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COMMENTARY

THE collection of statistics and their interpretation is now universal; the dangerous game of trend-spotting on charts whose horizontal axis is time is played by many who ought to know better. Nevertheless, statistics used to examine aspects of an industry, a business or the national economy have a very real value if they are handled with common sense. This is particularly so in our rapidly changing electronics industry. In 1964, therefore, the Little Neddy (EDC) for the electronics industry set up a working group to consider available statistics and their use, and they have now published their report*.

It ought to be emphasized at the outset that the EDC's idea of statistics is as an apparatus for decision-making and not a rather esoteric form of puzzle for experts. For statistics to be of practical value, there must be more knowledge of production figures, sales, exports etc., so that marketing at the level of the individual company can be properly based instead of the empirical approaches which so often have to be made.

The first task in preparing industrial statistics is to define the industry. The Committee has proposed, as a base, the use of the Standard Industrial Classification (s.i.c.) used by Government departments; this groups together products or authorities having features in common. Establishments, i.e. buildings in one location where these products are made, are allocated to set groups. The groups are known as minimum list headings, or m.l.h.s, and they may be subdivided. The s.i.c. definition of the electronics industry is thus contained within m.l.h. 364, subdivided into four: valves; broadcast-receiving equipment and gramophones; radio, radar and electronic capital goods; and radio and electronic components. It is convenient to add m.l.h. 363 to this, to take in telegraph and telephone apparatus. The EDC has proposed a revision of these groups into five separate m.l.h.s. One is to be for components, subdivided into active, positive and microcircuits, one m.l.h. each for capital goods, telephone equipment and consumer goods, and a fifth for companies manufacturing audio products and components. In addition to this sort of categorization, which defines the industry fairly well, it may be that a product list in which each type of product is shown may be of great value in forming the basis of systematic collation of statistical data and subsequent classification into m.l.h. groups. The publication of such a product list would, the EDC maintains, give a better understanding and more consistency to the provision of figures by those whose job this is.

* Electronics Economic Development Committee: Statistics of the Electronics Industry. 1967, June. National Economic Development Council. Prepared by John Howard of Mullard Ltd.

Having defined the electronics industry to its approximate satisfaction, the Committee has next considered the needs of users of statistics. The industry itself appears largely to be interested in output and sales of particular types of products in terms of numbers and current values. The Government's requirements are more general and embrace the whole electronics industry, its manpower, exports, research etc. The Electronics EDC looks at growth and performance of the industry and their relationship with labour requirements, exports etc., and is in some respect viewing from the standpoint of both the industry and the Government.

A consideration of the collation of statistics in foreign countries leads to the perhaps foreseeable result that they have a more systematic approach. The EDC, for example, considers that the American system of linking the census of production with regularly published statistics is sound and, furthermore, suggests that the census sales classification should be the same as that for the regular figures.

The larger part of the report is naturally devoted to the production and use of output and sales figures, and they are discussed conveniently under the five proposed m.l.h. categories and their subdivisions described earlier. Similar critiques are included for exports and imports, labour, capital spending, research and development contracts and prices. A general impression derived from these parts of the report is that those statistics which are published are in a fairly chaotic state and anyway mainly inferential. Those statistics which are *not* published create great gaps which ought to be filled. This situation does not, of course, arise from incompetence on the part of the collectors of data but rather from the fact that human organizations do not evolve expressly for the statistician to enumerate them—they overlap (and get counted twice) or are modest (and do not get counted at all). But once suitable and accurate figures are obtained that is not the end. The Committee sees that the lesson to be learnt from its report is of the continuous change taking place in the electronics industry, necessitating a concomitant change in the methods of, and approaches to, the collection of statistics.

It must finally be said that the EDC report is a powerful and constructive critique of electronics statistics, using common sense and laced with humour. We cannot resist quoting a paragraph where a plea is made for publication of figures for sale of equipment for defence purposes. It has been suggested that such a publication might help foreign powers; in riposte the report says '....a spy who cannot get close to the right figures after several hours work on the published information should take up pig-breeding!'

Amplifiers Combining Bipolar and Field-Effect Transistors

by W. Gosling*

Combinations of bipolar and field-effect transistors can yield amplifiers having high gain and low output impedance. Two families of circuits can be distinguished: those in which the f.e.t. is operated pinched off and those in which it is not. The former can have superior electrical characteristics but are usually more complex than the latter, and the choice between them is not entirely straightforward. Some circuits exhibit inherent compensation of temperature-dependant effects.

(Voir page 533 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 534)

FIELD-EFFECT transistors have extremely high current gain, since virtually no current flows in the input circuit in the usual common-source amplifier configuration; the voltage gain obtainable, however, is rather modest. By contrast, bipolar transistors have a lower current gain but can have high voltage gain. A combination of the two might thus be expected to give high power gain and very substantial amplification of both voltage and current. The field-effect device has the desirable property of high input impedance, while the bipolar transistor can have fairly low output impedance and also bottoms to a lower voltage than that typical of an f.e.t., and so has a higher possible power efficiency as an output stage. It is thus reasonable to direct attention to 2-stage amplifiers using an f.e.t. in the first stage and a bipolar device in the second.

To obtain a high input impedance, the f.e.t. stage should be in the c.s. (common source) or c.d. (common drain) configuration, while the second (bipolar) stage may be c.e., c.c. or c.b., although the last will give a rather high output impedance in many cases. Thus there are six circuit configurations of intrinsic interest: c.s.-c.e., c.s.-c.c., c.s.-c.b., c.d.-c.e., c.d.-c.c., c.d.-c.b. The properties of the amplifiers are very sharply dependant on the configuration of the second stage, and it will therefore be convenient to discuss them in three groups divided on this basis.

Amplifiers with c.e. output stages: introduction

The c.e. amplifier has high current gain and reasonably large voltage gain; its output impedance is intermediate between those of c.c. and c.b. amplifiers. These characterestics largely determine the properties of 2-stage amplifiers using c.e. output stages.

Considering first the c.d.-c.e. amplifier, a typical circuit is that of Fig. 1(a). With this circuit configuration, shunt voltage-derived negative feedback is easily applied, as is series current-derived feedback (by inclusion of a resistor in the bipolar-transistor emitter lead), but shunt current feedback and series voltage feedback cannot easily be arranged. Provided that the rise in output impedance resulting from the use of current feedback can be tolerated, the introduction of a resistor in series with the bipolar-transistor emitter will give improved gain stabilization and tend to reduce the apparent input capacitance of the amplifier.

In calculating the gain (without feedback) of the c.d.-c.e. amplifier, it is most useful to begin by calculating the overall transfer conductance $g_{f,1}$. The voltage gain of the c.d. stage is

$$a_1 = \frac{-g_{fs}r_{in}}{1 - g_{fs}r_{in}}$$

provided that $r_{in} \ll r_{DS}$, where r_{in} is the input resistance of the c.e. amplifier, g_{fs} is the transfer conductance of the f.e.t., and r_{DS} is the incremental drain-source impedance of the f.e.t. Hence it follows that

$$g_{f,1} = \frac{h_{fe}g_{fs}}{1 - g_{fs}r_{in}} \quad \ldots\ldots\ldots\ldots\ldots (1)$$

The value of r_{in} may, of course, be obtained quite readily from the hybrid parameters of the bipolar device:

$$r_{in} = h_{ie} - \frac{h_{fe}h_{re}}{(1 + R_L h_{oe})} R_L \quad \ldots\ldots\ldots (2)$$

The exact significance of equation (1) depends on the relative magnitudes of g_{fs} and r_{in}. Two limiting cases are of particular interest. For f.e.t.s of low g_{fs}, particularly when operating into bipolar transistors having large load resistors, the product $g_{fs}r_{in}$ may be small compared with unity. Here

$$g_{f,1} \to h_{fe}g_{fs}$$

This is the value that would be obtained with the c.s.-c.e. configuration, as will be shown subsequently; and in fact the properties of the c.d.-c.e. configuration are, in this case, closely similar in respect of gain to those of the variant with the f.e.t. connected in common source.

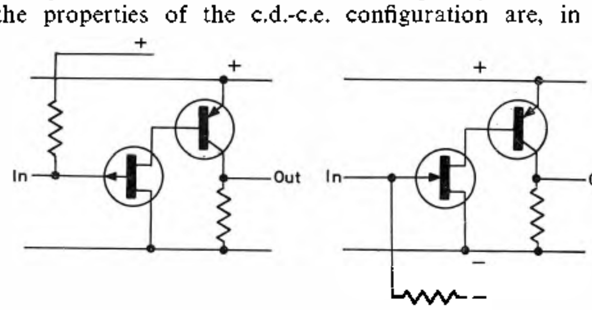

Fig. 1 (a) The c.d.-c.e. amplifier
(b) The c.s.-c.e. amplifier

Modern f.e.t.s, however, have g_{fs} large enough to make $g_{fs}r_{in}$ of the order of, or even larger than, unity. In this case, from equation (1),

$$g_{f,1} \to h_{fe}/r_{in}$$

This represents a limiting value of transfer conductance, beyond which it is impossible to go by increasing g_{fs}. The maximum transfer conductance for the amplifier as a whole ranges from a few tens of millimhos up to a few mhos, depending on the transistors chosen. This seems to suggest that high voltage gains should be attained, particularly as the output impedance of the amplifier is high. Since the output impedance of the c.d. stage is approximately $-1/g_{fs}$, the output impedance of the c.e. bipolar device is

$$r_{out} = \frac{h_{ie} - 1/g_{fs}}{\Delta h - h_{oe}/g_{fs}} \quad \ldots\ldots\ldots\ldots (3)$$

which tends to $1/h_{oe}$ for small g_{fs}.

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However, where the c.d.-c.e. amplifier works into a resistive load, high voltage gains are not easily attained. Although the forward transfer conductance has been greatly increased relative to g_{fs} by the addition of the c.e. stage, the standing direct current has also been increased from I_D to I_C, which, with the simple direct-coupled circuits described, will be of the order of $h_{FE}I_D$. For many amplifiers with a resistive load, the factor determining gain is the ratio of forward transfer conductance to standing direct current, assuming high output impedance; and in the present case this has only improved in a ratio less than h_{fe}/h_{FE}.

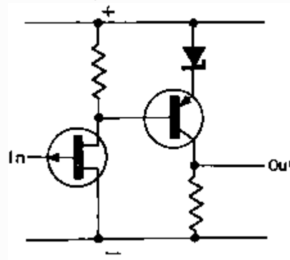


Fig. 2 Zener-diode coupling network

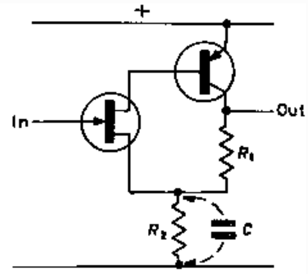


Fig. 3 Series voltage-derived feedback

Simple circuit modifications can effect a substantial improvement. For example, the Zener-diode coupling network of Fig. 2 will reduce I_C with only a minimal reduction of $g_{f,i}$. However, the drawback with this circuit is that variations in drain current of the f.e.t. due to temperature changes or other causes appear undiminished at the base of a bipolar device, and thus correspond to a much larger fractional change of collector current. Where d.c. response is not required, some kind of overall d.c. stabilizing feedback circuit provides a solution to the problem, which is all the more effective because of the relatively large forward-path gain of the amplifier.

The c.s.-c.e. amplifier in Fig. 1(b) presents very similar features, except that the forward transfer conductance is

$$g_{f,2} = g_{fs}h_{fe} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

and does not reach a limiting value for large g_{fs}, as in the c.d.-c.e. case. Thus, for transistors of large g_{fs}, the c.s. configuration for the first stage will often result in a higher transfer conductance than would be obtained with a c.d. stage. The output impedance must now be calculated in terms of a source impedance to the bipolar transistor of r_{DS} (instead of $1/g_{fs}$ as in the c.d. case) and is, therefore, very close to $1/h_{oe}$, since r_{DS} is usually large.

The same problem of high collector current arises as for the c.d.-c.e. form, and it is treated in the same way. The amplifier is non-phase-inverting, so that negative feedback is best provided from the bipolar-device collector to the source terminal of the f.e.t. In this way, series voltage-derived feedback results (Fig 3), with the consequent high input impedance and low output impedance which are common design goals. Alternatively, if high a.c. gain is required, the feedback path can be bypassed for signal frequencies by a capacitor connected in shunt with the resistor R_2, and in this way d.c. stabilisation is retained.

Frequency response

As far as high-frequency limitations are concerned, three factors are important in amplifiers of this kind:

(a) capacitative loading of the amplifier output

(b) combined effect of amplifier input capacitance and nonzero signal source impedance

(c) fall in forward transfer conductance at high frequencies, due to the large equivalent input capacitance of the bipolar device.

The first of these is not peculiar to amplifiers of the present type: its effect can be calculated in an entirely conventional manner, and it need not be discussed further here. As far as the second is concerned, the c.d. and c.s. input stages differ in their input capacitances. In the c.d. amplifier, the gate-source capacitance is bootstrapped, and the input capacitance is thus

$$C_{in,1} = C_{GS} + C_{GD}\,\frac{1}{1 - g_{fs}r_{in}} \quad \ldots\ldots\ldots \quad (5)$$

By contrast, in the c.s. amplifier, the component of input capacitance derived from C_{GD} is increased by Miller effect, so that the input capacitance is

$$C_{in,2} = C_{GS} + C_{GD}\,(1 - g_{fs}r_{in}) \quad \ldots\ldots\ldots \quad (6)$$

Remembering that g_{fs} is negative, it will be seen that $C_{in,2}$ is invariably larger than $C_{in,1}$; so that the c.s. amplifier will be worse than the c.d. in respect of high-frequency response, unless the signal source impedance is negligibly small.

With regard to the high-frequency characteristics of the bipolar transistor, it is convenient for the purposes of discussion to represent this device by its hybrid-π equivalent circuit. Then the interstage coupling between the two transistors will look, in the c.d.-c.e. case, like the equivalent circuit shown in Fig. 4, assuming that the output impedance of the c.d. stage may be approximated as $1/g_{fs}$.

If, as is usual for high g_{fs} f.e.t.s,

$$1/g_{fs} + r_{BB'} \ll r_{B'E}$$

the cut-off frequency for the interstage coupling is approximately

$$\omega_2 = \frac{g_{fs}}{(1 + g_{fs}r_{BB'})C_{in}'} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (7)$$

where $C_{in}' = C_{B'E} + (1-a_2)C_{B'C}$

and a_2 is the voltage gain of the intrinsic bipolar transistor. For the c.s.-c.e. amplifier, the output impedance of the first stage is now not $1/g_{fs}$ but r_{DS}. Since r_{DS} is much larger than $r_{B'E}$ an approximation comparable with Equation (7) is not possible: but instead, provided that

$$r_{DS} \gg r_{B'E}$$

it follows that, in this case,

$$\omega_2' = \frac{1}{r_{B'E}C_{in}'} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (8)$$

Comparing equations (7) and (8)

$$\omega_2/\omega_2' = \frac{r_{B'E}g_{fs}}{1 + r_{BB'}g_{fs}} \quad \ldots\ldots\ldots\ldots \quad (9)$$

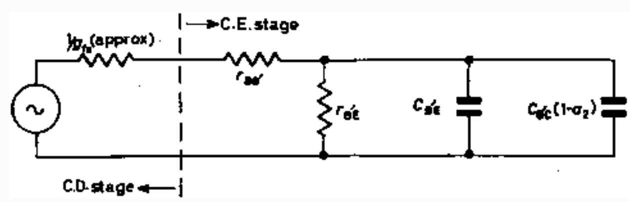


Fig. 4 Equivalent circuit of the interstage coupling between the two transistors of the c.d.-c.e. amplifier

This ratio is invariably greater than unity and tends to $r_{B'E}/r_{BB'}$ for large g_{fs}. Thus the c.d.-c.e. amplifier has an advantage in high-frequency response which becomes more marked for larger g_{fs}.

Effect of a base-emitter resistor

An interesting variant of the c.d.-c.e. and c.s.-c.e. amplifiers results when the voltage of the Zener diode in the coupling circuit of Fig. 2 is reduced to zero, so that the circuit becomes the simple direct-coupled configuration of Fig. 1, but with a resistor connected between base and emitter of the bipolar device. Since some of the f.e.t.

drain current flows through this resistor, the collector current will be reduced.

In the c.d.-c.e. circuit, the output impedance of the f.e.t. is so low that substantial diversion of current from the base of the bipolar device is possible without much loss of signal. If the base-emitter voltage of the bipolar device is V_{BE} and the resistance is R,

$$R = \frac{V_{BE}}{I_D - I_B} \ll -1/g_{fs} \quad \ldots\ldots\ldots\ (10)$$

for negligible signal loss

but

$$g_{fs} = -\frac{2I_{DSS}}{V_p}\{1 - V_{GS}/V_p\}$$

for a junction f.e.t.

Hence $\quad g_{fs} = -2/V_p \sqrt{(I_D I_{DSS})}$

Thus the condition for negligible signal loss is

$$\frac{I_D V_{BE}^2}{(I_D - I_B)^2} \ll V_p^2/I_{DSS}$$

or $\quad I_D/I_{DSS} \gg (V_{BE}/V_p)^2 \quad \ldots\ldots\ldots\ldots\ (11)$

if $I_B \ll I_D$.

This condition sets a lower limit on I_D, relative to I_{DSS}, which is, however, easily met, since V_{BE} (at about $\frac{1}{2}$V for a silicon device) is usually at least half an order of magnitude less than V_p. Thus the inclusion of a resistor between base and emitter of the bipolar device in the c.d.-c.e. amplifier reduces the collector current, while producing no significant reduction in overall power gain.

With the c.s.-c.e. amplifier, the effect of the resistor is quite different. Since the output impedance r_{DS} of the c.s. amplifier is very high, a substantial reduction in gain as well as collector current will result. If the base-emitter resistor is R_B, the forward transfer conductance is reduced to g_f', where

$$g_{f,2}' = \frac{R_B g_{fs} h_{fe}}{r_{in} + R_B} \quad \ldots\ldots\ldots\ldots\ldots\ (12)$$

But, if $I_B \ll I_D$, to reasonable approximation,

$$R_B = V_{BE}/I_D$$

Hence

$$g_{f,2}' = \frac{V_{BE} g_{fs} h_{fe}}{V_{BE} + r_{in} I_D} \quad \ldots\ldots\ldots\ldots\ (13)$$

This result shows that, provided that $r_{in}I_D$ can be kept no larger than V_{BE}, the loss in gain will be only 6dB. Since V_{BE} is virtually constant, this implies an upper limit on I_D for a given r_{in}. However, r_{in} decreases with increasing I_C, owing to the working point dependence of the transistor parameters. Thus a limit is implied on the extent to which I_C can be reduced for given I_D. Once I_C has been chosen, r_{in} can easily be calculated, and then a value of I_D chosen so that $I_D r_{in}$ is not too large compared with V_{BE}, thus minimizing gain loss.

It will be noted that, when the resistor R_B is included, the signal source impedance at the input of the c.e. stage falls from r_{DS} to a little less than R_B. This fall in impedance results in an improvement in the frequency at which $g_{f,2}'$ falls owing to input-capacitance effects.

In fact the introduction of the base-emitter resistor tends to reduce the differences between the characteristics of the c.s.-c.e. and c.d.-c.e. amplifiers. Without the resistor, the c.s. circuit shows higher gain but poorer frequency response than the c.d. circuit. With R_B present, the gain of the c.s.-c.e. amplifier falls but its frequency response improves (other circuit parameters being assumed to remain unchanged).

Temperature effects

The effects of temperature on unipolar- and bipolar-transistor parameters have been discussed elsewhere[1]. When transistors are cascaded, the individual temperature-dependant effects combine in a perfectly straightforward way. Only two special points need to be mentioned.

Circuits in which the standing current in the second stage is reduced without significant reduction in small-signal gain present, as has already been indicated, considerable stabilization problems, as the changes of f.e.t. drain current cause very much larger fractional variations in collector current. The effect can be minimized by using overall d.c. negative feedback or, particularly where amplifier response must extend down to zero frequency, by using a compensation technique in the first stage, such as so-called zero-drift biasing[1].

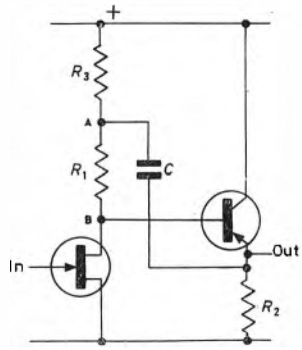


Fig. 5 Bootstrapping the load resistor of the c.s.-c.e. amplifier

The second point of interest is that thermal compensation for the overall amplifier can sometimes be achieved by balancing the thermal characteristics of the first stage against those of the second. For example, in the c.s.-c.e. amplifier, if the f.e.t. is operated at or near zero gate-source voltage, the main temperature-dependant effect is a fall in the drain current with increasing temperature, owing to a decline in charge-carrier mobility. The effect of gate-channel contact potential will be negligible provided that V_p is of the order of at least a few volts; effects due to gate-current variation may also be disregarded provided that the resistance in the gate circuit is not excessive. If a negative coefficient of drain current is assumed, the base-emitter voltage of the bipolar device will fall with rising temperature, thus tending to offset the fall in internal contact potential of the base-emitter junction and reducing the dependence of collector current on temperature[2].

Amplifiers with c.b. output stages

When the bipolar device is operated in the common-base configuration its current gain is less than unity. Thus the forward transfer conductance of a c.d.-c.b. or c.s.-c.b. amplifier is very substantially less than that of comparable amplifiers having the output transistor in the c.e. configuration.

The main function of the c.b. transistor is to give a degree of isolation between input and output circuits and an impedance transformation. Since the input impedance of the common-base amplifier is small, the forward transfer conductance of the two amplifiers is not very different, being in both cases less than that of the f.e.t. alone by the factor h_{fb}, which is only slightly less than unity. The output impedance of both amplifiers is high, approximating closely, for the c.s.-c.b. (and also, provided that g_{fs} for the f.e.t. is not too large, for the c.d.-c.b.) to $1/h_{ob}$, which is tyically of the order of megohms for a small transistor.

An amplifier of this type, in which the forward transfer conductance is rather modest but the output impedance high, can only give high voltage gain when working into very large loads. It is thus not well suited to operation with resistive loads and seems best suited to use as a tuned narrowband amplifier. The main problem with amplifiers of this kind is to produce designs which are stable and not significantly regenerative, preferably without the use of neutralization. Early f.e.t.s had relatively low g_{fs} and were fairly easy to stabilize in either of the configurations considered, but more modern units with much higher transfer conductance may present substantial problems. The c.s. configuration presents particular difficulties. With available units the drain-gate feedback capacitance C_{GD} is of the order of picofarads, and thus regeneration and even oscillation is readily provoked by an inductive component in the drain load impedance. This will be present when the parallel resonant circuit in the collector circuit of the bipolar device is tuned slightly above the signal frequency.

By contrast, the c.d. amplifier can become unstable on a capacitative load, but the low output impedance is in shunt with this and consequently a marked capacitative load phase angle is essential before significant regeneration occurs. As a result it is easier to design stable c.d.-c.b. amplifiers using high-g_{fs} f.e.t.s than the c.s. variant.

The detailed design procedure for tuned c.d.-c.b. amplifiers is too extensive to present here. The approach is quite conventional but, to ensure stability, capacitative loading of the c.d. stage must be minimized.

Amplifiers with c.c. output stages

The two amplifiers using c.c. output stages have very different properties. Where the c.s. input stage is adopted, the voltage gain is subtsantially that of the c.s. stage alone, but the output impedance is much reduced. When the input transistor is in the common-drain configuration the output impedance is even lower, but the overall voltage gain of the 2-stage amplifier is less than unity.

The c.s.-c.c. amplifier

The c.s.-c.c. amplifier presents no very unexpected features. The c.c. stage acts merely as an impedance transformer, reducing the output impedance by a factor very near to $1/(1 + h_{fe})$ relative to that of the c.s. stage alone. Provided that the c.c.-stage load resistor is of the same order as the c.s.-stage drain resistor, the output impedance is well approximated by

$$r_{out} = \frac{R_1 r_{DS}}{(R_1 + r_{DE})(1 + h_{fe})} \quad \ldots\ldots\ldots (14)$$

where R_1 is the c.s.-stage drain resistor, while the voltage gain is very close to A_V, where

$$A_V = \frac{g_{fs}}{1/r_{DS} + 1/R_1 + \dfrac{1}{(1 + h_{fe})R_2}} \quad \ldots\ldots (15)$$

where R_2 is the c.c.-stage load resistor.

In many cases $R_1 \ll r_{DS}$ and $R_1 \ll (1 + h_{fe})R_2$, in which event,

$$r_{out} = R_1/1 + h_{fe} \quad \ldots\ldots\ldots\ldots (16)$$

and

$$A_V = g_{fs}R_1 \quad \ldots\ldots\ldots\ldots\ldots\ldots (17)$$

to good approximation.

Owing to the low output impedance, c.s.-c.c. amplifiers give a large gain-bandwidth product when cascaded. Otherwise, however, the large input capacitance of amplifiers of this kind (due to the substantial voltage gain, and hence Miller effect, of the c.s. stage, which results from its operation into the relatively large input impedance of the c.c. bipolar amplifier) militates against high-frequency operation from signal sources of high internal resistance.

Higher voltage gain can be achieved with the c.s.-c.c. amplifier by 'bootstrapping' the load resistor (Fig. 5). If the capacitor C is of the negligible impedance at signal frequencies, resistors R_3 and R_2 are effectively connected in parallel as the load resistors of the c.c. stage. If the voltage gain of the c.c. stage is A_2, the load seen by the c.s. stage is thus R_L, where (assuming r_{DS} large and $1/h_{ob} > r_{DS}$)

$$1/R_L = \frac{1}{(1 - A_2)R_1} + \frac{R_2 + R_3}{(h_{fe} + 1)R_2R_3} \quad \ldots. (18)$$

Usually R_1 is chosen large enough to make the first of these two terms negligible, and also R_2 can be made large compared with R_3, thus, to fair approximation,

$$R_L = (h_{fe} + 1)\,R_3$$

The improvement in gain due to bootstrapping (assuming g_{fs} unchanged) is thus

$$\frac{R_L}{R_1 + R_3} = (h_{fe} + 1)\frac{R_3}{R_1 + R_3} \quad \ldots\ldots\ldots\ldots (19)$$

Increase in voltage gain by a factor of 10 is easily achieved. A disadvantage of this circuit is that the output impedance of the amplifier rises substantially, up to a limit of $r_{DS}/(1 + h_{fe})$. Use will be limited to low-frequency applications where the relatively high input capacitance (due to Miller effect) is not too embarrassing.

The c.d.-c.c. amplifier

Since both the c.d. and c.c. amplifiers have a voltage gain less than unity, that of the c.d.-c.c. combination is also less than unity. It is used as an impedance transformer when the reduction in impedance afforded by the c.d. stage alone is insufficient.

Analysis of the c.d.-c.c. amplifier is entirely straightforward[1] and will not be treated here. The main design point of interest is the use of bootstrapping to increase the effective input impedance. A typical arrangement uses bootstrapping to the low-potential end of the gate-continuity resistor, to reduce input conductance, and to the drain, to reduce the component of input capacitance due to C_{GD}. Since C_{GS} is already bootstrapped, the input capacitance and conductance can be very small, typically down to a fraction of picofarad and a few hundred picomhos.

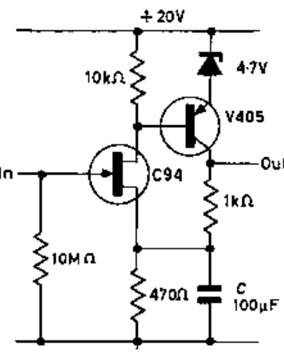


Fig. 6 The c.s.-c.c. amplifier circuit

2-Stage amplifiers in which the f.e.t. is not pinched off

In all the amplifiers previously considered, the field-effect transistor has been operated with a sufficient drain-source voltage to ensure that the conducting channel is pinched off. If this is not the case, the output impedance and, to a lesser extent, the forward transfer conductance of the f.e.t. (whether junction or insulated-gate type) will fall; however, if the transistor is operating into a relatively low-impedance load, such as the input of a c.b. or c.e. bipolar- transistor amplifier, the fall in output impedance does not produce any significant loss of power gain.

Operation with low drain-source voltage has certain marked advantages. Power dissipated in the f.e.t. is much reduced, and hence very high-current switching-type f.e.t.s, which have correspondingly high transfer conductances, can be used without overdissipation. If the drain-source voltage is reduced to the base-emitter voltage of the following stage, a particularly simple arrangement results. Finally, in junction-gate f.e.t.s, but possibily not in enhancement-mode insulated-gate devices, low-voltage operation seems to reduce noise level.

The detailed analysis of the operation of amplifiers of this kind has been described elsewhere[3]. Of the four possible configurations (c.d.-c.e. c.s.-c.b., c.d.-c.b. and c.s.-c.e.) the first two are of greatest interest, since they are phase inverting and thus lend themselves to d.c. stabilization by means of overall feedback.

The c.d.-c.e. amplifier has the important further property that the gate-channel and base-emitter contact potentials in the unipolar and bipolar devices, respectively, operate in opposite sense as far as the effect of their temperature dependence on the c.e. stage collector current is concerned. Although cancellation, at least in macrocircuits, cannot be close enough to be useful for any but the least critical d.c. amplifier applications, it certainly can simplify the design of a.c. amplifiers for satisfactory operation over a range of temperatures.

The choice between pinched-off and non-pinched-off operating conditions for the f.e.t. in a given amplifier design is not always straightforward. The small gain loss which results from non-pinched-off operation can usually be countered by changing the f.e.t. for a type of slightly larger channel cross-section, while the difference in frequency response is rarely significant. Where the c.d.-c.e. or c.s.-c.b. configurations are acceptable, in respect, for example, of the operating input and output impedances and types of feedback which can be applied, the circuit simplicity of the non-pinched-off version will commend it, particularly in integrated applications.

Experimental results

Some typical 2-stage amplifier circuits were designed and their experimental properties measured. The results obtained are not necessarily particularly representative, and certainly do not indicate the maximum performance which can be attained, but they may be regarded as general evidence for the correctness of the conclusions which have been drawn.

A c.s.-c.e. amplifier

A c.s.-c.e. amplifier was constructed as in Fig. 6. Performance characteristics were as follows:

(a) *Without the bypass capacitor C*
  Voltage gain (at 1kHz): 3 (+9·5dB)
  Input impedance (at 1MHz): 2·3pF in parallel with 10MΩ
  −3dB frequency: 14MHz (a peak of +6dB occurred at 8·4MHz)
  Signal-handling capacity (at output): 1·5V peak-to-peak
  Output impedance: 56Ω.

(b) *With bypass capacitor C*
  Voltage gain (at 1kHz): 39 (+32dB)
  Input impedance (at 1MHz): 4 pF in parallel with 10MΩ
  −3dB frequencies: 34Hz and 1·2MHz
  Signal-handling capacity (at output), 2V peak-to-peak
  Output impedance: 850Ω.

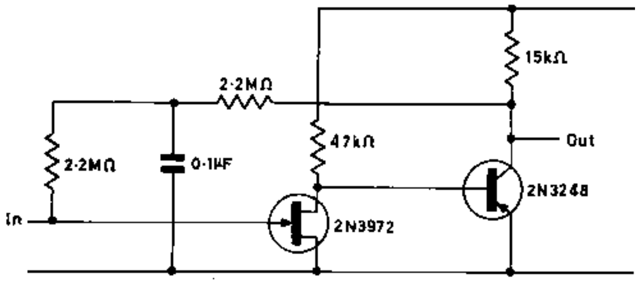


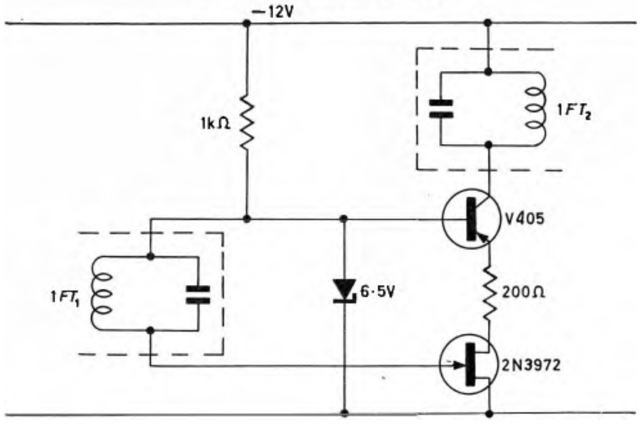
Fig. 8 A c.d.-c.e. amplifier in which the c.d. stage is operated with insufficient drain-source voltage to pinch off the conducting channel

It will be noted that a large gain peak occurred at 8·4MHz, when the amplifier was operated with maximum signal negative feedback, by omitting *C*. This is due, of course, to phase shift within the feedback loop, and was eliminated by connecting a 15pF capacitor between the collector of the c.e. stage and the source of the c.d. stage, in which case the upper −3dB frequency becomes 9·5MHz.

A c.d.-c.b. amplifier

A 465kHz i.f. amplifier is shown in Fig. 7. This was constructed using valve-type double-tuned i.f.t.s. A stage gain of 660 (+56dB) was obtained without any evidence of regeneration or instability, regardless of signal level. The stability was maintained despite ±30% changes in h.v. supply. A bandpass characteristic 20dB down at ±10kHz was observed.

Fig. 7 465kHz i.f. amplifier



Note that the 200Ω resistor between the source of the c.d. stage and the c.b. emitter stabilizes the circuit by modifying the phase angle of the input impedance of the second stage.

A c.d.-c.e. amplifier operating below pinchoff

The circuit of Fig. 8 shows a c.d.-c.e. amplifier in which the c.d. stage is operated with insufficient drain-source voltage to pinch off the conducting channel. The voltage gain was 232 (+47dB), falling by 3dB at 2·5Hz and 90kHz, when the amplifier was driven from a constant-voltage signal source. The input impedance was 2·2MΩ shunted by 30pF, and the output impedance was 15kΩ.

Conclusions

Amplifiers having great circuit simplicity, high gain and other useful properties can be designed using two stages, of which the first is a field-effect device and the second a bipolar transistor. Circuits of this type are particularly worthy of consideration at the present time, when bipolar transistors are available at substantially lower cost than f.e.t.s.

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Measurement of Parameters of Semiconductor Devices

by G. May*

The article outlines two methods for measuring the parameters of semiconductor devices: the scattering-parameter method and the transmission-parameter method. These avoid the difficulties resulting from the requirement of other methods for a perfect short- or open-circuit immittance test, which has been accentuated by the introduction of linear integrated circuits.

(Voir page 533 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 534)

THE methods in common use for measuring the parameters of semiconductor devices are mainly of the type where open- or short-circuit immittance tests are applied to the device undergoing characterization, leading to results in the form of z, y or h parameters.

The introduction of linear integrated circuits has accentuated such difficulties encountered when making this type of measurement as

(a) the need for perfect short-circuit or open-circuit measurement;
(b) instability and oscillation which may occur when carrying out such measurements: these effects can be difficult to detect;
(c) the large difference of values between the forward transfer admittance and the reverse transfer admittance.

Most of the difficulties are a function of the open- and short-circuit requirements. Ideally it would be more satisfactory to measure the device under typical working conditions where the test signals can be monitored. With the introduction of phase-sensitive voltmeters having a high input impedance, good signal sensitivity and a working range well beyond 100MHz, two further methods of parameter measurement are now available. One is to measure the scattering, S, parameters, the other is to measure the transmission, a or $ABCD$, parameters.

The Scattering-Parameter Method

This method has been fully described elsewhere[1], but it can be briefly summarized as a set of four voltage insertion ratio measurements which lead to a matrix set **P** where the values of the insertion ratio are in the form exp (ϕ), and ϕ is complex,

$$\mathbf{P} = \begin{bmatrix} e^{\phi 11} & e^{\phi 12} \\ e^{\phi 21} & e^{\phi 22} \end{bmatrix}$$

This can be converted to a scattering matrix S

$$\mathbf{S} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix}$$

where

$$\mathbf{S} = \begin{bmatrix} \dfrac{3 - 2e^{\phi 11}}{-1 + 2e^{\phi 11}} & e^{-\phi 21} \\ e^{-\phi 12} & \dfrac{3 - 2e^{\phi 22}}{-1 + 2e^{\phi 22}} \end{bmatrix}$$

Tables for converting the scattering parameters[4,5] or the matrix set[2] **P**, into h, y or z parameters are readily available.

Above 100MHz alternative methods[3,4] using directional couplers are available for measuring the **S** parameters.

The Transmission-Parameter Method

Let the transmission matrix of the device, including input and output terminations be denoted by $[\mathbf{A}^T]$ where

$$[\mathbf{A}^T] = \begin{bmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{bmatrix}$$

If we now connect for measurement purposes, a known resistance R_M in series with the output of the device we can form a new matrix $[A^M]$ as follows:

$$[\mathbf{A}^M] = [\mathbf{A}^T]\begin{bmatrix} 1 & R_M \\ 0 & 1 \end{bmatrix} = \begin{bmatrix} a_{11} & (a_{12} + a_{11}R_M) \\ a_{21} & (a_{22} + a_{21}R_M) \end{bmatrix}$$

and the matrix equation can be written

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = [\mathbf{A}^M] \begin{bmatrix} V_2 \\ -I_2 \end{bmatrix}$$

or

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} a_{11}^M & a_{12}^M \\ a_{21}^M & a_{22}^M \end{bmatrix} \begin{bmatrix} V_2 \\ -I_2 \end{bmatrix}$$

Writing the matrix equation in this form enables the positive inwards current convention to be retained when converting from the transmission parameters to the h, y or z parameters.

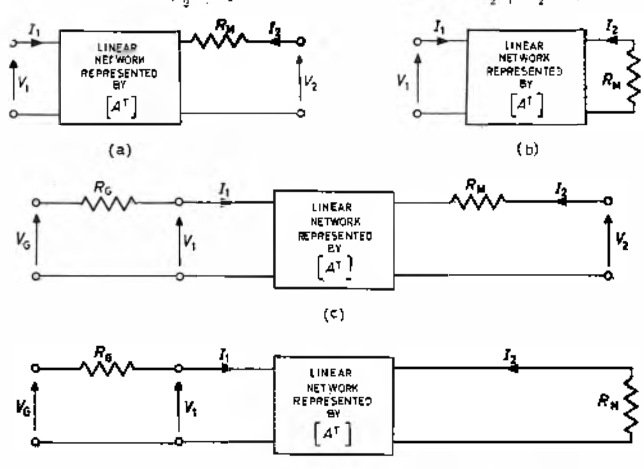
The purpose of the resistance R_M is to permit the output current I_2 to be found when R_M is connected in shunt across the output port (Figs. 1b and 1d). Likewise a series input resistance R_G will also be required in order to find the input current I_1 from the relationship $(V_G - V_1)/R_G = I_1$, where V_G is the generator voltage and V_1 is the potential at the input port (Fig. 1c and 1d).

Having determined the elements of the $[\mathbf{A}^M]$ matrix the elements of the $[\mathbf{A}^T]$ matrix can then be calculated. Although the determination of the $[\mathbf{A}^M]$ matrix elements demands both short-circuit and open-circuit measure-

Fig. 1 Determination of the elements of the (A^m) matrix

(a) $a_{11}{}^M = \dfrac{V_1}{V_2}\Big|_{I_2 = 0}$ (b) $a_{12}{}^M = \dfrac{V_1}{-I_2}\Big|_{V_2 = 0}$

(c) $a_{21}{}^M = \dfrac{I_1}{V_2}\Big|_{I_2 = 0}$ (d) $a_{22}{}^M = \dfrac{I_1}{-I_2}\Big|_{V_2 = 0}$



* Ministry of Technology, Royal Aircraft Establishment.

ments, the open-circuiting and short-circuiting of the semiconductor device is avoided.

The transmission-parameter method is more limited in its application than the scattering method but its relative simplicity is of particular value below 5MHz.

Acknowledgment

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Accurate a.c.-d.c. Convertor for Low Frequencies

by R. W. Gellie and A. G. Klein*

A design for an a.c.-d.c. convertor is presented which is suitable for the precision measurement of the rectified mean value of a low-frequency signal, using a digital voltmeter. It relies on a rectifier bridge inside the feedback loop of an operational amplifier, together with a 2-terminal transistor circuit which provides a large smoothing time-constant without recourse to unduly large capacitances.

(Voir page 533 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 534)

THE essential functions required for a.c. measurements using a digital voltmeter are rectification and smoothing. The accuracy of the measurements must not, however, be affected by nonideal rectifier characteristics or by leakage currents due to the use of large smoothing capacitors. The circuit shown in Fig. 1 satisfies the first requirement, provided that low-leakage silicon diodes are used in the bridge circuit. The alternating current through R_1, and hence also the mean direct current through the load, are forced by the operational amplifier to be accurately proportional to the alternating voltage at the input of the amplifier, irrespective of diode forward voltage drops.

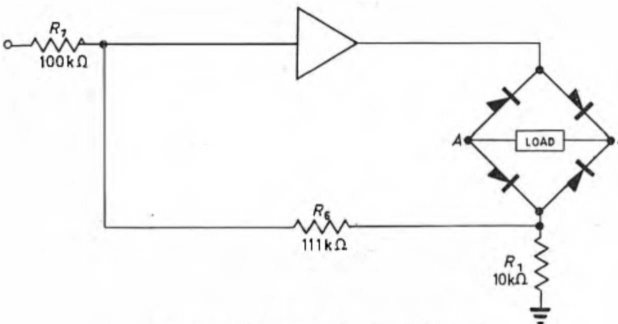

Fig. 1 Amplifier and bridge-rectifier circuit

A digital voltmeter with differential input may be connected across the load AB, provided that sufficient smoothing of the ripple is obtained by the RC time constant of the load.

A 2-terminal network is required between points A and B of Fig. 1, which can pass all the direct current but presents a high impedance to the a.c. ripple, so that a long time constant can be realised by a relatively small shunt capacitor. These conditions are satisfied by the circuit shown in Fig. 2. Capacitor C_1, chosen to have negligible leakage, bypasses almost all of the ripple and supplies sufficient d.c. bias for the transistors to operate. The current-sensing resistor R_1 is chosen so that the current drawn from the operational amplifier during the initial charging of C_1 is kept within the amplifier capabilities. R_2 should be kept small compared with the resistance of the bias network and the input impedance of the digital voltmeter, though large enough to allow a reasonable dynamic range. A suitable and convenient choice is to have R_2 equal in value to R_1. The bias resistors R_3 and R_4 may be large (of the order of $h^2_{fe}R_2$), and therefore the filter capacitor C_2 required to define a suitable smoothing time constant may be made sufficiently small to allow the use of a high-quality low-leakage capacitor.

