION

(Illustration à la page 493)

En plus de l'indication usuelle des tensions thermoélectriques entre 0 et 50 mV, le compensateur de tensions thermoélectriques "Thermokomp" permet la lecture directe de température sur trois échelles: 0 à 1112° F (fer-constantan), 0 à 2192° F (chrome-alumel) et 0 à 2912° F (PtRh-Pt). La température de jonction froide peut être réglée continûment entre 32°F et 122° F.

Une tension électroniquement stabilisée fournie par piles sèches est appliquée à un fil à contact glissant dont on obtient la tension de compensation. Cette dernière à la tension soumise au contrôle au moyen d'un galvanomètre et réglée jusqu'à ce que le galvanomètre indique zéro. L'indicateur de l'échelle est accouplé mécaniquement au fil à contact glissant. Aucune charge n'est appliquée à la source de la tension de compensation une fois l'équilibre réalisé et le résultat est donc sensiblement indépendant de la résistance interne de la source et de la résistance des conducteurs.

L'échelle mesure 200 mm de longueur, sa précision est de ±0,3 % sur la totalité de l'échelle et la durée des batteries est d'environ 100 heures. Les dimensions hors-tout de l'appareil sont d'environ 200 × 210 × 150 mm, son poids est à peu près de 6,5 kg, batteries comprises. Il se prête non seulement à la mesure des tensions thermoélectriques et des tensions continues mais aussi à l'étalonnage d'instruments utilisés avec des couples thermoélectriques sur place.

EE 71 767 *pour plus amples renseignements*

ENREGISTREUR POTENTIOMÉTRIQUE

(Illustration à la page 493)

Le "Servogor" est un enregistreur potentiométrique portatif à auto-compensation pour travaux de banc d'essai ainsi que pour les applications en laboratoire. La mesure des tensions continues est effectuée conformément à la méthode potentiométrique, c'est à dire que le fil à contact glissant d'un potentiomètre est déplacé simultanément avec le style d'enregistrement par un servo-moteur. L'appareil comporte onze gammes de mesure commutables de 0 à 2 mV et pouvant aller de 0 à 20 V. La déviation totale effective peut être réglée à n'importe quelle valeur intermédiaire entre 40 % et 100 % de la gamme choisie. Ainsi lorsque la gamme de mesure de 10 mV a été choisie, elle peut être réglée pour couvrir n'importe quelle gamme entre 0 et 4 mV et entre 0 et 10 mV. La déviation maxima du style est de 200 mm sur papier de 230 mm de large d'un rouleau de 25 mm de longueur, ou de 180 mm sur papier format A4 lorsque la longueur utilisable est de 250 mm. Les vitesses d'alimentation en papier sont de 30, 120, 600 mm/h 30, 120, 600 mm/min. Un système marqueur de temps fournit une échelle des phénomènes enregistrés. La limite d'erreur est de ±0,5 % de la totalité de l'échelle, mais pas supérieure à ±20 μV (±0,2 lorsque des valeurs intermédiaires sont fixées); le temps de réponse est de 1 sec. L'appareil mesure 360 × 350 × 100 mm et il pèse environ 8 kg.

EE 71 768 *pour plus amples renseignements*

Olympia Werke AG

B.P. 960, 2940 Wilhelmshaven, Allemagne

LECTEUR DE BANDE PERFORÉE

(Illustration à la page 494)

Le prototype d'un lecteur de bande perforée avec dispositifs d'enroulement et de déroulement a été exposé à la Foire. Cet appareil permet la lecture de bandes à des vitesses de 9,8 pieds/sec, soit 1200 caractères par seconde. Le pont comporte un dispositif de compensation à ressorts et il peut être arrêté à un caractère près.

Le déroulement vers l'avant et le réenroulement s'effectuent à la même vitesse et n'exigent qu'une instruction ou la pression d'un bouton poussoir. Le prototype a été montré avec le code de bande à 8 voies, mais des lecteurs peuvent être fournis également pour d'autres codes.

Chaque caractère est lu à deux reprises, soit une première fois dans une matrice à mémoire d'où il est transmis en parté avec la deuxième lecture.

Les deux rouleaux de bandes peuvent être aisément échangées contre une spacieuse cassette pour l'analyse des boucles de bandes ou pour des travaux analogues.

EE 71 769 *pour plus amples renseignements*

P-E-K Electronic

Dr.-Ing. Paul E. Klein

B.P. 80, 7992 Tettnang, Allemagne

INDICATEUR MULTIPLE ELECTRONIQUE

Il devient de plus en plus difficile de vérifier toutes les nombreuses mesures et valeurs de contrôle essentielles aux processus automatiques. Le "Multiscop

3122" électronique offre un grand écran cathodique qui rend visibles un grand nombre de valeurs mesurées en succession rapide, de sorte que près de 400 valeurs mesurées apparaissent sur l'écran comme si elles étaient présentées simultanément. Des valeurs son électriques sont converties en valeurs électriques au moyen de transducteurs et préamplifiées si nécessaire. Dans le modèle standard, les tensions appliquées sont injectées au contrôleur à travers un commutateur de relais piloté par transistors à intervalles de 20 msec.

L'indication prend habituellement la forme de colonnes lumineuses disposées l'une près de l'autre et ayant des déviations dans les directions correspondant aux amplitudes négatives ou positives. Lorsque les amplitudes se produisent dans une seule direction, la trace peut être deux fois aussi longue. Les lignes d'étalonnage sont enregistrées simultanément avec les valeurs mesurées dont la déviation peut être comparée avec la valeur d'étalonnage et réglée en conséquence. L'image mesure 280 × 200 mm sur un écran mesurant 320 × 220 mm. Toutes les valeurs mesurées peuvent être transmises à un bâti supplémentaire livrable pour la conversion en valeurs numériques pour l'emmagasinement rapide ou l'impression.

EE 71 770 *pour plus amples renseignements*

Rohde & Schwarz

15 Mühldorfstr., 8000 Munich 8, Allemagne

GÉNÉRATEUR DE FRÉQUENCE STANDARD

(Illustration à la page 494)

Le nouveau générateur de fréquence standard Rohde & Schwarz XUC fournit des fréquences pour les mesures de haute précision. La dérive de fréquence de l'étalon standard incorporé est inférieure à 2 × 10^{-9}.

La fréquence de sortie peut être fixée sur deux échelles entre 470 et 1000 MHz. Elle est formée d'un composant basé sur l'étalon de fréquence stabilisée par cristal et elle est réglable en plots de 10 MHz et la fréquence à variation continue d'un oscillateur d'interpolation. L'erreur du réglage à variation continue demeure au dessous de 5 kHz. La tension de sortie du générateur de fréquence standard est réglable en plots ou continûment entre 0,1 mV et 1,5 V (dans 50Ω).

Lorsqu'il est nécessaire de répondre à des conditions plus élevées, l'oscillateur d'interpolation peut être remplacé par le générateur de fréquence standard XUA qui emploie le même principe de fonctionnement de base. L'erreur de calage est ainsi réduite à 0,5 Hz. Il est également possible d'utiliser un étalon de fréquence extérieur d'une plus grande stabilité pour actionner le générateur. En employant, par exemple, l'étalon de fréquence XSD de cette manière, on peut obtenir une fréquence de 1000 MHz avec une précision absolue de 0,2 Hz.

L'équipement pèse 70 kg et ses dimensions hors-tout sont de 535 × 445 × 615 mm.

EE 71 771 *pour plus amples renseignements*

RMS Regelungs- und Messtechnik Dipl.-Ing. Schmidt KG

Gutenbergstr., 2057 Reinbeck bei Hamburg, Allemagne

CONTRÔLEUR DE VIBRATIONS

(Illustration à la page 494)

Des comités allemands et internationaux ont adopté des normes appropriées pour le contrôle des vibrations faisant partie des spécifications pour de nombreux éléments et composants sur une échelle croissante. Les contrôleurs de vibrations RMS répondent aux conditions de ces spécifications d'essai.

Notre photo montre l'appareil SW100 comprenant un générateur d'ondes sinusoïdales RC, un amplificateur de puissance BF et le vibrateur électrodynamique. Sur le haut du cylindre on voit une douille pour le montage de l'opération d'essai.

L'appareil SW100 est conçu pour le contrôle de pièces pesant environ 1 kg et il peut produire des vibrations entre 5 et 2000 Hz avec des amplitudes maxima de ±3 mm. Les valeurs correspondantes pour l'appareil SW1000 sont de 5 kg en ce qui concerne le poids de la pièce soumise à l'essai et de 5 à 800 Hz pour la gamme de fréquence.

EE 71 772 *pour plus amples renseignements*

Sadowski & Co.

7301, Esslingen-Berkheim, Allemagne

GÉNÉRATEUR D'IMPULSIONS DOUBLES

Le générateur de doubles impulsions DIG 121 est entièrement à transistors au silicium et il a été conçu pour l'usage universel. Son petit format et son poids réduit s'allient à une fiabilité exceptionnellement élevée.

L'appareil comprend les éléments suivants: générateur d'espacement, générateur d'impulsions 1, étage de retard, générateur d'impulsions 2. Le générateur d'impulsions produit des impulsions de déclenchement à espacement réglable entre 1 µsec et 1 sec et il peut être synchronisé avec une onde sinusoïdale extérieure ou une tension d'impulsion. La sortie du générateur d'espacement déclenche le générateur d'impulsions 1; sa sortie d'impulsions à ondes carrées a une largeur pouvant varier entre 0,1 µsec et 0,1 sec. Le générateur d'impulsions 2 est de conception analogue et il est déclenché au moyen d'un circuit à retard donnant des temps différés réglables entre 0,1 µsec et 0,1 sec. Le temps de montée des impulsions est de 0,02 µsec, la période d'extinction étant de 0,03 µsec.

L'amplitude des impulsions de sortie est à variation continue entre 0 et 10 V et la polarité est réversible. Les sorties des deux générateurs d'impulsions peuvent être fournies séparément ou à l'aide d'un commutateur sous forme de double impulsion. Une source de tension extérieure (ondes sinusoïdales ou impulsions) peut être utilisée à la place du générateur d'espacement pour déclencher les générateurs d'impulsions. Lorsqu'il est déclenché extérieurement, le DIG 121 peut aussi être employé comme générateurs de temps différés.

EE 71 773 *pour plus amples renseignements*

Sodeco

Société des Compteurs de Genève

70 Grand-Pré, Genève, Suisse

APPAREIL DE SOMMATION TRANSISTORISÉ

(Illustration à la page 495)

Le sommateur transistorisé Sodeco SC 58 est construit sous forme d'élément à fiches et peut être utilisé pour l'enregistrement d'impulsions électriques provenant de différentes sources. Cet enregistrement s'effectue par un compteur ou imprimeur commun et sans pertes.

Le sommateur, qui est à circuits imprimés, emmagasine automatiquement et sans erreur des impulsions reçues indépendamment les unes des autres dans le temps et de différentes sources. Le réservoir est immédiatement balayé par une haute fréquence correspondante. La sortie du sommateur peut soit être reliée directement au compteur d'impulsions à action rapide SC60 (jusqu'à 250 coups par seconde) soit au moyen du convertisseur d'impulsions 3Tul à tous les types de compteurs d'impulsions Sodeco, y compris ceux équipés d'un imprimeur.

On peut employer jusqu'à 24 voies d'entrée et les signaux d'entrée peuvent être obtenus de dispositifs photo-électriques ou consister en impulsions de différentes formes et largeurs. L'appareil est fourni pour le fonctionnement sur courant secteur de 110 V, 125 V, 220 V ou 250 V.

EE 71 774 *pour plus amples renseignements*

COMPTEUR D'IMPULSIONS IMPRIMEUR

L'équipement de base se compose de trois éléments, à savoir: un aimant de comptage qui actionne simultanément l'indicateur à six chiffres et l'imprimeur à six chiffres, un mécanisme d'impression électrique et un dispositif électrique de zéro. Deux des éléments de base peuvent être disposés l'un près de l'autre et pour l'impression sur le même rouleau de papier.

Trois différents compteurs peuvent être incorporés:

1. Un compteur avec report des dizaines.
2. Un compteur horaire avec indication en heures, minutes, secondes, dixièmes de seconde, ou en heures, minutes, secondes, et
3. Un compteur de données sans zéro électrique, actionné par impulsions minutées.

Les compteurs d'impulsions sont fournis pour alimentation en continu (secteur ou batteries) ou pour alimentation en alternatif et leur taux de comptage maximum est de 10 coups/sec.

EE 71 775 *pour plus amples renseignements*

Telefunken AG
Division des Installations HF
3 Elisabethenstr., 79, Ulm, Allemagne

CALCULATRICE NUMÉRIQUE

La société Telefunken a exposé pour la première fois la calculatrice numérique "TR10", qui n'est pas seulement une machine complémentaire à la grande et puissante calculatrice "TR4" mais également une calculatrice numérique des plus souples de son propre chef. Elle peut être utilisée dans toute la gamme des études techniques et scientifiques pour la solution de problèmes de traitement de données commerciales, ainsi que pour le contrôle des procédés de fabrication industrielle qui gagnent continuellement en importance.

La TR10, qui est une calculatrice numérique décimale avec commande de programme emmagasiné exigeant fort peu d'espace, se caractérise par son excellente performance et particulièrement par sa capacité de fonctionnement simultané entre le dispositif de traitement et ses éléments périphériques. L'appareil permet également le traitement des cartes et bandes perforées. Un plus grand nombre de données peuvent être rapidement inscrites par un imprimeur à action rapide. Les réservoirs à noyau intérieur peuvent être augmentés par des réservoirs extérieurs à bande magnétique numérique ainsi que par des réservoirs à disques magnétiques. Le mécanisme de traitement des données peut être relié à des systèmes de télex et la commande de trafic peut être réglée à l'aide d'éléments qui permettent le tranfert à partir de points de mesure, de même qu'à des points de commande dans les procédés industriels. Le programme de calcul peut aussi être interrompu par des moyens extérieurs afin de provoquer la réaction à des signaux de priorité.

Telefunken offre ainsi une calculatrice numérique de capacité moyenne en plus de sa grande installation de calcul.

Le "TR10" est une calculatrice décimale et les données soumises au traitement sont donc inscrites de la façon employée dans la vie courante. Les chiffres décimaux, 26 lettres et 28 symboles, sont lus à partir du réservoir de calcul et assemblés pour le traitement comme des mots de n'importe quelle longueur. La capacité des réservoirs de calcul peut être choisie en plots de 10 000, entre 10 000 et 80 000 chiffres. La durée du cycle du réservoir à noyau de ferrite est de 8 μsec.

Des transistors planaires sont principalement employés dans le TR10 et dans l'équipement auxiliaire. Le dispositif de traitement central est logé dans un séduisant coffret mesurant 1800 $\times$ 600 $\times$ 400 mm et il pèse environ 200 kg.

En dehors de la performance technique de cette installation, le TR10 offre également des possibilités de programmation des plus variées pour les applications les plus diverses. Les adresses peuvent être modifiées aussi souvent que nécessaire et indépendamment les unes des autres. Les instructions pour le maniement des éléments individuels auxiliaires par la calculatrice peuvent être changées pour chacun des éléments afin de répondre à des besoins particuliers. Les instructions sont transmises de la calculatrice aux éléments périphériques, ces derniers les exécutant indépendamment et en même temps que la calculatrice.

EE 71 776 pour plus amples renseignements

Telefunken AG
Division des Tubes, Service de Distribution
100 Söflingerstr., 79, Ulm, Allemagne

CIRCUITS CONSTITUÉS DE CORPS SOLIDES

(Illustration à la page 495)

Les circuits constitués de corps solides consistent en général de silicium dans lequel des composants actifs et passifs sont diffusés selon les techniques dites planaires. L'isolement des composants dans des tablettes de silicium, placées l'une contre l'autre, est effectué par la diffusion de couches d'arrêt dans le silicium. La surface de silicium est recouverte d'une couche d'oxyde silicium sur laquelle des conducteurs métalliques sont déposés par évaporation pour relier les composants.

Les résistances produites par la diffusion ont un coefficient de température de $3 \times 10^{-3}/°$ C, et toute modification dans la température de 100° C produit ainsi un changement de résistance d'environ 30%. Une tension d'arrêt de 1 V donnera une capacité d'environ 150 pF/mm² entre cette résistance et le matériau de base. Dans le cas d'une résistance typique de 10 kΩ cette capacité sera de 5 pF répartis sur l'ensemble de la résistance.

On peut éviter ces désavantages en employant les nouvelles méthodes de fabrication hybride pour circuits constitués de corps solides. Des éléments actifs sont produits à l'aide des méthodes planaires comme auparavant, tandis que les éléments passifs sont déposés par évaporation dans un vide poussé. Ce processus réduit le coefficient de température à la valeur sensiblement plus réduite de $2 \times 10^{-4}/°$C, et un changement de température de 100° C produira donc un changement de résistance d'environ 2% seulement. La capacité du matériau de base ne sera plus que de 1/10 ème de celle des résistances diffusées.

Notre photo montre un demi-registre consistant en 4 portes NI, chacun des côtés mesurant 9 mm de long, ce qui est comparable à une punaise de bureau.

EE 71 777 pour plus amples renseignements

Résumés des Principaux Articles

Dispositifs magnétiques à plusieurs ouvertures par D. J. Morris

Résumé de l'article aux pages 440 à 447

Les noyaux magnétiques toroidaux sont d'un emploi étendu dans les applications de la logique binaire. L'addition de plusieurs ouvertures disposées en diverses configurations géométriques résulte en une série de dispositifs offrant de nombreux avantages d'importance. Ces dispositifs sont connus sous le nom de Transfluxors ou Dispositifs à plusieurs ouvertures. La principale caractéristique de ces dispositifs est l'isolement du circuit d'entrée du circuit de sortie. On peut ainsi obtenir la lecture non destructive et effectuer des connexions entre dispositifs en fil métallique uniquement.

Cet article décrit la performance des dispositifs à plusieurs ouvertures puis indique des moyens d'améliorer leur fonctionnement en modifiant leur forme. L'article examine enfin diverses configurations géométriques pour différentes applications de ces dispositifs.

Réalisation systématique de structures Cohn pour filtres passe-bande HF par K. E. Brown

Résumé de l'article aux pages 448 à 452

Une modification de la technique actuelle de transformation des bandes étroites a permis de produire un circuit de filtrage passe-bande à composants entassés. Ce circuit offre de nombreux avantages pratiques car les filtres conçus selon ce principe peuvent avoir des réponses centrées sur des fréquences pouant atteindre 200MHz avec des rapports de largeur de bande allant de 1% à 20%. Toutes les valeurs d'inductance voulues peuvent être spécifiées et des formules de réalisation sont présentées en cas d'inductances égales. Les inductances ne se produisent que dans les bras de dérivation du réseau en échelle, chaque inductance ayant une capacité de pont. En outre, des capacités spécifiées d'entrée et de sortie sont prévues. Ces dernières fournissent un chemin capacitif vers la masse à chaque mode de l'échelle. Cette méthode peut être appliquée sans difficulté lorsque les résistances de terminaison ne sont pas égales (même dans des conditions de charge à circuit ouvert). L'auteur montre que la technique standard de prédistorsion pour compenser la dissipation dans les inductances peut s'appliquer avec succès pour cette transformation.