Except for a very small, and approximately constant, fraction of the mean rectified direct current that flows through R_3 and R_4, all the current is driven through R_2 by transistors VT_1 and VT_2, producing an accurate direct

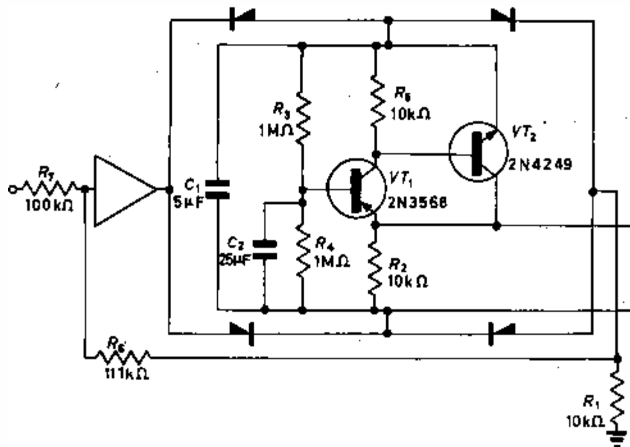

Fig. 2 Overall circuit giving filtered output

voltage across it. This voltage, being accurately proportional to the a.c. input, may be fed directly to a differential digital voltmeter, or to an ordinary digital voltmeter *via* an operational amplifier with differential input.

The operational amplifier used has a maximum voltage swing of 100V; so that, with the circuit parameters shown in Fig. 2, input signals with a peak voltage between 3 and 30V can be accepted. By changing the gain of the circuit, e.g. by changing the value of R_6, the range of measurement may be varied over wide limits. Furthermore, if the feedback resistor is chosen so that $R_6 = (2\sqrt{2}/\pi)\,R_7$, the output voltage will be equal to the r.m.s. of the input for sinusoidal signals. The circuit can be calibrated and tested with a d.c. input of either polarity.

A worthwhile improvement may be obtained by the use of a field-effect transistor in place of VT_1, allowing the use of much larger bias resistors and smaller values for C_2. This was not found necessary in the present application.

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New Shutdown Amplifier (s.d.a.) for Nuclear Reactors

(Part 1)

by I. G. Morris*, C.Eng., M.I.E.R.E., A.M.I.C.E., A.M.I.Mech.E., M.I.E.E.

Since its original application as a neutron-flux protective element over the upper decades of power and for continuous monitoring during full-power operation, the shutdown amplifier (s.d.a.), as employed in UK civil reactors, has remained almost as it was originally designed by the UKAEA, except for minor facilities which have been included to suit particular station requirements. Basically, however, the commercial equivalents have retained the original circuit form over many years, this consisting of a d.c. amplifier operating through a mechanical chopper of dubious reliability and relatively short life. Here, the low-current input signal, converted into a suitable voltage, is compared with a reference, amplified in various valve stages followed by a phase-sensitive output stage; meter indication and appropriate alarm and trip stages are also provided, of course.

Over the years, reactor-protection instrumentation has followed the general trend, and transistorized equipment has been developed. The first item in this field to be made solid state was the temperature-trip amplifier; this has been followed, more recently, by the neutron-flux instrumentation. This article discusses the basic design of one of a family of newly developed transistorized neutron-flux protection instruments, namely the shutdown amplifier which is to be installed as part of the Wylfa Nuclear Power Station equipment

(Voir page 533 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 534)

THE system employed for the shutdown amplifier is shown in Fig. 1. The heart of the device is a solid-state chopper d.c. amplifier comprising two a.c.-amplifier stages (*A* and *B*) of well defined gain which are connected in cascade, the gain of this section of the amplifier being stabilized by internal a.c. feedback. The input current from the ion chamber is compared with a reference derived from either an internal or an external source which is also employed in other stages.

The output of the first stage of this initial amplifier section is employed to drive a third a.c.-amplifier stage (*C*) which is utilized to provide the zero-margin-trip (z.m.t.) facility. This stage provides an output which is in

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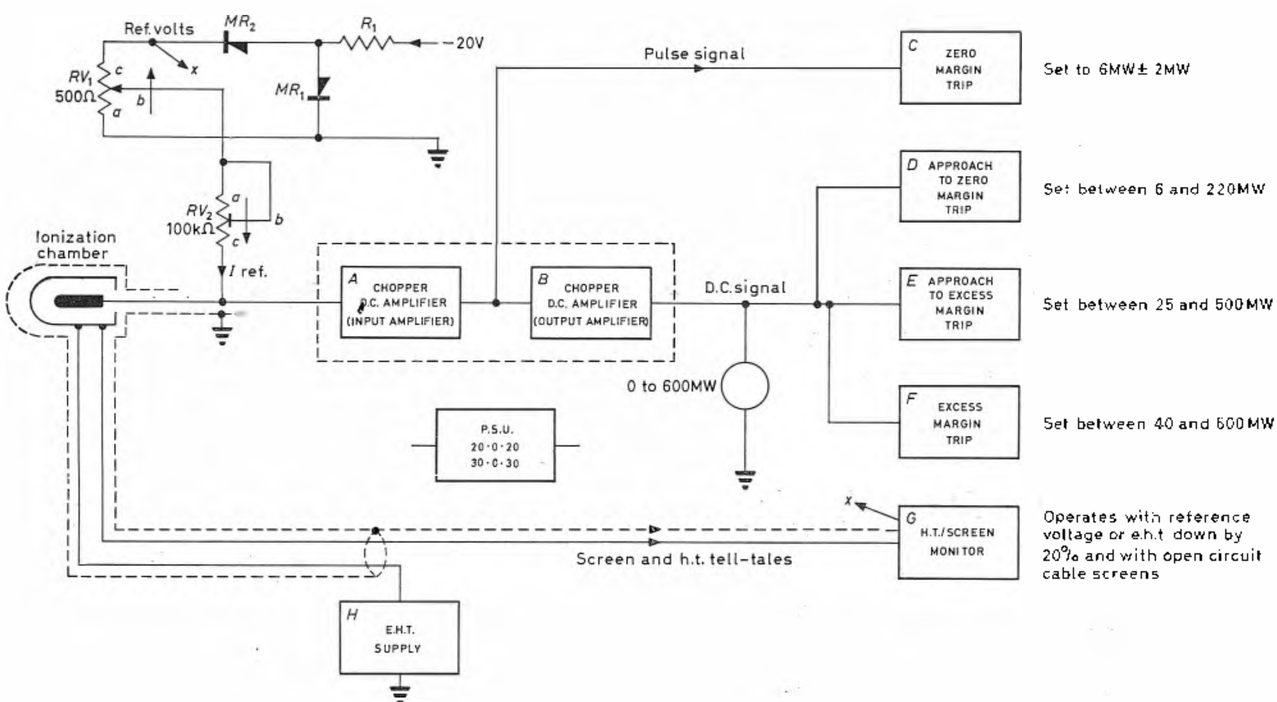
Fig. 1 Schematic diagram of the s.d.a.



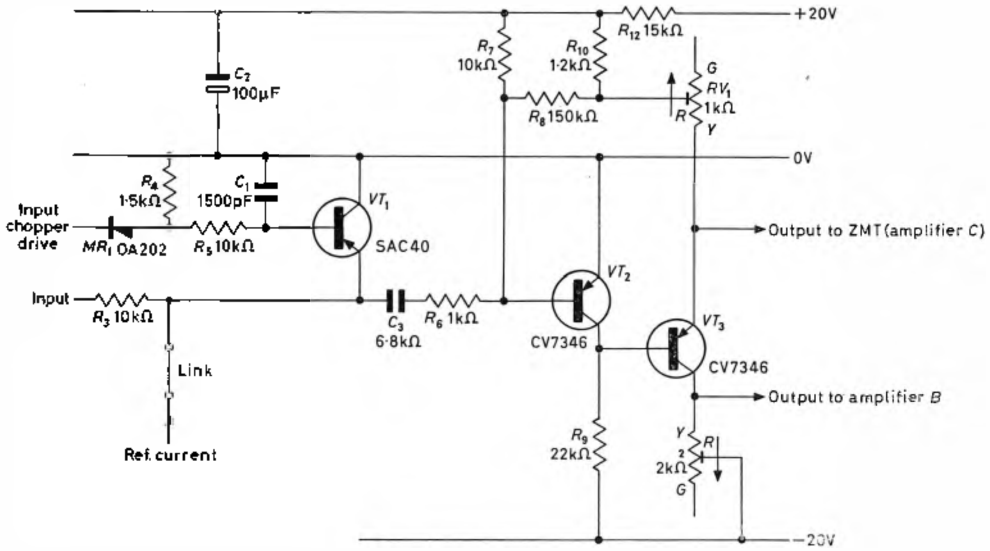


Fig. 2 Amplifier A

parallel with the second amplifier stage (*B*) and is approximately 100 times as sensitive. The zero-margin trip is derived from a d.c. signal output from this amplifier; this energizes an internal slave relay, which, in turn, controls the relays in the reactor safety lines.

A d.c. output is obtained from the main amplifier (*B*), which is employed in energizing internal and external meters and three trip circuits: approaching-zero-margin trip (a.z.m.t.), approaching-excess-margin trip (a.e.m.t.) and excess-margin trip (e.m.t.). These trip circuits are d.c. chopper amplifiers similar to the zero-margin-trip stage but with the addition of individual input choppers. An identical trip unit is utilized to monitor the continuity of the cable screening to the ionization chamber, the internal reference voltage and the e.h.t. voltage *via* a 'tell-tail' lead. The system is arranged so that a trip occurs in the event of an open-circuit in the cable screen or when the e.h.t. falls, or the internal reference voltage rises, by 20%. The margin to trip is indicated on a 0-600MW scaled meter, and the trip levels range between the highest figure quoted in Fig. 1 at maximum reactor output, say 2400MW, to the minimum quoted figure, when the reactor power is below 200MW, the transition between these two extremes being performed smoothly by the provision of a motorized potentiometer which varies the trip levels accordingly.

The range of ion-chamber currents that the instrument is designed to cater for is between 40 and 400μA, for, say, 2400MW full power. This range will, of course, be much reduced when the precise requirements for a particular application are known. The full-scale deflection of the s.d.a. margin meter (600MW) will, in this case, correspond to 10-100μA.

Circuit descriptions

Since the overall circuit is made up of individual wired boards, they are considered here as separate items.

Amplifier A

This amplifier (Fig. 2) is capable of accepting d.c. input signals in the range 10^{-7} to 10^{-4}A which are converted into a rectangular waveform having a 1:4 mark/space ratio by the input chopper stage. Amplification of this a.c. waveform is achieved using a ring-of-two amplifier VT_2 and VT_3, which gives an output signal to the z.m.t. of approximately 5V peak-to-peak. This output is taken from the emitter of VT_3 and its amplitude can be preset by the gain control RV_1. A secondary output signal is taken from the collector of VT_3 to the input of amplifier *B*, the output of which is used to energize the e.m., a.e.m. and a.z.m. trip circuits. It is possible for this secondary output to be preset over a range of 0 to −2·5V peak-to-peak by RV_2.

The circuit is arranged so that the reference current and the ionization-chamber current are of opposite polarity and are fed *via* a link and R_3 to the chopper d.c. amplifier input. If these two currents are equal (i.e. no margin), there is no current feed to the amplifier and the resultant

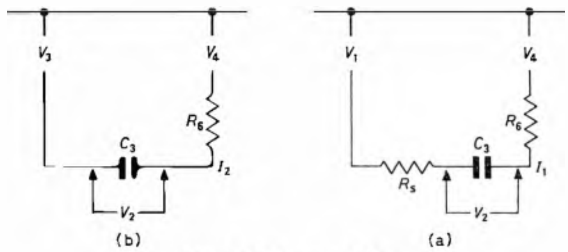


Fig. 3 (a) Chopper OFF
(b) Chopper ON

output is zero. Under normal conditions, however, the reference signal always exceeds the ionization-chamber signal, and the difference forms a negative input current from the junction of the link and R_3 to the chopper emitter. The output signal is, of course, proportional to this margin.

Input chopper

The operation of the chopper circuit can best be described with reference to Figs. 3(a) and (b), which illustrate the condition of the circuit with the chopper OFF and ON, respectively.

If

V_1 and R_5 are the current-source parameters to produce an input current of I_1 amperes

V_4 is the base bias of VT_2

V_3 is the offset voltage of VT_1

R_6 is assumed to be much greater than the resistance of the chopper in the ON state and the input resistance of the amplifier

t_1 is the duration of the OFF state

t_2 is the duration of the ON state

the time constant R_6C_3 is large compared with both t_1 and t_2; so that any voltage changes across C_3 are linear.

Consider the system when first switched on, when V_2 is zero:

$$I_1 = \frac{V_1 + V_4}{R_s + R_6}$$

and

$$I_2 = \frac{V_3 + V_4}{R_6}$$

In the static condition, the charges into and out of C_3 in each halfcycle must be equal:

$$I_1t_1 = I_2t_2$$

It would be extremely fortuitous if the currents I_1 and I_2 obtained on initial switching-on met this condition. However, any unbalanced state is corrected over the first few cycles by the building up of a standing charge on C_3, shown in Figs. 3(a) and (b) as V_2. Once V_2 is established,

$$I_1 = \frac{V_1 + V_2 + V_4}{R_s + R_6} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

$$-I_2 = \frac{V_2 + V_3 + V_4}{R_6} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

and

$$I_1t_1 = I_2t_2 \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

In equation (1), $V_1 \gg V_2 + V_4$, $R_s \gg R_6$ and I_1 is very nearly equal to V_1/R_s and is virtually independent of the chopper circuit; from equation (2), I_2 is dependent on V_3, V_4 and R_6, and it would be expected that variations in these should cause errors. However, should any change occur, owing to temperature changes for instance, V_2 changes to compensate for it and maintains the validity of equation (3). From this, it can be seen that the most important parameter in determining chopper stability is not the pedestal voltage but, in fact, the mark/space ratio of the chopper-drive signal.

The peak-to-peak a.c. signal which is seen by amplifier VT_2 is the sum of the charge and discharge currents, $I_1 + I_2$. In the circuit used, $I_1 = 10$ to $100\mu A$ (determined from the assumed range of input currents for f.s.d.), the output mark/space ratio is 4:1, and $I_2 = 40$ to 400μA (determined from the assumed range of input circuits). The peak-to-peak amplifier input is thus 50 to 500μA.

The chopper-drive signal is d.c.-coupled and balanced about earth, the transistor being driven into its ON state by a 10V negative pulse of 0·5ms duration and the OFF state being produced by a 10V positive pulse of 2ms duration. The stability of the mark/space ratio is better than 1% over a 20 deg C change in temperature.

The diode MR_1 has been included to reduce the effects of spikes from the chopper drive on the amplifier output, which normally pass through the base-collector impedance of the chopper transistor when it is cut off. MR_1 limits the amplitude of this cutoff voltage to manageable proportions and keeps the zero error of the output meter down to within approximately $\pm 0{\cdot}5\%$ of f.s.d.

Common amplifier

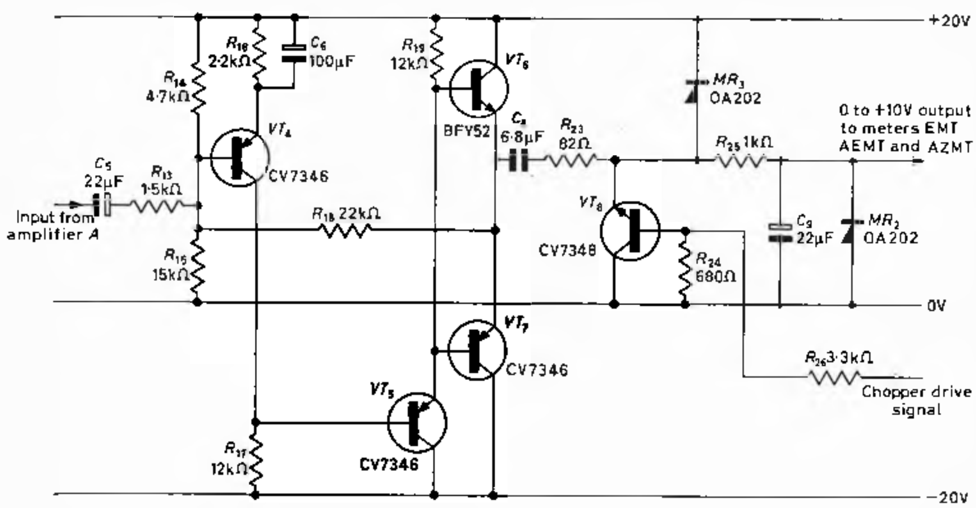
This is a ring-of-two circuit utilizing transistors VT_2 and VT_3, the d.c. working points being stabilized by virtue of an overall d.c. feedback loop from the emitter of VT_3 to the base of VT_2. This d.c. feedback has a ratio $R_{12}/(R_7 + R_{12})$ (15kΩ/25kΩ), which is approximately 60%, providing very stable d.c. conditions and reducing the possibility of the amplifier operating outside its linear range. This feedback path is decoupled to a.c. by C_2 and has no effect on the a.c. gain of the amplifier.

Additional a.c. feedback is also included, again from the emitter of VT_3 to the base of VT_2. Since both the amplifiers connected to the output require a voltage drive, it is better to define the gain of amplifier A as a voltage output for a particular current input. Gain to the emitter of VT_3 (z.m.t. input) is, when calculated on this basis,

$$R_{10} + RV_1 (R \text{ to } Y) \times 1/R_8 \text{ volts/ampere}$$

By suitable adjustment of RV_1 a total range of $\pm 25\%$ is available on the z.m.t. gain of the amplifier, which is considered adequate when all the parameters of reactor flux, chamber position and chamber sensitivity are known

Fig. 4 Amplifier B



for a particular application. At present, however, the range to be accommodated is 10:1, which is achieved by selecting the optimum value for R_8.

The gain to the collector of VT_3 is virtually independent of the setting of RV_1. The effect on collector gain of varying R_8 is the same as the effect at the z.m.t. output, giving the same total dynamic range of 10:1. It is arranged that the output from the collector circuit can be attenuated by RV_2, in order to provide the correct overall gain to the meter circuit.

A feature of this circuit is that any drift in the gain of an amplifier is reduced in the same ratio as that of the amplification. Taking the worst case, where a 50μA input is required to give a 5V peak output at the emitter (a 50μA pulse input is equivalent to a 10μA d.c. input to the chopper), with no feedback and minimum transistor gains of 35, the overall gain (taking into account shunt paths) is 300 to 400. For an output of 5V peak-to-peak into an emitter load of approximately 1·7kΩ, the output is 3mA and the input signal is 10μA. Reduction in gain, due to feedback and improvement in stability, is by a factor of 5, and at the other end of the range required this factor rises to 50.

it capable of providing a stable gain with large voltage swings at the output (up to 30V peak-to-peak); its voltage gain is determined approximately by the two resistors R_{13} and R_{18}. The open-circuit gain is, in fact, some ten times higher than this. The feedback thus gives an improvement of ten times on the gain stability and reduces any non-linearity of the amplification by the same factor.

The feedback provided is both d.c. and a.c., but, in addition, the stability of the d.c.-bias condition is improved by the a.c.-decoupled resistor R_{16} in the emitter of VT_4. Here the base voltage of VT_4 is set by the two resistors R_{14} and R_{15} and the current fed back to the input *via* R_{18}. This voltage fixes the potential across R_{16} and hence determines the emitter and collector currents of VT_4; the collector current develops a voltage across R_{17} which defines directly the bias settings for the base of VT_5 and indirectly the bias setting for VT_6 and VT_7 bases. Hence, by design, this circuit is capable of simultaneous limiting at both extremes of the highest peak-to-peak a.c. output signals.

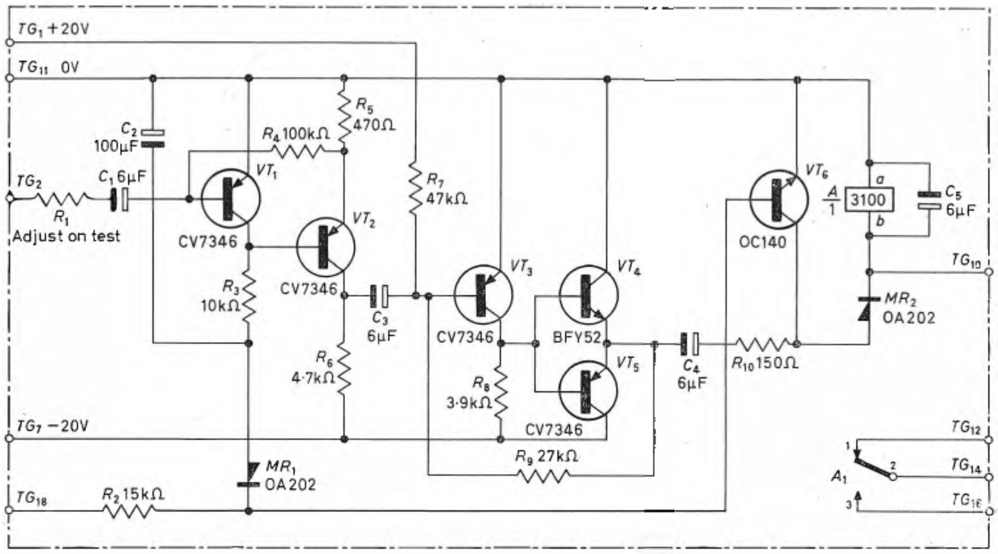


Fig. 5 Amplifier (z.m.t. board)

Amplifier B

The circuit for this amplifier is shown in Fig. 4. The input is a rectangular waveform of about 1·5 to 2V peak-to-peak received from the collector output of amplifier *A*. The a.c. output from amplifier *B* is rectified by an output chopper circuit, referenced to earth, which is driven by a signal in antiphase to the input chopper drive of amplifier *A*. The phase of the output is arranged to give a positive d.c. signal across the storage capacitor C_9. This output signal is within the range 0 ±10V and is capable of delivering 3mA load current, which supplies internal and external meters (10V giving f.s.d.), the signal for three trips (e.m., a.e.m. and a.z.m.) and, in addition, recorder and data-process signals.

A.C. amplifier

The a.c. amplifier provides all the gain in amplifier *B* and consists of a ring-of-three circuit having a complementary output stage. The design of the amplifier makes

Output chopper

This circuit is fairly conventional in its operation, which, in the ideal case, can be described as follows:

The waveform appearing at the emitters of VT_6 and VT_7 is out of phase with that on the collector of the input chopper VT_1 of amplifier *A*, and therefore consists of positive pulses of long duration and a negative base line of shorter duration (4 times shorter). During the negative period, VT_8 is turned on by a waveform in synchronism with, but of opposite polarity to, the input chopper-drive waveform. In this period, C_8 acquires a charge which is proportional to the negative swing of the VT_6 and VT_7 emitters. At the instant when the emitters of VT_6 and VT_7 go positive, the drive to VT_8 turns it off, both sides of C_8 being raised by the same voltage together (i.e. no charge or discharge of C_8 takes place in this 'ideal' case). The voltage across VT_8 is then equal to the full peak-to-peak voltage at VT_6 and VT_7 emitters, which, in the ideal case, would be transferred without loss to C_9 and there retained as a steady direct output voltage.

In practice, however, this voltage is reduced by many factors:

(a) Owing to phase changes and the finite bandwidth of the amplifiers, the amplifier output and the chopper are not in the correct phase relationship. This can produce large spikes of current at the edges of the pulses and cause errors in the output voltage. For example, VT_6 and VT_7 are driven positive while VT_8 is still in the ON state, C_8 is partially discharged *via* VT_9 (in the circuit, R_{23} is incorporated to limit this error). Admittedly, when current is drawn off C_8 to feed the load circuit, some error voltage is developed across R_{23}, but, since this error is stable, it is more easily corrected than that due to spikes.

(b) The storage capacitor C_9 loses some of its charge in the loading circuit (meters etc.) during the period when the output signal is negative; thus, when the output rises to its positive peak, this is in excess of the voltage across C_9. Charge has therefore to be taken from C_8 to recharge C_9, making the voltage across C_9 lower than that expected in the ideal case. The charge removed from C_8 every positive cycle is replaced through VT_8 on the negative excursion, when VT_9 is turned on.

(c) When the a.c. output is fully positive, the 'ideal' value of R_{25} is zero, since this allows for a free transfer of charge from C_8 to C_9. However, in the opposite case, when the output is negative, C_9 tends to discharge *via* R_{25} and the output chopper. The ideal in this case is a very large resistance for R_{25}, giving a low discharge current. It may be thought that a diode would give the best solution, and from the point of view of efficiency this is certainly the case, but it unfortunately leads to a nonlinear output signal. Such a signal would be permissible in trip circuits but it is not satisfactory where the output is intended to give a linear meter display. The value of R_{25}, chosen as a compromise solution, has the overall effect of reducing the d.c. output signal very considerably, by two means, while still permitting a linear 10V d.c. signal output. Firstly, during the period when VT_9 is on and R_{25} is in parallel with the load, it increases the charge drained from C_9; this increases the voltage drop, which has, of course, to be made up by further discharging C_8. Secondly, during the period when C_8 is charging up, C_9, R_{25} and the load resistance form a voltage-dividing network, and it is this network which steps down the voltage transferred and limits the maximum voltage that C_9 could attain.

The overall result of these departures from the ideal condition is to reduce the output from a 22V peak-to-peak signal to a 10V d.c. signal.

When the circuit is initially switched on, there is a possibility that the output will be taken well negative, and in order to protect C_9 from damage until C_8 acquires its standing d.c. bias, MR_2 has been included in the circuit. This diode conducts whenever the inverse polarity is applied across the electrolytic capacitor. MR_3 prevents the maximum positive swing of the output pulse from exceeding +20V, which is possible under conditions of excessively high inputs by maintaining the voltage across the output chopper within the transistor ratings.

Amplifier *C* (Zero-margin trips)

This item is the zero-margin-trip circuit (Fig. 5). The input is the a.c. output from amplifier *A*, which is amplified and rectified to produce a direct-current drive to energise the z.m.t. relay.

First amplifier stage

This amplifier is a ring-of-two stage using VT_1 and VT_2, which is similar in design to amplifier *A*. The d.c. conditions are determined and stabilized by d.c. feedback from the emitter of VT_2 to the base of VT_1, the working point of VT_2 collector being held to within ±20% to prevent the stage from operating under nonlinear conditions. The stage gain is determined, in part, by the values of R_4 and R_1, C_1 acting purely as a blocking capacitor. The value of R_1 is selected to ensure that, with the equivalent of 6MW (50mV peak-to-peak input), the relay is just de-energised. The collector supply for VT_1 is derived from the output chopper signal; MR_1 and C_2 are rectifier and smoothing capacitor, respectively.

The gain of the amplifier can be assessed as follows: the voltage gain to VT_2 emitter is R_4/R_1, the current output from VT_3 is equal to the emitter voltage of VT_2 divided by the emitter resistance:

$$V_{in}(R_4/R_1)/R_5$$

$$(R_5 \ll R_4,\ VT_3 \text{ input} \ll R_6)$$

$$\text{gain} = \frac{R_4}{R_1 + R_5} \text{ amperes/volt}$$

With values of R_1 that can vary between 12 and 33kΩ, the gain will be between 8·5 and 3mA/V. The current gain of the stage is R_4/R_5—approximately 200. Assuming transistor gains of 35, the open-circuit gain is 400 and, taking shunt paths into account, this gives overall improvement in stability by a factor of 2.

Output amplifier and chopper

The output amplifier is basically a ring-of-two stage, with a complementary emitter-follower output stage (VT_4 and VT_5). The d.c.-bias conditions are determined by R_9 and R_7, and the resulting large d.c. feedback ensures very stable d.c. operating conditions. The a.c. gain is determined by R_9 in conjunction with the output impedance of the first amplifier; the amount of feedback effected results in a stability factor of 8 over the open-loop case. The maximum output voltage without amplifier limiting is 10V, which is sufficient to drive the most insensitive relays of the type employed.

The output-chopper-circuit description is similar to that given for amplifier *B*, except that this is a trip circuit; therefore, in this case, R_{25} can be replaced by a diode (MR_2) and no charge is then lost from C_5 via the short-circuit chopper. The overall efficiency of this chopper is much better than that of the main d.c. amplifier but, as previously stated, the output is nonlinear at the lower end, However, since this stage incorporates a trip output, this is not a serious condition. R_{10} is again included to limit current surges in the possible event of amplifier output and chopper-drive signal not being exactly in phase.

Amplifier *D* (standard trip circuit)

This was designed as a general-purpose trip circuit. In th overall circuit, it is employed in monitoring for e.h.t. failure and signal and e.h.t. cable-screen continuity and as the e.m., a.e.m. and a.z.m. trips. The input in all cases is the sum of a d.c. signal current and a reference current of the opposite polarity, the resultant d.c. current of which is chopped by an input chopper VT_1 to produce a rectangular waveform in the manner described earlier. This is amplified and rectified by a circuit similar to that described for the e.m.t., but Synchronisation of its waveform with that used in the other cases is not essential.

To overcome certain fail-to-danger faults in this circuit, which occur with modes of use, two varieties have been designed and are discussed in the second part of this article 'Design for Fail-Safety.'

(*To be continued*)

Strip Transmission Lines

by J. Swift*, B.Sc., M.I.R.E.E.(Aust.)

A survey of strip transmission lines used in microwave circuits is given, including applications, design data and information on discontinuities, transitions and coupling between striplines.

(Voir page 533 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 534)

STRIPLINE is widely used in the design of microwave components in the frequency range 1-12GHz, including such devices as filters, directional couplers, mixers, parametric amplifiers, varactor multipliers and tunnel-diode circuits. Most coaxial and waveguide components may be manufactured in stripline, with the advantages of low cost and ease of construction, particularly in complex equipment. Stripline has small bulk and weight and quantity production is possible using printing techniques.

During the 1950s, considerable work was done on stripline and, at the time, enthusiasm among microwave engineers ran very high. However, over the years, it is clear that waveguides and coaxial lines have not been displaced, and stripline occupies a firm, but auxiliary, role. Nevertheless, its use is likely to increase, particularly in circuits using microwave solid-state devices.

Five common types of stripline are shown in Fig. 1. The shielded striplines shown in (a) and (b) consist of two metal sheets with a rectangular strip or cylindrical rod between them; air dielectric is most frequently used in a line of this type. The microstripline in (c) consists of a strip conductor spaced from the ground plane by a dielectric layer, the spacing being a small fraction of a wavelength. The unshielded stripline shown in (d) consists of an asymmetrical air-spaced conductor above a ground plane, while (e) shows a triplate line which is particularly useful when a high Q-factor is required. The metal parts of the lines are usually made of brass or copper, and, if a dielectric is used, a low-loss material such as Teflon or silicon-impregnated glass fibre is usual.

Although stripline configurations appear simple, the presence sometimes of longitudinal electric and magnetic fields at the dielectric interface make a rigorous analysis difficult. In these cases the TEM-mode solution offers only an approximation, and Maxwell's equations must be applied directly to the line in question. As far as wideband performance is concerned, stripline is superior to ordinary waveguide and comparable with coaxial line and ridge waveguide. It has the disadvantage that it is sometimes difficult to design a sufficiently rigid structure. In directional couplers and other devices where coupled lines are used, it is frequently difficult to control critical dimensions during manufacture. A further problem occurs where strip-line and other components have to be interconnected, and a transition to a coaxial-line connector is usually necessary. The attenuation of stripline is, in general, comparable with that of coaxial lines but poorer than that of waveguides.

Characteristics of stripline

Shielded stripline

A cross-section of this type of strip transmission line is shown in Fig. 2. The line[1] consists of a central metal strip of thickness t and width w enclosed by two metal planes of separation b. If the width of the planes is large compared with the plate thickness, propagation is in the TEM mode, with a phase velocity $C/\sqrt{\epsilon}$ and a guide wavelength $\lambda_0/\sqrt{\epsilon}$, ϵ being the relative permittivity of the medium between the planes.

An exact formula for the characteristic impedance Z_0 when $t = 0$ is available:[2]

$$Z_0 = \frac{30\pi\, K(k)}{K(k')} \quad \text{.............. (1)}$$

where $K(k)$ and $K(k')$ are complete elliptic integrals of the first kind, given by:

$$k = \operatorname{sech} \pi w/2b \;;\; k' = \tanh \pi w/2b \quad \text{........ (2)}$$

When $w/b \gg 1$, equation (1) reduces to

$$Z_0 \simeq 15\pi^2 [(\pi w/2b) + \ln 2] \quad \text{........... (3)}$$

and when $w/b \ll 1$, equation (1) reduces to

$$Z_0 \simeq 60 \ln (8b/\pi w) \quad \text{............. (4)}$$

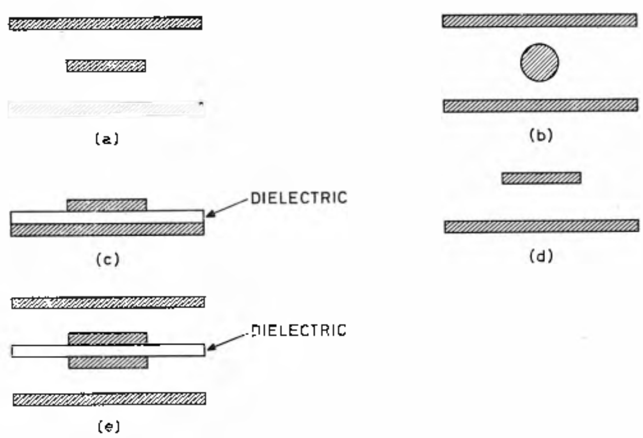
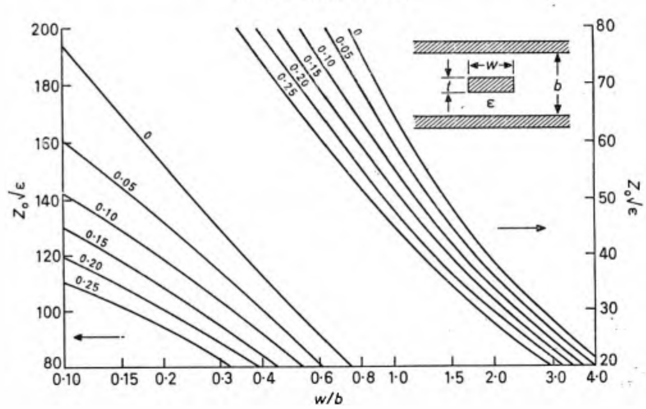


Fig. 1 (a) Shielded stripline (rectangular inner conductor)
(b) Shielded stripline (cylindrical inner conductor)
(c) Microstrip
(d) Single groundplane
(e) Triplate line

Fig. 2 Z_0 of shielded stripline, Z_0 as a function of w/t for various values of t/b (from Cohn[2])



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In any practical design, the strip thickness t is finite and may have a large effect on the characteristic impedance. The case where t is finite has been examined[3-7], and Cohn[3,4] gives expressions for wide and narrow strips that agree with the exact expression [equation (1)] to within 1·2%. The expressions hold to this accuracy for a strip thickness up to one-quarter of the plate spacing. A graph of Z_0 based on Reference 3 is given in Fig. 2.

The attenuation of the stripline comprises two components: the dielectric loss α_d and the conductor loss α_c. In most cases these losses are relatively small, and, if the dielectric completely fills the line, the total attenuation α is given by

$$\alpha = \alpha_c + \alpha_d \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (5)$$

For any TEM transmission line, including this type of stripline, α_d is given (in dB/m) by

$$\alpha_d = (27{\cdot}3/\lambda_0)\sqrt{\epsilon}\tan\delta \quad \ldots\ldots\ldots\ldots \quad (6)$$

where λ_0 is the free-space wavelength and $\tan\delta$ is the loss tangent of the dielectric.

Formulas for the conductor loss α_c are considerably more difficult in derivation. Cohn[4] has provided expressions for α_c for wide and narrow strips and given a resultant composite curve, accurate to within a few per cent, for t/b ranging from 0·001 to 0·1; the curve is given in Fig. 3. For a typical line of this type operating at 3GHz, α is of the order of 1·25dB/m.

Another particularly useful type of shielded stripline, using a cylindrical inner conductor, is shown in cross-section in Fig. 4. The characteristic impedance Z_0 is given approximately by Wheeler[8]:

$$\sqrt{\epsilon}\,Z_0 \simeq 60\,[\ln(1/\rho) - (2/9)\,\rho^4] \quad \ldots\ldots\ldots \quad (7)$$

where $\rho = \pi d/4b$

The error in equation (7) is less than 0·5% for $Z_0 \geqslant 50\Omega$. A curve of Z_0 based on equation (7) is given in Fig. 4.

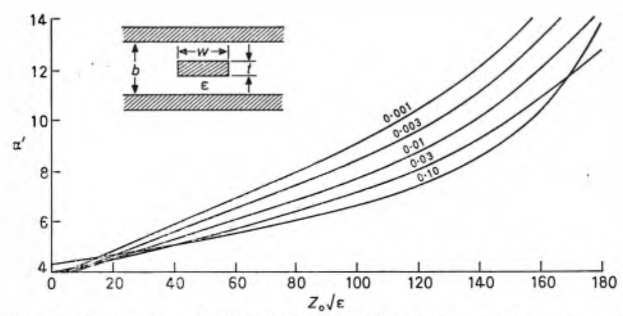
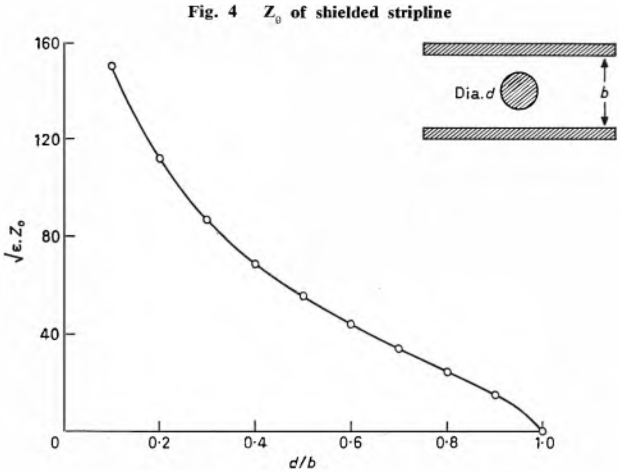


Fig. 3 Attenuation of shielded stripline. The curves are for copper and various values of t/b. $\alpha' = 10 \cdot \alpha_c b(\epsilon f)^{-1/2}$ dB/m (from Cohn)

Fig. 4 Z_0 of shielded stripline



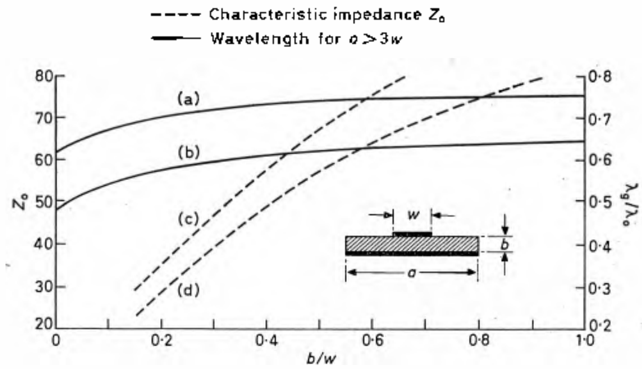


Fig. 5 Characteristics of solid-dielectric microstripline (from Dukes[16])
(a) and (c) P.T.F.E. Fibreglass, ϵ = 2.65
(b) and (d) Silicone Fibreglass, ϵ = 4·18

For more accurate work, a more exact expression has been given by Wheeler[8].

For a ground-plane spacing of less than a quarter-wavelength in the two shielded striplines described above, the transverse attenuation of the fields α_t is given (in dB/m) by

$$\alpha_t = 27/b \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (8)$$

Thus, in a transverse distance equal to the ground-plane spacing, the attenuation is 27dB. Hence several stripline circuits can be placed close together with common ground plants and there is negligible interaction.

Microstripline

A microstripline[9,10] consists of a strip conductor spaced from a ground plane by a dielectric layer, the spacing being only a small part of a wavelength. Microstripline is used less than the shielded types of line because of radiation and stray coupling effects, and its use is usually confined to low-Q-factor components. It may be produced readily by printed-circuit techniques, with considerable saving in cost, space and weight. A cross-section of this type of line is shown in Fig. 5.

Approximate[11] and exact[12,13] theories of microstripline propagation have been given, and an accurate calculation of the wavelength has also been carried out[14]. In general, propagation is not that of a pure TEM mode. For wide strips, however, where b/w tends towards zero, the wavelength in the microstripline approaches $\lambda_0/\sqrt{\epsilon}$, as in a coaxial line. The characteristic impedance Z_0 has been given by Bowness[15] for when $w > b$ and the strip conductor is thin compared with its width:

$$Z_0 = 10^4/3\,\sqrt{\epsilon}\,[7 + 8{\cdot}83\,(w/b)] \quad \ldots\ldots\ldots \quad (9)$$

A graph of Z_0 is given in Fig. 5, based on the work of Dukes[16], for the dielectric materials Teflon and silicone glass fibre. The wavelength in the microstripline λ_g in relation to the free-space wavelength λ_0, for $a > 3w$, is also given.

The attenuation in a microstripline is due to conductor loss α_c, dielectric loss α_d and a relatively small amount of loss by radiation. The dielectric loss, which is usually the largest component, is given by equation (6). The conductor loss α_c is given (in dB/m) by

$$\alpha_c \simeq 27{\cdot}3R/\pi b Z_w \quad \ldots\ldots\ldots\ldots\ldots \quad (10)$$

where R is the skin resistance (Ω/m) and Z_w is $377/\epsilon$, the wave impedance of the dielectric.

A typical $\frac{1}{8}$in microstripline operating at 6GHz has a total attenuation of 1·1dB/m (α_c = 0·3dB/m, α_d = 0·8dB/m) and an unloaded Q-factor of about 700[9].