Quelques amplificateurs à transistors à couplage direct de précision et leur réalisation approximative par C. W. B. Grigson

Résumé de l'article aux pages 454 à 459

L'analyse mathématique détaillée est de peu de valeur pratique durant les travaux de mise au point, à moins que les valeurs paramétriques des transistors individuels ne soient connues. Cet article indique des méthodes de réalisation permettant d'exécuter des circuits de précision tout en n'ayant qu'une connaissance approximative du transistor B, B_{be}, et V_{knee}.

Les circuits dont il est question sont: l'amplicateur de tension dit "ring-of-three", utilisé en tant qu'amplificateur de tension à dérive de $20\mu V/°C$, un amplificateur à gain stabilisé avec sensibilité totale de $3 \times 10^{-11}A$, une source à courant stabilisé de 0 à 20A avec une résistance de sortie de 300Ω, un amplificateur de courant classe B avec une capacité de 20A de crête à crête.

Un système logique NI à alimentation simple et à régime lent par G. Flanagan et L. Molyneux

Résume de l'article aux pages 460 à 463

Le système logique NI décrit dans cet article n'emploie qu'une seule valeur de résistance, le transistor le moins cher de sa gamme et un seul bloc d'alimentation. Il est à régime lent, son taux d'impulsions maximum étant de 500 coups/sec, mais il est approprié aux conditions d'utilisation où la vitesse des opérations est régie par les composants électromécaniques, comme c'est le cas dans le traitement des données.

Extraction de luminance du signal composé SECAM par G. Melchior et J. P. Doury

Résumé de l'article aux pages 464 à 465

Il est nécessaire pour certains usages d'extraire le signal de luminance du signal composé SECAM. Cet article traite d'une méthode permettant d'effectuer cette opération sans perte sérieuse de définition.

Un émetteur transistorisé de 85 MHz par A. E. Hilling

Résumé de l'article aux pages 466 à 469

Il s'agit ici d'un émetteur de 85 MHz à transistor au germanium à jonction par diffusion et alliage. Cet émetteur produit une puissance d'enveloppe de pointe de 1W lorsqu'il est modulé en amplitude à 100%.

Ce circuit conviendrait pour les émetteurs portatifs de faible puissance dans la bande de 71,5 à 88 MHz.

La mesure des paramètres de noyaux de ferrite à des courants de commutation très élevés par M. C. Stevens

Résumé de l'article aux pages 470 à 473

L'auteur décrit un appareil permettant de produire des impulsions de courant atteignant une amplitude de 10A avec un temps de montée d'un quart de nanoseconde. Cet appareil a été utilisé pour la mesure des paramètres de commutation des noyaux de ferrite à boucle d'hystérésis pour les trois applications où le mécanisme d'inversion est mis en oeuvre par mouvement mural, par rotation non uniforme et par rotation uniforme respectivement. Les paramètres mesurés pour huit types de noyau sont indiqués.

Perfectionnements dans la réalisation des résistances bobinées de précision par J. R. Price

Résumé de l'article aux pages 474 à 477

L'un des plus importants paramètres de la résistance bobinée de précision est sa stabilité à long terme. Une nouvelle formule de fabrication a été mise au point, qui permet d'obvier aux effets de la tension d'enroulement et du matériau de la bobine. Dans ces résistances l'enroulement est entièrement lâche et se meut librement sur la bobine. Des mesures de stabilité sur des résistances caractéristiques sont indiquées.

Un filtre simple de zéro avec fréquence variable par J. L. Douce et K. H. Edwards

Résumé de l'article aux pages 478 à 479

Cet article décrit comment une modification d'un réseau à "T" parallèle permet de régler la fréquence de transmission minima par un seul potentiomètre. Un analyse du filtre est donnée en appendice.

Un système de commande perfectionné pour stabilisateurs de courant alternatif par R. G. Ackland et N. M. Buckland

Résumé de l'article aux pages 480 à 485

Cet article traite d'un système simple d'entraînement de moteur asservi et sensible à la tension alternative pouvant, lorsqu'il est utilisé pour entraîner le Variac dans des stabilisateurs de tension de ligne alternative entraînés par moteur, maintenir la tension de sortie à un niveau constant à ±0, 1% près. Le taux de rétablissement pour les changements de points est de 18V/sec.

Transformateur de courant à large bande pour formes d'onde de courant par F. J. U. Ritson et J. Wood

Résumé de l'article aux pages 483 à 485

Cet article décrit un transformateur de courant et son amplificateur tels qu'utilisés pour l'examen des formes d'ondes de courants. L'appareil a une très grande largeur de bande et il introduit une impédance négligeable dans le circuit soumis à l'observation.

Elektronische Geräte auf der Hannover-Messe

Beschreibung einiger auf der Hannover-Messe vom 26. April - 5. Mai 1964 ausgestellten deutschen Erzeugnisse nach Angaben der Hersteller.

Übersetzung der Seiten 490 bis 495

Becker Flugfunkwerk GmbH
757 Baden-Baden-Oos, Flugplatz, Deutschland

BORD-SENDE-EMPFÄNGER

(Abbildung Seite 490)

Der ausserordentlich kompakte, quarzgesteuerte Sende-Empfänger AR380 hat im Bereich 117,0 ... 135,95 MHz 380 Sende- und Empfangskanäle mit einem Abstand von jeweils 50 kHz. Empfänger und Stromversorgungs-Modulatorteil sind volltransistorisiert; im Sender werden drei Röhren verwendet, um die Sendeleistung von 10 Watt zu erreichen. Die Transistorisierung sorgt nicht nur für das niedrige Gewicht des Gerätes, sondern gleichzeitig für seine Robustheit und den äusserst sparsamen Stromverbrauch.

Die Frontplatte entspricht mit 95 × 146 mm der ARINC-Norm; sie ist von innen beleuchtet und erlaubt auch bei Dunkelheit sichere Frequenzeinstellung durch grosse Zahlen und griffige Knöpfe. Gegen unbemerkte Funktionsstörung des Senders oder der Antennenanlage schützt der Mithörton, der in einem zusätzlichen Breitbandempfänger durch die eigene Sendung erzeugt und über einen Verstärker den Kopfhörern zugeführt wird.

Ein wesentliches Konstruktionsmerkmal des AR380 ist die Baugruppentechnik. Die Anlage entspricht allen gültigen Bestimmungen und Vorschriften; der Sende-Empfänger wiegt 2,5 kg, die Stromversorgung 2,0 kg.

EE 71 751 *für weitere Einzelheiten*

Bölkow-Apparatebau GmbH
7311 Nabern-Teck, Deutschland

MESSWERT-ERFASSUNGSSYSTEM

(Abbildung Seite 490)

Das Bölkow-System 2000 ist ein Messdaten-Erfassungssystem, das mit Hilfe von Messwertwandlern die unterschiedlichsten zu überwachenden Grössen wie Temperatur, Druck, chemische Zusammensetzung von Gasen, Beschleunigung, mechanische Schwingungen, Durchfluss u.ä. in Gleich- oder Wechselspannungen umsetzt. Die Messwertaufnehmer haben je nach Art und Ausführung eine Messunsicherheit von 0,1 bis 10%. Grundbausteine des Systems sind der Analog-Digital-Spannungsumsetzer und der Digital-Frequenzmesser, die Spannungen mit einer Unsicherheit von ca. 0,01 bis 0,2% und Frequenz und Zeit mit $10^{-5} \ldots 10^{-7}$ Unsicherheit messen und dann anderen Geräten des Systems zur Weiterverarbeitung und Registrierung zuführen. Als solche sind Digital-Uhren, Code-Wandler, Digital-Anzeigeeinheiten, Druckwerke, Digital-Vorwahlgeräte und Programmierbare Impulsgeneratoren lieferbar.

Bei der Durchführung komplizierter und umfangreicher Entwicklungs- und Forschungsaufgaben erweist es sich als besonders vorteilhaft, wenn für die Untersuchung von Teilproblemen dieselben Messgeräte eingesetzt werden können, die sich später bei der Prüfung des Gesamtprojektes zu einer programmgesteuerten Grossanlage integrieren lassen. Für kleinere Aufgaben oder Teilprobleme werden dafür mit normierten digitalen Baugruppen des Bölkow-Systems 2 000 spezielle Steuer- oder Rechengeräte zusammengestellt. Für Grossanlagen stehen Digitalrechner der Firma Compagnie des Compteurs (SETI) zur Verfügung, die sich direkt mit dem System 2 000 kombinieren lassen. Alle Bausteine des Bölkow-Systems 2 000 sind in 19"-Einschubtechnik ausgeführt.

EE 71 752 *für weitere Einzelheiten*

braun electronic gmbh
7808 Waldkirch/Breisgau, Lange Str. 109, Deutschland

BERÜHRUNGSLOSE TEMPERATUR-MESSUNG

Einzelgeräte des elektronischen Temperaturstrahlungs-Messystems TASTOTHERM-INFRA wurden erstmalig von der braun electronic gmbh vorgestellt. Eine Anlage zum berührungslosen Messen in serienmässigen Bereichen zwischen +100 und +3 000°C besteht jeweils aus einem Messkopf und einem angeschlossenen Grundgerät, das den Messverstärker, Stromvorsorgung und gegebenenfalls Grenzwertschalter enthält.

Das Messignal liegt als elektrische Grösse vor, Fernmessung und Registrierung des zeitlichen Temperaturverlaufes liegen deshalb in der Natur der Messmethode. Im Messkopf wird die einfallende Wärmestrahlung in ein elektrisches Signal umgesetzt und vorverstärkt. Die Grösse des Messfleckens und des Messabstandes sind innerhalb angegebener Grenzen frei wählbar. Ausserdem geht auch der Emissionsfaktor des Messobjektes, d.h. das Verhältnis der Strahlungsleistung des betreffenden Stoffes zur Strahlungsleistung eines schwarzen Körpers, und der Blendendurchmesser in die Messung ein. Anlagen werden entweder als Einzweck-Einrichtungen für bestimmte Stoffe oder als Mehrzweck-Geräte für mehrere Emissionsfaktoren geeicht geliefert.

EE 71 753 *für weitere Einzelheiten*

Brown, Boveri & Cie AG
68 Mannheim, Deutschland

MIKROWELLENGERÄT

(Abbildung Seite 491)

BBC stellte das neue volltransistorisierte Richtstrahlgerät FM12/7000 für 7 GHz vor. Durch ausschliessliche Verwendung der Halbleitertechnik wurde ein Gerät mit vollkommen neuer Konzeption geschaffen, das nur mit Silizium-Transistoren und -Dioden bestückt ist. Dadurch entfallen die komplizierten Mikrowellenröhren, die Leistungsaufnahme sinkt auf einen Bruchteil der früher üblichen Werte, und die Frequenzkonstanz wird durch Quarz-Steuerung so verbessert, dass ein Kanalabstand von 1 MHz eingehalten werden kann.

Das für Einsatz im Frequenzbereich 6400 ... 8400 MHz ausgelegte Richtstrahlgerät kann für 12 oder 24 Telefoniekanäle eingerichtet werden. Auch zur Übertragung eines hochwertigen Musikkanals ist das Gerät geeignet, das einen Verbrauch von nur 10 W hat. In Normalausführung werden Sende-, Empfangsgerät und Antennenspiegel als Einheit zusammengebaut. Zwischenschalten eines Wellenleiters bis zu 50 m Länge zur Montage unabhängig vom Antennenspiegel ist jedoch möglich.

EE 71 754 *für weitere Einzelheiten*

DeTeWe
Deutsche Telephonwerke und Kabelindustrie AG
1 Berlin 36, Wrangelstrasse 100, Deutschland

MEHRZWECK-MESS- UND PRÜFGERÄT

Mit diesem kombinierten Gerät können mehrere unterschiedliche Mess- und Prüfvorgänge auf den Gebieten der Fernsprech - Vermittlungstechnik sowie Steuerungs- und Regeltechnik durchgeführt werden. Es enthält einen universell verwendbaren Impulsgeber und einen Zeit- bzw. Frequenzmesser.

Die Dauer der Impulse und der dazwischenliegenden Pausen lässt sich an je fünf Wahlschaltern dekadisch einstellen. Die Anzahl der abzugebenden Impulse (1 ... 12 oder fortlaufend) ist ebenfalls einstellbar und automatische Wiederholung der Impulsserien möglich.

Die dem Gerät zugeleiteten Sinuswellen oder Impulse werden in einem Zeitintervall von 1 oder 10 s gezählt und das Ergebnis mit Angabe der Kommastelle auf der Leuchtziffernanzeige dargestellt. Der Messbereich ist 10...99 999 Hz.

Während einer zu messenden Zeit werden die mit grosser Konstanz erzeugten Schwingungen der Zeitbasis über eine Torschaltung auf den elektronischen Zähler gegeben, d.h. das Tor wird durch den Messeingang gesteuert, und die Leuchtziffernanzeige entspricht der Toröffnungszeit. Der Messbereich ist 1 ... 99 999 ms.

EE 71 755 *für weitere Einzelheiten*

Frieseke & Hoepfner GmbH
8520 Erlangen-Bruck, Postfach Nr. 72, Deutschland

STRAHLUNGSMESSGERÄTE

(Abbildung Seite 491)

Als Weiterentwicklung des Lieferprogrammes kernphysikalischer Messgeräte stellte Frieseke & Hoepfner erstmalig transistorisierte Bausteine des Transistor-Geräte-Programmes TGP aus. Die Bausteine sind in Einheitseinschüben UC/2 untergebracht, von denen zwei nebeneinander einen üblichen 19"-Einschub ergeben. Gleichzeitig wurden auch die erforderlichen Tischgehäuse für die verschiedenen Kombinationen geschaffen.

Die abgebildete Messanordnung besteht aus Hochspannungserzeuger FHT 101A1, linearem Ratemeter FHT 110B, Impulszähler FHT 120AZ und Zeitzählwerk FHT 120A3. Der netzbetriebene Hochspannungserzeuger gibt für Detektoren auf negative oder positive Polarität umschaltbare Spannungen zwischen 200 und 500 V bei max. 5 mA mit einer Konstanz von besser als 0,2‰ über 24 h ab. Die absolute Einstellgenauigkeit ist ±2%, die kleinste Schrittgrösse 1 V.

Das lineare Ratemeter zeigt den zeitlichen Mittelwert einer Impulsfolge in elf einstellbaren Messbereichen mit Endwerten zwischen 10^2 und 10^7 ipm an. Zwei Impulsrateneingänge können auch gegeneinander in Differenz geschaltet werden. Im Impulszählwerk werden die in den Strahlungsdetektoren erzeugten Impulse nach Verstärkung 6stellig auf Dekadenzählrohren angezeigt. Eine Zählsteuerung ist eingebaut, jedoch erlaubt ein Anschluss "Mehrzwecksteuerung" Betrieb mit beigeordneten Steuersystemen, z.B. Einkanal-Impulshöhenanalysator FHT 300A. Eine Ziffernvorwahl in Zehnern für automatischen Selbststopp ist vorhanden. Das quarzgesteuerte Zeitzählwerk steuert den zeitlichen Ablauf der Messung mit einer Konstanz von besser als 10^{-8} und einem Auflösungsvermögen von 1 µs. Die digitalangezeigten Werte lassen sich mittels eines Druckwerkes registrieren.

Das Programm umfasst auch alle erforderlichen Verstärker-Einschübe, und fast alle Bausteine sind für Netzanschluss oder für Batteriebetrieb lieferbar.

EE 71 756 *für weitere Einzelheiten*

fuba, Werk elektronischer Bauteile u. Geräte
Hans Kolbe & Co.
3371 Gittelde üb. Seesen, Deutschland

STROMSTABILISIERTES 400-A-NETZGERÄT

Die Grundentwicklung dieses Gerätes wurde vom Deutschen Elektronen Synchrotron (DESY) in Hamburg ausgeführt. Das Gerät dient zur Versorgung von z.B. Ablenkmagneten für Teilchen-Beschleuniger und andere induktive Verbraucher, die einen hohen konstanten Gleichstrom benötigen. Der Ausgangsstrom kann mit einem 4stelligen Ziffernumschalter in Stufen von 40 mA eingestellt werden, z.B. 384,8 oder 38,48 A. Der einmal eingestellte Strom ist weitgehend unabhängig von Änderungen der Eingangsspannung und der Last. Bei maximaler Ausgangsleistung von 16 kW kann die Ausgangsspannung 0 ... 40 V betragen.

In einem Schrank von 580 × 555 × 1640 mm sind der Netztransformator von 20 kVA und die Netzdrossel für 400 A fest eingebaut und die Funktionseinheiten als Einschübe in 19"-Norm ausgebildet. Der Forderung der Auftraggeber entsprechend wurde die ca. 400 kg wiegende Anlage für zusätzliche externe Bedienungs- und Überwachungsmöglichkeiten eingerichtet.

EE 71 757 *für weitere Einzelheiten*

DIGITALE EINSTECK-BAUSTEINE

(Abbildung Seite 491)

Die erstmalig in 1963 gezeigten Digitalbausteine des Systems Darmstadt werden nunmehr in drei Geschwindigkeitsklassen für Taktfrequenzen von 250 kHz, 5 MHz und über 10 MHz hergestellt. Der wesentliche Vorzug des Systems besteht in der strengen Normung der Anpassungsbedingungen der einzelnen Bausteine, durch die die Planung eines Gerätes bedeutend vereinfacht wird. Einheitliche Anschlussmöglichkeiten gestatten eine Kombination der verschiedenen Baureihen innerhalb einer Schaltung.

Lieferbar sind Flipflops, Inverters, Quarzoszillatoren, Impulsformer, Ausgabeverstärker und Nixietreiber.

EE 71 758 *für weitere Einzelheiten*

P. Gossen & Co. GmbH
8520 Erlangen, Nägelsbachstr. 25, Deutschland

GLEICHSPANNUNGSVORVERSTÄRKER

(Abbildung Seite 492)

Bei niedrigen Temperaturen von Thermoelementen abgegebene Spannungen sind zu klein, um auch hochempfindliche Drehspulmesswerke noch betriebssicher betreiben zu können. Es ist jetzt jedoch gelungen, durch Verwendung von Magnetverstärkern auch kleinste Thermospannungen mit normalen robusten Drehspulmessgeräten zu messen. In einem Schaumodell wurde gezeigt, wie die bei +60° C von einem Fe-Konst-Element abgegebene Spannung von 3,19 mV durch Zwischenschalten eines Gleichstrom-Vorverstärkers PMV 10 in einen Strom von 1 mA umgeformt wird, so dass der Temperaturbereich 0 ... +60° C betriebssicher auf einem unempfindlichen Drehspulstrommesser mit 250°-Zeigerausschlag abgelesen werden kann.