Single ground plane

This simple, unshielded type of stripline consists of a ground plane and an air-spaced rectangular strip[17] or cylindrical rod[10] (Fig. 6). For a strip of negligible thickness t and with $w \gg b$, Z_o is given approximately by[11]

$$Z_o \simeq Z_w \{(w/b) + (2/\pi)[1 + \ln(1 + \pi w/2b)]\}^{-1} \ldots (11)$$

where Z_w is the wave impedance of the air dielectric.

For t finite, but small,

$$Z_o \simeq Z_w \{2+(w/b)+(t/\pi b)[1+\ln(1+2b/t)]\}^{-1} \ldots\ldots (12)$$

Graphs of Z_o, based on these expressions and with $t = 0$ and $t/w = 0{\cdot}2$, are given in Fig. 6.

The application of this form of stripline to the head end of a 2GHz receiver has been described[17], including details of a variable attenuator, resonator and crystal mixer. Without special shielding, the radiation from the ends of the halfwave resonator was considerable and the unloaded Q-factor was only about 100. Typical line dimensions for a characteristic impedance of 50Ω where $w = {}^5/_{16}$in, $b = t = {}^1/_{16}$in.

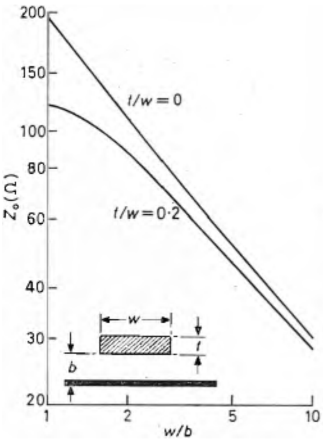


Fig. 6 Z_0 **of single ground plane line (from Assadourain and Rimai[11])**

Triplate line

This type of line has little dielectric or radiation loss and is particularly useful in high-Q-factor microwave components such as resonators and filter elements. The losses are comparable with coaxial line and waveguide, and the stripline may be etched or printed in one operation. A cross-section of triplate line is shown in Fig. 7. Apart from the symmetry obtained, the main reason for using two strips of metal for the inner conductor is to ensure that only fringing fields pass through the dielectric, thus appreciably reducing this type of loss.

The propagation in triplate line is not a pure TEM mode and no unique characteristic impedance can be defined; also the phase velocity depends on the line dimensions. Approximate relations for these parameters have been found, however, using perturbation[18] and relaxation[19] methods, and the results are sufficiently accurate for design purposes.

Experimental results using triplate and other lines have been given by Fromm[20] for the frequency range 0·25 to 10GHz. Among the components described are transitions to coaxial line, attenuators, direction couplers and filters. The unloaded Q-factor in some of the lines ranged from 2000 to 6000. The transverse attenuation of triplate line is given by equation (8). Thus several triplate circuits can be placed close together with common ground planes, and there is negligible interaction.

Coupled lines, transitions and discontinuities

Coupled lines

In a number of applications, including the design of directional couplers, phase shifters and filters, it is necessary to couple striplines together[21-25]. The behaviour of coupled lines is expressed in terms of even- and odd-mode impedances Z_{0e} and Z_{0o} respectively. These are the impedances of one line with respect to ground with even (unbalanced) or odd (balanced) excitation of the line.

For example, in a directional coupler consisting of two striplines a quarter-wavelength long, the midband coupling factor k is given by[24]

$$k = \frac{Z_{0e} - Z_{0o}}{Z_{0e} - Z_{0o}} \ldots\ldots\ldots\ldots\ldots (13)$$

and the characteristic impedance of the coupler (Z_0) is given by

$$Z_0 = \sqrt{(Z_{0e} Z_{0o})} \ldots\ldots\ldots\ldots\ldots (14)$$

For a 3dB coupler, with $Z_0 = 50\Omega$ and $k = 1/\sqrt{2}$, $Z_{0e} = 120{\cdot}71\Omega$ and $Z_{0o} = 20{\cdot}71\Omega$.

In a given design, it is therefore necessary to be able to calculate Z_{0e} and Z_{0o}. For the configuration shown in Fig. 8, the following expression may be used in the case of loose coupling[21]:

$$Z_{0e} \simeq 60 \ln[\coth(\pi s/2b)\coth(\pi d/4b)] \ldots\ldots (15)$$

$$Z_{0o} \simeq 60 \ln[\tanh(\pi s/2b)\coth(\pi d/4b)] \ldots\ldots (16)$$

For the configuration shown in Fig. 9(a), Z_{0e} and Z_{0o} have been found for the case $t = 0$[26,27], and for $t \neq 0$ a thickness correction has been given by Cohn[6,27]. For the lines shown in Figs. 9(b) and (c), values of Z_{0e} and Z_{0o} have been given by Cohn[28] and Hachemeister[26].

A new stripline configuration that allows very close coupling without the usual critical dimensions has been described by Getsinger[29].

Transitions

In order to connect stripline components to coaxial line and waveguides, wideband transitions are needed. In addition, it may be necessary to use suitable baluns to connect balanced and unbalanced striplines together.

A number of coaxial-line stripline transitions have been described, including transitions to triplate[30,31], dielectric-sandwich line[13], microstrip[9], and a stripline with a rectangular-strip inner conductor[31]. Levy[32] has described a transition to triplate line which has a v.s.w.r. of better than 1·02 for frequencies up to 11GHz, while other transitions have been described by Fubini[33] and Muser[34]. The very useful transition, shown in Fig. 10, is suitable for a shielded stripline with a round inner conductor (Fig. 4) and a ground plane spacing of ½in. The v.s.w.r. is less than 1·05 up to a frequency of 10GHz.

At higher frequencies, a waveguide/stripline transition may be desirable. Several such devices have been described, including a waveguide/triplate-line transition[30], a microstrip transition[9] and one involving the use of an intermediate section of coaxial line[16].

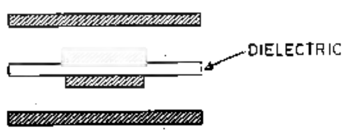


Fig. 7 Triplate line

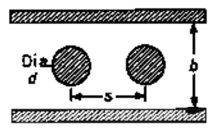
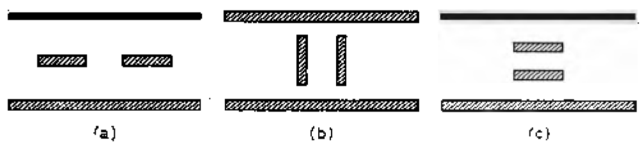


Fig. 8 Coupled lines

Fig. 9 Coupled lines



Where balanced and unbalanced striplines are to be interconnected, several designs for suitable stripline baluns have been given[35-37].

Discontinuities

A variety of discontinuities in the stripline shown in Fig. 2 have been studied by Oliner[38]. Discontinuities discussed include a round hole, a gap in the inner conductor, sharp bends, a radiating slot and a vertical post. The theoretical results are compared with available experimental data. In general, such discontinuities possess purely reactive equivalent networks if they contain no dissipative elements. The equivalent circuit of a step discontinuity in the stripline conductor is an ideal transformer, the square of the turns ratio being given approximately by the ratio of the characteristic impedances.

Further data on stripline discontinuities have been given by Altschuler and Oliner[39,40].

Applications of stripline

Stripline has been widely used in the design of microwave components in the frequency range 1-12GHz. Most waveguide and coaxial-line devices have their counterparts in stripline, which is particularly useful in the design of complex components.

A stripline directional coupler is shown in Fig. 11. The device uses a quarter-wavelength coupled line and operates in the frequency range 0·4 to 0·5GHz. Capacitive slugs are used at each end to compensate for bends in the stripline. Several stripline directional couplers have been described[16,20,21,24,30]. Filters using striplines usually use triplate line in order to obtain a high Q-factor[21,22,41-43]. Direct-coupled filters have been described by Cohn[44], quarter-wavelength coupling by Wanselow and Tuttle[45] and parallel-coupled-resonator filters by Cohn[22]. Directional filters have also been designed[45,46].

Stripline has been used in the design of a precision slotted line[47]. Among other devices described are fixed and variable attenuators[30], microwave switches[48], phase shifters[23], hybrid rings[9,16,30,49,50], and mixers[16,51]. Ferrites have been used in a stripline in isolators[52,53] and circulators[54-56], while complete transmitter-receiver assemblies using stripline components have been designed[16,20].

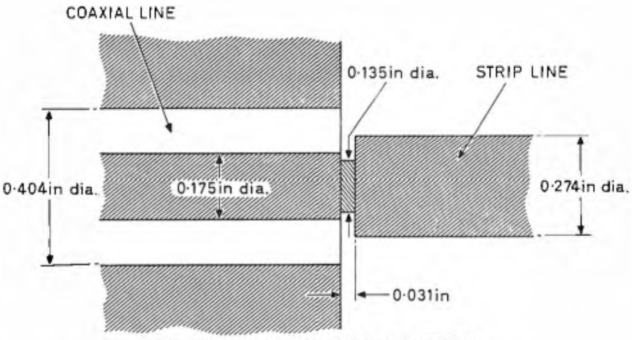
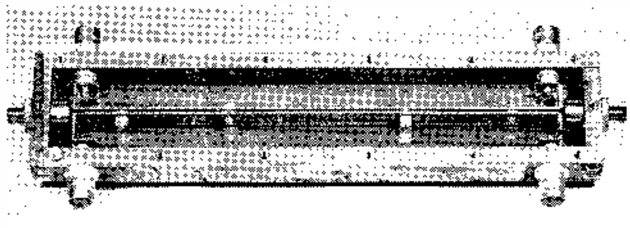


Fig. 10 Coaxial to stripline transition

Fig. 11 Stripline directional coupler



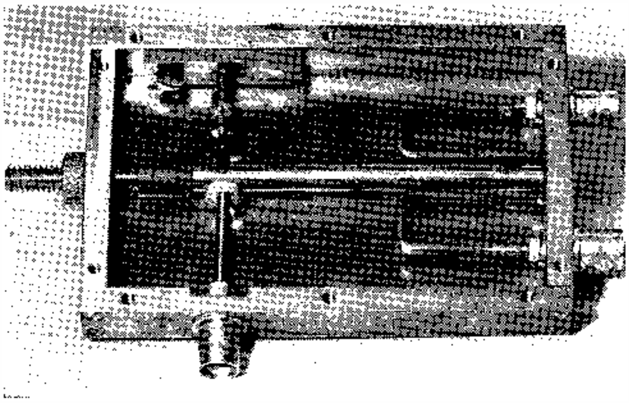


Fig. 12 Stripline parameter amplifier

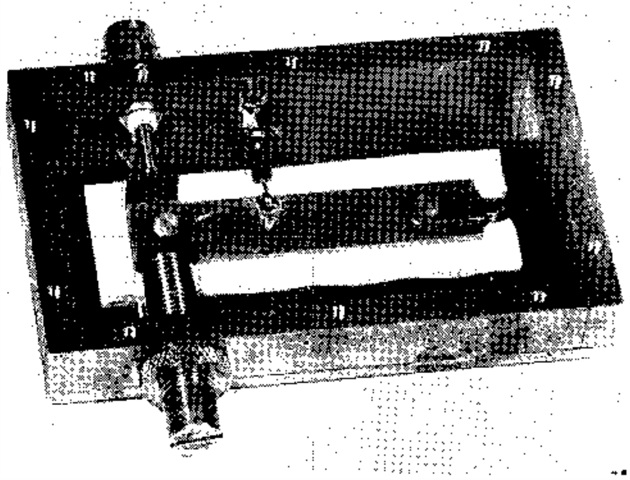


Fig. 13 2GHz tunnel-diode oscillator

Perhaps the biggest potential application of stripline is its use with small microwave semiconductor devices. A simple parametric amplifier using stripline is shown in Fig. 12. This device operates between 0·4 and 0·5GHz and the signal and idler circuits use different modes of the same resonator. A tunnel-diode oscillator is shown in Fig. 13. This device operates at 2GHz and uses microstripline for the resonator[57]. Stripline is also very useful in u.h.f. transistor circuits and varactor multipliers where a compact light unit may be designed.

Acknowledgment

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First Colour-Television Outside Broadcasts

The beginning of the colour-launching period on BBC2, preceding the full colour service which will open on December 2, coincided with the Wimbledon Lawn Tennis Championships. The opportunity was taken to televise some Centre-Court matches in colour, using the newly commissioned 4-camera outside-broadcast unit. The unit consists of the colour mobile control room (c.m.c.r.) and the following ancillary vehicles: rigging tender, camera van, 15cwt runabout van, and 40kVA power van. Only the c.m.c.r. was at Wimbledon, where mains power was used. Four Peto-Scott 3-plumbicon cameras were used, each fitted with a Rank-Taylor-Hobson zoom lens; three cameras viewed the Centre Court and the fourth was used for interviews.

The c.m.c.r. is the first of three to be equipped during 1967-8 by the BBC Studio Planning and Installation Department. Behind the driving cab is the apparatus area and at the rear is the production and vision-control area. The latter part is fully air-conditioned from equipment in the upper part of the cab.

The layout of the production and vision-control area follows conventional o.b. practice in having the producer, secretary, sound mixer and engineering manager at their control desks on a raised area behind the vision-control engineers and the camera-control units. Above the c.c.u.s, and visible to all personnel, are the picture monitors. The BBC vision mixer has 10 channels; it is similar to the type fitted in the latest monochrome m.c.r.s, with modifications for colour operation. The colour-reference burst is maintained at constant amplitude and phase when the output signal is faded. No burst is transmitted when the mixer is handling a monochrome signal. Each camera-control unit is in a self-contained console having a hinged lid giving access to all controls needed for setting up the camera. With the lid closed, the normal operating controls available to the operator are: beam target, black level and gain for each of the red, green and blue channels. There is a master gain control, a master black-level and a lens-iris control. Fitted experimentally is a contours-out-of-green unit which utilises the green signal to provide vertical and horizontal aperture correction and a linear matrix to improve colorimetry.

The picture-monitor rack contains two 19in colour and six 14in monochrome monitors. One colour monitor is coded and displays the transmission or preview picture; the other is switchable to r-g-b or coded operation and is used for colour matching on the camera pictures. Four monochrome monitors are used for channel preview (cameras 1-4); one for a compatibility check on the transmission picture and one is a switchable preview monitor. Feeds are provided to one colour and two monochrome monitors at the commentator's position.

Colour splitting to the three plumbicons is effected by a Philips dichroic prism, resulting in a small assembly with no exposed surfaces in the optical path. Apart from the plumbicons, the cameras are fully transistorized; the head amplifiers were designed and made by the BBC Research Department. The zoom lens is a Rank-Taylor-Hobson Varotal 9, and is servo-operated. This lens has a zoom range of 10:1, covering viewing angles from 45° to 4·5° with a constant aperture of f2; the minimum focusing distance is 6ft. A range extender can be used; this narrows the zoom angle range to 22·5°-2·25° and reduces the aperture to f4. It is intended to fit later the Varotal 16 zoom lens; this has a range of viewing angles from 29° to 1·8° and a minimum object distance of 10ft. The camera carries a zoom-lens-viewing-angle indicator, consisting of a yellow plate which rises out of the top of the camera as the viewing angle is increased, primarily for the guidance of microphone boom operators. The camera viewfinder has a 6½in monochrome c.r.t., on which red, green or blue signals, or any combination of them, may be displayed. Each camera uses two BICC Mk-IV cables and can be used up to 2000ft from the c.c.u.

Large-Deviation Phase and Frequency Modulators

by D. J. Comer*, Ph.D.

This article discusses large-deviation phase and frequency modulators that can be used in control applications. The basic element of these semiconductor circuits is a nonlinear resistor, which is used to offset circuit nonlinearities. The deviations obtained are related linearly to the amplitude of the modulating signal.

(Voir page 533 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 534)

THE circuits to be considered here find their major application in the field of automatic control. Large-deviation frequency modulators are not often used in communication systems, since the frequency deviation is limited to a very small range. However, in control systems requiring phase or frequency correction, large deviations are necessary to construct a usable system.

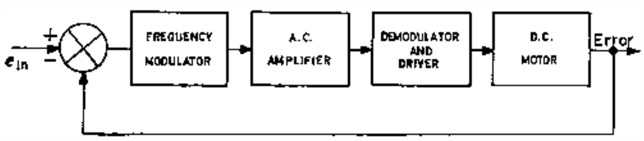
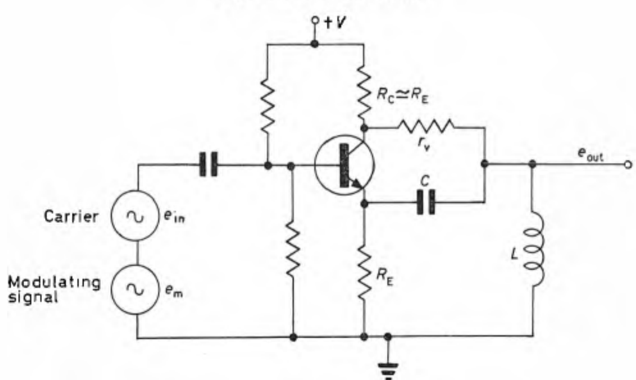


Fig. 1 Hybrid control system

The hybrid control system of Fig. 1 shows one application of a frequency modulator. The error signal is a direct voltage which varies the frequency of the modulator. The a.c. modulator output is amplified and demodulated; the demodulator is designed to have no output when the input frequency is equal to the carrier frequency. Frequencies below the carrier frequency cause a negative demodulator output, while positive outputs result from input frequencies above the carrier frequency. The hybrid system allows the use of a.c., rather than d.c., amplification. A.C. stages are usually more easily designed, cheaper and more stable than d.c. stages; hence, if the modulator and demodulator are relatively uncomplicated, the hybrid system can be profitably used[1].

Fig. 2 Phase modulator



Frequency-modulation systems have found only limited application in the past, because of the complexity of the modulating unit. Conventional modulators do not allow a large enough frequency deviation to function properly

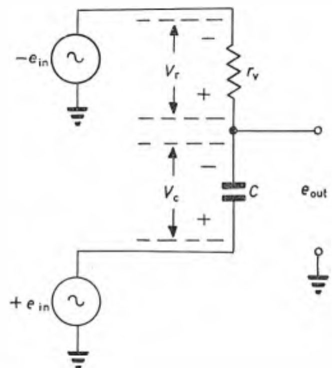


Fig. 3 Equivalent circuit of phase shifter

over the necessary dynamic range of input signal. Unless the frequency deviation is large, undue sensitivity requirements are imposed on the demodulator.

The modulators discussed in this article were developed for the purpose of obtaining large phase and frequency deviations, primarily for use in control applications. The deviations are linearly dependent on the amplitudes of the modulating signals, allowing linear analysis to be applied to the control systems. Furthermore, the modulators are simple and inexpensive to construct.

Phase modulator

The nonlinearities inherent in semiconductor phase modulators limit the maximum phase deviation to relatively

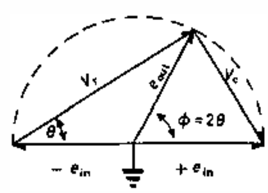


Fig. 4 Phasor diagram used to calculate phase shift

* *University of Calgary, Alberta, Canada.*

small values. At present, a linear phase deviation of $\pm 25°$ per stage is considered to be quite large[2]. The phase modulator of Fig. 2 uses a nonlinear element to offset the circuit nonlinearities and extend the range of phase deviation. Phase deviations of $+50°$, varying linearly with the amplitude of the modulating signal, are obtained with this circuit.

The operation of the circuit is best explained by first considering the phase shift of the carrier signal as a function of the resistance r_v, assuming $e_m = 0$. If the resistors R_C and R_E are small compared with r_v, and so chosen that the gain of the circuit is unity, the equivalent circuit of Fig. 3 applies. The vector diagram of Fig. 4 aids in evaluating e_{out}. The input voltage is taken as the reference and this is also equal to the voltage between emitter and ground. The collector voltage is equal in magnitude to, but 180° out of phase with, the input signal. A total voltage of 2 e_{in} appears across the series combination of resistance and capacitance. The current through this combination will lead the applied voltage by an angle θ:

$$\theta = \tan^{-1}(X_c/r_v) = \tan^{-1}(1/\omega C r_v)$$

where ω is the radian frequency of the input signal. The voltage drop across r_v will bear this same phase relationship to the input signal. The voltage drop across C will lag the current by 90°. Summing voltages around the loop requires that the vectors form a closed path, resulting in the diagram of Fig. 4. As r_v is varied, the length of the vector v_r varies also. Since v_c is always 90° out of phase with v_r, the locus of points traced by the tip of v_r is a semicircle. The output voltage as shown in the diagram has a phase angle of ϕ, which is obviously

$$\phi = 2\theta = 2\tan^{-1}(1/\omega C r_v) \quad \ldots\ldots\ldots \quad (1)$$

A plot of phase shift as a function of r_v is shown in Fig. 5. A power-series expansion of $\tan^{-1}(1/\omega C r_v)$ allows equation (1) to be written as

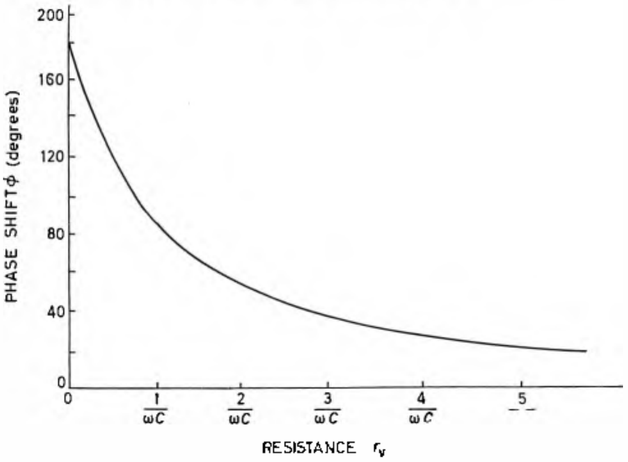
$$\phi = 2\left[\frac{1}{2\pi f C r_v} - 1/3\left(\frac{1}{2\pi f C r_v}\right)^3 + 1/5\left(\frac{1}{2\pi f C r_v}\right)^5 - \ldots\ldots\right]$$

If $\dfrac{1}{2\pi f C r_v}$ is somewhat smaller than unity,

$$\phi = \frac{1}{\pi f C r_v} \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

Equation (2) indicates values of ϕ that are in error by less

Fig. 5 Phase shift as a function of r_v



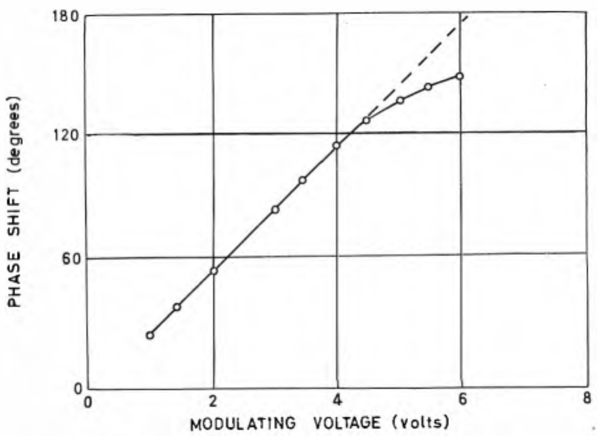


Fig. 6 Phase shift as a function of modulating voltage

than 10% if ϕ lies between zero and 1·1 rad. The phase shift between input and output signals varies inversely with r_v over this range. From Fig. 4, it is seen that the amplitude of the output signal remains constant, independent of ϕ.

To convert the phase-shift circuit to a phase modulator, a silicon-carbide voltage-variable resistor (v.v.r.) is used for the resistor r_v. The small-signal resistance of r_v depends on the direct voltage applied to the v.v.r.[3]. This dynamic resistance is approximately

$$r_v = K/V \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

Equation (3) indicates the resistance presented to the carrier frequency by the v.v.r. if the amplitude of this signal is small. The quiescent phase shift of the carrier signal will be determined by the resistance r_v, which is determined by the direct voltage across the v.v.r. Since the inductance (or transformer) offers a short-circuit to d.c. and low-frequency signals, the voltage across the v.v.r. is equal to the collector voltage. The quiescent phase shift is then (assuming the amplitude of the carrier signal is small enough not to affect r_v)

$$\phi_0 = \frac{1}{\pi f C r_v} = \frac{V_{CQ}}{\pi f C K} \quad \ldots\ldots\ldots\ldots \quad (4)$$

where V_{CQ} is the quiescent collector voltage. Applying the modulating voltage causes the collector voltage to vary as

$$V_C = V_{CQ} - V_m \sin \omega_m t$$

where V_m is the amplitude and ω_m is the radian frequency of the modulating voltage. The phase is then

$$\phi = \frac{V_{CQ}}{\pi f C K} - \frac{V_m \sin \omega_m t}{\pi f C K} = \phi_0 - \frac{V_m \sin \omega_m t}{\pi f C K} \quad \ldots\ldots \quad (5)$$

Equation (5) indicates that the phase shift is a linear function of the amplitude of the modulating signal subject to the limitation imposed on equation (2). In the circuit, errors associated with equations (2) and (3) offset each other, and the resulting linearity of phase shift extends over a larger range than expected. Fig. 6 shows a plot of the measured phase shift of a 13kHz carrier frequency as a function of a modulating direct voltage. The phase shift is linear to within 4° over the range $\phi = 30°$ to 130°. If the circuit is biased to give a quiescent phase shift of 80°, a linear phase deviation of $\pm 50°$ can be obtained. Furthermore, several stages can be cascaded and modulated by a common source to give larger values of phase deviation.

The operation of the linear phase modulator is based on the fact that the nonlinear resistor (v.v.r.) offsets the

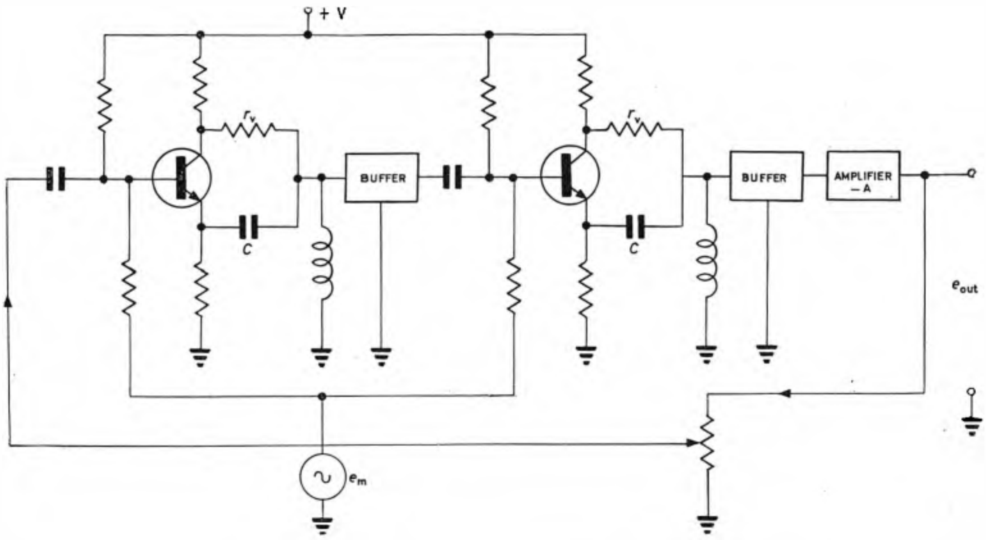


Fig. 7 Frequency modulator

nonlinearities in the phase-shift/resistance curve, giving a linear variation of phase shift with modulating voltage.

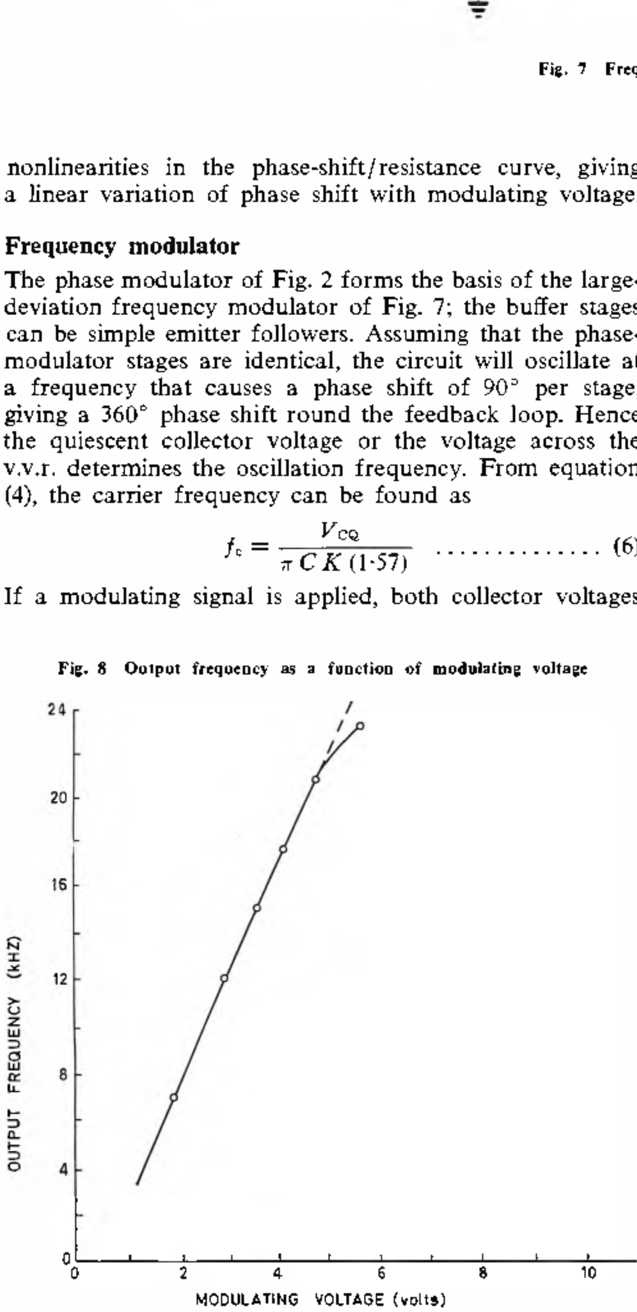
Frequency modulator

The phase modulator of Fig. 2 forms the basis of the large-deviation frequency modulator of Fig. 7; the buffer stages can be simple emitter followers. Assuming that the phase-modulator stages are identical, the circuit will oscillate at a frequency that causes a phase shift of 90° per stage, giving a 360° phase shift round the feedback loop. Hence the quiescent collector voltage or the voltage across the v.v.r. determines the oscillation frequency. From equation (4), the carrier frequency can be found as

$$f_c = \frac{V_{CQ}}{\pi C K (1{\cdot}57)} \quad \ldots\ldots\ldots\ldots\ldots (6)$$

If a modulating signal is applied, both collector voltages

Fig. 8 Output frequency as a function of modulating voltage



will be

$$V_C = V_{CQ} - V_m \sin \omega_m t$$

The oscillation frequency must change as V_C changes, keeping a 90° phase shift for each stage. Since the gain of each stage is frequency-independent, the amplitude of the output voltage remains constant:

$$f = \frac{V_{CQ}}{\pi C K (1{\cdot}57)} - \frac{V_m \sin \omega_m t}{\pi C K (1{\cdot}57)} = f_c - \frac{V_m \sin \omega_m t}{\pi C K (1{\cdot}57)} \quad \ldots\ldots\ldots\ldots\ldots (7)$$

Fig. 8 shows a plot of frequency as a function of a direct modulating voltage. The frequency change with voltage is quite linear from 5 to 21kHz. Choosing a carrier of 13kHz will allow a frequency deviation of $\pm$8kHz, controlled by the modulating signal. The amplitude of the output voltage is constant, as expected.

Both the phase and frequency modulator have excellent dynamic response. The response time is limited by the time constant of the v.v.r., which is normally less than a microsecond[4]. For control applications, the response times of the modulators are negligible.

Conclusions

The modulators described here can be quite useful in control-system applications, for the following reasons:

1. The phase and frequency deviations are quite large and vary linearly with the modulating signal.
2. The amplitude of the output is independent of frequency or phase.
3. The response times are very low.
4. The circuits are relatively simple and inexpensive.

Control systems that might use these modulators are hybrid systems and systems requiring phase or frequency correction.

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Some Novel *RC* Oscillators for Radio Frequencies

(Part 1)

by R. S. Sidorowicz*, Dipl. Ing., D.I.C., A.M.I.E.E.

Various RC oscillators are briefly reviewed and it is shown that they have some disadvantages when used at radio frequencies. An oscillator which avoids most of these drawbacks is suggested. This requires a special active phase-shift circuit of the all-pass type, in which the phase-shifting capacitance or resistance can be earthed. Such a circuit is proposed and analyzed in detail. The circuit is used in the design of RC oscillators based on silicon n-p-n transistors. One of the oscillators requires three transistors and is tuned by means of a 2-gang capacitor which can be earthed. Frequencies up to 5MHz can be generated easily in this circuit, and even higher frequencies can be obtained without difficulty. The second oscillator consists of two active shift stages and requires five transistors. If the oscillator is tuned by a 2-gang capacitor, it produces two output signals which are in quadrature. When only one output is required, the circuit can be tuned by a single variable capacitor. This oscillator can generate frequencies of up to (and above) 5MHz. An additional oscillator with two active phase-shift stages is also briefly described; in this circuit the resistive parts of frequency-determining networks are earthed.

(Voir page 533 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 534)

MOST of the known oscillators with *RC* frequency-determining networks can be divided into two distinct groups: (a) single-phase oscillators employing a passive network and a single amplifier unit, and (b) polyphase oscillators with several active stages. The main representatives of group (a) are:

(i) multi-element *RC* or *CR* phase-shift oscillators

(ii) Wien-bridge oscillators

(iii) twin-*T* or bridged-*T* oscillators.

In oscillators of group (b), several output signals with prescribed phase differences between them are required. When an oscillator of this type is tuned, the phase relationships must be maintained accurately; this is achieved by employing a number of active stages which separate the individual sections of a *CR* or *RC* network. The following circuits are typical:

(i) oscillators based on active all-pass circuits

(ii) oscillators with buffers between the sections of a frequency-determining network.

It appears that, owing to their comparative simplicity, the oscillators of group (a) are much more widely used than those of group (b), the use of the latter being confined to rather specialized branches of electronics and communications. However, from several points of view, the oscillators of group (a) have a number of drawbacks, and those of group (b) have some advantages. It was felt, therefore, that an oscillator which could usefully combine the best features of both groups should be well worth while. An attempt was made to develop some such oscillators and the results achieved will be described.

Before presenting the circuits, an outline of the theoretical considerations leading to their development will be given.

Characteristics of *RC* oscillators

In the following paragraphs, some of the salient features of the oscillators of both groups will be discussed and compared. Those readers who may be interested in various details of the oscillator circuits are referred to the Bibliography given at the end of the article.

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A block schematic of one of the best known oscillators of group (a) is shown in Fig. 1. The circuit consists of a 3-section *RC* network and an inverting amplifier which may comprise not only an amplifying stage but also an emitter follower (or cathode follower). The circuit oscillates at the frequency at which the *RC* network produces a phase shift of 180°, provided that the amplifier gain A_0 is large enough to counteract the attenuation of the network at that frequency. The oscillator can be tuned by simultaneously varying three of its frequency-determining elements which are of the same kind. In the circuit of Fig. 1, the tuning is normally done by means of a 3-gang capacitor, the three sections of which should be tracked quite accurately. It is not possible to tune the circuit (except incrementally) by varying only one or two elements, since the gain required for oscillation will vary. The alternative oscillator with a *CR* network is also well known; here the tuning is done by a 3-gang potentiometer.

One important advantage of these phase-shift oscillators is that their tuning elements are connected to a common point and can be earthed. This means that stray capacitances of the elements are added in parallel to them and their effect on the performance of the oscillators can be predicted. They are normally used for subaudio and audio frequencies; they appear to be rather unsuitable for radio frequencies, since their frequency-determining elements become impracticably small.

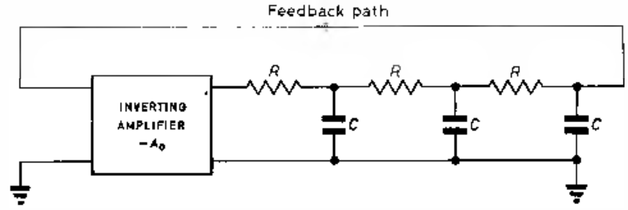


Fig. 1 Phase-shift oscillator with a 3-stage *RC* network

A Wien-bridge oscillator is shown in Fig. 2. The circuit uses a series-parallel *RC* network which produces a zero phase shift at the frequency of oscillation. The gain required to compensate for the attenuation of the network is provided by a noninverting amplifier. The oscillator is tuned by varying simultaneously either two network capa-

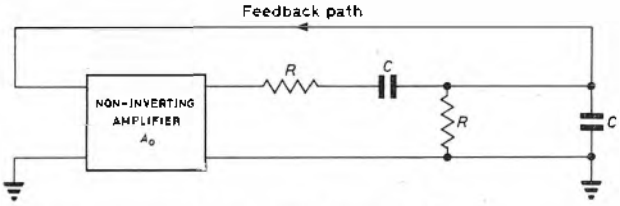


Fig. 2 Wien bridge oscillator

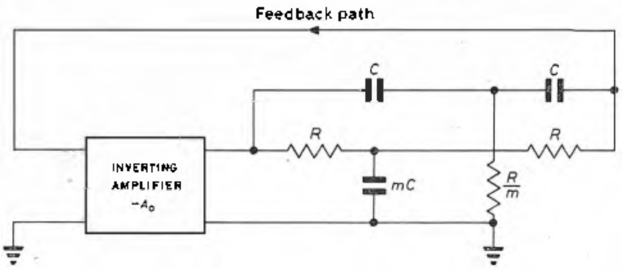


Fig. 3 Oscillator with a twin-T network

citors or two resistors. It is seen that only one of the tuning elements can be earthed, while the other must remain 'hot'. The physical size of a tuning element is such that its stray capacitance to earth may amount to 15-30pF. This capacitance will restrict the upper frequency limit of the oscillator to, say, about 0·5MHz.

An oscillator based on a twin-T network is shown in Fig. 3. It employs an inverting amplifier of fairly high gain and can be tuned by simultaneously varying either both of its 'upper' capacitors or both resistors; neither of the tuning elements can be earthed.

A polyphase oscillator with three outputs can be arranged as shown in Fig. 4 or 5. In the circuit of Fig. 4, the oscillation occurs at a frequency at which the active network consisting of three *CR* stages and separating emitter followers produces an overall phase shift of 180°. The gain required for oscillation is fairly low, and the circuit can be tuned by simultaneously varying either its three resistors or three capacitors. It is not possible to tune it by changing only one or two of its frequency-determining elements. An oscillator with *RC* stages between emitter followers is also possible; the minimum number of stages in this type of oscillator is three. It is seen that in such oscillators all the tuning elements can be earthed.

The circuit of Fig. 5 consists of three inverting amplifiers of modest gain and three simple *CR* networks; an alternative arrangement with three *RC* networks is also possible. If more than three phases are required, the number of stages can be increased, but such circuits have few practical applications. Again, the oscillator of Fig. 5 can be tuned by means of either a 3-gang capacitor or a 3-gang potentiometer. Normally, the second method of tuning is used so that all the tuning elements can be earthed. Oscillators of the types shown in Figs. 4 and 5 can be used at frequencies up to 10-15MHz. However, they have the disadvantage that the number of tuning elements is at least three.

Another type of polyphase oscillator is illustrated in the block schematic of Fig. 6. This consists of two all-pass phase-shift stages based on electronic phase splitters, and an inverting amplifier. The oscillator provides two quadrature voltages at the cathodes of the two splitter valves; two additional voltages which are inverted relative to the cathode outputs can be derived from the anodes of the valves. The oscillator is thus effectively a 4-phase device, although it is usually regarded as a 2-phase circuit. By introducing additional all-pass stages, the number of phases generated by the oscillator can be increased.

The oscillator of Fig. 6 is very unusual, in that the gain required for oscillation is independent of the values of individual capacitors or resistors. This is due to the fact that the all-pass circuits have a constant-amplitude response over a wide frequency range. Hence the oscillator can be tuned by varying only one frequency-determining element; it is equally possible to tune it by simultaneously changing all the elements. When it is tuned by more than one element, no tracking difficulties arise; on the other hand, it is seen that none of the frequency-determining elements can be earthed. The oscillator is therefore usually employed only at audio and subaudio frequencies and is not suitable for generating radio-frequency signals.

The above discussion is summarized in Table 1, where various relevant characteristics of the oscillators are compared. The features which are regarded as important from the point of view of this article are emphasised.

Two additional oscillators should perhaps be mentioned. One of these is described by Owens[1], and it is tuned by a 2-gang capacitor whose two sections are earthed. This circuit was primarily designed for audio frequencies and its usefulness at radio frequencies is not immediately obvious. The second oscillator was recently described by Pasupathy[2]. This is essentially a modification of the

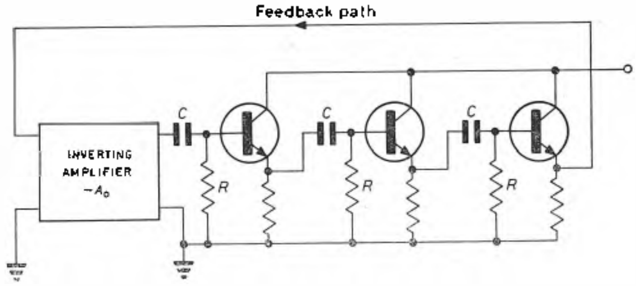


Fig. 4 Three-phase oscillator with emitter followers

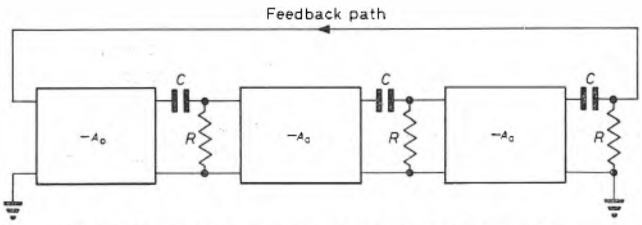


Fig. 5 Three-phase oscillator with three identical amplifiers

Fig. 6 Two-phase oscillator with all-pass circuits

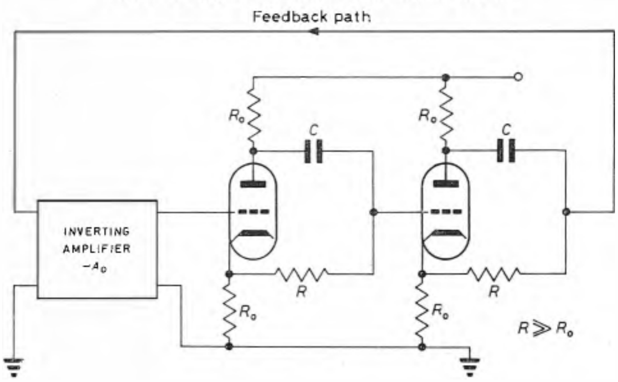


TABLE 1

TYPE OF OSCILLATOR	MINIMUM NUMBER OF TUNING ELEMENTS	GAIN REQUIRED	POSITION OF TUNING ELEMENTS RELATIVE TO EARTH	APPROXIMATE UPPER FREQUENCY LIMIT, MHZ
Phase-shift *RC* or *CR*	3	High	Can be earthed	1
Wien-bridge or T-network	2	Medium	At least one element cannot be earthed	0.5
Phase-shift *RC* or *CR* with buffer stages	3	Medium or low	Can be earthed	10
Oscillator with *RC* or *CR* all-pass circuits	1	Low; independent of tuning elements	Cannot be earthed	0.5

Wien-bridge circuit, in which the tuning capacitors can be earthed. Again, it appears that this circuit is primarily intended for low frequencies.