Der Magnetverstärker wird von einem transistorisierten Rechteckspannungs-Oszillator mit einer 1-kHz-Hilfsspannung gespeist. Durch die Wahl dieser Frequenz erreicht man kleine Abmessungen, kleine Zeitkonstante (< 0,1 s) und geringe Restwelligkeit des Ausgangsgleichstroms. Der mit 24 V Gleichspannung oder 220 V, 50 Hz Wechselspannung betreibbare Vorverstärker kann von hinten auf die Grundplatte von Messgeräten mit Flansch 144 × 144 mm montiert werden.

EE 71 759 *für weitere Einzelheiten*

Gebr. Grieshaber
762 Wolfach, Postfach 30, Deutschland
Vertrieb: VEGA GmbH, 78 Freiburg i. Breisgau, Postfach 1649, Deutschland

FÜLLSTANDMESSGERÄTE

(Abbildung Seite 492)

Die von Grieshaber ausgestellten neuen transistorisierten Füllstandmessgeräte sind ausschliesslich oder bevorzugt mit Silizium-Planar-Transistoren bestückt, um der Forderung nach grösstmöglicher Zuverlässigkeit Rechnung zu tragen. Die Geräte arbeiten nach dem kapazitiven Messprinzip, jedoch wird statt der Spannung hier der Strom des Kondensators gemessen, woraus sich zwangsläufig eine elektrische Grösse ergibt, die sich linear proportional zur veränderlichen Kapazität verhält.

Durch Verlegung des Oszillators in den Elektrodenkopf sind lange, unabgeschirmte Verbindungsleitungen zulässig. Sowohl der Elektrodenkopfeinsatz, wie auch das eigentliche elektronische Gerät können ohne Auswechslung irgendwelcher elektronischer Bauteile in einem

Bereich von 10 ... 10 000 pF Verwendung finden, trotzdem die Grenzstandanzeiger dieser Serie bereits auf Kapazitätsänderungen von weniger als 0,5 pF ansprechen. Besondere Aufmerksamkeit wurde auch der thermischen Stabilität gewidmet.

Das Gerät ist in Baukastentechnik ausgeführt; die Abbildung zeigt den allen Ausführungen gemeinsamen Netzteil und den neuen Doppelgrenzschalter Vegator/Vegasel einmal für Wandmontage und in Einbauausführung.

EE 71 760 *für weitere Einzelheiten*

Grundig Werke

8510 Fürth, Deutschland

NUMERISCHE MASCHINENSTEUERUNG

(Abbildung Seite 492)

Das erstmalig vor sechs Monaten in Mailand gezeigte numerische Positioniersystem für Werkzeugmaschinen wurde in ausgebauter Form mit 8-Kanal-Lochstreifeneingabe vorgestellt. Die in Bausteintechnik ausgeführte Positioniersteuerung mit Schaltinformation kann je nach den gestellten Forderungen für eine breite Reihe von Maschinen zusammengesetzt und auch nachträglich ergänzt werden. Unter Schaltinformationen wird das Auslösen von Zusatzfunktionen an der Maschine verstanden, wie z.B. verschiedene Vorschubgeschwindigkeiten und Drehzahlen, Wechseln von Werkzeugen oder Zuführen von Kühlmitteln. Die Steuerung kann für ein bis drei Koordinaten geliefert werden, und für die Positionierung in zwei Koordinaten können jetzt bereits 20 Schaltinformationen in vier voneinander unabhängigen Gruppen programmiert werden.

Das in Einschubtechnik ausgeführte Steuergerät ist mit einer rasch auswechselbaren Lochstreifenkassette mit Endlosband ausgestattet. Die durchgehende Satznummer des Programmes wird am Bedienungspult mit Glimmlicht-Ziffern angezeigt. Das System ist nicht nur für Bohren, sondern auch für einfaches Fräsen verwendbar.

Innerhalb des Verfahrweges von fast 10 m kann der Nullpunkt frei gewählt werden, und das Ergebnis der über einen lichtelektrischen Weggeber gemessenen Strecken wird mittels eines Vor-Rückwärtszählers digital angezeigt.

In der einfachen Ausführung nur für Handeingabe auf zwei Koordinaten enthält der Steuerschrank anstelle des Streifenlesers einen Blindeinschub für den späteren Ausbau.

EE 71 761 *für weitere Einzelheiten*

LICHTSPRECHGERÄT

(Abbildung Seite 492)

Eine leicht tragbare Lichtsprechanlage besteht aus zwei Geräten LiG3, die je eine Sende- und Empfangseinrichtung für moduliertes Licht enthalten und Gegensprechverkehr über Entfernungen bis zu 2 km gestatten. Die Geräte können auf handelsübliche Kamerastative mit Kinoneiger montiert und durch eingebaute Zielfernrohre aufeinander eingestellt werden.

Zum Sprechen und Hören dient ein Fernsprechhandapparat mit Sprechtaste, jedoch können auch Zusatzgeräte wie Verstärker und Lautsprecher oder Tonbandgeräte angeschlossen werden. Eine Fernbedienung enthält auch eine vollständige Rufeinrichtung mit Summer und Ruftaste. Es ist also möglich, der Gegensprechstelle durch einen Summerton über Lichtstrahl die Sprechabsicht anzukündigen.

Die Lichtwellen werden auf direktem Wege über den Betriebsstrom der als Lichtquelle verwendeten Glühlampe moduliert. Der sichtbare Spektralanteil lässt sich ohne merkbare Beeinträchtigung der Übertragungseigenschaften durch ein schwenkbares optisches Filter ausschalten. Als photoelektrischer Wandler auf der Empfangsseite dient eine Photodiode. Vier Monozellen mit insgesamt 6 V speisen das Gerät, das einen Frequenzbereich von 300 ... 2 500 Hz hat und einschliesslich Fernbedienung mit 10 Transistoren und 3 Dioden bestückt ist. Bei Abmessungen von 315 × 240 × 115 mm wiegt es ca. 2,5 kg.

EE 71 762 *für weitere Einzelheiten*

Hartmann & Braun AG

6 Frankfurt (Main) West, Gräfstr. 97, Deutschland

GLEICHSTROM-VIELFACHMESSGERÄT

(Abbildung Seite 492)

Aus dem Programm der ELIMA GmbH zeigte Hartmann & Braun ein Gleichstrom-Vielfachmessinstrument für höchste Ansprüche, das ELAVITRON. Durch einen betriebssicheren mit Transistoren bestückten Messverstärker (Transistor-Gegentaktzerhacker) wird eine hohe Messempfindlichkeit gewährleistet. Insgesamt 24 Messbereiche gestatten Spannungs- und Strommessungen zwischen 0,1 mV bis 1 000 V bzw. 0,1 μA bis 1 A. Durch aufsteckbare Nebenwiderstände 100 mV kann man die Strombereiche bis auf 100 A erweitern. Die Güteklasse 1,5 gewährleistet eine gute Anzeigegenauigkeit, die grosse übersichtliche Spiegelskala mit nur 2 Teilungen eine hohe Ablesegenauigkeit. Anzeigeinstrument und Verstärker sind durch Sicherungen gegen grobe Überlastungen geschützt. Das ELAVITRON gestattet auch den Anschluss eines Registrierinstrumentes.

EE 71 763 *für weitere Einzelheiten*

Herfurth GmbH

2000 Hamburg-Altona, Beerenweg 6-8, Deutschland

LUFTÜBERWACHUNGSGERÄT

Das Luftüberwachungsgerät H 1346 (System KFA Jülich) gestattet die Ermittlung der effektiven Halbwertszeit durch automatische Messung der Abklingkurve. Von den verwendeten Detektoren misst der erste die Aktivität direkt während der Anreicherungszeit, der zweite unmittelbar danach und der dritte schliesslich mit grosser Verzögerung, die frei wählbar ist bis zu mehreren Tagen. Bei einem Anreicherungszyklus von 3 Stunden und einer natürlichen Aktivität von 300 pC/m³ ergeben sich folgende Nachweisgrenzen für künstliche radioaktive Aerosole:

nach 18 Min. 500 pC/m³
nach 6 Std. 5 pC/m³
nach 48 Std. 0,3 pC/m³

Die Elektronik besteht aus 3 volltransistorisierten logarithmischen Ratemetern mit einem Messbereich von 10 ... 10 000 Imp/min. Um eine möglichst universelle Einsetzbarkeit des Gerätes zu gewährleisten, wurde das Gerät so ausgelegt, dass sich an ihm alle handelsüblichen Detektoren verwenden lassen. Der Aufbau erfolgt in 19" Einschubtechnik.

EE 71 764 *für weitere Einzelheiten*

Hewlett-Packard Vertriebs GmbH

6 Frankfurt/Main, Kurhessenstr. 95, Deutschland

20-MHZ-OSZILLOGRAF

(Abbildung Seite 493)

Ein neuer, in Deutschland hergestellter Oszillograf von Hewlett-Packard ermöglicht durch seine vielseitige Einschubtechnik universelle Verwendung im Labor; so dient er wahlweise als 20 MHz Oszillograf oder als sehr empfindliches 10 μV Gerät für die Darstellung langsamerer Vorgänge. Für die Mikrowellentechnik wurde ein Laufzeit-Reflektometer mit 90 psec Anstiegszeit (Sampling-Technik) entwickelt. Damit können noch kleinste Fehlstellen in Koaxial-Übertragungssystemen gemessen werden.

Die beiden Einschübe des 140A sind übereinander angeordnet. Sie haben die gleichen Ausmasse und arbeiten direkt auf die Ablenkplatten der Bildröhre. Zwei identische Vertikal-Einschübe ergeben einen X-Y-Oszillografen. Innenraster vermeidet Parallaxenfehler auf der blendfrei mattierten 10 × 10 cm Bildröhre. Fast alle Einschübe sind als Differential-Verstärker konstruiert mit wählbaren gleich- oder wechselspannungsgekoppelten Eingängen.

Folgende Einschübe sind lieferbar: 1400A: 0 ... 400 kHz, einstellbare Bandbreite, 100 μV/cm, 1 Kanal; 1401A: 0 ... 450 kHz, 1 mV/cm, 2 Kanäle; 1402A: 0 ... 20 MHz, 5 mV/cm, 2 Kanäle; 1403A: 0,1 Hz ... 400 kHz, einstellbare Bandbreite, 10 μV/cm, 1 Kanal; 1415A: Sampling-Einschub, 90 ps, 1 mV/cm; 1420A: Zeitbasis, 0,5 μs...5 s/cm, automatische Triggerung; 1421A: Zeitbasis mit Verzögerungsgenerator, 0,1 μs ... 1 s/cm.

EE 71 765 *für weitere Einzelheiten*

Dr. Lehfeldt & Co. GmbH

6148 Heppenheim/Bergstr., Postfach 449, Deutschland

ÜBERSCHALL-PRÜFGERÄT

(Abbildung Seite 493)

Das Echoskop MPT1 ist ein transistorisiertes, batteriegespeistes Ultraschall-Prüfgerät für das Impuls-Echo-

Verfahren, das aber auch mit zwei getrennten Schallköpfen nach dem Druchstrahlungs- bzw. Reflexionsverfahren arbeiten kann. Aus Gründen der Stromersparnis ist das Gerät mit einem kleinen Kathodenrohr ausgerüstet, dessen Schirmbild durch eine vorgesetzte Lupe vergrössert wird. Ausser diesem Kathodenrohr besteht die Bestückung nur aus Transistoren.

Der Empfänger ist ein Breitbandverstärker für den Bereich 1 ... 6 MHz, und Schallköpfe mit beliebiger Frequenz innerhalb dieses Bereiches können angeschlossen werden. Vorzugsweise arbeitet man mit 2 oder 4 MHz.

Dickenmessung ist ab 2 mm möglich, und die maximale Durchdringungstiefe bei Stahl beträgt ca. 2 m für Longitudinalwellen. Die Batterie reicht für 8 Stunden Betrieb aus und kann zusammen mit einem Ladegerät auch als Pufferbatterie benutzt werden. Das Gerät wiegt einschliesslich Batterie nur ca. 4,8 kg.

Ein aufsteckbarer Signalzusatz wird aus der gleichen Batterie gespeist. Er ermöglicht, in einem beliebig einstellbaren Tiefenbereich auftretende Echoimpulse von ebenfalls einstellbarer Mindesthöhe durch Aufleuchten einer Signallampe zur Anzeige zu bringen.

EE 71 766 *für weitere Einzelheiten*

Metrawatt AG

8500 Nürnberg, Schopperhofstr. 50-54, Deutschland

THERMOSPANNUNGSKOMPENSATOR

(Abbildung Seite 493)

Der Thermospannungskompensator "Thermokomp" erlaubt neben der üblichen Ablesung der Thermospannungen von 0 ... 50 mV das direkte Ablesen in °C auf drei Skalen: 0...600°C (Fe-Konst), 0 ... 1 200°C (NiCr-Ni) und 0 ... 1 600°C (PtRh-Pt). Die Vergleichsstellentemperatur kann stufenlos zwischen 0° und +50°C eingestellt werden.

Eine aus Trockenbatterien entnommene elektronisch geregelte Spannung liegt an einem Schleifdraht, von dem die Kompensationsspannung abgegriffen wird. Diese Kompensationsspannung wird mit der zu messenden Spannung über ein Nullgalvanometer verglichen und verändert, bis Null angezeigt wird. Der Skalenschieber ist mechanisch mit dem Schleifdrahtabgriff verbunden. Da nach dem Nullabgleich die Messspannungsquelle nicht mehr belastet wird, ist das Messergebnis von deren Innenwiderstand sowie vom Widerstand der Zuleitungen weitgehend unabhängig.

Die Skalenlänge ist 200 mm, die Messunsicherheit ±0,3% des Bereichendwertes und die Lebensdauer der Batterien ca. 100 Stunden. Bei Abmessungen von 300 × 210 × 150 mm wiegt das Gerät einschliesslich Batterien ca. 6,5 kg. Anwendungsgebiete sind nicht nur das Messen von Thermo- und Gleichspannungen, sondern auch das Eichen von Betriebsmessgeräten zum Anschluss an Thermopaare.

EE 71 767 *für weitere Einzelheiten*

POTENTIOMETERSCHREIBER

(Abbildung Seite 493)

Der "Servogor" ist ein tragbarer selbstabgleichender Kompensationsschreiber im Tischformat für Anwendungen im Labor. Die Messung der Gleichspannung erfolgt nach dem Potentiometermessverfahren, wobei ein Servomotor den Abgriff eines in einer Brücke liegenden Potentiometers verstellt und gleichzeitig die Schreibfeder bewegt. Er hat elf umschaltbare Messbereiche von 0 ... 2 mV bis zu 0 ... 20 V. Als tatsächlicher Bereichendwert kann jeder Zwischenwert von ca. 40 ... 100% des gewählten Bereiches eingestellt werden, so dass z.B. bei gewähltem Messbereich 10 mV ein Bereich von 0 ... 4 mV bis 0 ... 10 mV einstellbar ist. Die Schreibbreite ist 200 mm auf einer 25 m langen und 230 mm breiten Papierrolle oder 180 mm auf einem Blatt A4 mit 250 mm nutzbarer Vorschublänge. Der Papiervorschub erfolgt mit den umschaltbaren Vorschubgeschwindigkeiten 30-120-600 mm/h und 30-120-600 mm/min. Der zeitliche Ablauf kann von Zeitmarkiersystemen registriert werden. Die Fehlergrenze ist ±0,5% vom Messbereichendwert, jedoch nicht besser als ±20 µV (bei Einstellung von Zwischenwerten ±0,2%), die Einstellzeit beträgt 1 s. Abmessungen 360 × 350 × 100 mm, Gewicht 8 kg.

EE 71 768 *für weitere Einzelheiten*

Olympia Werke AG

2940 Wilhelmshaven, Postfach 960, Deutschland

LOCHSTREIFENLESER

(Abbildung Seite 494)

Der Prototyp eines fotoelektrischen Lochstreifenlesers mit Auf- und Abwickelvorrichtung wurde gezeigt, in dem der Lochstreifen mit einer Geschwindigkeit von 3 m/s läuft, d.h. 1200 Zeichen je Sekunde ausliest. Die Ausführung hat eine federnde Kompensation und kann innerhalb eines Zeichens gestoppt werden.

Vor- und Rücklauf erfolgt mit derselben Geschwindigkeit und erfordert nur einen Befehl oder Drücken einer Taste. Der Prototyp wurde mit 8-Kanal-Code vorgeführt, jedoch kann das Gerät auch für andere Codes ausgeführt werden.

EE 71 769 *für weitere Einzelheiten*

P-E-K Electronic
Dr.-Ing. Paul E. Klein

7992 Tettnang, Postfach 80, Deutschland

ELEKTRONISCHES VIELFACHANZEIGEGERÄT

Das elektronische "Multiscop 3122" hat ein Grossbild-Sichtgerät, auf dem eine grosse Zahl von Messwerten in rascher Folge sichtbar gemacht werden, so dass bis zu 400 Messgrössen dem Beobachter scheinbar gleichzeitig erscheinen. Nichtelektrische Werte werden mittels Messwertaufnehmer in elektrische Werte umgewandelt und gegebenenfalls vorverstärkt. In der Normalausführung werden die zugeführten Spannungen mit einem transistorgesteuerten Relais-Umschalter in Abständen von 20 ms dem Sichtgerät zugeführt.

Die Anzeige erfolgt in der Regel in nebeneinander angeordneten Leuchtsäulen, die bei positiven und negativen Amplituden einen entsprechenden Ausschlag ergeben. Bei Amplituden in nur einer Richtung ist der Ausschlag doppelt so gross. Zusammen mit den Messwerten werden Eichlinien geschrieben, deren Höhe man mit Eichwerten vergleichen und entsprechend einstellen kann. Bei einer Schirmgrösse von 320 × 220 mm ist das Bildformat 280 × 200 mm. Alle Messwerte können herausgeführt und mit einem zusätzlich lieferbaren Messgestell in Digitalwerte umgeführt und Schnellspeichern oder Druckern zugeführt werden.

EE 71 770 *für weitere Einzelheiten*

Rohde & Schwarz

8000 München 8, Mühldorfstr. 15, Deutschland

NORMALFREQUENZGENERATOR

(Abbildung Seite 494)

Der neue Normalfrequenzgenerator XUC von Rohde & Schwarz erzeugt Frequenzen für Messungen hoher Präzision. Die Frequenzänderung des eingebauten Frequenznormals ist kleiner als 2×10^{-9}.

Ein aus der Frequenz des quarzstabilisierten Frequenznormals gebildeter, in 10-MHz-Schritten rastbarer Anteil und die zwischen diesen Schritten stetig durchstimmbare Frequenz eines Interpolations-Oszillators bilden die zwischen 470 und 1 000 MHz auf zwei Skalen einstellbare Ausgangsfrequenz. Bei der kontinuierlichen Einstellung bleibt der Fehler unter 5 kHz. Die Ausgangsspannung des Normalfrequenzgenerators lässt sich in Stufen oder kontinuierlich von 0,1 mV bis 1,5 V (an 50 Ω) einstellen.

Für höchste Ansprüche kann der Interpolations-Oszillator durch den nach dem gleichen Prinzip arbeitenden Normalfrequenzgenerator XUA ersetzt werden. Dadurch vermindert sich der Einstellfehler auf 0,5 Hz. Ausserdem besteht die Möglichkeit, den Generator mit einem externen Frequenznormal höherer Konstanz auszusteuern. Mit dem Frequenznormal XSD ergibt sich z.B. bei einer Einstellung von 1 000 MHz eine absolute Frequenzgenauigkeit von 0,2 Hz.