Requirements of the new *RC* oscillators

It is seen from Table 1 that the *RC* oscillators which are of interest in this work should possess the following features:

(i) satisfactory operation at radio frequencies

(ii) tuning elements earthed

(iii) gain required for oscillation fairly low and independent of the values of frequency-determining elements

(iv) tuning by varying one, or not more than two, of the frequency-determining elements.

The second requirement is an almost natural consequence of the first, and the fourth requirement is automatically satisfied if the third one is met. In addition, the first requirement necessitates the achievement of operation at radio frequencies without using using prohibitively small *R* or *C*. These must be such that the load imposed by them on the amplifier used in the oscillator does not reduce its gain below the figure required for oscillation. Secondly, the minimum practical values of the capacitors are dictated by the input capacitances of the active devices used in a circuit and by the minimum settings of tuning capacitors.

It may be added that it appears that, in oscillators tuned by means of two or more ganged resistors or capacitors, the frequency of oscillation is given by

$$f_0 = d/2\pi RC \quad \ldots\ldots\ldots\ldots\ldots\ldots (1)$$

where d is a constant. For the Wien-bridge circuit of Fig. 2, the twin-T oscillator of Fig. 3 and the 2-phase system of Fig. 6, this constant is unity. On the other hand, oscillators of the type shown in Figs. 1, 4 and 5 have $d > 1$ if they are based on *RC* stages (as in Fig. 1) and $d < 1$ if *CR* stages are used. This suggests that it should be preferable to employ *RC* rather than *CR* frequency-determining networks at radio frequencies.

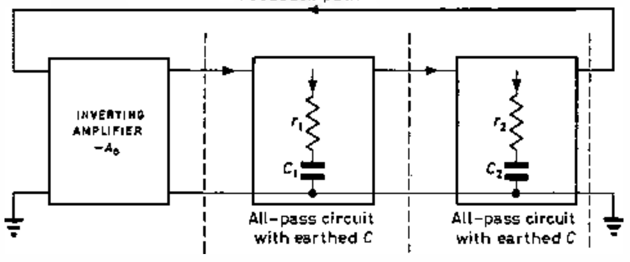


Fig. 7 Proposed oscillator circuit

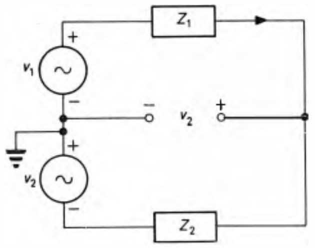


Fig. 8 Usual all-pass phase-shift circuit

Table 1 shows that none of the oscillators discussed meets all the above requirements simultaneously. However, it appears that this can be achieved comparatively simply by adopting the system illustrated in Fig. 7. This consists of an inverting amplifier and two phase-shift stages of the all-pass type, the latter being unusual in that capacitors should be earthed.

Development of a special active all-pass circuit

An all-pass phase-shift circuit is usually in the form of the network shown in Fig. 8. The two input voltages required by the circuit can be derived either from the secondary of a centre-tapped transformer or an electronic phase splitter (such as those in Fig. 6). If the circuit is to be an all-pass network, Z_1 and Z_2 should meet certain requirements. However, only the simplest case, when one of the impedances is a pure resistance and the other is a pure capacitive reactance, will be considered here. The voltage transfer function of the circuit is given by[3]

$$v_2/v_1 = \frac{Z_2 - Z_1}{Z_2 + Z_1} = \frac{1 - (Z_1/Z_2)}{1 + (Z_1/Z_2)} \quad \ldots\ldots\ldots (2)$$

If $Z_1 = 1/j\omega C$ and $Z_2 = r$, equation (2) becomes

$$v_2/v_1 = \frac{1 - 1/j\omega Cr}{1 + 1/j\omega Cr} = \frac{j\omega Cr - 1}{j\omega Cr + 1}$$

$$= -\frac{1 - j\omega Cr}{1 + j\omega Cr} = -e^{-j2\phi} \quad \ldots\ldots\ldots\ldots (3)$$

where $\tan\phi = \omega Cr$.

On the other hand, if $Z_1 = r$ and $Z_2 = 1/j\omega C$, equation (2) yields

$$v_2/v_1 = \frac{1 - j\omega Cr}{1 + j\omega Cr} = e^{-j2\phi} \quad \dots\dots\dots\dots (4)$$

where, again, $\tan\phi = \omega Cr$.

Equation (4) can be rearranged as follows:

$$v_2/v_1 = \frac{1 - j\omega Cr}{1 + j\omega Cr} = \frac{(1 + j\omega Cr) - 2j\omega Cr}{1 + j\omega Cr}$$

$$= 1 - \frac{2j\omega Cr}{1 + j\omega Cr} \quad \dots\dots\dots\dots\dots\dots (5)$$

so that

$$v_2 = v_1 - \frac{2v_1 j\omega Cr}{1 + j\omega Cr} = v_1 - 2v_{1x} \dots\dots\dots\dots (6)$$

Alternatively, equation (4) can be rewritten as

$$v_2/v_1 = -\frac{j\omega Cr - 1}{j\omega Cr + 1} = -\frac{(1 + j\omega Cr) - 2}{1 + j\omega Cr}$$

$$= -1 + \frac{2}{1 + j\omega Cr} \dots\dots\dots\dots\dots\dots (7)$$

so that

$$v_2 = -v_1 + \frac{2v_1}{1 + j\omega Cr} = -v_1 + 2v_{1y} \quad \dots\dots (8)$$

Examination of equations (6) and (8) shows that the transfer function given by equation (4) can be realized by adopting either the system of Fig. 9 or that of Fig. 10. These arrangements have the advantage that one of their phase-shifting elements (either R or C) is earthed. On the other hand, they appear to be rather complicated, but this complexity is more apparent than real.

It will be shown now that the system of Fig. 10 can be realized by the circuit of Fig. 11, provided that the resistances R_1, R_2, R_{k1} and R_{k2} are suitably chosen. The circuit is based on two triodes, but it is equally possible to employ two n-channel field-effect transistors. Operation of the circuit can be explained approximately as follows.

When a signal v_{g1} is applied to the input of the circuit, it produces a voltage v_{a1} at the anode of V_1, which is inverted relative to v_{g1}; an inverted signal is also produced across R_2. By choosing suitable values of R_1 and R_2 it is possible to obtain $v_{a1} = -2v_{g1}$, and the voltage drop across R_2 can be made equal to $-v_{g1}$. Now the grid of V_2 receives a signal

$$-2v_{1y} = v_{a1}/(1 + j\omega Cr)$$

This signal is then inverted by V_2 and added to the voltage developed across R_2. The resultant output signal in R_2 is

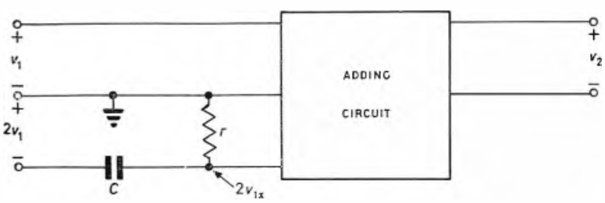
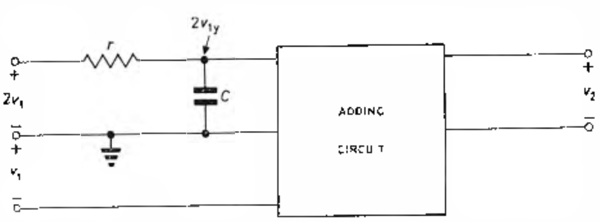


Fig. 9 All-pass circuit realizing equation (6)

Fig. 10 All-pass circuit realizing equation (8)



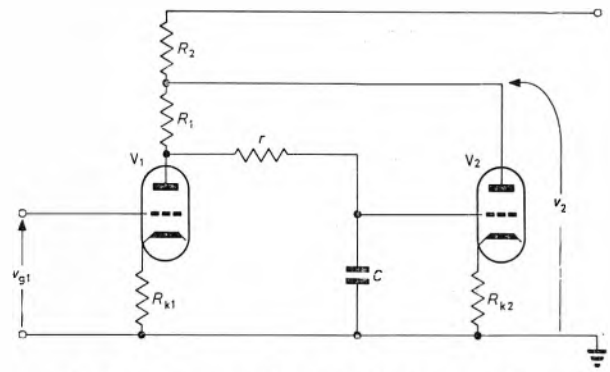


Fig. 11 All-pass phase-shift stage with two active devices (biasing details are not shown)

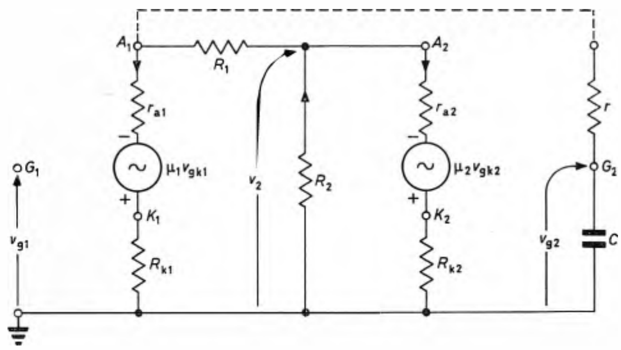


Fig. 12 Small-signal equivalent circuit of the phase shifter of Fig. 11

thus a combination of an inverted voltage and a phase-shifted signal.

In order to determine the values of R_1, R_2, R_{k1} and R_{k2} to ensure that the circuit of Fig. 11 will operate as an all-pass device, it is necessary to analyse it in detail.

A small-signal equivalent circuit of the phase-shift stage of Fig. 11 is illustrated in Fig. 12. It is seen that each valve is represented by an equivalent voltage generator and that all the interelectrode capacitances are neglected The equivalent circuit will be valid if field-effect transistors are used instead of valves, provided that their internal capacitances can be disregarded.

For the purpose of analysis, it will be assumed that $r \gg R_1$ and $r \gg R_2$. Basic equations for the circuit of Fig. 12 can be written as

$$\mu_1 v_{gk1} = \mu_1(v_{g1} - i_1 R_{k1}) = i_1(R_1 + r_{a1} + R_{k1}) + (i_1 + i_2) R_2 \quad \dots\dots\dots\dots (9)$$

$$\mu_2 v_{gk2} = \mu_2(v_{g2} - i_2 R_{k2}) = i_2(r_{a2} + R_{k2}) + (i_1 + i_2) R_2 \quad \dots\dots\dots\dots (10)$$

which can be rearranged as

$$\mu_1 v_{g1} = i_1 [R_1 + r_{a1} + R_2 + (1 + \mu_1) R_{k1}] + i_2 R_2 \quad \dots\dots (11)$$

$$\mu_2 v_{g2} = i_1 R_2 + i_2 [r_{a2} + R_2 + (1 + \mu_2) R_{k2}] \quad \dots\dots (12)$$

However, v_{g2} is not an independent quantity. It is given by

$$v_{g2} = v_{a1}/(1 + j\omega Cr) = v_{a1}\hat{b} \quad \dots\dots\dots\dots (13)$$

where v_{a1} is the voltage at the anode of V_1 and $\hat{b} = 1/(1 + j\omega Cr) = 1/(1 + jX)$.

Since

$$v_{a1} = -[i_1 R_1 + (i_1 + i_2) R_2] = -[i_1(R_1 + R_2) + i_2 R_2] \quad \dots\dots (14)$$

the grid voltage of V_2 is given by

$$v_{g2} = -\hat{b}\,[i_1(R_1 + R_2) + i_2 R_2]$$

Equation (12) can be written as

$$-\mu_2 \hat{b}\,[i_1(R_1+R_2)+i_2R_2] = i_1R_2+i_2\,[r_{a2}+R_2+(1+\mu_2)R_{k2}]$$

or

$$0 = i_1\,[R_2+\mu_2\hat{b}\,(R_1+R_2)]+i_2[r_{a2}+R_2(1+\mu_2)R_{k2}+\mu_2\hat{b}\,R_2] \quad \ldots\ldots\ldots\ldots (15)$$

Equations (11) and (15) can now be written in an abbreviated form as

$$\mu_1 v_{g1} = i_1R_A + i_2R_2 \quad \ldots\ldots\ldots\ldots (16)$$

$$0 = i_1\,[R_2 + \hat{b}_2\,R_B] + i_2\,[\hat{b}_2\,R_2 + R_C] \quad \ldots\ldots (17)$$

where $R_A = R_1 + r_{a1} + R_2 + (1 + \mu_1)\,R_{k1}$

$R_B = R_1 + R_2$

$R_C = r_{a2} + R_2 + (1 + \mu_2)\,R_{k2}$

$\hat{b}_2 = \mu_2\,\hat{b}$

Output voltage of the circuit is

$$v_2 = -R_2\,(i_1 + i_2) \quad \ldots\ldots\ldots\ldots (18)$$

Expressions for i_1 and i_2 as functions of v_{g1} are found by solving equations (16) and (17). If these are then substituted into equation (18), the following formula is obtained:

$$= (\mu_1R_D/R_F)\,\frac{(B_1 - jX)}{(B_2 + jX)} = A_1\,\frac{(B_1 - jX)}{(B_2 + jX)} \quad \ldots\ldots\ldots (21)$$

where

$A_1 = \mu_1R_D/R_F$

$B_1 = \mu_2R_1/R_D - 1$

$B_2 = 1 + \mu_2R_E/R_F$

It can be seen by examining the explicit expressions for the constants B_1 and B_2 of equation (21) that it is possible to obtain $B_1 = 1$; but B_2 must be greater than unity. However, the circuit of Fig. 11 will behave as an all-pass device even if B_1 and B_2 are not unity, provided that $B_1 = B_2 = B$. This condition is satisfied if the circuit resistances are chosen so that

$$\mu_2R_1/R_D = B + 1 \quad \ldots\ldots\ldots\ldots (22)$$

$$\mu_2R_E/R_F = B - 1 \quad \ldots\ldots\ldots\ldots (23)$$

Only two of the four resistances of the circuit of Fig. 11 can be determined from equations (22) and (23). The remaining two resistances can be chosen on a different basis to the all-pass requirement, e.g. choice of a suitable biasing point or the required frequency bandwidth.

TABLE 2

B	1·5	2·0	3·0	4·0
R_2, kΩ	0·42	0·71	1·1	1·35
R_{k2}, kΩ	0·59	0·46	0·3	0·2
A_1	0·33	0·55	0·82	0·98

$$v_2 = -\mu_1 v_{g1}\,\frac{R_2\,[R_C - R_2 + \hat{b}_2\,(R_2 - R_B)]}{R_AR_C - R_2^2 + \hat{b}_2\,(R_A - R_B)R_2} \quad \ldots\ldots (19)$$

If the denominator of equation (19) is divided by R_2 and, in addition, the following notation is adopted:

$R_D = R_C - R_2 = r_{a2} + (1 + \mu_2)\,R_{k2}$

$R_E = R_A - R_B = r_{a1} + (1 + \mu_1)\,R_{k1}$

$R_F = \dfrac{R_AR_C}{R_2} - R_2 = (R_1 + R_E)\,(1 + R_D/R_2) + R_D$

the voltage transfer function of the circuit can be written as

$$v_2/v_{g1} = \mathrm{f}(\omega) = -\mu_1\,\frac{R_D - \hat{b}_2R_1}{R_F + \hat{b}_2R_E} \quad \ldots\ldots (20)$$

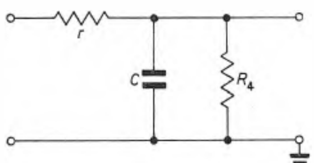


Fig. 13 *RC* **network with a parallel load resistance**

If the expression for $\hat{b}_2$ is now substituted into equation (20), this becomes

$$\mathrm{f}(\omega) = -\mu_1\,\frac{R_D - \dfrac{\mu_2R_1}{1 + jX}}{R_F + \dfrac{\mu_2R_E}{1 + jX}} = -\,\frac{\mu_1(R_D + R_DjX - \mu_2R_1)}{R_F + R_FjX + \mu_2R_E}$$

$$= \mu_1\,\frac{[\mu_2R_1 - R_D - jR_DX]}{\mu_2R_E + R_F + jR_FX} = \frac{\mu_1R_D\{[(\mu_2R_1/R_D) - 1] - jX\}}{R_F\{[(\mu_2R_E/R_F) + 1] + jX\}} =$$

If it is assumed that the circuit is based on two identical valves (or field-effect transistors) having $r_{a1} = r_{a2} = 10\text{k}\Omega$, $g_{m1} = g_{m2} = 5\text{mA/V}$ and $\mu_1 = \mu_2 = 50$, and if $R_1 = 2\text{k}\Omega$ and $R_{k1} = 1{\cdot}0\text{k}\Omega$, R_2 and R_{k2} can readily be calculated. Some of these are shown in Table 2 (for four different values of B). The Table also gives values of the circuit gain parameter A_1.

It may be worth mentioning that the all-pass condition can be satisfied by the circuit of Fig. 11, even if its capacitor C is connected in parallel with a resistor R_4 whose value is comparable with that of the series resistance r. However, in this case it is necessary to modify the values of R_2 and R_{k2}. Modified transfer function of the circuit can be determined by considering the network of Fig. 13. The function $\hat{b}_2$ for equation (20) is written as follows:

$$\hat{b}_2 = \mu_2\,\frac{R_4/(1 + j\omega CR_4)}{r + R_4/(1 + j\omega CR_4)} = \frac{\mu_2R_4}{R_4 + r\,(1 + j\omega CR_4)} =$$

$$= \mu_2\,\frac{1}{1 + j\omega Cr + r/R_4} = \frac{\mu_2}{a + jX} \quad \ldots\ldots\ldots\ldots (24)$$

where $a = 1 + r/R_4$.

The transfer function is again given by

$$v_2/v_{g1} = \mathrm{f}(\omega) = A_1\,\frac{(B_1 - jX)}{(B_2 + jX)} \quad \ldots\ldots\ldots (25)$$

where

A_1 has the same value as in equation (21), but

$$B_1 = (\mu_2R_1/R_D) - a \quad \ldots\ldots\ldots\ldots (26a)$$

$$B_2 = (\mu_2R_E/R_F) + a \quad \ldots\ldots\ldots\ldots (26b)$$

(To be continued)

Digitizing-Pulse Generator

by S. C. Acharyya, Ph.D., B.E.E.

The article deals briefly with the design considerations for an electronic pulse generator that generates calibrating pulses for digitizing error pulses at varying frequencies. The analysis of a direct-voltage-controlled oscillator circuit that constitutes the essential part of the pulse generator is presented to facilitate analytical approach to the oscillator design.

(Voir page 533 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 534)

THE application of digital techniques to the industrial problems of metering and control often requires the generation of digitizing pulses. These pulses may be produced by sensing movement which may be rotational, reciprocating or oscillating. The sensing may be either by photoelectric devices or by toothed wheel and magnetic pickups. However, a problem arises when the digitizing pulses are required at a frequency that is many times the pulse frequency that can be obtained from the movement by direct means, e.g. stepping up the speed of the movement by appropriate gears and then sensing. Although, theoretically, any speed of sensing may be realized by using combinations of gear trains, large disks with peripheral pinholes and photoelectric devices, practical design considerations impose a speed limit, particularly in the higher-speed range.

The problem can be dealt with by electronic means, and the design considerations of an electronic circuit that produces pulse-repetition frequencies (p.r.f.s.) much higher than those produced by direct means are discussed below.

Functional Requirements

Normally, a digitizing-pulse generator is necessary in a digital control system where an error measured in terms of the number of digit pulses causes a proportional correction signal for application to a servomotor in the control system. The function of the generator is to produce digitizing pulses to calibrate an error pulse ΔT which has been generated by appropriate electronic devices to represent an error of Δl over L (inches) of a movement at a linear velocity of V (ft/min).

If the movement is a one-speed device, the digitizing pulses (commonly known as clock pulses) may be obtained from an oscillator; but when the speed of the movement varies over a wide range, the width of the error pulse ΔT generated for the measurement of the same error of length Δl will vary according to the equation

$$\Delta T = \frac{\Delta l \times 60}{V \times 12} \text{ seconds} \quad \dots\dots\dots\dots \quad (1)$$

The equation shows that the error pulse ΔT varies inversely with the velocity of the movement and follows the curve shown in Fig. 1. In order to calibrate the error pulse at different movement speeds, the changes in the digitizing-pulse periods have also to follow the same curve. If a high-resolution error measurement is required, the pulsewidth ΔT may be of the order of a few tens of microseconds, while the pulses that may be generated from the movement by direct means may be of the order of some tens of milliseconds. In such cases, the generator will provide the necessary transformation of the p.r.f. from the lower to the required higher value.

Principle of Operation

The principle of operation of the generator consists in generating a direct voltage proportional to the movement and using this voltage to control an oscillator circuit for generating the necessary digitizing pulses, as shown in Fig. 2. A photoelectric device or a toothed-wheel and magnetic-head technique is used to generate pulses at a low p.r.f. corresponding to the velocity of the controlled element. Since these pulses may vary in width and amplitude with speed, they are converted into constant-amplitude and -width pulses by a monostable circuit, the mark period of which is smaller than the smallest period of the pulses, corresponding to the highest speed of the movement. A linear frequency-to-voltage convertor converts the output of the monostable circuit into a frequency-proportional direct voltage. This voltage is then used to control the p.r.f. of an oscillator to produce the digitizing pulses. Since the p.r.f. of the output of the specially designed oscillator varies inversely with the applied voltage, the output of the frequency-to-voltage convertor is inverted by an amplifier with a gain of -1, and this inverted output voltage is applied to the oscillator.

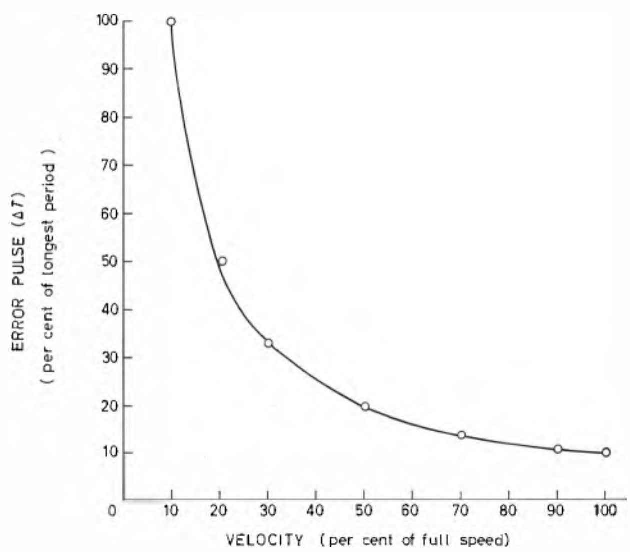
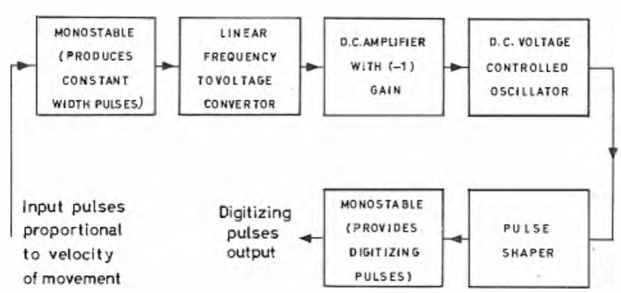


Fig. 1 Error-pulse/velocity relationship

Fig. 2 Functional diagram of digitizing-pulse generator



Design Considerations

MONOSTABLE CIRCUIT

This is a conventional multivibrator circuit with one stable state. When triggered by a pulse at its input, the circuit goes into a quasistable state, at the end of which it returns to its stable state. The quasistable period is controlled by an RC timing circuit and is designed to be always less than the periods of the input triggering pulses, so that the monostable circuit can always return to its stable state between any two successive pulses.

FREQUENCY-TO-VOLTAGE CONVERTOR

The pulse train from the monostable circuit is applied to the input terminals of a linear frequency-to-voltage convertor, as shown in Fig. 3. The principle of operation and design of such a convertor have been discussed in detail in Reference 1. Basically, the circuit operates as follows:

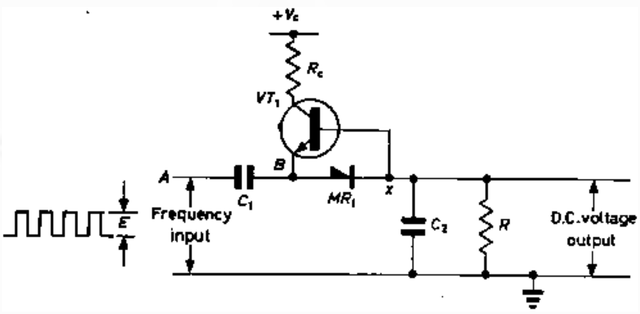


Fig. 3 Linear frequency-to-voltage convertor

For each positive pulse E appearing at point A, the voltage at X is given by

$$V_X = E\frac{C_1}{C_1 + C_2} \quad \ldots\ldots\ldots\ldots\ldots (2)$$

When the pulse returns to earth, transistor VT_1 becomes forward biased and C_1 discharges through it until point B assumes nearly the same potential as X, neglecting the 'cut-on' voltage of the transistor. Under this condition, charges on C_1 and C_2 balance one another out and any further voltage of a pulse at A will be divided as in equation (2) and will add linearly to the voltage already across C_2, which, in the pulse period, discharges partially into resistor R. The process of building up voltage across C_2 continues until an equilibrium is reached, when, during each cycle, the decrease in voltage across C_2 is equal to the voltage added according to equation (2), and, under this condition, the direct component of the output voltage is given by[1]

$$V_{d.c.} = EC_1Rf \quad \ldots\ldots\ldots\ldots\ldots (3)$$

VOLTAGE-CONTROLLED OSCILLATOR

This is an oscillator circuit specially developed for generating the digitizing pulses fulfilling the requirements discussed in a previous section. In this circuit, shown in Fig. 4, as the supply voltage V_c is switched on, the potential at the emitter of VT_1 rises, owing to the charging of C_T. Depending on the potential V_1 at the base, VT_1 starts conducting as its emitter voltage V_T exceeds that at the base and causes a voltage drop across its collector resistor R_1. When the voltage drop across R_1 is high enough, transistors VT_2 and VT_3 conduct. With conduction of current through VT_3, V_1 drops and this causes an increase in the conductivity of VT_1, so that C_T now rapidly discharges through R_1. As the capacitor discharges to a level V_T (0^+) when VT_1 is cut off, V_T rises again and the same cycle of operation takes place. Each time the capacitor discharges, a negative-going pulse is generated at V_1. The pulse period T_p is given by

$$T_p = C_T R_T \ln \frac{V_c - V_T(0^+)}{\left[V_c - \left(\dfrac{R_{B2}}{R_{B1}+R_{B2}}\right)V_{BB} - V_{eb1}\right]} \quad \ldots\ldots\ldots\ldots (4)$$

The derivation of the equation is shown in the Appendix.

INVERSION OF CONVERTOR OUTPUT VOLTAGE ($V_{d.c.}$)

Examination of equation (4) shows that, as the control voltage V_{BB} increases, the pulse period increases (i.e. the p.r.f. of the oscillator decreases), and the reverse is the case when the control voltage decreases. If the output voltage $V_{d.c.}$ [equation (3)] of the convertor is applied directly as the control voltage to the oscillator, the p.r.f. of the oscillator output tends to decrease with increase in the p.r.f. of the input to the convertor. This is just the opposite of the requirements of the oscillator—that digitizing pulses are produced at a higher p.r.f. with increase in the input p.r.f. to the convertor. This effect is taken into account by interposing a 3-stage high-gain d.c. amplifier, with appropriate feedback, between the output of the frequency-to-voltage convertor and the input control voltage to the oscillator, and obtaining the inverted output of $V_{d.c.}$ for application as control voltage V_{BB} to the oscillator. The d.c. amplifier provides a gain of -1 by appropriate feedback.

PULSE SHAPING

The amplitude of the output pulses at V_1 of the oscillator varies widely with the control voltage V_{BB}, which varies between nearly 2 and 11V. The output of the oscillator is followed by a buffer stage and then amplified by a.c. coupled invertor stages. The output from the final stage of the amplifier is then applied to a monostable circuit, which provides the required digitizing pulses. The mark periods of the output pulses from the monostable circuit are adjusted to a few microseconds.

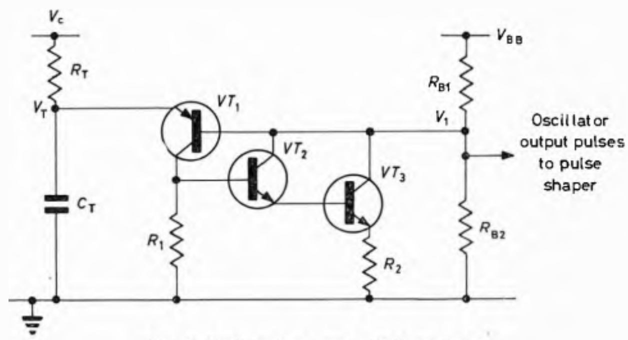


Fig. 4 D.C. voltage-controlled oscillator

Results

The circuit diagram of an oscillator which forms part of a digitizing-pulse generator is shown in Fig. 5. The input pulses to the generator have a p.r.f. varying between 400 and 40 pulse/s and allow the control voltage V_{BB} to vary between approximately 2 and 11V. The input/output characteristics are shown in Fig. 6.

APPENDIX

Analytical Approach to the Oscillator Design

SYMBOLS USED

V_c = supply voltage

V_T = voltage at the junction between the timing resistor R_T and timing capacitor C_T

$V_T(0^+)$ = voltage on the timing capacitor C_T at the end of its discharge period

V_{BB} = control voltage to the oscillator

In the circuit diagram of Fig. 4, as the supply is switched on, the potential V_T at any instant is given by

$$V_T(t) = V_c - i(t)R_T \quad \ldots\ldots\ldots\ldots\ldots (5)$$

$$\text{where } i(t) = C_T \frac{dV_T(t)}{dt}$$

Substituing for $i(t)$ in equation (5),

$$C_T R_T \frac{dV_T(t)}{dt} + V_T(t) = V_c \quad \ldots\ldots\ldots (6)$$

By applying Laplace transformation to equation (6), and after manipulation, the equation may be written as

$$V_T(S) = \frac{V_c}{C_T R_T} \frac{1}{S(S + 1/C_T R_T)} + \frac{V_T(0^+)}{S + 1/C_T R_T} \ldots (7)$$

where $V_T(0^+)$ represents the voltage on C_T at the start of the charging period.

Let $1/R_T C_T$ be represented by τ; equation (7) may be written as

$$V_T(S) = \frac{\tau V_c}{S(S+\tau)} + \frac{V_T(0^+)}{S+\tau} \quad \ldots\ldots\ldots\ldots (8)$$

which, on manipulation, gives

$$V_T(S) = V_c/S + \frac{V_T(0^+) - V_c}{S+\tau} \quad \ldots\ldots\ldots\ldots (9)$$

Taking the inverse Laplace transformation gives $V_T(t)$ as

$$V_T(t) = V_c(1 - e^{-\tau t}) + V_T(0^+)e^{-\tau t} \quad \ldots\ldots (10)$$

which represents the waveform of the voltage at the emitter VT_1 in Fig. 4

Now V_1, the voltage at the base of transistor VT_1, is given by

$$V_1 = V_{BB}\frac{R_{B2}}{R_{B1} + R_{B2}} = K_1 V_{BB} \quad \ldots\ldots\ldots (11)$$

$$\text{where } K_1 = \frac{R_{B2}}{R_{B1} + R_{B2}}$$

When $V_T(t) > V_1$ by V_{eb1} (the emitter-to-base voltage drop), VT_1 starts conducting, and under this condition the equation may be written as

$$V_1 + V_{eb1} = V_T(t) \quad \ldots\ldots\ldots\ldots\ldots (12)$$

i.e.

$$K_1 V_{BB} + V_{eb1} = V_T(t)$$

Substituting for $V_T(t)$ from equation (10), in equation (12) and solving for t gives

$$t = C_T R_T \ln \frac{V_c - V_T(0^+)}{V_c - (K_1 V_{BB} + V_{eb1})} \quad \ldots\ldots (13)$$

$$\text{i.e. } T_p = C_T R_T \ln \frac{V_c - V_T(0^+)}{V_c - [R_{B2}/(R_{B1}+R_{B2})]\,V_{BB} - V_{eb1}} \quad \ldots (14)$$

where T_p represents the pulse period at the end of which C_T discharges through VT_1 to a voltage $V_T(0^+)$, when the transistor is cut off and the capacitor starts charging according to eqn. 14.

Eqn. 14 may be simplified by assuming that

$$V_c \gg V_T(0^+) \text{ and } \left(\frac{R_{B2}}{R_{B1} + R_{B2}}\right) V_{BB} \gg V_{eb1}$$

whence

$$T_p = C_T R_T \ln \frac{V_c}{V_c - [R_{B2}/(R_{B1} + R_{B2})]\,V_{BB}} \quad \ldots\ldots (15)$$

When the control voltage V_{BB} and the ratio $R_{B2}/(R_{B1}+R_{B2})$ are low, V_{eb1} may be comparable with the term $V_{BB}[R_{B2}/(R_{B1}+R_{B2})]$ and cannot be neglected in computations for T_p.

The circuit diagram of an oscillator designed on the basis of the above analysis is shown in Fig. 5.

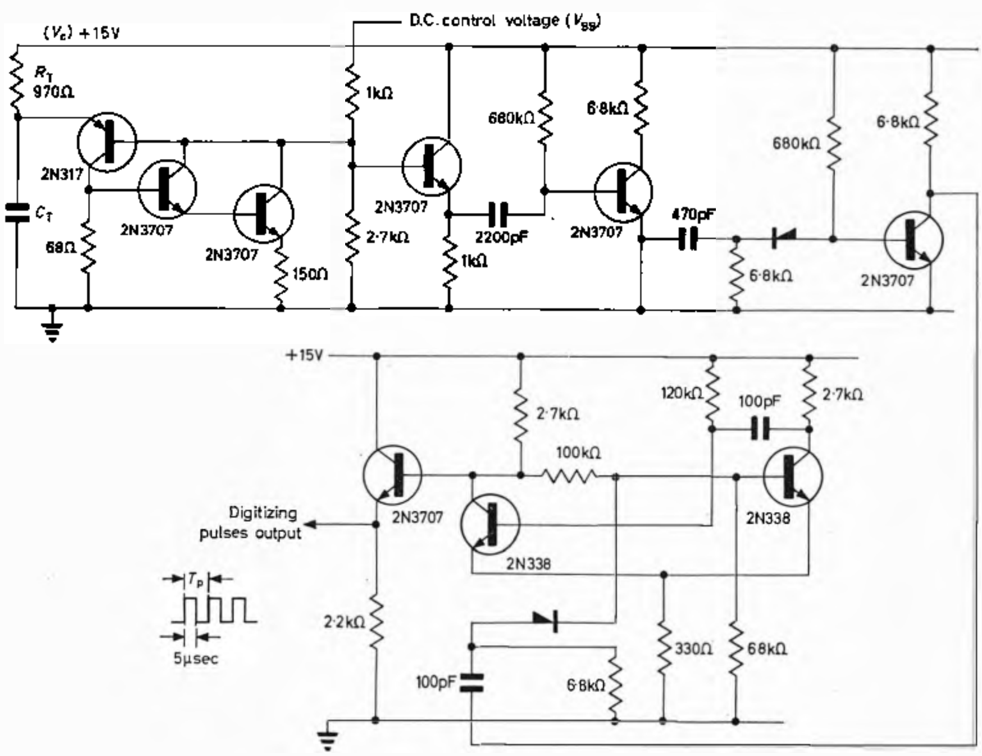


Fig. 5. Oscillator circuit diagram with amplifiers and pulse shaper

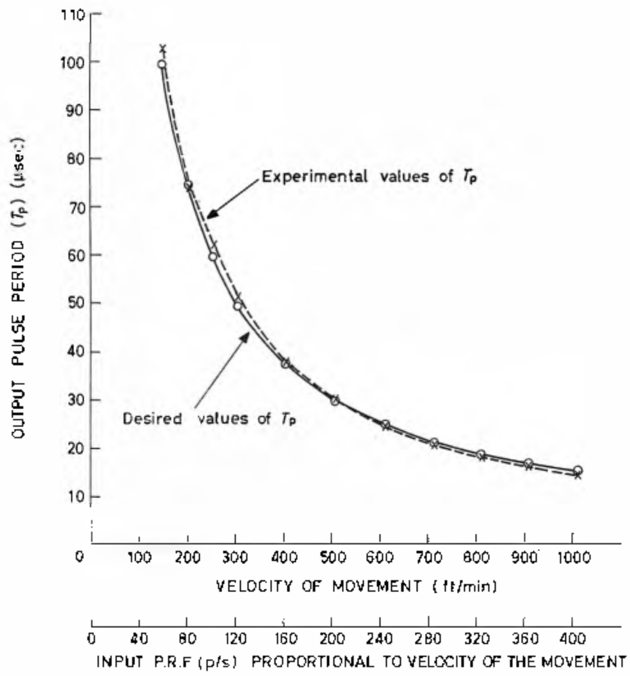


Fig. 6 The pulse generator input/output relationships

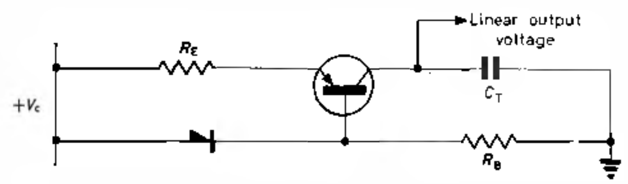


Fig. 7. Constant-current source for linear-ramp generation

EFFECT OF CONSTANT-CURRENT CHARGING OF C_T

It may be of interest to investigate theoretically the effect of charging C_T from a constant-current source (Fig. 4). A circuit that may be used to provide a constant-current source is shown in Fig. 7. In this circuit, a transistor is connected as an emitter follower with its base at a fixed voltage and the collector connected in series with C_T. The voltage $V_T(t)$ across the capacitor may then be applied to the emitter of VT_1 in Fig. 4.

For a capacitor being charged by a constant current I, the equation may be written as

$$I = C_T \frac{dV_T(t)}{dt} \quad \ldots\ldots\ldots\ldots \quad (16)$$

By solving the equation,

$$V_T(t) = (I/C_T)t + V_T(0^+) \quad \ldots\ldots\ldots \quad (17)$$

Substituting this value of $V_T(t)$ in equation, the following equation is obtained:

$$K_1 V_{BB} + V_{eb1} = (I/C_T)t + V_T(0^+) \quad \ldots\ldots \quad (18)$$

Solution of the pulse period T_p from the above equation gives

$$T_p = [K_1 V_{BB} + V_{eb1} - V_T(0^+)]\, C_T/I \quad \ldots\ldots \quad (19)$$

Assuming $K_1 V_{BB} \gg [V_{eb1} - V_T(0^+)]$,

$$T_p \simeq \left(\frac{R_{B2}}{R_{B1} + R_{B2}}\right)(C_T/I)\, V_{BB} \quad \ldots\ldots\ldots \quad (20)$$

Equation (20) shows that the pulse period of the oscillator varies linearly with the applied control direct voltage to the oscillator. Since V_{BB} is obtained as the inverse of the voltage $V_{d.c.}$, which is proportional to the input p.r.f. to the pulse generator, the output p.r.f. which is the inverse of the pulse period T_p is linearly proportional to the input p.r.f.

Thus it can be seen that the oscillator output frequency may be made to follow V_{BB} linearly at the expense of using extra components to build the constant-current source. Although the oscillator designed to operate from a constant-current source should theoretically provide better linearily of input/output response of the pulse generator as compared with that with an $R_T C_T$ timing circuit, greater design efforts have to be directed towards achieving linearity of input frequency-to-control-voltage conversion for the oscillator to achieve this better linearity. Depending on the accuracy of measurement required, either of the two ways of charging C_T discussed above may be used.

Acknowledgments

The author wishes to express his gratitude to Aviation Electric Ltd, Canada, for permission to publish this article.

REFERENCE

1. O'GRADY, W. P. A Linear Frequency-to-Voltage Convertor. *Electronic Engng.* 36, 776 (Nov. 1964).

St. Thomas's Hospital Intensive-care Monitoring System

A system has been developed to provide monitoring facilities at the bedside in the new Intensive-Therapy Ward of St. Thomas' Hospital. The objective was an apparatus occupying a minimum of space in the bedside area, providing for display, measurement and recording of up to three blood-pressure waveforms and e.c.g., and for remote display (e.g. at a nurse's station) of e.c.g. only. In deciding to proceed with the development of a new system, it was planned to make use of commercially available subassemblies where possible. Preliminary design study established the main tasks: the development of miniaturised electronics for the amplification, display and measurement of signals, and the design of a pressure-transducer/tap table and tap system suitable for use at the intensive-care bedside.

A decision to use strain-gauge pressure transducers led to the design of a simple and very compact pressure-signal preamplifier based on integrated-circuit microamplifiers. The unit incorporates an electrical calibration system which automatically generates a sequence of simulated pressure levels appropriate to the range in use. This facility provides both a check on system linearity and unambiguous range identification on strip-chart recordings. In the e.c.g. channel, a parametric amplifier is used. The fully isolated input circuit provides excellent common-mode rejection, and has inherent immunity from damage by transient high-voltage inputs which can occur when a d.c. defibrillator is used.

A raster multitrace c.r.t. display technique has been developed, using a 20cm-diagonal magnetic tube. The system incorporates an electronically generated calibration graticule to facilitate measurement of the signal waveforms. Display of either two, three or four signals may be selected, each making maximum use of the available tubeface area. The x-timebase can be operated in a triggered mode which gives a stationary display of waveforms.

The need for a transducer tap simple to operate and capable of leakfree operation over long periods led to the design of a disk-type tap in which all required functions (flushing, measurement, calibration etc.) are selected by a single control. The tap table carries up to three transducers and taps, together with facilities for flushing and calibration. It is adjustable in height, sliding on two vertical guide rods, and is counterbalanced by a unit containing the pressure and e.c.g. preamplifiers. The extensive use of integrated-circuit devices has permitted the realisation of a very compact system occupying a total volume of approximately 0·04 m.

A 10MHz 1A 50V Pulse Generator

by R. W. Alsford* and R. E. Hayes*, M.Sc., Ph.D., M.I.E.E.

The article describes a high-power pulse generator originally designed for testing thin magnetic films. The generator is variable in frequency and will deliver its rated output at a 50 : 50 duty cycle up to a maximum of 10MHz. Isolated trigger pulses are provided, together with an accurate pulse delay achieved by using a tapped delay line. The output pulsewidth is variable between 10 and 100ns. The generator was designed for use with 50Ω cables and the output current is continuously variable up to a maximum of 1A.

(Voir page 533 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 534)

THE current-pulse generator to be described was designed for use in thin-magnetic-film research. A large-amplitude 1A drive pulse was required to feed into standard 50Ω coaxial cable. The use of 50Ω cable adds considerably to the versatility of the generator. In contrast to existing generators using valves[1], a high repetition rate of 10MHz with a 50:50 mark/space ratio is obtained.

The generator starts with a variable-frequency multivibrator which produces an isolated prepulse for triggering an oscilloscope and a pulse which is fed into a short-circuited delay line of variable length. The reversed-polarity reflected pulse, after being 'sharpened', is used to trigger a monostable circuit in which pulsewidth is controlled by variable inductances. This pulse is taken, via a complementary emitter-follower pair, to the final output stage of two transistors in cascode. The first transistor, driven in common-emitter mode, provides the current gain and defines the output current at 1A; the second transistor, in common-base mode, provides voltage gain up to the required 50V.

Clock generator

An emitter-coupled multivibrator (Fig. 1) is used as the basic clock generator; the operation is as follows. Assuming that VT_1 is switched off, the base of VT_2 is near +6V. The VT_2 emitter current is defined by resistors R_3 and R_2 in parallel and is arranged not to bottom VT_2. The current in R_3 discharges C_1 until VT_1 emitter start to conduct. The collector voltage of VT_1 falls, causing VT_2 emitter to follow it down initially. This regenerative action finally results in the base of VT_2 being at +4V with its emitter near +5V. VT_1 is now switched on with its emitter current defined by R_2 and R_3 in parallel. R_3 now discharges C_1 until the emitter of VT_2 is about to conduct; this reduces the current in VT_1, causing its collector voltage to rise and regenerate VT_2 on. This rise in base voltage of VT_2 is transmitted via its emitter and C_1 to the emitter of VT_1. Thus R_2 begins to discharge C_1 and the cycle recommences. Using silicon planar transistors, care must be taken not to exceed the reverse emitter-base voltage.

The important feature of this circuit is that the charge on the timing capacitor C_1 is not significantly altered during the regeneration process. Thus the circuit is not required to produce large transient charging currents and as a result the switching is extremely fast and independent of the value of C_1. It is for this last feature that the circuit was chosen, as it enables the generator to be operated over the range from 10MHz to less than 1Hz without any change in the output pulse. Transformer T_1 gives additional positive feedback and also provides an isolated prepulse. The timing waveforms at the emitter of VT_1 and VT_2 are linear ramps. They are thus susceptible to interference pulses just before switching, which will cause jitter. To limit this, since feedback can occur from the large final output pulse, the circuit is totally enclosed and carefully decoupled, and the output pulse is taken *via* an isolation transformer.

The repetition rate is varied from 10MHz to 5kHz by switching in fixed values for C_1. Lower speeds are available by adding external capacitors.

Delay stage

To assist with the viewing of traces on an oscilloscope, it is desirable to have a variable delay between the prepulse and the output pulse of a pulse generator. In this unit it is provided by a zig-zag printed-circuit delay line, the length of which is altered by a sliding contact; this contact

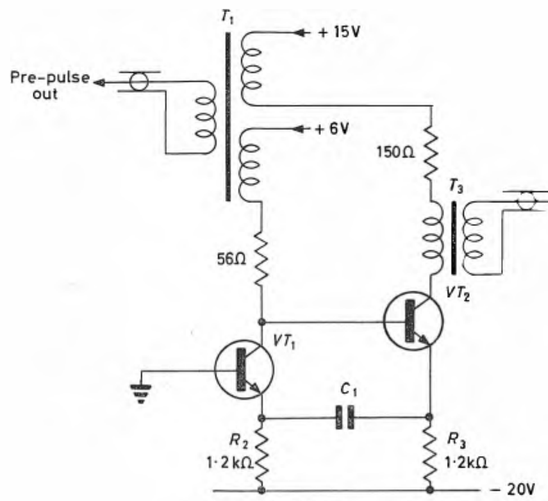
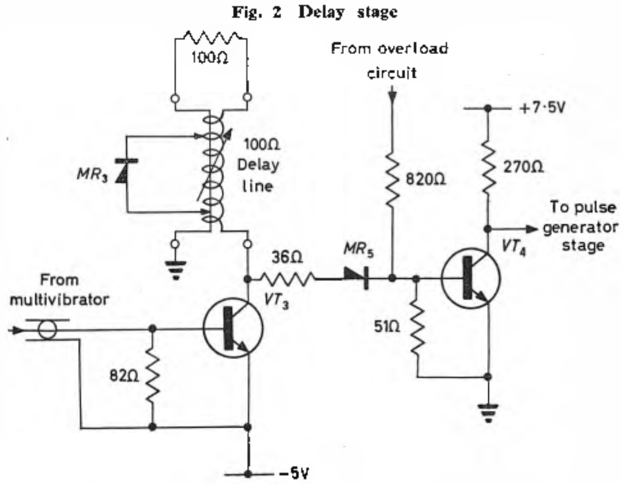


Fig. 1 Multivibrator

Fig. 2 Delay stage



* *The Plessey Co. Ltd.*

contains a diode to act as a short circuit. Its position is controlled by a large-pitch screw thread connected to a knob on the front panel. The circuit is shown in Fig. 2. A positive pulse from the clock multivibrator turns VT_3 on, putting a negative edge into the delay line. This is returned from the diode as a positive pulse which is coupled via MR_5 to turn on VT_4. The diode MR_5 allows VT_4 to be held permanently on by the overload circuit, thus preventing any further pulses from being fed to the pulse-generator stage. Exact matching for the delay line is not possible between VT_3 and VT_4; so the impedance is made greater than the characteristic impedance of the line. Positive reflexions from this point now ignore the diode MR_3 and are absorbed in the exact termination at the far end of the delay line.

A 100Ω characteristic impedance was chosen for the delay line as it is easily obtained on a printed-circuit board and is sufficiently high to make the diode effective. The actual electrical delay is 50ns, giving an overal delay variable from 0 to 100ns.

The output from VT_4 is differentiated and triggers the pulse-generator stage.

Pulse-generator stage

The pulsewidth generator uses a collector-to-base cross-coupled *n-p-n*/*p-n-p* pair, made stable in the off state by the use of inductances which also serve to time the duration of the pulse. The circuit is shown in Fig. 3 and this unit, comprising transistors VT_5 and VT_6 can be treated as one *p-n-p* transistor with the output taken from the *n-p-n* emitter.

The circuit is triggered on by a negative pulse generated by buffer transistor VT_4, applied to the base of transistor VT_5. The regeneration applied round the *p-n-p*, *n-p-n* loop then rapidly turns both transistors hard on. After the initial transients have died down the timing circuit is formed by VT_6 and L_2. The current in VT_6 is limited to the emitter current defined by the 680Ω resistor R_{21} and that required by the driver stages, which, towards the end

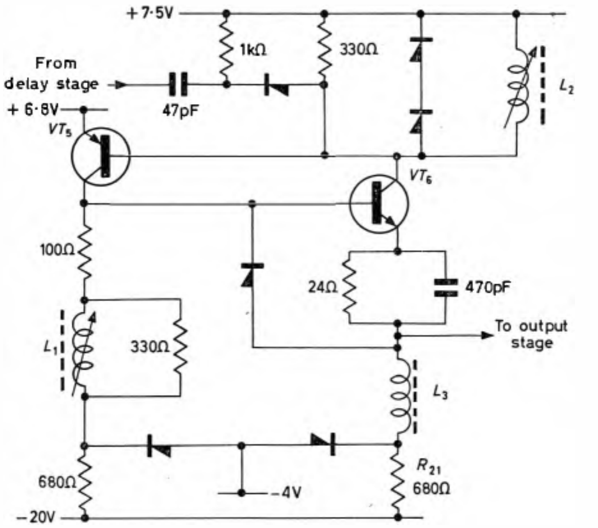


Fig. 3 Pulse-generator stage

of a pulse, has fallen to a value not greater than that in R_{21}. Thus when the current in L_2 reaches this value it turns off VT_5, which then switches off VT_6 helped by L_1. The value of this inductance is thus not critical, since the current in VT_5 collector does not appreciably alter, via its base, that in L_2. For the range of widths between 10 and 100ns, two fixed values were found sufficient. However, since L_2 was made variable, L_1 was ganged in sympathy, to obtain the fastest turn-off for transistor $VT6$.

The trigger input is applied to the base of VT_5, since this gives the minimum switch-on delay, because the internal regeneration is applied to the base in the same sense. If an emitter trigger is used, the resultant emitter current flowing through the necessary series diode causes the final emitter voltage to be in an opposite sense to that of the trigger, increasing the voltage swing necessary for complete turn-on.

Output stage

The output stage uses transistors in common-base connection, to obtain the maximum possible working voltage. The transistor chosen was the 2N2218 which has a minimum BV_{CBO} of 60V, an f_t of 400MHz and an output capacity of 4pF. To switch 1A it was necessary to use four

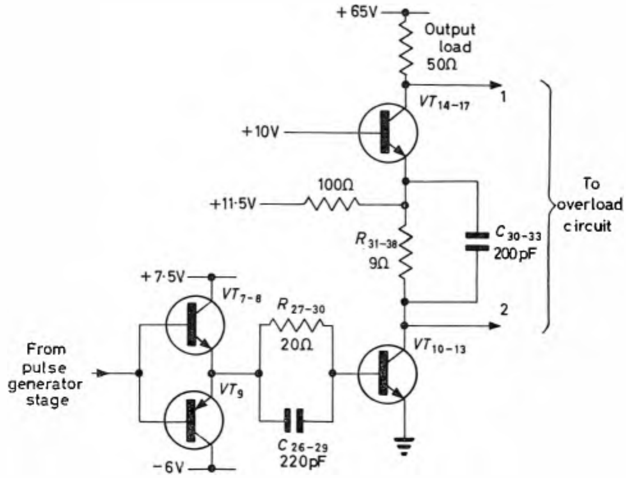


Fig. 4 Output stage

transistors in parallel. Even so, the output capacity and speed is better than any comparable single transistor. A simplified circuit of the output stage is shown in Fig. 4. Transistors $VT_{14\text{-}17}$ are the common-base stages and each forms with $VT_{10\text{-}13}$ a cascode pair. The bases of $VT_{14\text{-}17}$ are held at +10V and each emitter is fed with a pulse of current defined by the resistances $R_{31\text{-}38}$ in the collectors of $VT_{10\text{-}13}$ when they are bottomed.

Additional advantages of the cascode circuit are the elimination of feedback via the collector-base capacitance in $VT_{14\text{-}17}$ and, since the current is fixed, the output collector voltage can be controlled to minimise carrier storage.

As the transistors $VT_{10\text{-}13}$ work at comparatively low collector voltages, the faster type 2N2368 is used. Four in parallel switch 1A in 2·5ns. To ensure the equal division of current, each $VT_{14\text{-}17}$ emitter is individually fed from its corresponding collector via a separate resistor and speed-up capacitor ($R_{31\text{-}38}$, $C_{30\text{-}33}$). To assist the turn-off of $VT_{14\text{-}17}$ the emitter-base is reverse biased to 1·5V by returning the emitters through 100Ω resistors to +11·5V. A diode is also included on the emitter to restrict the amplitude of any overshoot at the end of the pulse.

The inductance of the path of the output current, including the cascode circuit, must be kept as small as possible. The bases of $VT_{10\text{-}13}$ are driven hard on and off by a complementary emitter-follower pair, $VT_{7\text{-}8}$ and VT_9 (Fig. 4). These transistors can be connected directly between their supply voltages, as the common-base and emitter connections prevent them from both being turned on simultaneously. To equalise the base drive between the four output stages, a separate resistance-capacitance network ($R_{27\text{-}30}$, $C_{26\text{-}29}$) is used.

High-current Zener diodes must be used to supply the 7·5V rail to $VT_{7\text{-}8}$. The +10V supply to $VT_{14\text{-}17}$ must also be held constant and this is supplied from a 2N706 emitter follower. These precautions are necessary to make sure that $VT_{10\text{-}13}$ bottom and that $VT_{14\text{-}17}$ do not. They also ensure the operation of the overload circuit.

Current probe

A current probe is provided to monitor the output current (Figs. 5 and 6). It consists of two Ferroxcube A5 rings, type FX2073. The primary is of one turn, the two rings being slipped over a convenient wire. The secondary has three parallel windings of six turns shunted with three 22Ω resistors which gives a sensitivity of 1V/A. Provided that the 50Ω output cable is correctly terminated, the total load presented by the probe is 0·2Ω. Fig. 7 shows that the risetime is less than 1ns and the time constant is 1·25μs.

Output attenuator

The pulse generator is designed to supply 1A pulses continuously into a 50Ω load. Smaller outputs are obtained *via* an attenuator which is designed to give a continuous nominal 50Ω load to $VT_{14\text{-}17}$ (Fig. 11). The maximum variation which can be tolerated is determined by the need to prevent $VT_{14\text{-}17}$ bottoming or exceeding their dissipation limit and by the operation of the overload circuit (Fig. 10). A variation of 6V, or ±3Ω, can be tolerated. This is just achieved with an attenuator consisting of ganged 500Ω log and 500Ω antilog potentiometers. To make the matching simpler, the attenuator consists of similar 500Ω log tracks mounted back-to-back. To assist in the dissipation of heat, a cooling fin is mounted between the two tracks.

When the attenuator is turned to zero, the output is diverted into a dummy load. The maximum dissipation occurs when the attenuator is in the midway position. The output power is then equally divided between the load, the dummy load and the two attenuator tracks. At the maximum 50:50 mark/space ratio this is 6·5W each.

A fan is used to cool the whole equipment and the attenuator and fin are placed in this air stream. Some dissipation and temperature measurements taken on various transistors during the operation of the generator are given in Fig. 10 and Table 1.

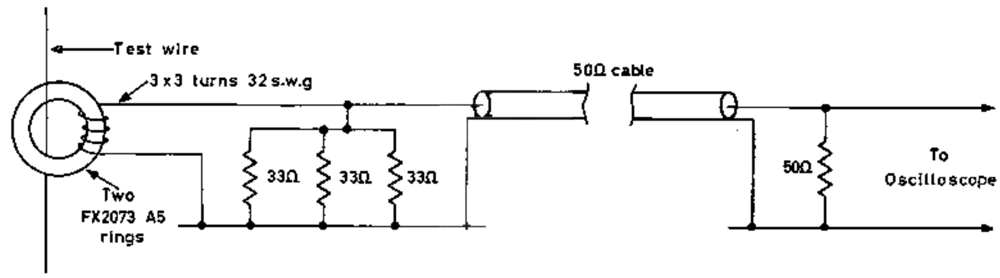


Fig. 5 Current-probe circuit

Fig. 6 Current probe

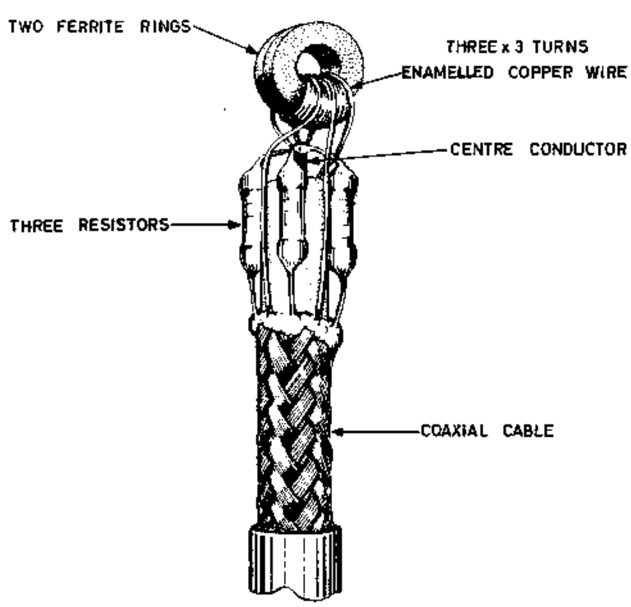
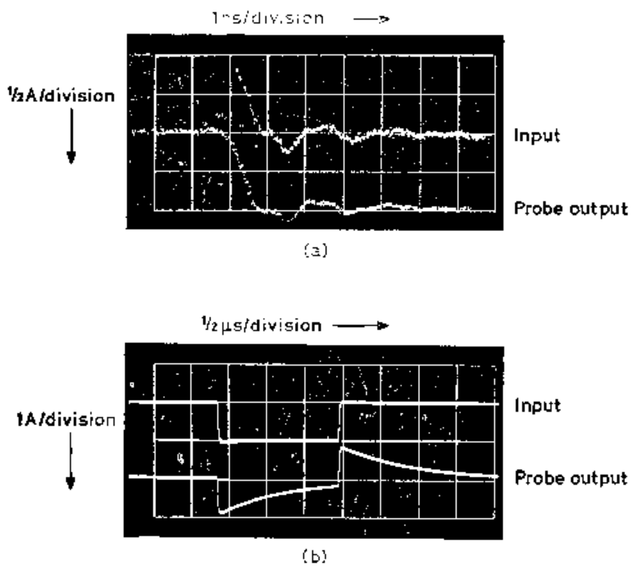


Fig. 7 Current-probe waveforms



Overload-protection circuit

Without special protection, an excessive dissipation in the output transistors could occur if the output terminal were short-circuited or even if a load less less than 50Ω were used. A protection circuit was thought necessary to monitor the output collector voltage during the pulse. The first method tried was by gating the output collector current and voltage using diodes and tunnel diodes. This circuit proved to be critical in adjustment.

The arrangement finally adopted is shown in simplified

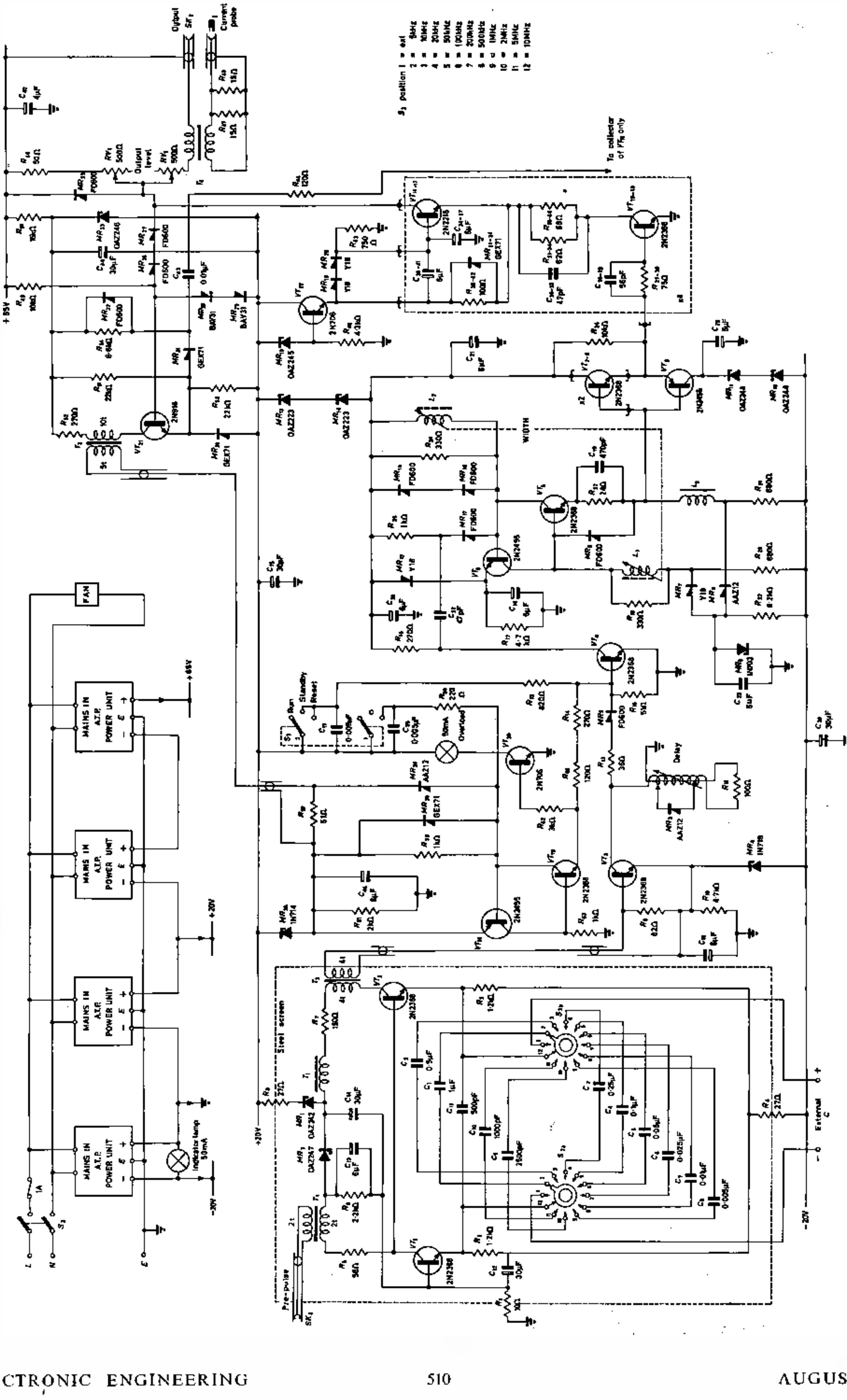


Fig. 8 1A pulse-generator circuit

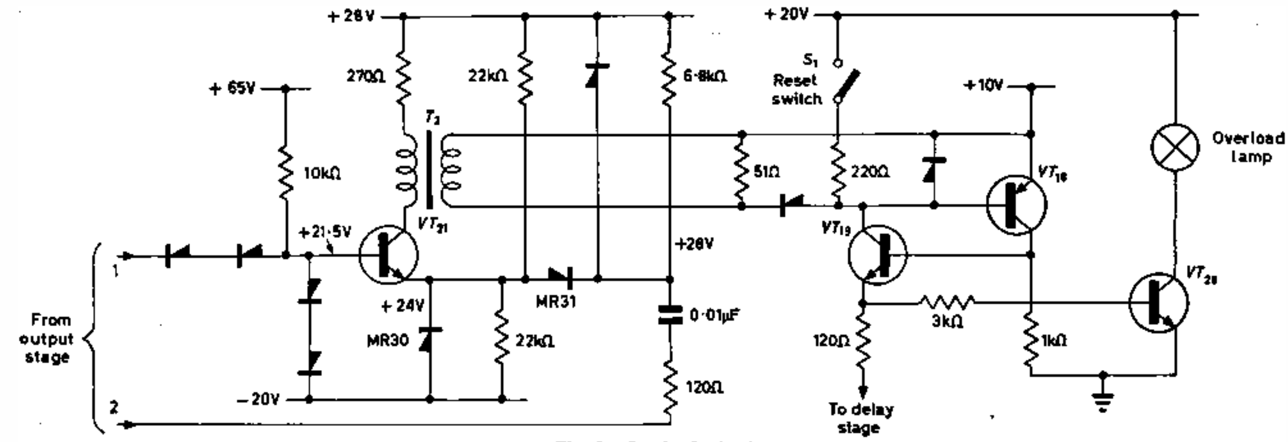


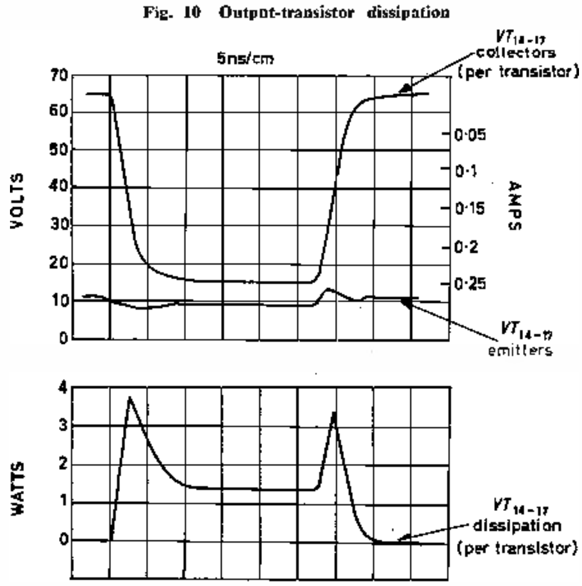
Fig. 9 Overload circuit

TABLE 1 Transistor dissipations

TRANSISTOR	AVERAGE DISSIPATION IN EACH TRANSISTOR DURING A 30ns PULSE	AVERAGE DISSIPATION IN EACH TRANSISTOR AT 10MHz WITH 50ns WIDTH	TRANSISTOR-CASE TEMPERATURE AT 24°C AMBIENT		HEAT SINK
			AT 2MHz 100ns WIDTH	AT 10MHz 40ns WIDTH	
	mW	mW	deg C	deg C	
$VT_{7\text{-}8}$	350	175	29	31	Small copper block
$VT_{10\text{-}13}$	35	17·5	29	34	None
$VT_{14\text{-}17}$	2000	1000	33	41	Large copper block

form in Fig. 9; the steady-state voltage levels are shown in small circles. This arrangement works as follows: during a normal pulse, $VT_{14\text{-}17}$ collectors drop to +15V and pull down the base of transistor VT_{21}. At the same time, the emitter is pulsed by a negative drive taken from the collectors of one of the drive transistors $VT_{10\text{-}13}$ (Fig. 4). It is arranged, by catching the waveform on MR_{30}, that this pulse is insufficient to turn on VT_{21}. This emitter pulse is derived slightly in advance of the final output and from a faster source, and is therefore delayed by biasing back the drive point using diode MR_{31}.

Fig. 10 Output-transistor dissipation



However, if the output collector voltage is too positive, VT_{21} is turned on by its emitter pulse, providing a trigger for the bistable lock-out circuit using transistor VT_{18} and VT_{19}. When this circuit is switched on, it lights an overload-indicator lamp and blocks the subsequent trigger pulses from transistor VT_4 by turning it permanently on. The bistable circuit is reset by the manual operation of the switch S_1.

Power supplies

Each of the three main power supplies is derived from standard voltage-stabilized current-limited units, bought commercially. Their interconnexion is shown on the main diagram (Fig. 8).

Performance specification

Repetition rate:	10MHz to 5kHz or lower with external capacitors
Isolated prepulse output:	±1·4V into 50Ω, 3ns risetime
Delay between prepulse and output:	25-125ns in 4ns steps
Pulsewidths available:	10-100ns

Mark-space ratio limited to 50:50 at high repetition rates
Output 0·1A into a nominal 50Ω load
Average rise and fall time 4ns (3ns minimum)

Acknowledgments

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Analogue Switches for Computer Mode Control

by N. W. Bellamy*, B.Eng., Ph.D., and S. R. Styler*

A solid-state integrator-mode-control circuit to directly replace present relay control is described. By using a complementary-transistor series switch with a rechargeable-battery power supply, it is shown that offset voltages of less than 50μV and leakage currents of less than 10^{-10}A can be achieved easily. Test results show that the performance of the mode-control circuit is as near to ideal as present-day solid-state devices permit.

(Voir page 533 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 534)

UNTIL recently, the majority of analogue computers have used relays to switch their integrators into the required mode of operation. To overcome the inherent problems of reliability and speed of action of relays, a number of solid-state mode-control circuits have been suggested and used[1]. In each case, the analogue signals in an integrator are switched at virtual earth levels by logic signals, to give RESET, COMPUTE and HOLD modes.

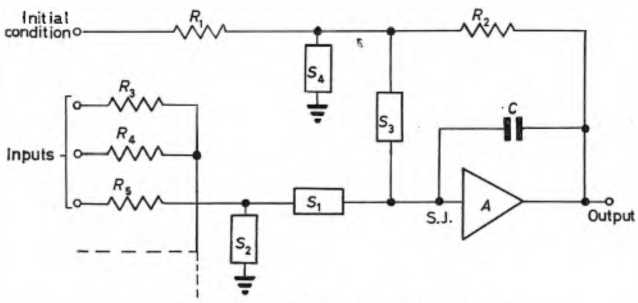

Fig. 1 Integrator-mode-control circuit

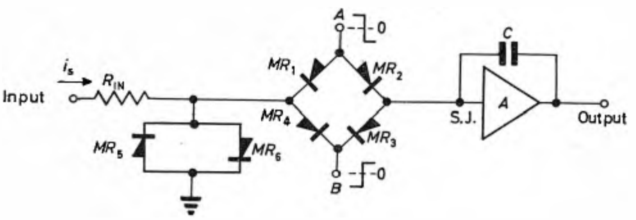

Fig. 2 Diode-bridge mode switch

The basic requirements of an ideal analogue switch are zero closed resistance, infinite open resistance, no spurious current or voltage offsets and instantaneous operation. Unfortunately, semiconductor devices are not as close an approximation to an ideal switch as relays, since they exhibit a lower resistance and greater leakage current when 'open' and an appreciably higher resistance when 'closed'. Furthermore, it is difficult to prevent interference with the analogue signal by the control signal of the switch.

A commonly used form of integrator mode control is shown in Fig. 1, in which two series-shunt combination switches are used. On examination, it can be seen that the two series switches S_1 and S_3 are subject to different specifications. Theoretically, S_1 requires a zero closed resistance if computing errors are to be avoided (as shown in the Appendix), whereas the only effect of the closed resistance of S_3 is to slightly increase the reset charging time of the feedback capacitor C. Both series switches require balanced or isolated drive circuits, and in some cases isolated power supplies, whereas the shunt switches S_2 and S_4 can usually be directly coupled to their logic drives. One point to note, however, is that, since the logic drive of the individual series switches is complementary to their associated shunt switches, isolation is only necessary when a series switch is closed, since, in its open state, a path to ground is provided by the closed shunt switch.

Types of solid-state analogue switch

The three semiconductor devices generally used as analogue switches are junction diodes, transistors and field-effect transistors. Each of these devices may be used as series, shunt or series-shunt-combination switches. The choice of switching device is determined by which particular advantage the application demands.

The junction diode has long been used as an analogue switch. The bridge connection shown in Fig. 2 is in common use in mode-control circuits, but suffers the disadvantage of having a fairly high nonlinear closed resistance[2]. Difficulties are also experienced in matching diodes to eliminate offset voltages and in providing balanced or isolated bias supplies.

Fig. 3 shows a simple version of a series-shunt-combination transistor switch which uses the transistor in the inverted mode of operation to produce low offset voltages. This type of switch has a closed resistance of about an order less than that of the diode-bridge type, but has the disadvantage that the series component of the switch requires an isolated power supply[3].

The most promising analogue switch at the present time is the field-effect transistor. This device provides inherent isolation of the signal and drive circuits and has no offset voltage in its closed condition[4]. Unfortunately, the closed resistance is relatively high and the device is expensive. However, field-effect transistors are being developed rapidly and in the near future neither of these two disadvantages may hold. Fig. 4 shows a field-effect transistor used as a series element in a shunt-series-combination switch.

Complementary-transistor series switch

The Appendix shows that, in order that the summing errors are to be tolerable, the closed resistance of the series switch S_1 has to be less than 10Ω. This rules out diode and field-effect-transistor switches and necessitates the use of a series transistor operating in its inverted mode. An investigation into the characteristics of a number of common transistors resulted in the Mullard OC201 *p-n-p* transistor and the Fairchild C400 *n-p-n* transistor being

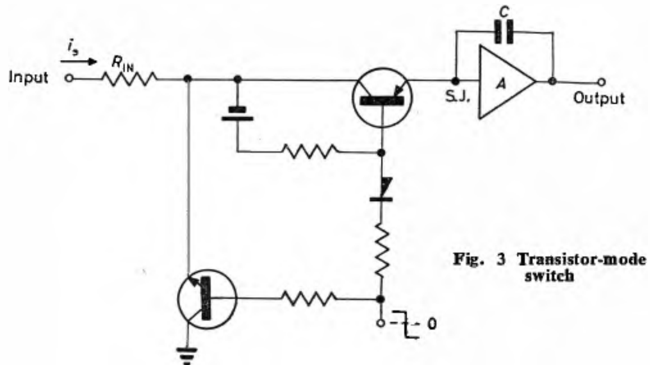

Fig. 3 Transistor-mode switch

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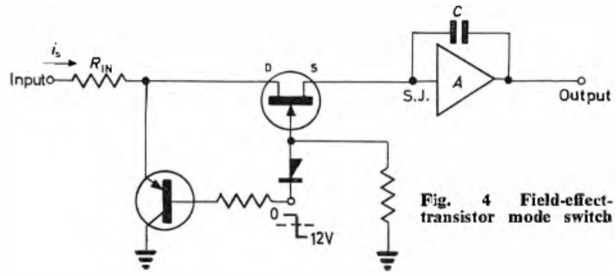


Fig. 4 Field-effect-transistor mode switch

selected as the most suitable. The ON characteristics of an OC201 and a C400 operating in their inverted modes are shown in Figs. 5(a) and (b). It will be observed that the off-set voltage of the OC201 is of the order of -1mV over the range of base currents shown and the offset voltage of the C400 is of the order of $+0{\cdot}5$mV. These offset voltages compare well with specially manufactured chopper transistors but are not really low enough for high-accuracy computing.

Several methods of offset-voltage compensation have been suggested and used[5], but they have usually resulted in more complex drive circuitry. However, one method which is particularly suitable for mode-control switching is the complementary-transistor combination. If the collectors and emitters of an OC201 and a C400 are connected in a complementary manner, as shown in Fig. 5(c), three advantages are achieved. First, the offset voltages of each transistor, being of opposite polarity, tend to cancel, and offsets of less than 100μV are achieved. Secondly, the closed resistances of the two transistors are in parallel, giving a lower overall closed resistance. Finally, a convenient control-current path is provided through the bases of the two transistors.

Since a complementary-transistor series switch of this type requires an isolated power supply to provide the base current to hold the transistors on, a compromise has to be made between its closed resistance and the economy of power-supply current. From the characteristics in Fig. 5(c), a base current of 2mA, giving a closed resistance of 6Ω over a signal current range of several milliamperes, appears to be the most suitable.

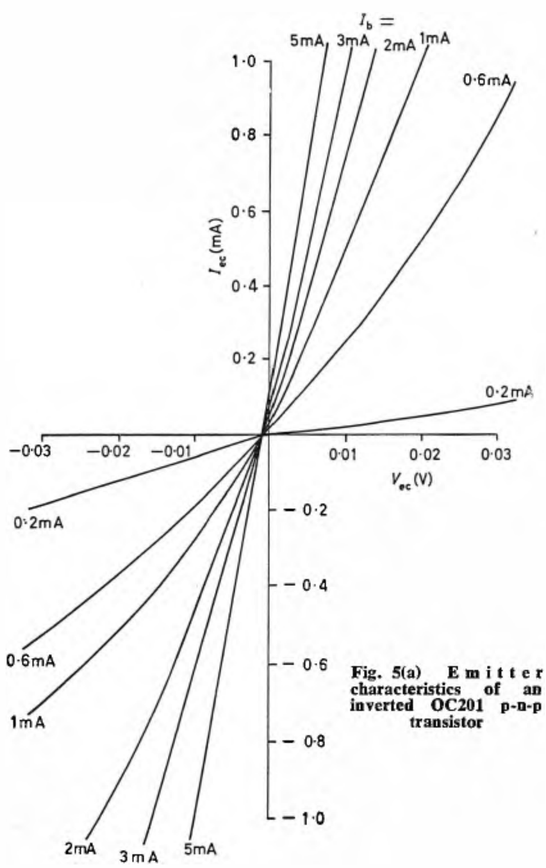


Fig. 5(a) Emitter characteristics of an inverted OC201 p-n-p transistor

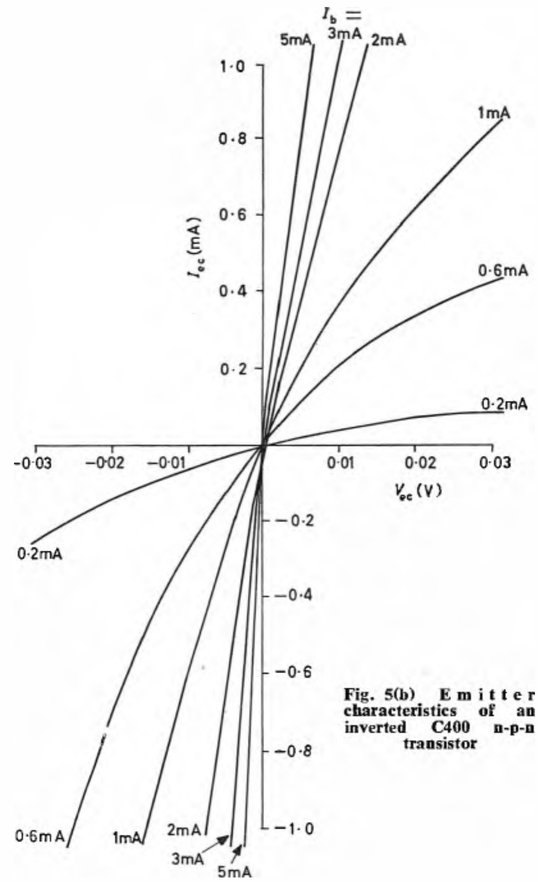


Fig. 5(b) Emitter characteristics of an inverted C400 n-p-n transistor

Isolated battery power supply

To provide the base control current of the complementary-transistor series switch, some form of balanced or isolated power supply is required. Balanced power supplies are possible but are sensitive to h.v. variations, although compensation circuits can be incorporated[5]. The majority of isolated power supplies require transformers and associated rectifying and smoothing circuits. Care has to be taken to ensure sufficient isolation and to prevent interference with the analogue signal by the carrier frequencies. Nearly perfect isolation is provided by photovoltaic cells, but several are required in series, and appreciable power and heat is dissipated in the lamp source.

The ideal isolated power supply is the rechargeable battery, provided that the duty cycle permits convenient recharging. On examination of the duty cycle of the series switch S_1, the only time the switch is closed (hence, requiring an isolated power supply) is in the COMPUTE mode. During the RESET and HOLD modes, the switch S_1 is open, giving an opportunity for charging the battery if this is the form of power source used. In general, analogue-computer integrators are not kept in the COMPUTE mode for more than a small part of their working day; so that the demand for battery capacity is quite modest.

Isolated series switch S_1

Fig. 6 shows how the complementary-transistor series switch can be driven from antiphase logic sources, which also enable the battery to be charged when the switch is in the open condition. If the logic drives at the points A

and B are such that the diodes MR_1 and MR_2 are reverse biased, the battery will provide a base current through the transistor, holding them in the ON condition. If the logic drives at the points A and B are inverted, MR_1 and MR_2 will be forward biased and will divert the base currents, switching the complementary transistors off. A battery-charging current path will also be provided through the diodes MR_3 and MR_4.

Three nickel-cadmium batteries in series give an output of 3·6V, which, in the circuit shown, provide a 2mA base current in the closed condition. Under these conditions 100mAh batteries are able to run for 50h in the COMPUTE mode, a condition not likely to be encountered. If physical size of battery is critical, the very small 20mAh cells could be used and would still give 10h COMPUTE time. By suitable choice of logic drive levels at points A and B, the battery-charging current can be set to suit the probable duty cycle.

A general criticism of the use of rechargeable batteries in electronic equipment is their limited life. When used under the circuit conditions discussed, the manufacturers give a conservative estimate of life of not less than 18 months' continuous use. This indicates that the best method of maintenance is probably by periodic replacement of the batteries, which is quite economical as they only cost a few shillings.

Complete integrator mode control circuit

As previously indicated, the specifications of the series switch S_3 are not as stringent as those of S_1. The closed resistance of S_3 is not critical and hence the field-effect transistor, with its simple drive requirements, is ideal. With regard to the shunt switches S_2 and S_4, no particular difficulty is presented, and the simple inverted transistor switches suffice.

The complete circuit of an integrator-mode-control switch is shown in Fig. 7. The two logic drives (1 and 2) enable the integrator to be put into either RESET, COMPLETE

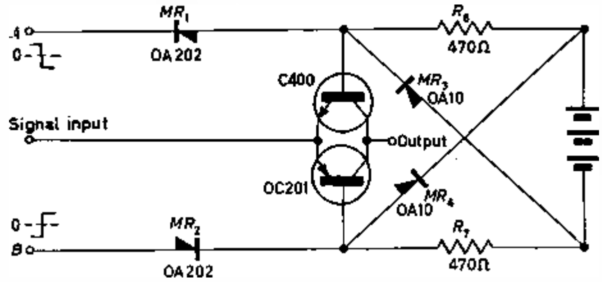


Fig. 6 Complementary transistor switch

or HOLD modes, as shown in the truth table. Logic line 1 is phase-split by VT_4 and VT_5, which drive the level-setting resistors $R_{12\text{-}15}$ *via* the emitter followers VT_6 and VT_7. The two resulting antiphase logic levels are then applied to the complementary transistors VT_1 and VT_2 of the series switch S_1 and also to VT_3 of the shunt switch S_2. Logic line 2 drives VT_8 of the shunt switch S_4 direct, but is inverted by VT_9 to provide the correct drive for the field-effect transistor FET_1 of the series switch S_3. To prevent discharge of the batteries when the analogue computer is switched off, a simple relay energised direct from the logic power supply operates the RL_1 in the battery circuit.