Das Gerät wiegt 70 kg und hat Abmessungen von 535 × 445 × 615 mm.

EE 71 771 *für weitere Einzelheiten*

RMS Regelungs- und Messtechnik
Dipl.-Ing. Schmidt KG

2057 Reinbek bei Hamburg, Gutenbergstrasse, Deutschland

SCHWINGUNGSANLAGEN

(Abbildung Seite 494)

Für die Durchführung von Schwingungsprüfungen haben deutsche und internationale Ausschüsse entsprechende

Normen ausgearbeitet, die in die Pflichtenhefte für zahlreiche Geräte und Bauelemente in wachsendem Masse aufgenommen werden. Die RMS-Schwingungsanlagen dienen zur Erfüllung solcher Bedingungen.

Die Abbildung zeigt eine Anlage SW 100, bestehend aus einem RC-Sinusgenerator, einem NF-Kraftverstärker und dem elektrodynamischen Schwinger. An der Oberseite des Zylinders ist ein Stutzen erkennbar, an dem der Prüfling befestigt wird.

Die Anlage SW 100 ist für ein Prüflingsgewicht von max. 1 kp ausgelegt und kann Schwingungen im Bereich von 5 bis 2 000 Hz bei einer Amplitude von max. ±3 mm erzeugen. Für die Ausführung SW 1 000 betragen die entsprechenden Werte 5 kp und 5 bis 800 Hz.

EE 71 772 *für weitere Einzelheiten*

Sadowski & Co.

7301 Esslingen-Berkheim, Deutschland

DOPPELIMPULS-GENERATOR

Der ausschliesslich mit Silizium-Transistoren bestückte Doppelimpuls-Generator DIG 121 wurde besonders für universelle Verwendung entwickelt und weist bei kleinen Abmessungen und niedrigem Gewicht eine aussergewöhnlich hohe Betriebssicherheit auf.

Das Gerät besteht aus den folgenden Einheiten: Abstandsgenerator, Impulsgenerator 1, Verzögerungsstufe und Impulsgenerator 2. Der Abstandsgenerator erzeugt Triggerimpulse mit einem regelbaren Abstand zwischen 1 μs und 1 s. Er ist mit einer externen Sinus- oder Impulsspannung synchronisierbar. Der Ausgang des Abstandsgenerators steuert den Impulsgenerator 1, der Rechteckimpulse mit einer regelbaren Breite von 0,1 μs bis 0,1 s erzeugt. Der gleichartig aufgebaute Impulsgenerator 2 wird über eine Verzögerungsschaltung, deren Verzögerungszeit zwischen 0,1 μs und 0,1 s einstellbar ist, angesteuert. Die Anstiegszeit der Impulse ist 0,02 μs, die Abfallzeit 0,03 μs.

Die Amplitude der Ausgangsimpulse ist zwischen 0 ... 10 V kontinuierlich regelbar, die Polarität umschaltbar. Die Ausgänge der beiden Impulsgeneratoren können entweder getrennten Buchsen entnommen oder mittels eines Schalters zu einem Doppelimpuls addiert werden. Statt des eingebauten Abstandsgenerators kann man eine externe Spannungsquelle (Sinus- oder Impulsspannung) zur Triggerung des Impulsgenerators verwenden. Durch Fremdtriggerung von aussen ist der DIG 121 auch als Verzögerungsgenerator zu verwenden.

EE 71 773 *für weitere Einzelheiten*

Sodeco

Société des Compteurs de Genève

Grand-Pré 70, Genf, Schweiz

TRANSISTORISIERTES SUMMIERGERÄT

(*Abbildung Seite 495*)

Das als Einschub ausgelegte transistorisierte Summiergerät Sodeco SC58 ist überall dort die richtige Lösung, wo aus verschiedenen Quellen stammende elektrische Impulse ohne Verluste auf einem gemeinsamen Zählwerk oder Druckwerk registriert werden sollen.

Mit dem in Druckschaltungstechnik gebauten Summiergerät werden die Impulse, die zeitlich unabhängig von verschiedenen Quellen eintreffen können, automatisch und fehlerfrei gespeichert. Das Speicherwerk wird sofort mit einer entsprechend höheren Frequenz abgefragt. Am Ausgang des Summiergerätes kann man unmittelbar den Impulsschnellzähler SC 60 (bis zu 250 Imp/Sek) oder über den Impulsumformer 3Tul sämtliche Sodeco-Impulszählertypen, einschliesslich der druckenden, anschliessen.

Bis zu 24 Eingangskanäle lassen sich anschliessen, und die Eingangssignale können von Lichtschranken oder elektromechanischen Kontakten abgegeben werden und elektrische Impulse verschiedener Form und Dauer sein. Das Gerät wird mit Netzanschluss für 110 V, 125 V, 220 V oder 250 V geliefert.

EE 71 774 *für weitere Einzelheiten*

DRUCKENDE IMPULSZÄHLER

Das Grundelement besteht aus drei Teilen, und zwar aus einem Zählmagneten, der gleichzeitig die 6stellige Anzeige und das 6stellige Druckwerk treibt, einer elektrischen Druckmechanik und einer elektrischen Nullstellung. Zwei dieser Grundelemente können nebeneinander und auf denselben Papierstreifen druckend angeordnet werden.

Drei verschiedene Zählwerke können eingebaut werden:

1. Zählwerk mit Zehnerübertrag,
2. Zeitzählwerk mit Anzeige in Stunden-Minuten-Sekunden-Zehntelsekunden oder Stunden-Minuten-Sekunden und
3. Datumzählwerk ohne elektrische Nullstellung durch zeitabhängige Impulse betrieben.

Die Impulszähler sind in Ausführungen für Gleichstrom (Netz oder Batterien) oder Wechselstrom lieferbar und haben eine maximale Zählgeschwindigkeit von 10 Imp/Sek.

EE 71 775 *für weitere Einzelheiten*

Telefunken AG

Fachbereich Anlagen Hochfrequenz

79 Ulm, Elisabethenstr. 3, Deutschland

DIGITALRECHNER

Der erstmals vorgestellte Digitalrechner "TR 10" von Telefunken ist nicht nur eine konsequente Ergänzung des schnellen und leistungsfähigen Grossrechners "TR 4", sondern stellt auch für sich allein betrachtet einen vielseitig einsetzbaren Digitalrechner dar. Das Spektrum seiner Einsatzmöglichkeiten reicht von den technisch-wissenschaftlichen Aufgaben bis zur Lösung von kommerziellen Daterverarbeitungsproblemen und zu den immer mehr in den Vordergrund tretenden Anwendungen zur Steuerung industrieller Produktionsprozesse.

Der "TR 10" ist ein speicherprogrammierter, dezimaler Digitalrechner mit geringem Raumbedarf, der sich aber dennoch durch eine gute Leistung auszeichnet und insbesondere das simultane Arbeiten zwischen dem Rechner selbst und den Peripheriegeräten gestattet. Es gibt Einrichtungen zum Verarbeiten von Lochstreifen oder Lochkarten; grössere Datenmengen werden in kurzer Zeit über einen Schnelldrucker ausgegeben. Die internen Kernspeicher können durch externe Magnetband-Digitalspeicher und durch Plattenspeicher ergänzt werden. Über den Fernschreibverteiler wird der Rechner mit Fernschreibeinrichtungen verbunden, und ein Verkehrsverteiler bildet den baukastenartig kombinierbaren Übergang zu Messstellen und Eingriffspunkten in industriellen Produktionsprozessen. Der Rechner kann auch von aussen in seinem Programm unterbrochen werden, um auf wichtige Signale zu reagieren.

Der "TR 10" ist ein Dezimalrechner; für die von ihm verarbeiteten Daten wird somit die auch im täglichen Leben übliche Schreibweise benutzt. Die Dezimalziffern, 26 Buchstaben und 28 Zeichen, werden stellenweise aus dem Arbeitsspeicher geholt und als Wörter von beliebiger Länge verarbeitet. Die Kapazität des Arbeitsspeichers kann je nach Bedarf in Stufen von je 10 000 Stellen zwischen 10 000 und 80 000 Stellen gewählt werden. Die Zykluszeit des Ferritkernspeichers beträgt 8 μs.

Im "TR 10" und den Peripheriegeräten werden vorwiegend Silizium-Planartransistoren verwendet. Der Zentralrechner ist in einem modernen Schrankgehäuse mit den Abmessungen 1800 × 600 × 400 mm untergebracht und wiegt ungefähr 200 kg.

Neben den technischen Eigenschaften der Geräte bietet der "TR 10" auch bei der Programmierung gute Anpassungsmöglichkeiten an die verschiedenartigen Anwendungen. Es können Adressenmodifizierungen beliebig oft und voneinander unabhängig ausgeführt werden. Die Befehle für den Umgang des Rechners mit den einzelnen Peripheriegeräten können von Gerät zu Gerät verschieden festgelegt und den speziellen Verhältnissen angepasst werden. Die Befehle werden vom Rechner an die Peripheriegeräte übermittelt, die dann selbständig und simultan zum Rechner die Durchführung übernehmen.