The calculation of the level-setting resistors $R_{12\text{-}15}$ involves the fulfilment of two conditions. First, when the complementary-transistor switch is closed, the drive diodes MR_1 and MR_2 must be just reverse biased, to keep leakage to a minimum; secondly, when the switch is open, the drive circuit must overcome the battery-circuit resistance and charge the battery by the required amount. It will be observed that some degree of charging regulation is inherent, owing to the effect of the decrease in battery voltage during prolonged discharge.

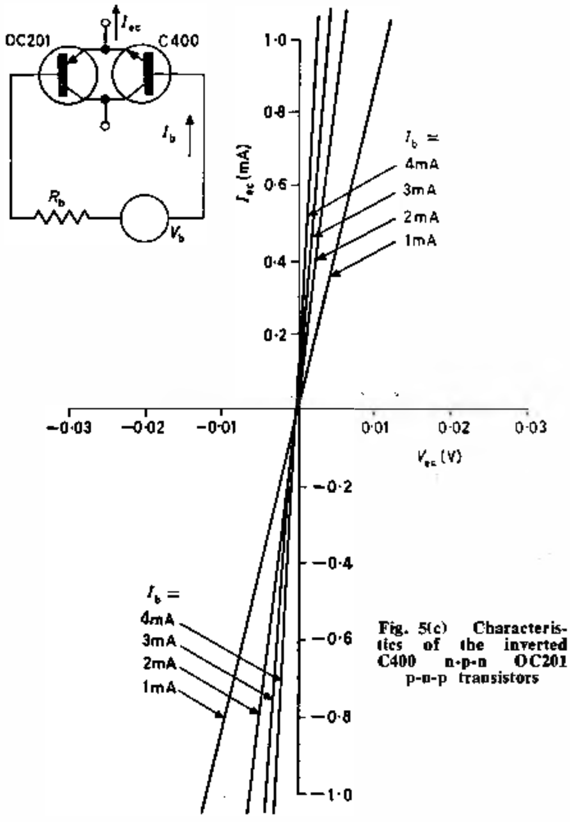


Fig. 5(c) Characteristics of the inverted C400 n-p-n OC201 p-n-p transistors

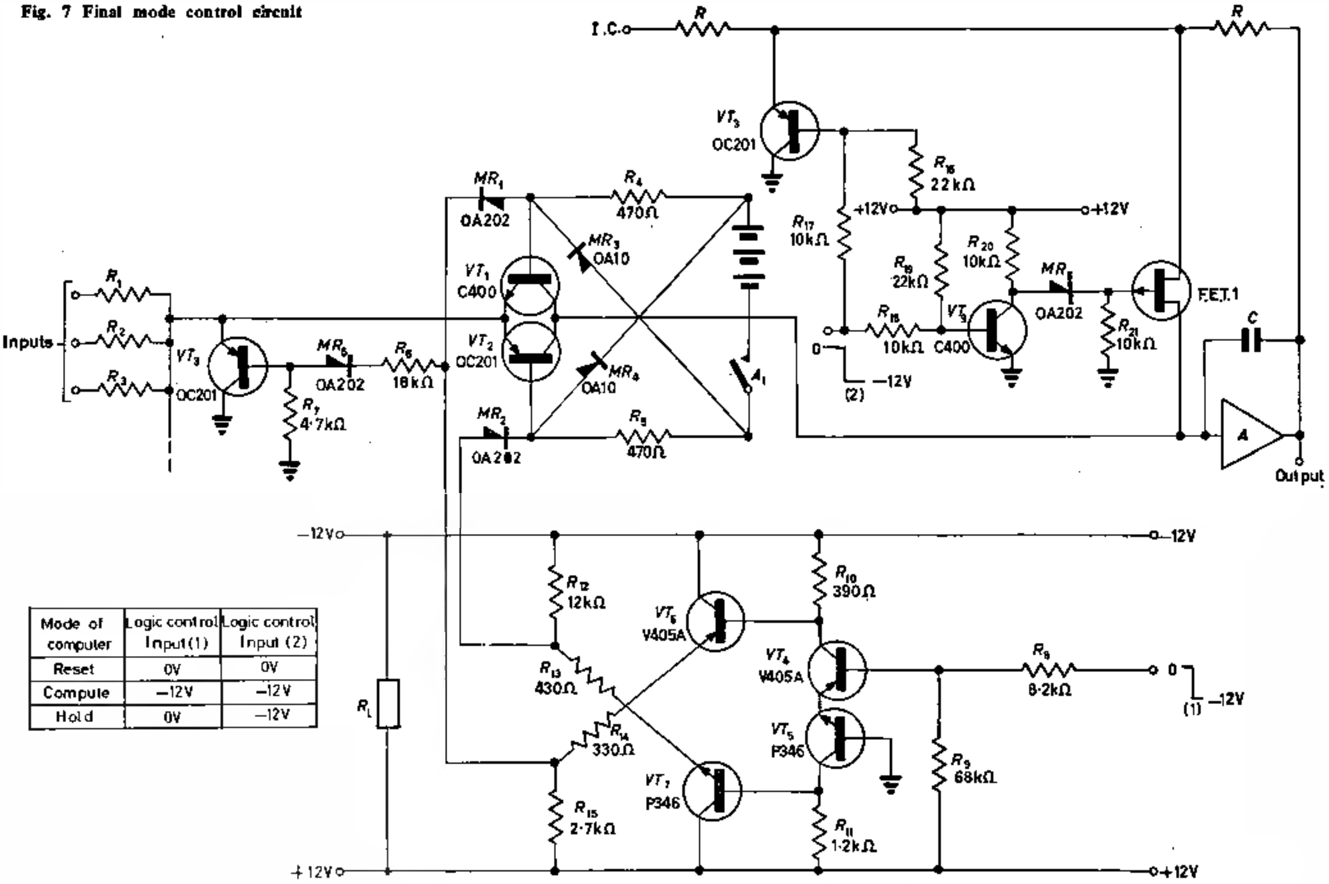
Performance of the mode-control circuit

The performance of an integrator is normally determined by its drift, which can originate from mode-control circuits in two ways. First, the voltage of the series switch S_1 causes a spurious input current to flow through the computing input resistors, which produces drift in the COMPUTE mode only; secondly, drift in both the COMPUTE and HOLD modes is produced by the leakage currents in both series switches S_1 and S_3.

High-quality computing amplifiers have input currents of 10^{-11}A or less, which give unity-time-constant drift rates of 10μV/s or less. In practice, however, when amplifiers are used as relay-mode-controlled integrators in general-purpose machines, drift rates of 100μV/s are more realistic. This suggests that a suitable target specification of a mode-control circuit of 100μV offset and 10^{-10}A leakage current would not appreciably deteriorate practical integrator performance.

To test the mode-control circuit of Fig. 7, a breadboard was constructed using unselected components and connected to Solartron type AA1054·2 amplifier used as an integrator in a Solartron 247 installation. A number of tests were devised to test the performance of the integrator in all possible conditions with its own relay mode control and also with the solid-state mode control. Analysis of the tests showed very satisfactory results, which is evident from the fact that, in some cases, the leakage introduced by the solid-state mode-control circuit was swamped by the existing integrator leakage. The results showed that the mode-control switch S_1 exhibited an offset of approximately +47μV with a resistance of 6Ω in all closed conditions, while the spurious leakage current of the whole circuit was $1{\cdot}4\times10^{-10}$A in the worst case. The

Fig. 7 Final mode control circuit



Mode of computer	Logic control Input (1)	Logic control Input (2)
Reset	0V	0V
Compute	−12V	−12V
Hold	0V	−12V

switching time of the circuit was of the order of 1μs and the injected charge when changing from COMPUTE to HOLD, or from HOLD to COMPUTE, was not more than 100pC.

Conclusion

A low-cost solid-state mode-control circuit has been put forward which incorporates a complementary-transistor series switch with low closed resistance and with a rechargeable battery for its isolated power supply. Replacement of existing integrator-mode-control relays by this circuit should overcome relay reliability and time-delay errors, while not noticeably impairing general performance.

A number of improvements and variations are possible. Transistor switches in place of R_4 and R_5 would considerably reduce the drive current required during the time in which the battery is being charged. The use of an enhancement-mode field-effect transistor in place of RL_1 would be more convenient, while high-speed low-leakage diodes for MR_1 and MR_2 should decrease charge injection. For the commonly used 3-relay resetting mode circuits, two low-impedance series switches could easily be adapted.

Appendix

The effect of closed resistance of series switch S_1 (Fig. 8)

(A) $R_s = 0$ (ideal switch)

$$X_{OA} = -\left\{\frac{X_1 R_f}{R_1} + \frac{X_2 R_f}{R_2} + \frac{X_3 R_f}{R_3}\right\}$$

(B) $R_s \neq 0$

$$X_{OB} = -I_{in} R_f$$

$$I_{in} = X'/R_s$$

and $I_{in} = \dfrac{X_1 - X'}{R_1} + \dfrac{X_2 - X'}{R_2} + \dfrac{X_3 - X'}{R_3}$

It can be shown that

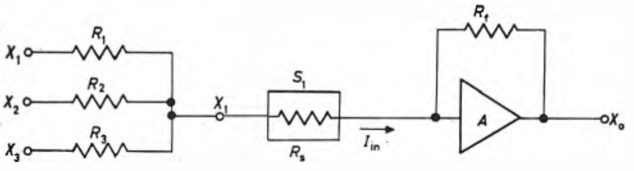
$$X_{OB}/X_{OA} = \frac{1}{1 + R_s(1/R_1 + 1/R_2 + 1/R_3)}$$

the error term being represented by

$$R_s(1/R_1 + 1/R_2 + 1/R_3)$$

It can be seen that the closed resistance of the switch S_1 introduces the greatest error when R_1, R_2 and R_3 are at their minimum values.

Fig. 8 Diagram showing the effect of closed resistance of series switch S_1



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LETTERS TO THE EDITOR

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

Suppression of radio interference in thyristor controllers

SIR,—In recent years, thyristor proportional controllers have become widely used for temperature control, motor-speed control etc., and phase control of the thyristors has generally been adopted as the method for obtaining proportional control. This form of control, however, also generates bursts of wideband interference at 100/s, posing a difficult intereference problem when sensitive electronic monitoring or data-logging equipment has to be used near the thyristor controller.

Using phase control, the alternating current through a load may be thyristor-controlled by supplying trigger pulses between the gate and the cathode of the thyristor, causing it to switch on and remain conducting while the anode is at a positive voltage. The proportion of the halfcycle during which the thyristor conducts (and therefore the power in the load) may be governed by controlling the delay between the start of the cycle and the arrival of the trigger pulse. It is often convenient to use a single thyristor to switch both halfwaves, by connecting it across the output of a bridge rectifier which is connected to the a.c. supply through the load. Such an arrangement is shown in Fig. 1. While this circuit gives smooth control of the output and may respond within one halfcycle, the rapid switching into conduction of the thyristor will cause interference unless it is suppressed by a filter between the load and the control circuit. Interference does not occur if the thyristor is only switched when the current is momentarily zero, which occurs twice during each cycle.

Where there is a slow response to a change in the electrical power supplied to the load, e.g. from motors and heaters, it is often acceptable to control the power by switching the power on and off for a number of complete halfcycles and controlling the proportion of time for which the power is switched on.

A circuit which enables the thyristor to be switched on only when the current is close to zero, at the beginning of each halfcycle, is shown in Fig. 2. When the transistor VT_2 ceases to conduct, i.e. when the switch S_1 closes on position *a*, positive pulses occur at the collector of VT_1 each time the base current falls to zero, which is at the end of each halfcycle. Therefore the thyristor begins to conduct when the first pulse to follow the closure of S_1 occurs, and the thryistor is not switched on until the current is close to zero. When the switch S_1 is closed on position *b* and transistor VT_2 is made to conduct, pulses are prevented from reaching the gate, and at the end of the halfcycle, when the current becomes zero, the thryristor will switch off.

Tests using a radio receiver show that the circuit given in Fig. 2 will not cause interference in the medium or long wavebands.

The circuit should be particularly useful for use in temperature controllers where a thermostat with contacts is already available. For manual control, a squarewave with a duty cycle which could be varied continuously would have to be generated by additional circuits and applied to the base of VT_2. The main advantage of this circuit is that it removes the requirement to use a filter which is capable of carrying the current through the load and is effective over a wide band.

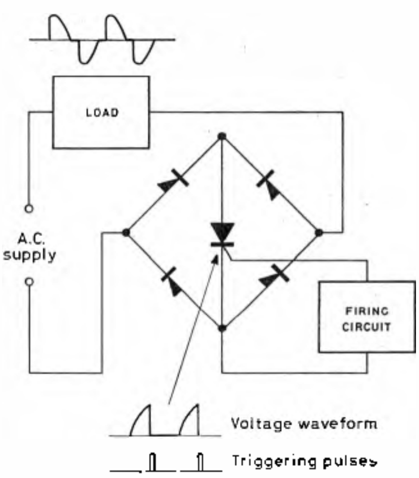
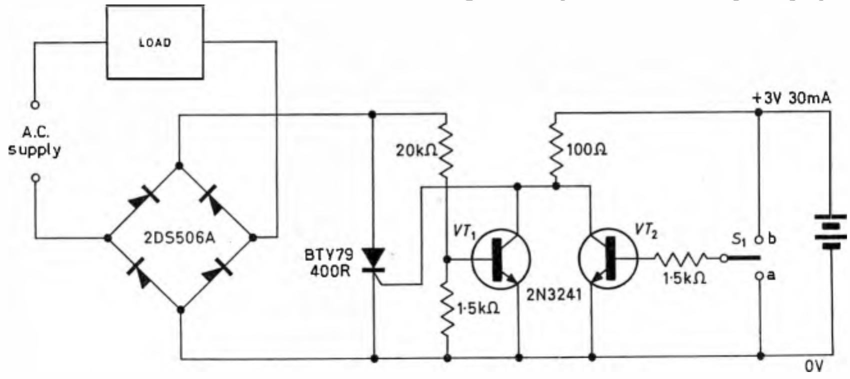


Fig. 1 Thyristor across bridge-rectifier-circuit output

Fig. 2 Complete circuit showing firing system



It is necessary to provide a low-voltage supply, but this will most probably be needed for other circuits, e.g. a thermistor bridge, amplifier and trigger circuit in a temperature controller, and will not have to provide specially for the thyristor control circuit.

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Electrical resistance of evaporated Nichrome films

SIR,—A study using the Fuchs-Sondheimer theory has been carried out for interpreting the behaviour of the resistivity of vacuum-deposited films obtained by the evaporation of Nichrome. Starting with these assumptions:

evaporant: 80Ni-20Cr alloy
vaporization temperature: 1500°C
pressure: 10^{-4}torr
substrate temperature: 250°C
source-substrate distance: 10cm

the following results were obtained: emission speeds = $9{\cdot}22 \times 10^4$cm/s (chromium) and $8{\cdot}68 \times 10^4$cm/s (nickel); mean free paths through the rarefied gas = 81·32cm (chromium) and 81·84cm (nickel); condensation rate in conditions of uninterrupted deposition = 41·66Å/s, corresponding to 16·66 monoatomic layers per s; film composition = 77·46% Ni, 22·54% Cr; oxide content = 0·006%; the oxidation is almost exclusively of chromium, this being more electropositive than nickel[1].

How the resistivity of the film ρ_f depends on its thickness s was then determined[2,3]. The following equations could be applied:

$$\rho_f = \rho_b(1 + 3\lambda/8s), \text{ for } s > \lambda$$
$$\rho_f = \rho_b(4\lambda/3s)\{1/[\log(\lambda/s) + 0{\cdot}4228]\}, \text{ for } s < \lambda \quad \ldots\ldots\ldots\ldots (1)$$

where ρ_b denotes the resistivity of bulk material (107·9$\mu\Omega$cm) and λ indicates the mean free path of the electrons at the surface of the Fermi distribution. Such equations, derived from Fuchs[4]-Sondheimer[5] theory assuming that the scattering of the electrons by the film boundaries is entirely diffuse, recall Planck[6]-Weale[7] and Lovell-Appleyard[8] formulas, respectively.

Equations (1) require the determination of λ, which depends on the electron concentration N by the relation

$$\lambda = (h/e^2\rho_b)(3\pi^2/N^2)^{1/3} \quad \ldots\ldots (2)$$

obtained by expressing the mean free path as the product of the relaxation time multiplied by the average electron velocity ($\lambda = \tau v$), where v is derived from the Fermi energy and the effective mass:

$$v = (2E_F/m^*)^{1/2}$$

and representing[9] both E_F and τ as

TABLE 1

FILM THICKNESS	SHEET RESISTIVITY (CALCULATED VALUES)	SHEET RESISTIVITY (MEASURED VALUES)
Å	Ω/square	Ω/square
100	111·53	235·80
200	54·86	105·75
300	36·37	60·35
400	27·20	37·10
500	21·73	28·25
600	18·08	23·15
800	13·54	17·20
1000	10·83	13·70
1500	7·21	9·10
2000	5·40	6·80
3000	3·60	4·55
4000	2·70	3·40
5000	2·16	2·75

TABLE 2

Cr_2O_3 CONTENT	TEMPERATURE COEFFICIENT
%	parts/10^6 per degC
0	+215·00
0·01	+185·23
0·03	+125·70
0·05	+ 66·18
0·07	+ 6·65
0·09	− 52·87
0·11	−112·40
0·13	−171·93
0·15	−213·45

functions of N:

$$E_F \simeq (h^2/2m^*)\,(3\pi^2 N)^{2/3}$$

$$\tau = m^*/(Ne^2\rho_b)$$

Examining the distribution of the charges between the $3d$ and $4s$ bands, and considering that the carriers can be trapped by unfilled states of the inner $3d$ shells, $N = 4{\cdot}77 \times 10^{22}\text{cm}^{-3}$ was obtained; inserting this in equation (2) gave $\lambda = 8{\cdot}96$Å. Employing equation (1) and applying Om's second law to square-shaped films, the results given in Table 1 were obtained.

The experimental values listed in the last column were obtained by averaging the sheet resistivities of many films deposited on very smooth glass surfaces, according to the above conditions. The measured resistances are higher than those calculated; the differences increase as the films become thinner. Such anomalies can be attributed to oxidation and structural alteration of the films, respectively.

Another important question concerns the exigency that the film resistance does not change with temperature. Therefore the temperature coefficient (for $s > \lambda$, this can be expressed by the bulk value[10] 215 parts/10^6 per degC) must be reduced by a suitable oxidation of chromium atoms. Employing the formula

$$\rho(Cr_2O_3) = \rho_0(Cr_2O_3)\exp[E_g(273\text{-}T)/546kT] \quad \ldots\ldots\ldots\ldots (3)$$

(which expresses the resistivity of chromium oxide in relation to absolute temperature by means of the energy gap $E_g = 4{\cdot}1$eV), expanding the exponential in series, stopping at the second term of the summation and assuming $T = 293°$K, the temperature coefficient of chromium oxide ($-297 \times 10_3$ parts/10^6 per degC) was calculated.

It was deduced how the temperature coefficient of the oxidized films changes with Cr_2O_3 contents, deriving the values listed in Table 2. The temperature coefficient is nearly zero when the oxidation affects 0·072% of the atoms: consequently the oxygen molecules absorbed by the film during its formation must become 12 times more numerous than those trapped in conditions of continuous deposition. For satisfying such an exigency the substrates are fixed to a turntable, using a screening system which hinders the deposition during 11/12ths of the rotation period: most of the obtained films (85% at s = 5000Å, 78% at s = 1000Å, 72% at s = 500Å, 64% at s = 100Å) show temperature coefficients whose absolute value is lower than 25 parts/10^6 per degC. The films that are worst in respect of expectations are the thinnest ones, whose temperature coefficients are frequently negative (−100 to −200 parts/10^6 per degC) since their discontinuous structure allows the activation of a conduction mechanism based on electron tunnelling between islands[11]. Finally, it can be seen that the lowering of temperature coefficient associated with modifications of the overlapping $3d$ and $4s$ bands[10] or with the formation of NiO semiconducting oxide can be neglected.

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Binary-quinary decade counter

SIR,—A recent paper by Parshad and Suri 'A binary-quinary decade counter using resistance logic' (December issue) pointed out the economies which may be achieved by using binary-quinary code for decade counting. Binary-quinary counting has been used previously in a number of applications, as has resistance logic, although not in the form proposed by Parshad and Suri; but I would like to point out that a binary-quinary readout system effects further economies.

In Fig. 3, outputs from each stage of the counter are brought out to seven lamp-drive transistors arranged in binary-quinary form, and these translate the readout into denary code. The number of active devices used is thus 14, as apposed to 20 for a denary ring, and 18 plus and unspecified number in the decoding matrix for a *BCD* counter. Resistance logic can certainly be used in a very simple form for decades with up to 100kHz rates, and with slight additional complications up to 10MHz rates; these decade systems should effect major economies in the design of counters. The matter will be the subject of a future paper.

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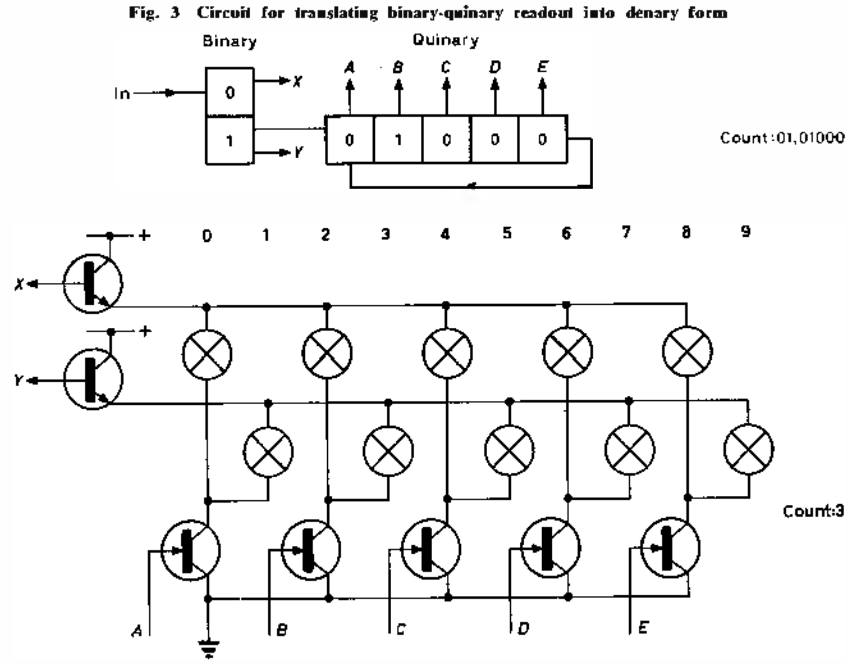
Fig. 3 Circuit for translating binary-quinary readout into denary form



NEW BOOKS

Fundamental Electronics

By A. S. P. Ledger and N. H. Roche. 594 pp. Blackie & Son Ltd. 1967. Price 75s. (board edition), 40s. (limp edition)

THIS textbook is intended for students pursuing the subject of electronics for the Higher National Certificate and Diploma, for CNAA degrees and for Part III of the IEE graduateship examination. It is divided into two parts: Part I (about 32% of the book) contains all the basic theories of physics and circuit theory necessary for a full appreciation of the principles behind the applications discussed in Part II.

Physical fundamentals are first considered; a wide range of topics is dealt with, including the electromagnetic spectrum, atomic models, energy levels, the periodic table, electron emission, conduction through gases, semiconductor theory and devices and electron ballistics. The remainder of Part I is concerned with circuit theory and includes accounts of self and mutual inductance, vector representation of sinusoidal quantities, use of complex numbers, properties of d.c. and a.c. circuits, network theorems, coupled circuits and transformers.

Part II, covering techniques, begins with discussions of the fundamental applications of thermionic and semiconductor diodes, triode and transistor characteristics, parameters and equivalent circuits, tetrodes, pentodes, h.f. transistors etc. Power supplies are next considered, with descriptions of rectification, smoothing circuits, voltage stabilization, thyratrons, thyristors, inversion and d.c. conversion.

Sections are then devoted to cathode-ray tubes, feedback, amplifiers, *RC*, *LC* and negative-resistance oscillators, frequency stability of oscillators and use of crystals, modulation and detection, trigger circuits, multivibrators and light-sensitive devices. Appendices give a list of relevant British Standards, prefixes, preferred values and colour codes, physical constants and definitions, data and formulas relating to semiconductor theory, mathematical formulas and some valve and transistor characteristics.

Problems (totalling 94) with answers are given at the end of each chapter. The book is liberally illustrated with 324 figures and gives an up-to-date presentation of the subject. It will be valuable to those students for whom it is intended. However, it contains very little which is novel; most of the material presented is adequately described in many well established textbooks. An attempt to cover a wide field has led (particularly in Part I) to a superficial treatment of some topics, with some wrong or misleading statements.

F. A. BENSON

Elementary Electronics

By D. Hywel White. 172 pp. Harper and Row. 1966.

THE basis of an elementary electronics course given to undergraduates at Cornell University is given in this book. In parallel with this text, a laboratory class runs at Cornell which enables students to try for themselves the adequacy and inadequacy of the ideas presented. The author believes that these practical extensions are imperative and has tried to provide equipment that allows a student to make demonstrations to himself without excessively detailed supervision.

The book begins with the first principles that are commonly used in understanding the operation of electric circuits, mentioning Kirchhoff's and Ohm's laws, Thévénin's theorem and the principle of superposition, and their application to d.c. circuits. A.C. theory and circuits, transient phenomena, potential dividers and differentiating and integrating circuits are discussed. Components commonly used in electronic circuits (insulators, resistors, capacitors, inductors, transformers, transmission lines) are then described and limitations met with frequently are pointed out. The properties of several measuring instruments are discussed. The devices covered are volt-ohm milliammeters, valve voltmeters, digital voltmeters, scalers, oscilloscopes and recording instruments.

Active circuit elements are next treated. The author now regards use of valves as a specialized branch of electronics, so emphasis throughout is on various semiconductor diodes and transistors. A simple approach to circuits in their quiescent state is developed, with particular attention to load lines, leakage current and transistor-biasing networks. Sections are then devoted to transistor equivalent circuits, amplifiers, feedback, *LC* and *RC* oscillators and switching circuits, including the multivibrator, tunnel-diode univibrator and Schmitt trigger. A chapter on logic circuits describes diode and transistor gates, analogue-digital convertors and the magnetic-core memory element. Finally, power supplies are considered with comments on rectification, filters, regulation and d.c.-to-a.c. conversion.

There are good problems (totalling 69, but without answers) at the end of each chapter, and a supplementary reading list is given. Only one minor misprint was noticed. This stimulating and lucid text begins with first principles and proceeds, with much novelty, to a level of reasonable skill in practical circuit and device design. It can be recommended to all undergraduates or other students taking an elementary electronics course.

F. A. BENSON

Magnetism in Solids

By D. H. Martin. 452 pp. Royal 8vo. Iliffe Books Ltd. 1967. Price £6.

THIS extremely lucid book aims to introduce newcomers to the field and provide a general account to those specialists who wish to widen the scope of their knowledge. The author has refused to be side-tracked into a discussion of techniques, whether theoretical or experimental, and the result is a well-balanced, up-to-date account of modern magnetic thought.

The opening chapter provides a good general account of the empirical characteristics of all types of ordered magnetic material. It is very pleasant to find alloy systems, especially those of rare earth metals, covered by such a survey. Chapters II and III are fairly conventional but serve the purpose of gathering together the underlying quantum theory of magnetism and its application to dia- and para-magnetism in solids.

Chapter IV deals with the molecular-field models of ordered magnetic solids and is extremely good. The author, while refusing to be drawn into the controversy, presents the background and current state of thinking on both localized and collective electron approaches to the problem of ferromagnetism. The cornerstone of most theories of ferromagnetism is the concept of exchange. Chapter V introduces the reader to this in 'simple' two-electron systems, and chapter VI extends these basic ideas to real magnetic solids. The two chapters provide an extremely illuminating review of the fascinating concept and consequences of exchange interactions.

Finally there is a very welcome account of excited states and statistical mechanics of magnetic materials. This chapter will do much in bridging the gap between the opening paragraphs of research papers and the technicalities which follow—one of the author's aims.

How refreshing it is to find a book, free from a historical approach, concerned almost wholly with intrinsic magnetization and only slightly with domain theory, a complete reversal of most works on magnetism in solids. The result is a readable, up-to-date book likely to appeal not only to newcomers, but also to the research worker in the field of magnetism.

E. H. THOMAS

Advances in Microwaves, Volume 1

Edited by L. Young. 400 pp. Med. 8vo. Academic Press Inc. 1966. Price £7

THIS new serial publication will be generally similar to the publisher's other already well-known Advances which cover a wide spectrum of disciplines in physical and biological science. It will contain critical reviews covering a wide range of topics of interest to the microwave engineer and physicist. The contributions will reflect the most significant contemporary progress in microwaves and will be written by experts who have played a major role in the development of the specific subject being covered. Each topic will be treated thoroughly and, as the series is not intended necessarily as a textbook, purely review material will be kept to a minimum. The volumes will, it is expected, be published at about yearly intervals.

Volume 1, containing 191 illustrations and 255 references, begins with a chapter on the design and fabrication of the accelerating structure for the Stanford 2-mile accelerator (by R. P. Borghi *et al.*, Stanford Linear Accelerator Center) which includes a historical review and a brief general description of the accelerator. Next follows a section on optical waveguides (by A. E. Karbowiak, University of New South Wales) and a lengthy treatment of directional couplers (by R. Levy, University of Leeds). The use of singular integral equations in the solution of waveguide problems is discussed (by L. Lewin, Standard Telecommunications Laboratories) and also the basic concepts and theorems of Lie algebraic theory and its application to microwave networks (by M. Pease, Stanford Research Institute). Finally, partially filled waveguides and surface waveguides of rectangular cross-section are considered (by W. Schlosser and H. G. Unger, Technische Hochschule Brunswick).

Articles already planned for future volumes include those on tunnel-diode devices, varactor parametric amplifiers and harmonic generators, high-power millimetre-wave sources, millimetre-wave astronomy and a variety of topics concerned with transmission lines and aerials.

'Advances in Microwaves' provides a permanent record of major developments in the field. It is invaluable to all scientists and engineers working directly with microwaves as well as to those in related fields who require a good general review of the particular topic described.

F. A. BENSON.

Introduction to Cybernetics

By Victor M. Glushkov. 322 pp. Med. 8vo. Academic Press Inc., U.S.A. 1966. Price 94s.

CYBERNETICS, as it is understood at present, covers a wide area, and the aim of this book is to introduce the reader to some of the more important aspects. In producing this work the author has drawn to a large extent from his lectures on cybernetics and mathematical logic given in 1959-1967 at the University of Kiev and at the Kiev Institute of Dissemination of Scientific and Engineering Knowledge.

To keep within the confines of one volume it has been necessary to limit the choice of material, firstly, to that capable of digital representation and secondly to those aspects whose mathematical theory has been adequately formulated. A reasonable list of references is given at the conclusion, the main interest of which is perhaps the inclusion of several Russian authors whose names may be unfamiliar to the reader.

In its broadest sense, cybernetics subdivides into two main fields, the first is concerned with the general theory of information procesing, and the second with the structure of the devices which process the information. The theory of algorithms underlies both aspects and forms the subject of the first chapter. Here, the author covers algorithmic operators and normal algorithms. Various algorithmic systems including the Kolmogorov-Uspenski scheme are discussed and the chapter concludes by introducing the very important concept of algorithmic unsolvability.

The restriction of the information alphabet to two letters leads naturally into Boolean algebra and functions which are discussed in the next chapter. Starting with the foundations of Boolean algebra, the author goes on to consider the synthesis of logical networks, minimization techniques, and the extension to the calculus of propositions.

The book continues with a chapter on the theory of automata. From the definition of an abstract automaton as an information processor which produces an output signal from an input signal it introduces the concept of internal state or memory and draws the distinction between Mealy and Moore automata. Then follow sections on the analysis, synthesis, and minimization of abstract automata. The chapter concludes with a discussion of the structural synthesis of finite automata making use of elementary circuit elements.

An automaton with internal states or memory exhibits many of the properties of learning and self modification in the sense that its output is dependent on the past history of its inputs as well as on the present input. The author introduces this aspect in chapter 4 and goes on to formulate a quantitative measure of self organization based on information theory concepts. A discussion of automata with random transitions and initial states then follows. The Markov chain arises quite naturally at this stage as a special case, and some of its more important properties are outlined. Other topics covered are pattern recognition leading to the concept of the α-Perceptron and logical classification or conditional probability machines. The chapter concludes with a discussion of a number of optimization problems related to self adaptation and control utilizing the methods of steepest descents and linear programming.

The electronic digital computer is introduced in the next chapter as an example of a general purpose automaton which may be programmed to carry out any arbitrary computational algorithm. A discussion of the structure of a typical modern machine leads to the concept of the microprogram unit. Programming is treated first of all in terms of elementary machine instructions and secondly in terms of the Universal Algorithmic Language Algol 60. The suitability of the language for describing self organizing algorithmic systems is illustrated by applications to the α-perception and to a model of the process of biological evolution leading to the formation of new species. As the author points out, the fact that these systems can be described in Algol implies the very important conclusion that they may be simulated on the general purpose digital computer.

The final chapter commences by extending the elementary mathematical logic of chapter 2 with the introduction of the predicate calculus to give a more powerful tool for the expression of complicated thought processes. The problem of decidability has been shown too insoluble for this calculus by Turing and Church. Discussion of this aspect leads to Gödel's theorem on the incompleteness of arithmetic.

The concluding section is perhaps the most stimulating in the whole book and deals with the automation of the process of scientific discovery. Gödel's arithmetic formalization of a mathematical theory shows that it is possible to write a digital computer program containing all the axioms and derivation rules of a theory. By means of a random search process with certain restrictions to limit the possibilities it is possible to get the computer to produce the proof of any given theorem. Other formalizations are possible and Hao Wang has used a Gentzen formal system for automating (by means of a digital computer) a large number of theorems both in the propositional and narrow predicate calculi. Although it is not possible to have a universal decision procedure the experiment shows that significant results may be obtained from partial decision procedures. An extension of the program to generate new and interesting theorems has so far not met with complete success owing to the lack of a suitable non-triviality criterion.

This book is aimed at workers in the field of applied cybernetics who wish to obtain an understanding of the underlying theory. The various mathematical disciplines are described in detail and it is felt that the book will be of more value as a work of reference rather than as a broad picture of cybernetics.

P. L. Owen

Einführung in die Anwendung kontaktloser Schaltelemente (Introduction to the Application of Contactless Switching Elements)

By H. Bühler. 158 pp. 122 ill. & tables. Royal 8vo. Birkhäuser Verlag, Basle and Stuttgart. 1966. Price sFr.32

This is volume 4 of a series of text books on electrical engineering and is intended as a survey of design and operation of switching elements without moving parts and their application in industrial electronics. It is aimed at engineers engaged in planning, construction and maintenance of industrial controls using such elements.

Machine Translation

Edited by A. D. Booth. 529 pp. Med. 8vo. North Holland Publishing Co. 1967. Price £9

The twelve chapters in this book consist of separate contributions by leading specialists in machine translation and mechanical linguistic data processing.

The contents enable the reader to compare the philosophical ideas and the relative merits of the approaches to this subject by the authors who come from Russia, Israel, Canada, Italy, Great Britain and America.

Transistors for Audiofrequency

By G. Fontaine. 384 pp. Post 4to. Philips Technical Library. 1967. Price £3 8s.

This book is divided into three chapters, the first one of which deals with concentration curves as a method of explaining both static effects and dynamic ones such as the Early effect.

The remaining two chapters are devoted to the response of the transistor to small and large signals, respectively.

Radio & Electronic Components Vol. I—Fixed Resistors

By G. W. A. Dummer. 242 pp. Demy 8vo. Sir Isaac Pitman & Sons Ltd. 1967. Price £2 10s.

This is the first volume of the Radio and Electronic Components series which has now been completed with the publication of the sixth.

'Fixed Resistors' originally appeared in 1956 and a full-scale revision has now been carried out. The comparison charts of characteristics have been replaced by the inclusion of data in the appropriate chapter.

Two new chapters have been added; one on resistors for integrated circuits and one on resistor reliability.

Worked Examples in Modern Physics Vol. I

By P. Rogers and G. A. Stephens. 126 pp. Demy 8vo. Iliffe Books Ltd. 1966. Price 18s. 6d.

Worked Examples in Modern Physics Vol. II

By P. Rogers and G. A. Stephens. 227 pp. Demy 8vo. Iliffe Books Ltd. 1966. Price 18s. 6d.

Primarily design for general degree students, the two volumes of the book contain questions from several British universities and colleges. The majority of the questions are fully worked out and the remainder have the answers given.

The subjects dealt with include atomic and electron physics, nuclear physics, X-rays and crystal structure, solid-state physics, wave mechanics and the special theory of relativity.

Engineering Science at the University

Edited by D. Hutchings. 238 pp. Demy 8vo. University of Oxford. 1967. Price 10s. 6d.

Full details of all the engineering courses available in the forty universities in the United Kingdom are given in this handbook for the benefit of schoolteachers and parents —and through them sixth-form pupils.

Information is given for each engineering department and includes the number of staff and students, subjects covered in degree courses, research interests and facilities, post-graduate studies and entrance qualifications.

As part of the Government's campaign to increase the number of professional engineers the Ministry of Technology has given financial support for the publication of this book and sufficient copies have been produced to permit free distribution of one copy to all schools with a sixth form and to all Youth Employment Officers.

Modelle und Ersatzschaltungen von Halbleiterdioden (Models and Equivalent Circuits of Semiconductor Diodes)

By W. Wunderlin. 61 pp. + 60 ill. Royal 8vo. Birkhäuser Verlag, Basle and Stuttgart. 1966. Paperback.

This is volume 26 of a series of text books and is written by a member of the staff of the Swiss Technical University at Zurich. Setting himself the dual task of bringing together models and equivalent circuits for diodes operating with low currents, he also develops a new lumped model with a single charge carrier as approximation of the continuous model of the path region. The booklet is considered as additional reading in conjunction with the first semiconductor booklet of the series and includes a useful bibliography.

Electronic Weighing & Process Control

By G. W. Van Santen. 271 pp. Med. 8vo. Macmillan & Co. Ltd. 1967. Price £3 18s.

The use and design of electronic weighing, batching and process control systems forms the subject matter in this book.

Attention is given to the theoretical principles and practical design of associated equipment together with methods whereby this equipment can be used to construct both both simple and complex weighing installations.

Advanced Electronic Techniques: Proceedings of the Second Oxford Electronics Lecture Course

Edited by G. W. A. Dummer. 142 pp. Demy 4to. United Trade Press Ltd. 1966. Price £3 10s.

This book contains the nineteen papers presented at the second Oxford Lecture Course held in July 1966. Among the subjects covered in this course are microwave generation, motor car electronics, hot electron devices and lasers.

Instruments, Electronics, Automation Year Book and Buyers' Guide

Third Edition. 729 pp. Morgan Brothers (Publishers) Ltd. 1967. Price £3

The third edition of this guide maintains the high standard set in the first edition. The largest section is the Buyers' Guide, arranged alphabetically under products, of which there are some 4600. Addresses are not included in the Buyers' Guide but a separate section, Manufacturers' Addresses, combines all the companies mentioned. At the back of the book are 22 equipment surveys, each written by a specialist, listing and describing in detail relevant manufacturers' products. In this larger edition four new surveys have been added: analogue computers, electric voltmeters, microelectronics and preset informatographs. All other surveys were published in the previous edition but have been revised and brought up to date in this edition.

Technical Drawing

By F. L. Gresecke, A. Mitchell, H. C. Spencer and I. L. Hill. 882 pp. Collier-MacMillan Ltd. 1967. 5th Edition. Price £4

This is a revised edition of a standard book of American practice, and it has been brought into line with the various sections of A.S.A. Y14—American Standard Drafting Manual. In addition to the main text on the technique of mechanical engineering drawings, electronic diagrams, maps and other topographical drawing, the preparation of graphs and monograms is described and an appendix containing American standards and symbols is given. The final chapter develops the principles of the graphical solution of equations and methods of integration.

Microelectronics

By G. W. A. Dummer: Morgan Brothers (Publishers) Ltd. 1967. Price 10s

Reprinted from the 1967 edition of Instruments, Electronics, Automation Year Book and Buyers' Guide, this pamphlet is a survey of microelectronic elements available in the United Kingdom. It is prefaced by a brief description of the different types of construction and their uses and limitations. The survey is classified according to the manufacturer and the information for each type is under four headings: circuit construction, range and availability, packaging, configuration and remarks.

The Radio Amateur's Handbook

By the Headquarters Staff of the American Radio Relay League. 634 pp. 1967 44th Edition. Price $5.50 in the UK

The chapters on radio communication in this well-known book have been brought up to date and equipment construction details include some new designs. Transmitters and receivers are described in detail for all the usual American amateur frequency bands, including mobile and portable operation. The use of solid-state devices is shown wherever they are applicable and information on the characteristics of valves and semiconductors has been updated.

SHORT NEWS ITEMS

London's International Broadcasting Convention, which takes place on September 20-22, 1967, at the Royal Lancaster Hotel, London, is attracting convention papers and delegates from all over the world. The convention, which includes an exhibition, is organized by the Electronic Engineering Association and the Royal Television Society. Papers to be read during the convention cover the whole field of television and sound-broadcasting techniques, transmitting, studio and outside-broadcast equipment.

The convention is also supported by the Institute of Electrical and Electronics Engineers, United Kingdom and Republic of Ireland Section. More detailed information and registration forms can be obtained from: The Convention Secretary, International Broadcasting Convention, Royal Television Society, 166 Shaftesbury Avenue, London WC2. Telephone: 01-836-3788.

The Polish Radar Authority is to offer Solartron video maps as standard accessories with the equipments it manufactures. The first airport to be supplied under the new agreement will be Schonefeld, East Germany. Two SY 2042 dual optic video maps will be used in conjunction with an Avia-B radar system. The order is worth £17 000 to Solartron. Two of the West's largest radar manufacturers, Plessey of Britain and Selenia of Italy, have already standardized on these video maps for sale with their systems. The latest order brings the total sales of Solartron maps to 106 in two-and-a-half years. Valued at almost £1 000 000, 47% of total orders have come from export territories.

The report of the IEETE council for the year ended the 31st March 1967 shows that the Institution made further substantial progress; the lecture programmes, both in London and elsewhere, were improved and extended; regional development work was expanded (five Regional Centres have now been established, four more will be set up by the end of this year); the designatory initials denoting corporate membership became more widely used; establishments of further education included the IEETE qualifications in their lists of nationally recognized awards; and the membership strength was increased by 1300.

Greece and Thailand are two of the most recent countries to order television cameras and transmitters from Pye TVT Ltd of Cambridge. The order from Greece is for 4½in image-orthicon cameras which will be the first professional television equipment to be imported by the Hellenic National Broadcasting Institute for use in their television station in Athens. In the case of Thailand, this represents a further phase in the development of its television network and follows the recent installation for the public relations department of a complete television station at Haadyai in the southern part of the country. The new contract calls for two 5kW band-III transmitters to be installed by the Army Television in the capital, Bangkok. These transmitters will augment the existing equipment supplied several years ago and will extend the range of the station in the Bangkok area.

More than 85 million trunk telephone calls were made in Britain in May, an increase of over 9% over the same month in 1966. Almost 60% of the calls were dialled.

The Mica Corporation of Culver City, California, have decided to open a factory at Newbridge, in Scotland. They will produce a range of epoxy-glass copper-clad laminates for conventional and multilayer printed circuits. The company will take over a 25 000ft^2 factory that is being built by Midlothian County Council, and hope to start production in Spring 1968. Initially, they will employ some 40 men who will be recruited locally for training. Mica spent several months examining potential sites, and finally selected Scotland after a rather intense evaluation of several potentially interested locations. Once having decided upon the UK, the growth of the electronics industry in Scotland provided a significant incentive for the company to locate north of the border.

The first full-length stereophonic play to be broadcast in Britain was heard on the 30th June in the BBC's third programme. The play, 'And now the barley', was well suited to stereophonic treatment, and the producer, Charles Lefeaux, showed considerable mastery of what is, in effect, a new medium. Obvious stereophonic 'tricks' were carefully avoided, and the result was a production which took advantage of, rather than being a vehicle for, the resources available. The occasional uses of the Radiophonic-workshop techniques and of stereo sound-effects were made much more effective by their relative infrequency. The play is to be the BBC entry for the Italia Prize (stereo section) in September.

The Postmaster General recently opened a 22 000ft^2 extension to Standard Telecommunication Laboratories Ltd at Harlow, Essex. The new addition brings the size of the laboratories up to 140 000ft^2, and makes STL one of the largest electronics research centres in the UK. As the PMG pressed a button to unveil a commemorative plaque, a laser-beam microcircuit-cutting machine cut a minute replica of the plaque on a piece of glass measuring only 120 × 70 thousandths of an inch in letters only 5 thousandths of an inch tall. It was afterwards presented to him at lunch by the director of the laboratories. The plaque unveiling and cutting of the replica was seen by the guests at the ceremony on closed-circuit television utilizing a 16 × 12ft large-screen projection system, and was also seen by the laboratory staff on monitors throughout the building. The new wing will be used for advanced circuit research of several types and the preparation of modern semiconductor materials. It will also house a completely new computer installation—an IBM 360/30F employing desk files—that will provide a scientific computing service to the whole of STL.

During May, the number of combined television and sound licences throughout Great Britain and Northern Ireland increased by 71 508, bringing the total to 14 463 221. Sound-only licences now total 2 540 904, including 803 463 sets in cars.

The White Fish Authority, in its annual report, states that, during the past year, the Industrial Development Unit, in conjunction with the Scottish Marine Laboratory, has tested a television camera in natural light at considerable depths, to observe the action of the trawl and the behaviour of fish in its vicinity. Also mentioned are further trials of equipment for acoustic telemetry of information from the trawl to the ship. The results so far obtained from this experimental work are considered to be encouraging.

A joint programme of research at Oxford and Reading Universities which may lead to the development of important new weather-forecasting instruments will be supported by a Science Research Council grant of £89 450. The grant has been made towards the cost of a world-wide survey of the temperature of the atmosphere up to about 30 miles (50 km). An instrument which has been developed and is being manufactured in its engineered form by

Elliott Brothers (London) Ltd, of Frimley, near Aldershot, Hampshire, will be mounted on the US satellite *Nimbus-D* as part of the international space-research programme of the US National Aeronautics & Space Administration (NASA). This satellite is due to be launched in late 1969. From a height of 600 miles, it will survey the greater part of the Earth and will measure the temperature of the atmosphere at various heights. Such information, continuously telemetered back to ground stations, will provide a better understanding of the nature of the atmosphere and should eventually lead to improved techniques of weather forecasting.

The instrument, known as a 'selective chopper radiometer', detects the infrared radiation from carbon dioxide in the atmosphere. Certain wavelengths in the infrared region can pass through the atmosphere more easily than others; so that, by choosing a suitable wavelength, one can see further down into the atmosphere below the satellite, and so select any layer of it for study. The *Nimbus-D* experiment will measure temperatures of six such layers, from cloud-top height up to about 30 miles, to an accuracy of about 1degC. The technique can be extended to the determination of the water-vapour content of the atmosphere, in addition to the measurement of temperatures. A further development of the basic radiometer aims at detecting clear-air turbulence (c.a.t.), which is the sudden and unexpected turbulence sometimes encountered by aircraft flying in perfectly clear air, and which cannot be detected by radar. For very fast jet aircraft, such as Concorde, c.a.t. may be potentially dangerous, so a reliable instrument carried on board must be capable of detecting it so far ahead that pilots have time to take evasive action.

The Boards of The English Electric Co. Ltd and Elliott-Automation Ltd 'have carried out a detailed examination as to the best method of co-operation for the purpose of exploiting and accelerating the very large potential growth in the field of industrial control, automation, computers and electronics in general.' They have decided that the industrial and commercial interest of both companies would be served to the best advantage if a complete merger were effected between them. Recognizing the importance to the national economy of rapid acceleration of growth in these areas, the Industrial Reorganization Corporation have confirmed their willingness to support it financially in a substantial manner. As a result of these discussions it has now been agreed that English Electric should offer to acquire the whole of the share capital of Elliott-Automation.

After exhaustive evaluation, Peto Scott plumbicon cameras were chosen for the BBC's first outside-broadcast colour units, because they give excellent colour performance and are smaller, lighter and easier to handle than the earlier 4-tube type. So far, eleven plumbicon colour cameras have been supplied to the BBC and six more are scheduled for early delivery. Fourteen others are already in use in Britain, including four each by ATV and Intertel, used mainly for recording plays and other programmes in colour, for sale abroad. In the plumbicon camera (its name is derived from the Philips plumbicon tube) the picture is focused onto the three pick-up tubes behind the lens, by means of a specially developed sealed prism system. This provides an efficient and compact means of separating the primary-colour components of the picture.

The performance of the camera shows that colour pictures of the highest quality are readily achievable with only three pick-up tubes. In place of the extra luminance tube, on which some designs rely to reach their registration performance, this camera uses an extension of special aperture-correction circuits by contour-addition techniques. Important features of plumbicon pick-up tubes are their fast response and small size. They have neither noticeable lag nor burn-in effects and give stable operation within minutes of switching on.

Radar units which fit together like a giant 'do-it-yourself' assembly kit are the prime feature of an entirely new series of radar equipment introduced by The Marconi Co. Almost any civil or military radar complex can be constructed from a large number of basic building blocks. These include aerials, transmitters and associated equipment, designed to make the maximum use of solid-state devices both conventional and microelectronic. Whatever the radar installation, the cost effectiveness of the scheme is superior to any other currently available, and Marconi believes it could mean tens of millions of pounds in export business for Britain in the next few years.

This new range of radar equipment incorporates 12 different types of aerial heads, five different transmitter/receivers and a complete selection of signal-processing equipment built in modular form. In addition, the Marconi Myriad computers can be integrated with this basic equipment, together with data displays, to provide advanced data-handling facilities to any system. The Secar secondary radar system can be incorporated, the aerial fitting onto any of the surveillance radar heads. Different combinations of the equipment can be used to form radar systems covering any of the following functions: ground control of interceptors (g.c.i.), tactical control for weapon systems, early warning and reporting, general air surveillance, military or civil air-traffic control, coast watching etc. Civil air-traffic-control systems, ranging from simple medium-range air-traffic surveillance to highly sophisticated air-traffic control by computer, can be supplied.

Considerable interest is being shown by the Australian dental profession in a new audio headrest which helps to relax patients and reduce discomfort. Developed by Audiorama Sales, South Australia, in conjunction with the Rola Division of Plessey Components (Australia), the device utilizes a technique known as audio analgesia, the principle of which is the use of music superimposed on, or mixed with, 'white sound'. This is a random combination of audio frequencies producing a noise similar to that of rushing water. Experiments have indicated that white sound combined with music can produce anaesthetic qualities in certain dental patients. Recorded on tape, the blend of music and white sound is amplified and transmitted to the patient. Previous systems have required the use of earphones or earpiece plugs; such systems introduce problems in regard to sterilization and inhibit normal conversation between dentist and patient.

The newly developed headrest incorporates two of the Rola division's latest model C3D ferrite magnet loudspeakers, providing maximum efficiency with minimum depth. Each speaker is incorporated in one wing of the headrest and provides effective sound transmission without restricting the patient's movements; sterilization problems are eliminated. Results indicate that the system minimizes pain and allows the patient to relax; this, in turn, enables the dentist to work more efficiently, with consequent reductions in treatment time.

The first edition of the International Directory of Translators and Interpreters (80s., Pond Press) lists 2120 addresses of technical and literary translators in 39 countries, offering more than fifty languages between them. The primary aim of this unique directory is to help in locating a 'mother-tongue translator' with special knowledge of any subject. This is done by a specially-devised 132-page index arranged under 22 main subject headings which shows at a glance each translator's native tongue, second languages and qualifications. No numbers are used. The index leads straight to a name that can then be looked up in an alphabetical list where full details of each translator are given. There are 224 entries under the heading 'Elec. Eng.'

Every country in Europe is now within reach of telephone users in Britain, since the 3rd July, when a round-the-clock telephone service began to Andorra (population 15 000). Andorra is the one European country to which a telephone service from Britain had not hitherto been available, although, for some years, it was possible to call Britain from Andorra by using the French network. The new service has been made possible by the connection of

Andorra's internal telephone network to the European grid through the French system. Any of the 700 subscribers in the Pyrenean principality may now be reached by telephone through the Continental exchange (in London, dial 104), where the operator will route the call onwards through Paris. Charges will be the same as for operator-connected calls to France.

The steady growth of Britain's fully-automatic Telex system continued over the past 12 months. At the end of May, 19 850 machines were connected and working—a 13% increase on the total of a year earlier (17 565).

The professional group on electronic valves and tubes of the Electronics Division of the Institution of Electrical Engineers, together with the electronics group and the optical group of the Institute of Physics & the Physical Society, are arranging a one-day colloquium on photoemission, to be held at the IEE, Savoy Place, London WC2, on Wednesday 20th September 1967. The colloquium, which will cover all aspects of photoemission, will be open to all who have an interest in the topic. Members of the Institution of Electrical Engineers or the Institute of Physics & the Physical Society may obtain tickets without charge; nonmembers should send a registration fee of 25s. with their applications for tickets. Applications for tickets (which will be available in late August) should be sent to the Secretary, The Institution of Electrical Engineers, Savoy Place, London WC2.

The GPO has awarded a contract in the region of £150 000 to GEC (Electronics) Ltd to supply, install and test intermediate-frequency and baseband equipment at their Goonhilly communication-satellite Earth station for association with the second aerial installation now in the course of construction. This follows the completion of a GPO development contract for prototype equipment by GEC at their Stanmore laboratories. The contract includes the supply of three modulator equipments for the transmission and reception of multichannel telephony. Any capacity from 12 to 252 channels may be accommodated on each equipment. Also being supplied are modulator and demodulator equipments for the transmission and reception of a television channel capable of handling colour programmes for 525- and 625-line standards, thus making possible the interchange of colour-television programmes between Britain and America.

A high degree of reliability is required, and each set of equipment is duplicated with facilities for automatic changeover in the event of any deterioration of quality in any of the channels. The equipments are completely transistorized to increase the reliability further.

Kent Controls Ltd (controls division of the George Kent Group) has been awarded the first instrumentation contract for the 1200 MWe advanced-gas-cooled reactor Dungeness-B nuclear power station now being constructed by Atomic Power Constructions Ltd for the Central Electricity Generating Board (Southern Project Group). This contract incorporates both the design and engineering work for boiler and gas-circuit instrumentation, reactor-coolant analysis, air monitoring and central-control-room panels and desks. Kent Controls have been associated with nuclear-power-station instrumentation in Britain from the inception of the commercial nuclear-power programme, including Calder Hall and each of the nuclear stations built or being built for the CEGB. In other countries, Kent have undertaken power-station instrumentation work at the Marviken nuclear power station in Sweden and nuclear stations in Italy and Japan, together with conventional stations in South Africa, Australia and Malaysia.

A look into the future of telecommunications was made by the Postmaster General when he opened in London a conference on European co-operation on telecommunications, organized by the British Council of the European Movement. He said that, by the end of the century, the telephone service would reach to practically every household and be a real maid of all work. All exchanges would be automatic, and computer techniques would be used to test and maintain the equipment, deal with operating difficulties, settle subscribers' queries and, especially in the international network, decide how calls should be routed. Push-button methods of dialling coupled with simplified selective codes for frequently called numbers, electronic switching and digital-code signalling would all speed up the service. 'The telephone will incorporate other new features: it could be a picturephone with loudspeaking facilities; and subscribers will be able to dial into computer libraries and get a playback of any audio or audio-visual tape available either on their picturephone or on their ordinary television sets. These libraries will cater for all shades of taste in education, entertainment and instruction and will play a vital part in helping to solve the problems of leisure time with which we will eventually be faced.'

Over 1700 members of the Institution of Electrical Engineers and their guests attended a Conversazione held on the 21st June at the Royal Festival Hall. The fullest use was made of the impressive amount of floor space at the Hall. In addition to two concerts of songs and piano music in the auditorium and dancing in the foyer, the promenades surrounding the auditorium were filled with a total of 69 exhibits and demonstrations showing recent developments in electrical science and engineering. Exhibits included automatic train-driving equipment, a machine for training electronic technicians in the logic of fault-finding, a recorded-commentary machine, a programme-effects generator, demonstrations of infrared optical-path communication and perspective drawing by computer. A collection of ten items featured medical applications of electronics, including an intensive-care monitoring system and various aids for the disabled.

A permanently installed system capable of providing simultaneous distribution of closed-circuit and received-broadcast television programmes throughout school buildings has been introduced by Teleng Ltd of Romford, Essex. Designed specifically for educational use, the system can carry, in one cable, live-camera programs originating within the school, together with nationally broadcast television and radio proprograms. If necessary teacher/pupil talkback can be provided using a separate cable.

The system allows each individual receiver outlet to be used also as a television-camera input point. This flexibility is possible because the system makes both ends of the cable accessible at any location. Whenever any one point is not in use for receiving or transmitting, a continuity plug is inserted to complete the circuit through the system and seal off the receptacle. This continuity plug can be replaced in seconds by any of the other plug-in devices to provide any desired function.

An airborne mapping radar is being used to provide information for the NASA Earth-science-resources programme. Developed for the US Army Electronics Command by the Aerospace Division of the Westinghouse Defense & Space Center, the mapping, or 'side-look,' radar has the ability to cover large areas in almost any kind of weather and at night. The aircraft does not have to fly directly over the area being mapped. Over the past two years, a large number of flights have been made to obtain data on remote-sensing techniques for the NASA programme. The radar provides high-quality radar images; by varying radar-signal characteristics and processing methods, scientists can use sidelook radar for such purposes as large-scale topographic mapping, agronomy investigation and making profiles of geological characteristics. Some other studies under way or being considered are sea-ice mapping, sea-state determination, polar-ice-cap profiling, subsurface profiling, soil moisture-content analysis and determination of undulations in the geoid.

Radar for geoscience research enables scientists to utilize the unique characteristics of another part of the electromagnetic spectrum—the microwave region—in addition to the visible and near-visible regions.

NEW EQUIPMENT

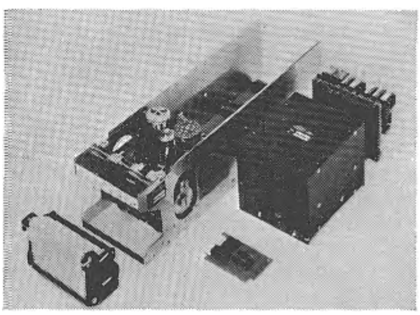
A description, compiled from information supplied by the manufacturers, of new components, accessories and test instruments.

MINIATURE POTENTIONOMETRIC RECORDER

Radiatron, 7 Sheen Park, Richmond, Surrey

(Illustrated below)

The Jaquet miniature potentionometric recorder is now available from Radiatron; it uses a unique form of construction. All the major component parts of the instrument are now produced in plug-in or clip-on form, so that they can be easily exchanged either in one instrument or between different instruments. These component parts include the chart magazine, chart-speed gears, dotting-speed gears, amplifier and measuring-circuit unit, range cards, range and channel-distribution unit etc. A single plug-in 'black-box' contains not only the amplifier section, but also all the measuring circuits required for the instrument to record from various



different types of input. This allows the Jaquet instrument to record mixed inputs simultaneously with up to six channels.

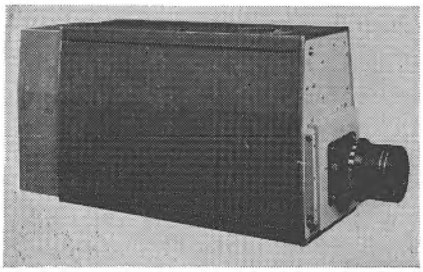
For more information circle **No. 281**

NEW CAMERA

Marconi Instruments Ltd, St. Albans, Hertfordshire

(Illustrated above right)

A revolutionary new British camera is announced by Marconi Instruments Ltd. The equipment employs television principles to present details, which, owing to low incident light, are difficult or impossible to discern with the human eye; this information is reproduced on monitors for direct viewing or recording. The system, which is a further development of a medical image intensifier



manufactured by the company, incorporates an entirely new type of highly sensitive camera tube—the Isocon—designed by the English Electric Valve Co. Ltd. The equipment has important military and industrial applications, and is capable of operating under full night-time conditions. It is an extension of Marconi Instruments' research since 1958 on medical and industrial image intensifiers.

The standard equipment (type-OA1705) comprises the following three units: image-detection unit (TF1706), master control unit (TF1702) and monitor (TF1703). The output from the image-detection unit, which incorporates the Isocon tube with its excellent light-sensitive qualities, is fed to the master control unit, where it is processed into a complete video signal conforming to CCIR standards. Either 625 lines or 1025 lines (50 fields/s) are selected by means of links in the control unit. On 60Hz operation, the line standards are 525 or 875. Optical systems to suit particular requirements can be supplied to order.

Two outputs are provided for monitors TF1703, which have been specially designed to ensure good geometry and stability of picture. The raster is normally 1:1 ratio, providing a square picture. Cine and still-photograph recording units are available, as well as video-tape facilities.

For more information circle **No. 282**

CLOSED-CIRCUIT TELEVISION

Thermionic Products (Electronics) Ltd, Hythe, Southampton, Hampshire

(Illustrated above right)

A comprehensive range of closed-circuit television systems and video-tape-recording equipment is announced by Thermionic Products (Electronics) Ltd. Manufactured by the Shiba Electric Co., Japan, the system, known as Shibaden equipment, was developed to meet requirements in industry, education, military, medicine, studio, domestic and other applications where a high standard of performance is demanded. Thermionic have been appointed sole UK representatives for Shibaden equipment.

All the Shibaden units are interchangeable, which allows many combinations of circuit and video-recording systems. A notable feature of the Shibaden video recorder, compared with most other recorders, is its complete compatibility between units, enabling tapes recorded on one machine to be replayed on any other unit. Such compatibility will allow the fullest use of video-tape

libraries for supplying prerecorded video tapes in the same way as film libraries, but with a reduction in cost, owing to the cheaper production processes involved. $\frac{1}{2}$in-wide tape is used, which makes running costs much lower than with 1in tape, but definition or clarity of the picture is equal or superior to many 1in-tape systems.

For more information circle **No. 283**

SWITCHED RESISTOR UNIT

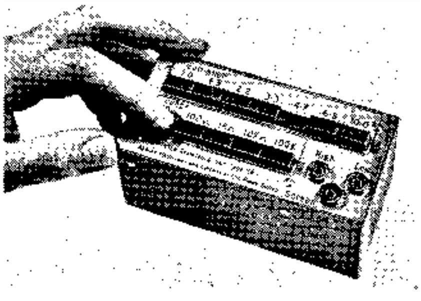
Applied Electronics and Automation Ltd, High Street, Ripley, Surrey

(Illustrated above right)

Applied Electronics and Automation Ltd are introducing a new range of inexpensive multi-instruments, designed especially for use in laboratories, workshops and educational establishments.

These are very compact and robust, with a durable stove-enamel finish, and the first in the range now available is a push-button switched resistor unit, which is being marketed under the trade name 'Pressistor'.

In the space of 3½in × 2½in × 5in (90mm



× 63mm × 127mm) there are 565 different resistances available instantly at the press of a button or buttons. Power dissipation is up to 2W and accuracy ±5% or ±2%. Voltage rating is 350V d.c. and current rating ¼A, with a noise level less than 0·4μV/V.

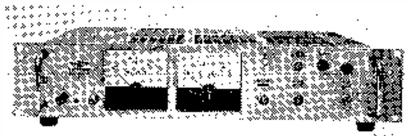
For more information circle **No. 284**

POWER SUPPLY

Livingstone Laboratories Ltd, Livingstone House, Greycaine Road, Watford, Hertfordshire

(Illustrated below)

A new series of wide-range power supplies specially designed for use in industrial and military systems calling for ultra-reliability has been developed by Trygon Electronics Inc., US specialists in precision power supplies and power modules. Known as the Liberator series, the new units—one of which is illustrated—are available in the UK from Livingstone Components Ltd. They represent a breakthough in compactness. Features of the units include all-silicon design, with matching differential amplifiers and integrated circuitry for extremely low noise and high stability. The units have a performance of 0·005% regulation and 0.01% stability. No temperature derating is necessary as there is full power to maximum temperature. They provide up to 70A in a 5¼in panel height, and 40A in a 3½in panel. The standard units are available in a wide range of models with constant voltage and automatic adjustable current limiting.



For applications that demand parelleling of supplies, an automatic 'load share' is provided as a standard feature. This permits interconnexion of the supplies so that each delivers an equal amount of current to the load, with single-knob control of up to four units.

For more information circle **No. 285**

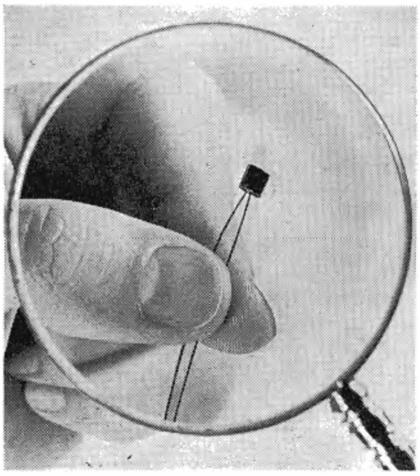
SUBMINIATURE TEMPERATURE TRANSDUCER

Thermal Systems Inc., Gardenia, California, USA

(Illustrated below)

A precision ceramic-encapsulated platinum temperature transducer for surface-temperature measurement over a range of −260°C to +400°C, with extremely fast response time and high standards of accuracy, is available from Thermal Systems Inc. Designated the 5001-19, the transducer measures only 0.2in square by 0·025in thick. It is constructed of reference-grade platinum wire wound strain-free onto a temperature-compensated ceramic mandrel to form the sensing element. Leads are 0.012in platinum wire which exit from a complete ceramic encapsulation of the mandrel and sensor, permitting use of the transducer in virtually any medium.

Resistance at 32°C is 500Ω (±1%) and the resistance/temperature relationship is linear over the entire temperature range. Coupled with an extremely fast response time of 250ms for a 63.2% change in resistance, corresponding to a



change in temperature, is a repeatability factor of ±0.1 degC after 20 consecutive thermal shocks from liquid nitrogen to ambient trichlorethylene.

The subminiature size of the 5001-19 makes it possible to make discrete temperature measurements on several points of extremely small surfaces. Its ceramic encapsulation provides a rugged housing for bonding to surfaces without danger of damage to the sensing element. Standard exit leads are Teflon insulated. Glass-fibre lead insulation is optional.

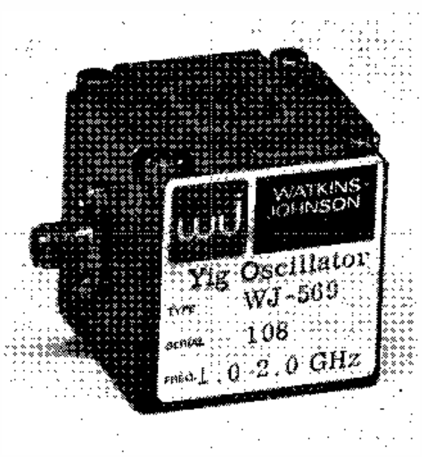
For more information circle **No. 286**

TRANSISTOR OSCILLATOR

Dale Electronics Ltd, 109 Jermyn Street, London SW1

(Illustrated above right)

Watkins-Johnson Co. has introduced the y.i.g.-tuned transistor oscillator for L-band as a new kind of compact solid-state microwave signal course. Designated the WJ-569, the oscillator's y.i.g.-tuning feature allows ±3% linearity over the full octave, while maintaining high temperature stability. Power output is 10mW minimum and residual f.m. is 10kHz maximum. A y.i.g.-driver supply is available for voltage-tuning from a



high-impedance low-voltage source. For applications requiring maximum frequency stability, a control circuit can be supplied to provide constant temperature at the y.i.g. resonator over a large ambient-temperature range. Solder terminals have r.f.i. shielding. The WJ-569 is small in size (1.4in cube) and weighs 12oz.

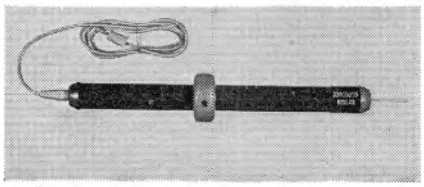
For more information circle **No. 287**

HIGH-VOLTAGE PROBE

Salford Electrical Instruments Ltd, Barton Lane, Eccles, Manchester

(Illustrated below)

Salford Electrical Instruments are now marketing two high-voltage probes which are intended to extend the range of their Minitest and Selectest Super-50, or any other make of 20 kΩ/V multi-range test instrument to 25kV or 30kV d.c. The probes are suitable for the measurement of e.h.t. voltages on both black-and-white and colour television sets and for similar electronic equipment where the source impedance is high. Each probe consists of an insulated housing and a lead provided with a 4mm plug. The maximum power consumption at full voltage on the 30kV resistor will be 1½W, the unit being rated at 9W. Temperature coefficient is less



than 0·15 per degC over the range 20-70°C, with an accuracy between 2·5 and 5% at full scale, depending on the tolerance of the instrument being used.

A minitest with a special 30kV scale is available for use with the probes. Dimensions are 13½in long with a 36in lead; barrel 15/16in diameter; guard 1¾in diameter.

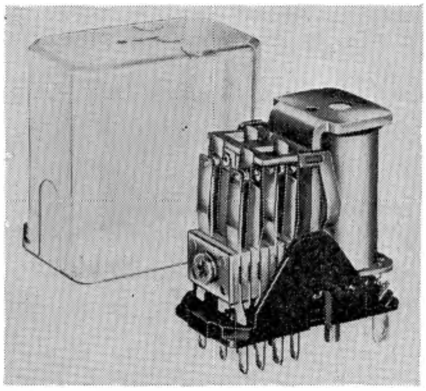
For more information circle **No. 288**

MINIATURE RELAY

Magnetic Devices Ltd, Newmarket, Suffolk

(Illustrated below)

A new addition to the off-the-shelf service operated by Magnetic Devices Ltd, is the Series-301/302 miniature telephone and control relay, which is now carried as a stock item. Incorporating improved design features, the relay is fully interchangeable with other types, both British and Continental, and special features include a full bridged mounting frame which overcomes base distortion, low-reluctance core/yoke junction giving improved sensitivity margin, and positive location of cover on yoke end, as well as base mounting. Series-301/302 is offered with up to four changeover gold-flashed fine-silver contacts and a range of coils up to 110V. It is available with socket and clip assembly or printed-circuit socket, or it can be stud-mounted for conventional wiring.



For more information circle **No. 289**

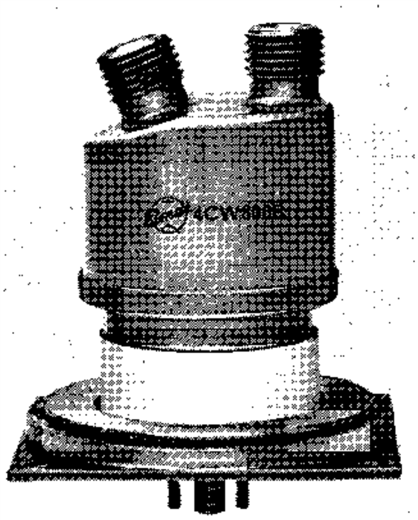
POWER TETRODES

Eimac Division of Varian, 301 Industrial Way, San Carlos, California, USA

(Illustrated above right)

Two new power tetrodes, the first designed specifically for distributed-amplifier applications, have been introduced by the Eimac Division of Varian. The 4CW800B and 4CW800F valves feature four individual circuit connectors to the active part of the grid, plus large cathode areas. Low-inductance characteristics typical of circuit requirements for distributed amplifiers are met in the rugged new valves by low-inductance supports for cathodes and screens; the 4CW800B and 4CW800F tetrodes are available with Eimac type-Y440 low-inductance screen-bypass capacitors.

Filament-voltage ratings of the water-cooled valves are 6·0V for the 4CW800B and 26·5V for the 4CW800F. The ceramic/metal valves are capable of 1000W output in linear amplifiers in class-AB_1 service. They incorporate ruggedized



electrode structures, and may be mounted directly onto the amplifier chassis for environments involving severe shock and vibration. Low inductance of the control and screen grids contribute to a minimum self-resonant frequency in excess of 900MHz.

For more information circle **No. 290**

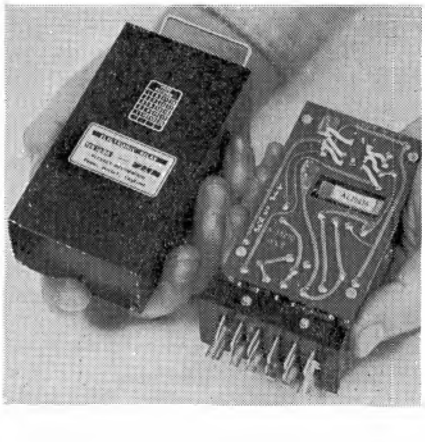
SOLID-STATE RELAY

Plessey Automation Group, Ilford, Essex

(Illustrated below)

A fully electronic solid-state relay (TER 10) has been developed by the Data-Handling Division of the Plessey Automation Group for use in the company's TDMS-5 series of telegraph equipment. The new relay has been designed primarily to replace the Carpenter electromechanical relay, type 3N1Z, in the TDMS-5 series. However, it can also be used for electronic conversion of voltage levels and as a general replacement for electromechanical relays under specific input and local conditions.

The TER 10, which has no moving parts and needs no maintenance or adjustment, eliminates distortion caused by contact bounce, bias and transit time. The switching elements comprise two thyristors which are switched sequentially by the input pulses. An autotransformer sends a reverse voltage pulse to switch



off one thyristor as the other is switched on, thus establishing a bistable condition. Should both thyristors switch on simultaneously, an internal relay switches them off, leaving the output to line 'open-circuit'. The next input transition restores correct signalling conditions to the line. Complete isolation of input and output is provided by an input transformer. The relay will operate with external telegraph supplies of 75-100V.

For more information circle **No. 291**

DIGITAL-OSCILLOSCOPE SYSTEM

Tektronix UK Ltd, Beaverton House, Harpenden, Hertfordshire

The Tektronix type-568 oscilloscope and type-230 digital unit comprise a new high-speed solid-state digital-oscilloscope system that provides digital readout of measurements that are displayed in analogue form on the c.r.t.. The type-568/230 has flexible measurement capabilities and features up to 50 measurements/s, easy external programming, BCD data outputs and solid-state circuitry, with extensive use of integrated circuits. The type-230 digital unit can make a wide variety of repetitive pulse measurements on the signals displayed on the type-568. The digital presentations can designate voltage measurements, time-difference measurements between percentages or voltages of pulse amplitudes.

The type-230 is designed to be externally programmed for use in high-speed measurement systems with up to 100 measurements/s possible with proper programming techniques. All of its measurement functions can be programmed by means of ground closures or logic levels. The programming is achieved with 104 program lines using negative logic, with true being ground or < 2V and false being open or > 6V. Suitable programming devices include card readers, block readers, computers etc.

Data outputs are available on the rear panel of the type-230 that permit the recording of measurement polarity, displayed digits, units of measure, decimal point and measurement-limit results. The information is in BVD code (1 2 4 8; true: ground; false: +12V). The type 230 can be synchronized to the data recorder, and regulated power is available for use in systems applications.

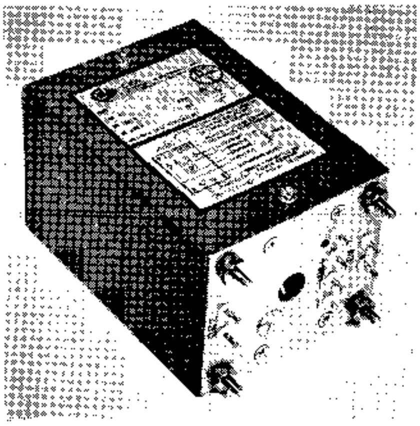
For more information circle **No. 292**

MODULAR POWER SUPPLIES AND VOLTAGE REFERENCES

B & K Instruments Ltd, 59 Union Street, London SE1

(Illustrated above right)

The ranges of modular d.c. power supplies and voltage references manufactured by the CEA Division of Berkleonics Inc. of the USA are now obtainable in the UK from B & K Instruments Ltd. These units offer total regulation, including stability, temperature coefficient and ripple, that permits use as a voltage reference. Standard features include regulation to 0·0005%, choice of exact centre voltage, 20dB isolation (up to 100dB available optionally), floating output terminals to ±500V and unique short-circuit protection. To meet various requirements, four classes of regulation are available. Guaranteed 40h stability of a class-D unit is better than ±0·0005% of the output voltage. For laboratory and standards applications etc. which require an absolute minimum ripple, modules can be provided with r.m.s. readings as low as 0·003%. Any output voltage from 3·5 to 500V may be selected at rated currents of 50mA to



25A. Ambient operating-temperature range is 0 to 80°C for the CEA-6 series and 0 to 65°C for the CEA-5, both series being capable of extensions to −55°C optionally.

For more information circle **No. 293**

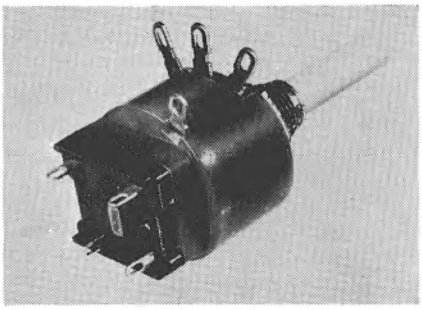
MOULDED-TRACK POTENTIOMETER

Davall Group Ltd, Gear Works, Potters Bar, Hertfordshire

(Illustrated above right)

The type-51 moulded-track potentiometer which has recently been introduced by Davall Electronics Ltd is a switched version of their existing type-50 potentiometer. It is fitted with a double-pole switch and various switching arrangements can be provided. The type-51 is specially suitable for high-surge-current applications including transistorized television equipment. The switch is an integral part of the potentiometer and it meets DIN and SEMKO specifications, as well as the new IEC draft specifications for potentiometer switches. The type-51 can be fitted with a screening plate between the track and switch if it is to be used for mains applications.

The specification of the type-51 is as follows: resistance range (preferred values) 50Ω-2MΩ (linear), 1kΩ-1MΩ (nonlinear); ±20% tolerance; power rating at 70°C 1W (linear), 0·5W (nonlinear); insulation resistance greater than 5000MΩ at 500V d.c.; maximum working voltage 500V d.c.; switch rating 2A at 250V a.c. The type-51 is 15/16in in dia-



meter and its length is 1 1/32in from mounting face to switch tags.

For more information circle **No. 294**

TRANSISTORIZED VEHICLE TRANSCEIVER

GEC (Electronics) Ltd, Communications Division, Spon Street, Coventry, Warwickshire

(Illustrated below)

The latest addition to the range of v.h.f. equipment available from GEC (Electronics) Communications Division is an all-transistor a.m. mobile transceiver small enough to fit into the normal radio cut-out in a vehicle dashboard. The unit (RC/660/TR) requires the minimum of connexions to a vehicle and operates from the normal 12V battery. It is extremely simple to use and is available in standard frequency ranges of 68-100MHz and 156-174MHz: other ranges are available to order. In addition to its application to vehicles, the unit is also available as a completely portable transceiver. Weighing only 10 lb, it is complete with rechargeable-battery power unit, loudspeaker and extendable whip aerial. In its portable form it is expected to meet many of the needs of commerce and industry as well as providing a useful addition to police, ambulance and security communications. Both units are the first all-solid-state equipment of this type to receive GPO approval.

For more information circle **No. 295**

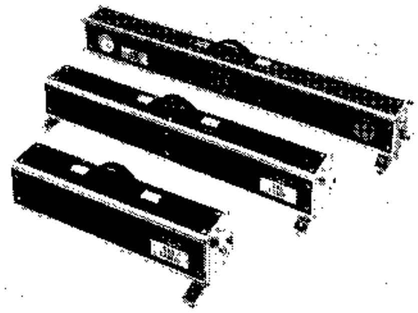
GAS LASERS

Scientifica & Cook Electronics Ltd, 148 St. Dunstans Avenue, Acton, London W3

(Illustrated below)

Scientifica & Cook Electronics announce the introduction of three new off-the-shelf gas lasers. The B18 gas laser has been developed by Scientifica to meet the specific requests for a laser that produces a high output at 6328Å, in preference to high outputs in the infra-red. This, however, does not preclude its use in infrared, for outputs at these wavelengths are obtainable provided that suitable mirrors are used. The B18 is normally supplied for use at 6328Å, although if desired it can be supplied for operation at 11 523Å or 33 912Å.

The B19 is a medium power uniphase d.c. gas laser whose power is adequate for the majority of applications where uniphase output is required. These fields include reflection and transmission holography, pattern identification, stress analysis, microdisplacement measurements, interferometry and numerous more. Its robust construction and freedom from alignment makes it particularly suitable for industrial applications. In this respect the laser has built-in dust seals that protect the unit from contamination; so that it may be operated under very adverse conditions.



As is common with the highly successful Scientifica B16/2 laser, the B20 has the unique feature of being demountable. This in practice means that the laser can be used as a standard self-contained unit with mirrors integrally mounted, or the laser may be operated so that access is available to the resonator cavity. The increased power of the B20 is obtained by using a higher-operating-voltage discharge tube altogether, with higher-transmission mirrors.

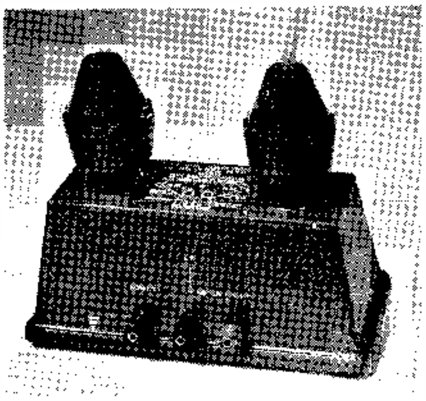
For more information circle **No. 296**

D.C. MULTIPLIER

Avo, Avocet House, Archcliffe Road, Dover, Kent

(Illustrated overleaf)

A new d.c. multiplier introduced by Avo (MI Group), for low-power high-voltage systems, is a combined 10kV and 30kV unit designed for use with the Avometer models 8 and 9 and the Avo testset (high-sensitivity) No. 1. It can also be used with any other multimeter having a range of 10V full-scale deflection and a sensitivity of 20 000Ω/V. The multi-



plier is particularly suited to the inspection and servicing of both colour and black-and-white television receivers, as it is specifically designed for use with the e.h.t. supplies encountered in r.f., e.h.t. and line-flyback units. A protective cap is supplied with the multiplier which must be used to cover whichever high-voltage terminal is not in use. Protection is provided by a resistor which limits the voltage across the terminals to 20V maximum should the meter be disconnected.

An electrostatic screen is provided inside the base; this is connected to a separate terminal which enables the user either to correct it or not, as required. A high-voltage lead is provided, the end of which is left free, enabling the user to select his own termination. A pair of meter leads and a low-potential lead are also supplied.

For more information circle **No. 297**

MINIATURE POTENTIOMETER

Plessey Components Group, Resistor Division, Cheney Manor, Swindon, Wiltshire

A high power rating (1W at 70°C) coupled with extremely small size are the outstanding features of a new wire-wound miniature preset potentiometer, designated type-WMP, from the Plessey Components Group. The element of this control is wound on a metal insulated former. This allows heat to be distributed throughout the body of the element, improving stability and giving smoother operation and a lower level of electrical noise. Standard resistance values are from 33Ω to 4·7kΩ at 10% tolerance. Acetal-resin knobs, available in a range of colours, enable the maximum ambient operating temperature of 85°C to be reached. Type-WMP potentiometers are available with centre termination for direct soldering into printed circuits.