EE 71 776 *für weitere Einzelheiten*

Telefunken AG

Fachbereich Röhren, Vertrieb

79 Ulm, Söflingerstr. 100, Deutschland

FESTKÖRPERSCHALTUNGEN

(*Abbildung Seite 495*)

Die bekannten Festkörperschaltungen bestehen im allgemeinen aus Silizium, in das die aktiven und passiven Bauelemente nach den Verfahren der Planartechnik eindiffundiert sind. Die Isolation der jeweils in einem einzigen Stück Silizium befindlichen Bauelemente voneinander erfolgt durch eindiffundierte Sperrfronten. Auf die mit einer Siliziumoxydschicht versehene Siliziumoberfläche werden zur Verbindung der Bauelemente untereinander Leitbahnen aus Metall aufgedampft.

Im Diffusionsverfahren hergestellte Widerstände haben einen Temperaturkoeffizienten von etwa $3 \times 10^{-3}/°C$, d.h. bei 100°C Temperaturschwankung ändert sich der Widerstand um ca. 30%. Bei einer angelegten Sperrspannung von 1 V beträgt die Kapazität eines solchen Widerstandes gegenüber dem Grundmaterial etwa 150 pF/mm², woraus sich z.B. für einen typischen 10-kΩ-Widerstand eine auf den ganzen Widerstand verteilte Kapazität von ca. 5 pF ergibt.

Die nach dem neueren Herstellungsverfahren der Hybridtechnik hergestellten Festkörperschaltungen vermeiden diese Nachteile. Aktive Elemente werden weiterhin in Planartechnik hergestellt, passive Elemente im Hochvakuum aufgedampft. Mit $2 \times 10^{-4}/°C$ ist der Temperaturkoeffizient der Widerstände wesentlich niedriger, d.h. 100°C Temperaturänderung bewirken nur ca. 2% Widerstandsänderung. Die Kapazität gegenüber dem Grundmaterial ist nur 1/10 der Kapazität diffundierter Widerstände.

Abgebildet ist ein aus 4 NOR-Gates bestehendes halbes Schieberegister mit einer Seitenlänge von 9 mm im Vergleich zu einem handelsüblichen Reissnagel.

EE 71 777 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Magnetische Elemente mit Vielfachöffnungen von D. J. Morris

Zusammenfassung des Beitrages auf Seite 440-447

Ringmagnetkerne finden bei der Anwendung der Binärlogik weitgehend Verwendung. Die Anordnung zusätzlicher Öffnungen in den verschiedensten geometrischen Anordnungen ergibt eine Auswahl von Elementen, die viele bedeutsame Vorteile haben. Diese Elemente werden als Transfluxor oder Elemente mit Vielfachöffnungen (M.A.D. = Multi-Aperture Devices) bezeichnet. Ihr hervorstechendstes Konstruktionsmerkmal ist die Isolierung der Eingangsschaltung von der Ausgangsschaltung, die zerstörungsfreies Lesen des Speicherinhalts ermöglicht und nur aus Draht bestehende Verbindungen zwischen Elementen erlaubt.

Der Beitrag beschreibt die Eigenschaften der Grundelemente mit Vielfachöffnungen und zeigt dann, auf welche Art ihre Arbeitsweise durch Formänderung modifiziert werden kann. Danach werden die verschiedenen geometrischen Anordnungen für verschiedene Anwendungsmöglichkeiten dieser Elemente besprochen.

Systematische Entwicklung der Cohn-Struktur für RF-Bandfilter von K. E. Brown

Zusammenfassung des Beitrages auf Seite 448-451

Die Abwandlung einer bestehenden Schmalband-Transformationstechnik, die eine Bandfilterschaltung mit punktförmig verteilten Komponenten ergibt, wird beschrieben. Auf dieser Basis entworfene Filter haben Frequenzgänge mit Mittenfrequenzen bis zu 200 MHz und relative Bandbreiten zwischen 1% und 20%. Die Induktivitätswerte können durchgehend vorgeschrieben werden; für den Fall gleicher Induktivitäten werden Konstruktionsformeln gegeben. Die Induktivitäten treten nur in den Nebenschlusszweigen des Leiternetzwerkes auf, und jede Induktivität hat eine Überbrückungskapazität; ausserdem sind vorgesehene Eingangs- und Ausgangskapazitäten vorhanden, so dass für jede Leiterform ein kapazitiver Weg zur Erde besteht. Dieses Verfahren kann überall dort, wo die Abschlusswiderstände nicht gleich sind (einschliesslich Leerlauf-Belastungszuständen), ohne Schwierigkeiten angewendet werden. Für diese Transformation ist die Standardtechnik der Vorverzerrung zum Ausgleich der Verluste in den Induktivitäten zulässig.

Galvanisch gekoppelte Präzisions-Transistorverstärker und ihr näherungsweiser Entwurf von C. W. B. Grigson

Zusammenfassung des Beitrages auf Seite 454-459

Eingehende mathematische Analysen haben ohne Kenntnis der Parameter der einzelnen Transistoren für die praktische Entwicklungsarbeit wenig Bedeutung. In diesem Beitrag werden Entwurfsmethoden besprochen, die die Entwicklung von Präzisionsschaltungen mit nur näherungsweiser Kenntnis der Transistor-Parameter B, U_{be} *und* $U0_{CE_K}$ *ermöglichen.*

Folgende Schaltungen werden behandelt; (1) der Ring-Spannungsverstärker; (2) seine Anwendung als Spannungsverstärker mit einer Drift von 20 µV/°C; (3) ein verstärkungsstabilisierter Verstärker mit einer Endwertempfindlichkeit von 3×10^{-11} *A; (4) eine stromstabilisierte Quelle für 0 . . . 20 A mit 300 Ω Ausgangswiderstand; (5) ein B-Stromverstärker mit einem Leistungsvermögen von 20* A_{ss}*.*

Ein langsames NOR-Logiksystem mit einer Stromversorgung von G. Flanagan und L. Molyneux

Zusammenfassung des Beitrages auf Seite 460-463

Das beschriebene NOR-Logiksystem benötigt nur einen Widerstandswert, den billigsten Transistor seiner Serie und nur eine Stromversorgung. Es arbeitet langsam—Höchstimpulsfrequenz 5000 Hz—ist jedoch für Bedingungen geeignet, in denen elektromechanische Bauelemente die Arbeitsgeschwindigkeit bestimmen, wie z.B. in der Datenverarbeitung.

Die Absonderung der Helle-Information vom SECAM-Videosignal von G. Melchior und J. P. Doury

Zusammenfassung des Beitrages auf Seite 464-465

Für bestimmte Zwecke ist die Absonderung des Helligkeitssignals vom SECAM-Videosignal notwendig. In diesem Beitrag wird ein Verfahren beschrieben, mit dessen Hilfe diese Aufgabe ohne ernsthafte Beeinträchtigung der Bildschärfe erzielt werden kann.

Ein transistorisierter 85-MHz-Sender von A. E. Hilling

Zusammenfassung des Beitrages auf Seite 466-469

Ein mit diffusionslegierten Germanium-Transistoren AFY19 bestückter 85-MHz-Sender wird beschrieben. Dieser Sender erzeugt bei 100%iger Modulation eine Spitzenleistung von 1 W.

Die Schaltung wäre für tragbare Kleinsender im Frequenzband 71,5 . . . 88 MHz geeignet.

Messen von Ferritkern-Parametern bei sehr hohen Schaltströmen von M. C. Stevens

Zusammenfassung des Beitrages auf Seite 470-473

Einrichtungen für die Erzeugung von Stromimpulsen mit einer Amplitude von bis zu 10 A und einer Anstiegszeit von einem Drittel einer Nanosekunde werden beschrieben. Das Gerät wurde zum Messen der Schaltparameter von Ferritkernen mit rechteckiger Hystereseschleife für die drei Gebiete eingesetzt, in denen die Umkehr durch Wandverschiebung, ungleichförmige, bzw. gleichförmige Drehung erfolgt. Die an acht Kerntypen gemessenen Parameter werden aufgeführt.

Konstruktionsverbesserungen in Präzisions-Drahtwiderständen von J. R. Price

Zusammenfassung des Beitrages auf Seite 474-477

Einer der bedeutendsten Parameter der Präzisions-Drahtwiderstände ist ihre Langzeitkonstanz. Eine neue Bauart wurde entwickelt, durch die Einflüsse von Wickelspannung und Wickelkörperwerkstoff beseitigt werden, und zwar ist in diesen Widerständen die Wicklung vollkommen lose und im Wickelkörper frei beweglich. Messresultate für die Konstanz typischer Widerstände werden gegeben.

Ein einfaches Nullfilter mit regelbarer Sperrfrequenz von J. L. Douce und K. H. Edwards

Zusammenfassung des Beitrages auf Seite 478-479

Der Beitrag beschreibt, wie eine Abwandlung der Doppel-T-Schaltung die Einstellung der Sperrfrequenz mittels eines Einzelpotentiometers ermöglicht. Eine Analyse des Filters wird im Anhang gegeben.

Ein verbessertes Regelsystem für Wechselspannungs-Konstanthalter von R. G. Ackland und N. M. Buckland

Zusammenfassung des Beitrages auf Seite 480-482

Dieser Beitrag beschreibt ein einfaches Wechselspannungsfühl- und Stellmotortreibsystem, das bei Einsatz als Treiber des Variacs eines motorgetriebenen Wechselstrom-Netzspannungs-Konstanthalters die Ausgangsspannung innerhalb ±0, 1% konstant halten kann. Die Wiederkehrrate für sprunghafte Änderungen ist 18 V/s.

Breitband-Stromwandler für Stromwellenformen von F. J. U. Ritson und J. Wood

Zusammenfassung des Beitrages auf Seite 483-485

In diesem Beitrag wird ein Stromwandler und zugehöriger Verstärker für die Untersuchung von Stromwellenformen beschrieben. Das Gerät hat eine sehr grosse Bandbreite und führt nur eine vernachlässigbar kleine Impedanz in die beobachtete Schaltung ein.