For more information circle **No. 298**

TELEVISION-RELAY CABLES

British Insulated Callender's Cables Ltd, 21 Bloomsbury Street, London WC1

British Insulated Callender's Cables have introduced new ranges of television-relay cables, called 'pod' cables, which are the subject of a patent application. These distribute television programmes in the high-frequency bands to subscribers on their networks. The 'pod' cables incorporate a basic unit which consists of a video pair of cores accurately laid up with a second smaller pair of cores laid in the interstices; the smaller pair of cores can be used for audio-distribution purposes. This construction ensures minimum interference between the video pairs in the cable and results in a remarkably good crossview performance.

Pod cables are obtainable with four or six basic units unscreened, semi-screened with conducting polythene or fully screened with copper tape, and are manufactured by the BICC Wiring and General Cables Division, Helsby, Warrington, Lancashire. Full technical information will be available shortly in the form of technical data sheet TD WGCD 22.

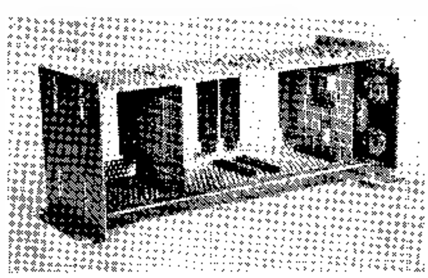
For more information circle **No. 299**

ELECTRONIC-CIRCUIT KITS

Shirehall Electronics Ltd, Borough Green, Sevenoaks, Kent

(Illustrated below)

The Duapack is the first item in the Shirehall Dualine series—a rationalized kit for the electronic or systems engineer, offering a range of basic parts to cover a wide field of application. The pack is designed to house plug-in cards with appropriate edge-connectors and is constructed in the form of two trays of perforated metal supported between end-brackets. These are slotted for adjustment to cover a range of card sizes and are available in different forms to suit rack or box mounting. Card guides are plastic-moulded with small studs which snap into the holes of the perforated metal. By virtue of the close-tolerance perforation, the guides and connectors are readily aligned, with a choice of fixing-centres on 0·2in (5·1mm) pitch, to suit the cards involved.



Initially, packs are available in four standard lengths: 16, 12, 8 or 6in (40·7, 30·5, 20·3 or 15·3cm), with cover-card sizes ranging from 2¾ to 4½in (7 to 11·5cm) in width and up to 3½in in depth. Other sizes will be added to the range later. The use of perforated metal combines strength and rigidity with good ventilation and is ideally suited to 'breadboarding' or subsequent modification. The pack is fabricated entirely in light aluminium alloy and supplied with an anodized finish. It should find wide application in industrial electronic and systems engineering, being equally suited to development and production requirements.

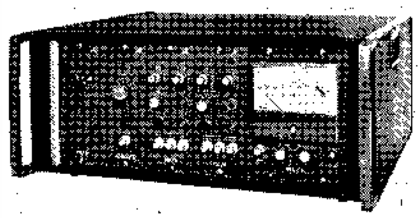
For more information circle **No. 300**

INSTRUMENTATION MODULES

Associated Engineering Ltd, Cawston, Nr. Rugby, Warwickshire

(Illustrated below)

New fully transistorized electronic instrumentation modules are announced by Associated Engineering Ltd for use with inductive transducers. Higher-frequency measurements of displacement and pressure can now be made with modules grouped to form either single- or multi-channel systems. For a multichannel



system, the modules may be mounted in a 19 × 7in rack unit, for which a carrying case is available if required. Similarly there is a smaller portable case for single-channel systems. All interconnecting leads are fitted in the rear of both units. A power-supply module feeds a complete single- or multichannel unit with all the necessary voltages. One oscillator and one detector/filter module give a single channel of measurement, while in the multichannel unit up to four detector/filter modules can be fed from a single oscillator. A 50 to 175kHz carrier signal is generated by the oscillator. When this is supplied to displacement or pressure transducers in full-, half- or quarter-bridge configuration, via the detector/filter unit, an output of ±2·0V with a frequency response from d.c. to 10kHz is obtained. For higher frequencies the output voltage is then fed into an oscilloscope or u.v. recorder. For lower frequencies, a meter unit is available which gives direct readout of static displacement or pressure and may also be used for bridge balancing, phase setting and calibration.

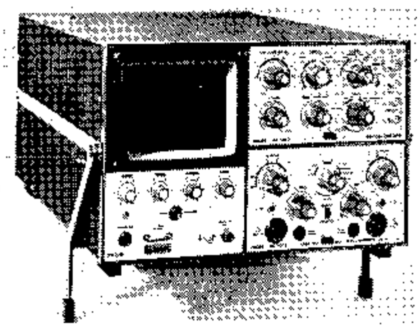
For more information circle **No. 301**

SAMPLING OSCILLOSCOPE

MEL Equipment Co. Ltd, Manor Royal, Crawley, Sussex

(Illustrated above right)

A new Philips high-performance sampling oscilloscope, which is claimed to be as simple to use as a conventional real-time instrument, is being introduced in Britain by the MEL Equipment Co. Ltd. The oscilloscope is designated type-PM3410. It is a dual-trace instrument with a sensitivity of 1mV/cm and a risetime of 0·35ns (1000MHz bandwidth). A 35ns delay line is incorporated in each channel and, with the exception of four Nuvistors, the circuitry is completely solid-state. The cathode-ray tube has a



10 × 8cm rectangular screen, all of which is available for display. Among the features which make for easy operation is a single trigger control which covers the whole range up to 1000MHz. Another unusual facility is a symmetrical, as opposed to the customary asymmetrical, sweep-expansion control. With this the trace can be conveniently expanded by up to 100 times about the centre of the screen. Timebase sweep speeds are from 1ns/cm to 10μs/cm, and the sampling rate is 10 000 or 1000 samples/cm. Provision is also made for scanning to be effected manually or by an external voltage. In addition, an internally generated slow ramp adjustable from 1 to 60s sweep allows a permanent trace to be made by an ordinary pen recorder. The instrument can be used in its case or be mounted in a standard rack. Dimensions are 8½ × 13¾ × 20½in.

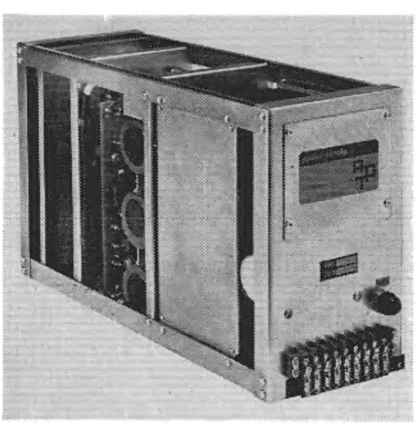
For more information circle **No. 302**

COMPACT H.V. POWER SUPPLY

APT Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

(Illustrated below)

APT has added to its range of power supplies a new high-voltage model employing all-silicon transistors. The new d.c. subunit, designated model-17580, is rated at 250V, 1A, and is of modular construction, making it very compact for its output. It measures only 8½in high × 8¼in wide × 14⅝in long. Claimed to have a stabilization ratio of greater than 10 000 : 1 with a temperature coefficient of 0·0005% per degC, the output resistance of the unit is not greater than 0·5mΩ. Its output impedance is less than 0·1Ω to 200kHz and 0·5Ω to 500kHz, and ripple and noise are less



than 1mV. A safety feature is built into the power unit which will cause the voltage to trip if more than 120% of the full-load rating is drawn. Model-17580 is rated for continuous use under worst conditions up to 60°C and its outer casing is finished in passivated bright zinc.

For more information circle **No. 303**

MONOLITHIC MULTIVIBRATOR

Raytheon Overseas Ltd, Lexington, Mass., USA

Raytheon's latest monolithic silicon integrated circuit is a threshold-triggered monostable multivibrator designed for use with zero-temperature-coefficient timing components. Designated the RM288, the device has excellent pulse-width stability from −55°C to +125°C. Its temperature coefficient is guaranteed at 0·005ms/deg C (±2%) for an output pulsewidth of 50μs. The RM288 is available in four package configurations: 10-pin O, 12-pin T, 14-pin dual-in-line, and 14-pin J. It is compatible for 200-series and 930-series DTL (digital transistor logic) integrated circuits.

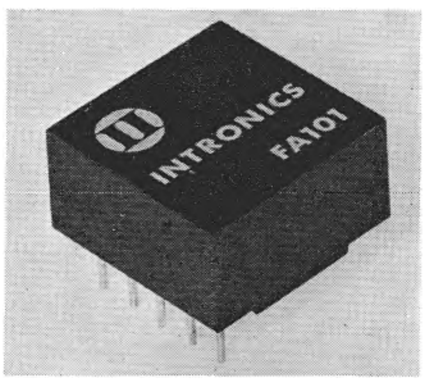
For more information circle **No. 304**

UNITY-GAIN AMPLIFIER

Intronics Incorporated, 57 Chapel Street, Newton, Mass. 02158, USA

(Illustrated below)

This encapsulated precision d.c. to 200kHz field-effect amplifier features extremely high input impedance, very



low output impedance, very high accuracy and excellent temperature stability. It is specifically designed for noninverting unity - gain impedance - transformation applications. The design includes protection against output short circuits and accidental supply-voltage reversal. No zero adjustments are required. Specifications are $10^{10}\Omega$ minimum input impedance, 50pA maximum-input offset current, output ±10V at 5mA maximum, output impedance less than 10mΩ, linearity better than 0·005%, maximum zero-output offset error ±300μV, output-voltage drift 50μV/degC, operating temperature −25 to +85°C, full output to 200kHz, power requirements ±15V d.c. at 11mA maximum, size 1·12 × 1·12 × 0·62in. It meets MIL standards.

Applications include: signal buffering, instrumentation read-in and read-out, sample-hold circuits and low-impedance signal transmission.

For more information circle **No. 305**

COMPACT I/C POWER MODULES

Powertec Inc., 9170 Independence Avenue, Chatsworth, California 91311, USA

(Illustrated below)

Powertec has introduced a new series of power modules, called Mini-Mod, which are one-fifth of the size of cur-

rently available devices with comparable output. These are designed especially for use in lightweight electronic chasses with I/C or other digital logic. The Mini-Mod's performance-to-size ratio (all modules are 4in × 2½in × 2¾in including heat-sink) is achieved through the elimination of all 60Hz magnetic components normally used in these applications. Many voltages are available: input 115V 47-440Hz; typical outputs are 3V at 6A, 5V at 5A, 7·5V at 3·6A, all at 0·1% regulation.

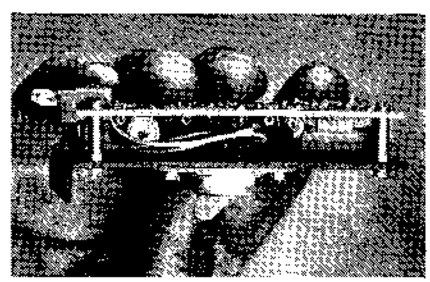
For more information circle **No. 306**

PLUG-IN SERIES-REGULATOR MODULE

Startronic Ltd, 117-119 Malden Road, New Malden

(Illustrated below)

A new easy-to-get-at series-regulator module, primarily designed for use with integrated circuits where close stacking of a multiple number of such units is required, has been introduced by Startronic Ltd, a member of the Stow Electronics Group, Hastings, Sussex. Standing only 3¾in high, this s.r. module M147 is claimed to be the first unit of its kind fitted with its own plug-in socket which allows for quick nonsolder mounting.



Provision is made in the base bracket for additional securing by screws, if desired. Nonencapsulated, the module can easily be adjusted or repaired, the accessibility of its components and its simple removal as a unit greatly facilitating the location of circuitry faults.

The maximum input voltage is 25V d.c. The minimum input voltage is 2V more than the required output voltage; these limits include peak ripple. The output is 9 to 15V d.c. at 0 to 1A. The output voltage is determined by an external resistor connected across the socket. A rectilinear potentiometer is incorporated in the module which gives a fine adjustment of approximately ±0·5V. the output current is limited at approximately 1·2A to protect the module against overload and short circuit. By changing one resistor (which can be mounted externally and wired to the socket), the output current may be limited to any value less than 1A, to provide protection for the circuitry being supplied by the module. The module is $2 \times 1^{1}/_{16} \times 1\frac{3}{4}$in.

For more information circle **No. 307**

FREQUENCY STANDARD

Rohde & Schwarz, Pressereferat, 8000 München 8, Muhldorfstrasse 15, Germany

The Rohde and Schwarz frequency and time standard type-CAQA, which is equipped exclusively with silicon planar transistors, contains a shock-mounted 5MHz harmonic-mode crystal of high precision (10^{-10}) and minimum drift in continuous operation. Drifts of $\Delta f/f < 5 \times 10^{-10}$ per day and a short-term stability of $\Delta f/f < 4 \times 10^{-11}$ for measurement times of 1s are guaranteed. The standard delivers sinusoidal voltages of 50/60Hz, 1kHz, 100kHz, 1MHz and 5MHz, as well as square-wave pulses between 1Hz and 10kHz in decade steps. The 5MHz output has a high spectral purity, the signal/noise ratio being better than 85dB. A goniometer phase shifter which provides digital readout serves for adjusting the phase of the frequencies from 1Hz to 10kHz, the seconds and minutes contacts and the clock movement, to whatever values are desired. The phase shifter is calibrated in units of 10µs. Type-CAQA is notable for its pulse outputs for 1Hz to 10kHz. Time markers in arbitrarily chosen decade steps can be generated by connecting the pulse outputs in parallel. The portable standard can be fed, without switching, either from an a.c. supply of 75-150V or 150-300V (47-1000Hz) or with a d.c. supply of 11-32V: in both cases the incorporated battery is charged. This battery of long service life permits operation up to 36h without external feeding.

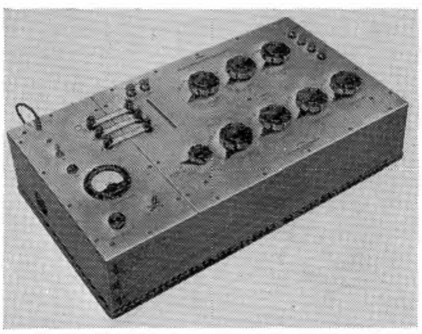
For more information circle **No. 308**

RESISTANCE BRIDGE

Croydon Precision Instrument Co., Hampton Road, Croydon, Surrey

(Illustrated below)

This bridge was designed to be as compact as possible and retain all of the features normally found in the larger bridges. It has seven dials with a measuring range of 111·1211Ω in steps of 0·00001Ω. When used with a 25Ω platinum resistance bulb, 10Ω is equal to 100°C. The absolute accuracy is better than ±0·01%. An accuracy of ±5 parts/10^6 is attainable if the bridge is calibrated with a National Physical Laboratory class-S resistance standard: under such conditions, temperature can be measured to 0·002°C. For comparative measurements the resolution of the bridge is 0·0001°C.



The bridge is mains-operated, having a built-in transformer rectifier and smoothing system. The total available bridge voltage is 15V d.c. with a polarity-reversing key which is provided with an OFF position. Thermometer current is indicated by a 0-5mA milliameter and an adjustment is provided. The thermostatic control of the bridge at 30°C is operated by means of a contact thermometer operating a relay *via* an amplifier circuit. Air is circulated through the bridge by means of a motor-driven fan blowing across the heating element. This is a recirculating system which is sealed from the atmosphere. In the event of the ambient temperature being higher than 30°C, a cooling coil is fitted which can be fed from a water main. A thermometer is fitted in the critical section of the bridge to satisfy the operator that the control system is functioning satisfactorily.

For more information circle **No. 309**

M-BAND REFLEX KLYSTRONS

Varian Palo Alto Tube Division, 611 Hansen Way, Palo Alto, California 94303, USA

Reflex klystron oscillators in the new VA-310 series, now available from the Palo Alto Tube Division, are designed primarily for use as pump tubes for parametric amplifiers and as transmitter tubes in point-to-point communication systems. These 7 oz tubes, operating in the 10-15GHz range, are available in two versions; one version offers an output of at least 1W (typical output is 1·25W) over a 500MHz mechanical tuning range. The other version offers a greater output of at least 1·5W over a 100MHz tuning range. External cavity tuning produces very stable operation. Life of these tubes is exceptionally long: they are warranted for 5000h. Cooled by forced air, the tubes mount in any position. The r.f. output flange mates with the flange for WR75 waveguide. Dimensions of the tube including the flange are 1·5 × 2 × 1·6in.

For more information circle **No. 310**

MULTIPLE TAPPED DELAY LINES

Johnson Matthey Metals Ltd, 81 Hatton Garden, London EC1

Johnson Matthey Metals Ltd announce the introduction of tapped Silver-Star delay lines for use in computer circuits and other specialized electronic applications. Each unit consists of interconnected delay lines encapsulated in a block of minimum size. They are manufactured to the same standard as other Silver-Star units. The delay-line leads are equally stepped at 0·2in (5·1mm) intervals to suit printed-circuit board. The standard range fulfils the requirements of most applications, but the company will produce other units and one-off prototypes at nominal cost, for assessment.

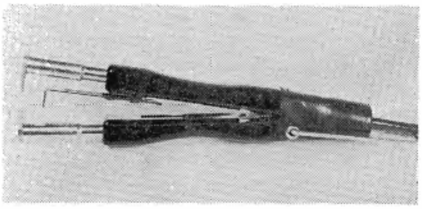
For more information circle **No. 311**

THERMAL WIRE STRIPPERS

Landis and Gyr Ltd, Victoria Road, North Acton, London W3

(Illustrated above right)

New thermal wire strippers are now available from Landis and Gyr Ltd. Manufactured by an associate company in Switzerland, model AXA-1 is suitable for operation from 230V a.c. supply and dissipates 13W. The jaws are operated by tweezer action and will accept wires from 2 to 25mm in diameter, while a readily adjustable incorporated stop allows for uniform length stripping. The



strippers comply with the Swiss standard SEV for quality and are ideal for stripping plastic insulation without damaging the conductor.

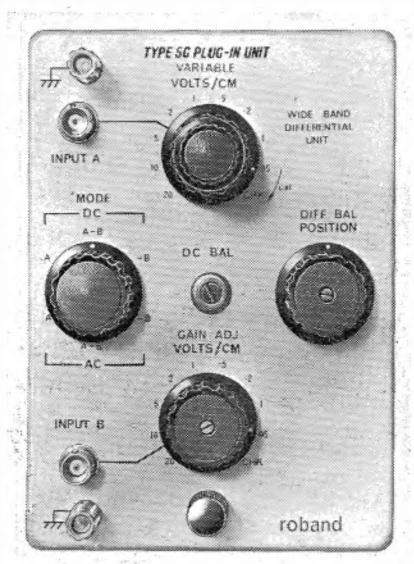
For more information circle **No. 312**

PLUG-IN DIFFERENTIAL AMPLIFIER

Roband Electronics Ltd, Lowfield Heath Road, Charlwood, Horley, Surrey

(Illustrated below)

Roband announce an addition to their already extensive range of plug-in units for their RO50A and RO55 oscilloscopes. The new plug-in unit, the 5G, is a wide-band differential amplifier unit. Used with the RO50A, it offers a common-mode rejection ratio of 100 : 1 at the full bandwidth of 23MHz and a basic sensitivity of 50mV/cm. The common-



mode rejection ratio at frequencies below 5MHz is better than 1000 : 1.

Costing only £65, the 5G offers good value. Used with the RO50A the total cost of instrument and plug-in 5G is £363. With the RO55 both instrument and plug-in cost £292. Full technical data is available on request.

For more information circle **No. 313**

NETWORK-ANALYZER SYSTEM

Hewlett Packard Ltd, 224 Bath Road, Slough, Buckinghamshire

(Illustrated above right)

A simple new system quickly sweep-tests for gain, phase, impedance, admittance and attenuation from 110MHz to 12·4GHz. Hewlett-Packard's network-analyser system will completely characterize active and passive devices, individually or in arrays. Magnitude and phase of all network parameters are presented, in any of three convenient forms: (1) phase and amplitude relations may be read directly by meter, (2) swept phase and amplitude responses simultaneously *versus* frequency, with 60dB dynamic range, on a conventional oscilloscope, and (3) alternately, on a cathode-ray plug-in for swept polar displays of phase and amplitude. Smith-chart overlays are provided for direct impedance or admittance readout on the polar display unit.

The new system coherently samples the signals to be compared, produces 20MHz replicas, and then measures their amplitude and phase relationships. The sampling principle used for hetero-

dyning, with a high-gain i.f. amplifier, results in high sensitivity. This linear-detection technique provides 60dB of dynamic range while eliminating square-law restrictions. No amplitude-response shaping adjustments are needed. R.F. power in the measuring channel may vary from −10 to −80dBm. The reference channel needs only −20 to −40dBm, well within the capability of readily available signal sources. The system easily attains resolution equal to that of manual measurements. Magnitude relations are resolved to 0·1dB, phase resolution is 0·1° for any angle over the full 360°. Furthermore, there is no ambiguity over phase lead or lag.

For more information circle **No. 314**

HIGH-CONTRAST STORAGE TUBE

Westinghouse Electric International Co., Zoo Park Avenue, New York 10017, USA

A new high-contrast display storage tube available from Westinghouse Electric International represents a significant advancement in the state of the art for modern aircraft and other display applications. The tube (WX-31016) uses a design that permits high-contrast operation; this means essentially that background brightness is made independent of erase duty cycle and persistence. The elimination of background brightness makes the tube ideal for new dual-mode radar displays, since it offers more halftones than conventional tubes under similar operating conditions. It has a minimum erase time of 7ms and is capable of displaying nine shades of grey. The increased halftone capability makes these tubes useful in applications requiring television displays in addition to radar.

Other applications include navigation, search and fire-control airborne radar displays, and ground-based or shipboard radar for air-traffic control. Two electrostatically focused and deflected writing guns in the WX-31016 afford a wide range of operating modes while offering an alternative television display.

For more information circle **No. 315**

SILICONE ENCAPSULENT

Dow Corning International Ltd, 25 rue de la Loi, Brussels 4, Belgium

A new flexible room-temperature-curing silicone encapsulating resin for electronic applications, with more than three times the tear strength of its predecessors, has been developed by Dow Corning. Tear strength of the pourable 2-part resin is typically 75 lb/in, compared with about 10-20 lb/in for older silicone resins of the same type. Elongation has also been improved greatly over previous products from 100 to 350%. Specific gravity of the resin is relatively low at 1.12, and tensile strength averages 700 lb/in². Designated Sylgard-187 encapsulating resin, the new material is useful over a temperature range of −65 to 250°C. It does not shrink when cured at room temperature, and is inherently flame resistant. It can be cured in unlimited thicknesses and requires no postcure to operate in confinement at high temperatures.

For more information circle **No. 316**

TUNABLE-FILTER MODULE

AIM Electronics Ltd, 71-73 Fitzroy Street, Cambridge

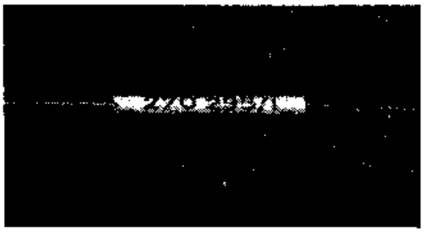
A tunable filter (TFO129) is the latest addition to the AIM Electronics range of instrumentation modules. These modules are 'black-box' building bricks by means of which anyone can assemble the most sophisticated test instruments. The modules fit side-by-side in standard cases, from which they draw their power, and they are interconnected by front connexions. Each module may be used in many different applications.

Applications of the TFO129 include coherent detection, distortion measurement, spectrum analysis, filtering of e.e.g. and e.c.g. signals, and sound-effect generation. In the educational field, the filter can demonstrate the components of electrical waveforms and the response of tuned circuits to step functions at very low frequency, so that the student can observe effects on a meter or galvanometer. The new slide-in module (2in wide × 7in high × 10in deep) will accept, reject or generate any frequency from 1·5Hz to 150kHz. As a filter, its Q-factor is continuously variable from 5 to more than 250 at all frequencies, and in fact users have experienced stable Q-factors or more than 5000 in normal use. A particularly valuable feature is the provision of a simultaneous quadrature output in the accept or oscillate modes. This enables the user to plot Nyquist diagrams directly on xy recorders, and to examine quadrature signals in phase-sensitive detection systems.

For more information circle **No. 317**

THERMAL FUSE
GE Industrial Supplies Ltd, Eardley House, 182/4 Campden Hill Road, London W8
(Illustrated below)

Wickmann Werke A.G. of Germany have developed a new miniature thermal fuse for the protection of any temperature-sensitive components. This device, which is in a special heat-resistant plaster sleeve, is 26 × 2·3mm; the fusing element is in the middle with a metal tube at each end into which the wires of the device to be protected are crimped. It is water-, vibration- and shock-proof and its function is not affected by the position in which it is used. The range consists, at present, of three temperature ratings, 155°C, 190°C and 210°C, all with a tolerance of ±5%.



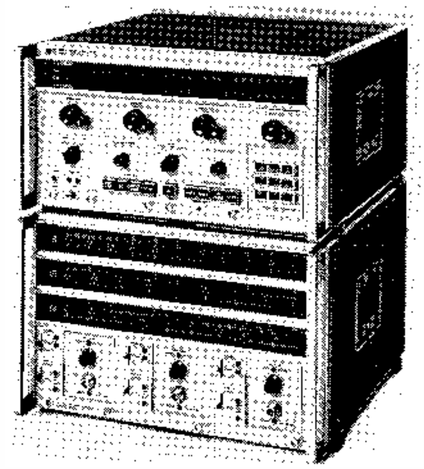
For more information circle **No. 318**

MICROWAVE SWEEP-OSCILLATOR CONTROL
Hewlett Packard Ltd, 224 Bath Road, Slough, Buckinghamshire
(Illustrated right)

The first simple and relatively inexpensive solution to the problem of broad-band-sweep capability (more than an octave) is offered by Hewlett-Packard's new model-8706 A control unit, with model-8707A r.f. unit holder. Used with model-8690A sweep oscillator and appropriate r.f. units, a compact bench-top multiband source is formed. The control unit, with its nine band-selector buttons, replaces the usual BWO r.f. unit as the plug-in for the sweep-oscillator main frame. It supplies power for, and controls, as many as three r.f. unit holders, each of which accommodates three r.f. units (single-band sweep oscillators).

Each r.f. unit holder provides coverage for three bands with three r.f. unit plug-ins. The sweep range of each plug-in is approximately an octave in width; available plug-ins can cover from 1 to 40GHz, and one has a range that extends from 100kHz to 110MHz. Units may be programmed, either by front-panel control-unit pushbutton selection, or sequentially by remote contact closure to ground. Switching time between bands is 1s. Between switching cycles, b.w.o.s are maintained on a standby basis, so that there is none of the degradation of b.w.o. life that otherwise results from fast start-up.

Flexibility of available sweep is broad; if normal sweep is selected on any of the r.d. units in the r.f. unit holder, the r.f. unit sweep is controlled by the sweep oscillator main frame and can be operated in any of the usual sweep modes. The frequency scale on the main frame is calibrated from 0 to 100 to show the percentage of the plug-in range that is being swept. If the preset mode is selected for one or more r.f. units, these will sweep between pre-adjusted frequency settings, made on screwdriver controls adjacent to each r.f.-unit plug-in, in the unit holder. All power is supplied by the sweep-oscillator main frame, and controlled by the control unit; cable interconnexions are part of the 8706A/8707A package. No modifications are required, either to the sweep oscillator main frame or any of the plug-in r.f. units, many of which are already in the hands of users.



For more information circle **No. 319**

Data Sheets and Catalogues

Photoelectric-cell-operated relays
Hird-Brown Ltd, Bolton, Lancashire

This brochure contains details of the 1967 range of photoelectric equipment, full information, diagrams and photographs of ten photorelays and a selection of time-delay relays. The brochure is referenced J135; apart from general information, it includes details of possible arrangements of photocells and relays in practical applications.

For more information circle **No. 320**

Contact-lubricant applications
Electrolube Ltd, Oxford Avenue, Slough, Buckinghamshire

Electrolube have issued an application guide showing how their various products can be used. There is an introduction to the theory of contact lubrication and methods of applying are also indicated.

For more information circle **No. 321**

Component catalogue
Welwyn Electric Ltd, Bedlington, Northumberland

The new Welwyn 224-page catalogue gives comprehensive details on all products; for the first time, a 50-page section is devoted to the range of hybrid integrated circuits. Film resistors and special-purpose resistors also receive full coverage. A general-information section gives a cross-reference of Welwyn products against national/international, Post Office and Defence specifications.

For more information circle **No. 322**

Microwave signal generators
Polarad Electronic Instruments, 34-02 Queens Boulevard, Long Island City, New York 11101, USA

A new bulletin from Polarad describes two microwave signal generators, one covering 0.95-2.4GHz and the other covering 2.0-4.6GHz. Complete specifications, descriptions and application information are included.

For more information circle **No. 323**

Introduction to teleprinters and punched-paper tape
Creed and Co. Ltd, Hollingbury, Brighton, Sussex

This booklet contains a brief introduction to the general principles underlying the operation of Creed teleprinters and punched-paper-tape equipment. It does not contain any details of the nontelegraphic applications that have been made of this equipment, nor does it give any engineering details of circuits or machines (i.e. it is concerned only with principles).

For more information circle **No. 324**

Precision potentiometers
Beckman Instruments Ltd, Glenrothes, Fife, Scotland

Helipot series-7360 and 7460 precision potentiometers are the subject of a new data sheet from Beckman Instruments. Sheet 67724 gives descriptive material, specifications and design diagrams of the compact 3- and 5-turn units.

For more information circle **No. 325**

Vidicons
English Electric Valve Co. Ltd, Chelmsford, Essex

This leaflet has been produced to summarize the range of English Electric Vidicons in a handy reference sheet. Full data sheets are available on any of the types mentioned in the leaflet.

For more information circle **No. 326**

Industrial valves and tubes: Equivalents guide 1967/68 Quick-reference guide 1967/68
Industrial semiconductors: Quick-reference guide 1967/68
Industrial components: Quick-reference guide 1967/68
Mullard Ltd, Industrial Markets Division, Mullard House, Torrington Place, London WC1

The Equivalents Guide presents in 93 pages an index of all Mullard industrial valves and tubes, together with a comprehensive guide to equivalents. The reference guides present quick-reference data on the products indicated. Product information has been abbreviated.

For more information circle **No. 327**

Résumés des Principaux Articles

Amplificateurs unissant des transistors bipolaires à des transistors à effet de champ par W. Gosling

Résumé de l'article aux pages 478 à 482

Les combinaisons de transistors bipolaires et de transistors à effet de champ peuvent produire des amplificateurs à gain élevé et à faible impédance de sortie. On distingue deux 'familles' de circuits: celles où le transistor à effet de champ fonctionne sans pincement et celles où il fonctionne avec. Dans le premier cas, les circuits ont des caractéristiques électriques supérieures mais, généralement, plus complexes que les autres. Le choix entre ces deux familles n'est donc pas particulièrement aisé. Certains circuits effectuent la compensation inhérente des effets dépendant de la température.

La mesure des paramètres des dispositifs semi-conducteurs par G. May

Résumé de l'article aux pages 483 à 484

L'auteur donne un aperçu de deux méthodes permettant de mesurer les paramètres de dispositifs semiconducteurs. Il s'agit de la méthode dite du paramètre de diffusion et de celle du paramètre de transmission. Ces deux méthodes obvient aux difficultés découlant des conditions exigées par d'autres méthodes, à savoir un contrôle parfait d'immitence de circuit ouvert ou de court-circuit, accentué par l'introduction de circuits linéaires intégrés.

Un convertisseur précis de tension alternative en tension continue pour basses fréquences par R. W. Gellie et A. G. Klein

Résumé de l'article à la page 484

Les auteurs présentent un projet de convertisseur de tension alternative en tension continue pour la mesure de précision de la valeur moyenne redressée d'un signal de basse fréquence à l'aide d'un voltmètre numérique. Ce convertisseur s'appuie sur un pont redresseur à l'intérieur du circuit réactif d'un amplificateur opérationnel, ainsi que sur un circuit à transistors à deux bornes qui fournit une constante de temps de filtrage étendue sans recourir à de trop grandes capacités.

Un nouvel amplificateur de fermeture pour réacteurs nucléaires par I. G. Morris

Résumé de l'article aux pages 485 à 489

Au cours des années, l'instrumentation de protection des réacteurs à suivi la tendance générale et le matériel transistorisé a été mis au point. Le premier dispositif dans ce domaine a être constitué de corps solides fut l'amplificateur de déclenchement de température. Ce dernier a été suivi, plus récemment, par l'instrumentation de protection contre le flux de neutrons. Cet article examine le dessin de base d'un de ces appareils transistorisés de conception récente assurant la protection contre le flux de neutrons, à savoir l'amplificateur de fermeture qui doit être installé prochainement à la centrale nucléaire de Wylfa.

Lignes de transmission de bande par J. Swift

Résumé de l'article aux pages 490 à 494

Cet article constitue une étude des lignes de transmission de bande utilisées dans les circuits micro-ondes. Elle porte sur les applications, les données de réalisation et les informations de discontinuité, ainsi que les transitions et le couplage entre les lignes de bande.

Modulateurs de fréquence et de phase à déviations étendues par D. J. Comer

Résumé de l'article aux pages 495 à 497

L'auteur analyse les modulateurs de phase et de fréquence à grandes déviations utilisés dans les applications de contrôle. L'élément de base de ces circuits semiconducteurs est une résistance nonlinéaire employée pour compenser les nonlinéarités de circuit. Les déviations obtenues sont rapportées linéairement à l'amplitude du signal modulateur.

Quelques nouveaux oscillateurs résistance-capacité pour hautes fréquences par R. S. Sidorowicz

Résumé de l'article aux pages 498 à 502

L'auteur donne un aperçu des divers types d'oscillateurs résistance-capacité et montre leurs désavantages aux hautes fréquences. Il indique un oscillateur qui obvie à la plupart de ces inconvénients. Il s'agit, en effet, d'un oscillateur doté d'un circuit de déphasage actif spécial du type passe-tout et dont la capacité de déphasage ou la résistance peuvent être mises à la terre. Un oscillateur supplémentaire à deux étages de déphasage actifs est également décrit brièvement.

Un générateur d'impulsions à contrôle numérique par S. C. Acharyya

Résumé de l'article aux pages 503 à 506

Cet article traite succinctement des particularités de réalisation d'un générateur électronique d'impulsions qui produit des impulsions d'étalonnage permettant d'effectuer un contrôle numérique des impulsions d'erreurs à diverses fréquences. L'analyse d'un circuit d'oscillateur à régulation de tension directe, constituant la partie essentielle du générateur d'impulsions est présentée de manière à illustrer l'étude analytique du dessin de l'oscillateur.

Un générateur d'impulsions de 50 V, 1A, 10 MHz par R. W. Alsford et R. E. Hayes

Résumé de l'article aux pages 507 à 511

Cet article décrit un générateur d'impulsions de grande puissance conçu pour le contrôle de filmes magnétiques minces. Ce générateur est de fréquence variable et assure une sortie nominale à un cycle de régime de 50 : 50 jusqu'à un maximum de 10 MHz. Des impulsions de déclenchement isolées sont prévues, ainsi qu'un retard d'impulsions précis réalisé par l'emploi d'une ligne à retard à prises. La largeur des impulsions de sortie varie entre 10 et 100 ns. Le générateur a été prévu pour l'utilisation avec des câbles de 50Ω et le courant de sortie est à variation continue jusqu'à un maximum de 1A.

Commutateurs analogiques pour le contrôle modulaire des calculatrices par N. W. Bellamy et S. R. Styler

Résumé de l'article aux pages 512 à 515

Les auteurs décrivent un circuit de contrôle modulaire intégrateur constitué de corps solides et prévu pour remplacer le système actuel de contrôle par relais. Ils démontrent que, par l'emploi d'un commutateur série à transistors complémentaire, on peut obtenir facilement des tensions de compensation inférieures à 50V et des courants de fuite inférieurs à 10^{-10}. Les résultats d'essais ont établi que le comportement du circuit de contrôle modulaire atteint le degré de quasi perfection qu'on peut en escompter, dans la mesure où les dispositifs actuels constitués de corps solides le permettent.

Zusammenfassung der wichtigsten Beiträge

Mit Bipolar- und Feldeffekttransistoren bestückte Verstärker von W. Gosling

Zusammenfassung des Beitrages auf Seite 478-482

Die Kombination von Bipolar- und Feldeffekttransistoren kann Verstärker mit hohem Verstärkungsgrad und niedriger Ausgangsimpedanz ergeben. Man kann zwei Schaltungsfamilien unterscheiden, je nachdem ob der FET gesperrt oder nicht gesperrt betrieben wird. Die ersten Schaltungen haben überlegenere elektrische Eigenschaften, sind aber im allgemeinen komplexer als die letzteren. Die Wahl zwischen den beiden Familien ist durchaus nicht einfach. Einige der Schaltungen haben eine Eigenkompensation für temperaturabhängige Effekte.

Das Messen der Parameter von Halbleitern von G. May

Zusammenfassung des Beitrages auf Seite 483-484

Dieser Beitrag beschreibt zwei Messverfahren fur die Parameter von Halbleitern: das Streuparameter- und das Durchlassparameterverfahren. Diese Verfahren vermeiden die durch die Anforderungen anderer Verfahren fur einen perfekten Kurzschluss- und Leerlaufimmittanztest entstehenden Schwierigkeiten, die durch die Einfuhrung linearer integrierter Schaltungen betont werden.

Präzisionskonverter Wechsel-Gleichspannung für niedrige Frequenzen von R. W. Gellie und A. G. Klein

Zusammenfassung des Beitrages auf Seite 484

Der vorgestellte Entwurf eines Konverters Wechsel-Gleichspannung ist für das genaue Messen gleichgerichteter Mittelwerte niederfrequenzer Signale mit einem Digitalvoltmeter geeignet. Die Schaltung beruht auf einer Gleichrichterbrücke in der Rückkopplungsschleife eines Funktionsverstärkers und einer Zweipol-Transistorschaltung, die ohne zu hohe Kapazitäten eine hohe Glättunsgzeitkonstante gibt.

Neuer Abschaltverstärker für Kernreaktoren von I. G. Morris

Zusammenfassung des Beitrages auf Seite 485-489

Die Instrumentierung für den Reaktorschutz folgte in den letzten Jahren dem allgemeinen Trend mit der Entwicklung transistorierter Ausrüstungen. Das erste in diesem Sektor in Festkörpertechnik ausgeführte Gerät war der Temperaturschutzverstärker, dem neuerdings die Neutronenflussinstrumentierung folgte. Der Beitrag beschreibt den Grundentwurf eines der Instrumente einer neuentwickelten Familie transistorierter Neutronenflussschutzinstrumente für das Kernkraftwerk Wylfa.

Bandleitungen von J. Swift

Zusammenfassung des Beitrages auf Seite 490-494

Ein Überblick über die Verwendung von Bandleitungen in Mikrowellenschaltungen mit Anwendungsbeispielen, Entwurfsdaten und Information an Sprungstellen, Übergang sowie Kopplung zwischen Bandleitungen wird gegeben.

Phasen- und Frequenzmodulatoren mit grossem Hub von D. J. Comer

Zusammenfassung des Beitrages auf Seite 495-497

Der Beitrag beschreibt Phasen- und Frequenzmodulatoren mit grossem Hub, die für Anwendung in der Regeltechnik geeignet sind. Das Grundelement dieser Halbleiterschaltungen ist ein nichtlinearer Widerstand, mit dessen Hilfe die Nichtlinearitäten der Schaltung ausgeglichen werden. Die erreichten Hübe stehen zur Amplitude des modulierenden Signals in linearer Beziehung.

Neuartige RC-Oszillatoren für Hochfrequenzen von R. S. Sidorowicz

Zusammenfassung des Beitrages auf Seite 498-502

In einer kurzen Übersicht verschiedener RC-Oszillatoren wird gezeigt, dass sie bei Betrieb im Hochfrequenzbereich Nachteile haben. Ein vorgeschlagener Oszillator vermeidet die meisten dieser Nachteile. Für diese Schaltung ist ein spezieller, als Allpass wirkender aktiver Phasenschieber erforderlich, dessen Phasenschieberkondensator oder -widerstand geerdet werden kann. Eine solche Schaltung wird vorgeschlagen und eingehend analysiert. Ein weiterer Oszillator mit zwei aktiven Phasenschieberstufen wird auch kurz beschrieben.

Ein Impulsgeber für Digitaldarstellung von S. C. Acharyya

Zusammenfassung des Beitrages auf Seite 503-506

Dieser Beitrag behandelt kurz Überlegungen, die für die Entwicklung eines elektronischen Impulsgebers, der die für die Digitaldarstellung von Fehlerimpulsen bei veränderlichen Frequenzen erforderlichen Eichimpulse erzeugt, angestellt werden müssen. Die den Kern des Impulsgebers darstellende Oszillatorschaltung mit Gleichspannungssteuerung wird analysiert, um eine analytische Stellungnahme zum Oszillatorentwurf zu erleichtern.

Ein Impulsgeber für 1 A, 50 V bei 10 MHz von R. W. Alsford und R. E. Hayes

Zusammenfassung des Beitrages auf Seite 507-511

Der in dem Beitrag beschriebene Hochleistungsimpulsgeber wurde ursprünglich für das Testen magnetischer Dünnfilme entwickelt. Der Generator ist frequenzveränderlich und gibt bei einem Tastverhältnis von 50:50 seine Nennausgangsleistung bis zu maximal 10 MHz ab. Es gibt isolierte Triggerimpulse sowie genaue Impulsverzögerungen, die von einer Laufzeitkette abgegriffen werden. Die Breite Ausgangsimpulses ist zwischen 10 und 100 ns regelbar. Der Generator wurde für Einsatz mit 50-Ω-Kabeln entworfen und hat einen bis zu maximal 1 A kontinuierlich regelbaren Ausgangsstrom.

Analogschalter für die Betriebsartsteuerung von Computern von N. W. Bellamy und S. R. Styler

Zusammenfassung des Beitrages auf Seite 512-515

Es wird eine Abfrage-Steuerschaltung in Festkörpertechnik beschrieben, die direkt die heute übliche Relaisschaltung ersetzt. Durch Anwendung von Serienschaltern mit Komplementärtransistoren und einem aufladbaren Element für die Stromversorgung können Verlagerungsspannungen von unter 50 V und Restströme von weniger als 10^{-10} A erreicht werden. Testergebnisse zeigen, dass die Eigenschaften der Betriebsartsteuerung dem Ideal so nahe kommen wie es die gegenwärtigen Festkörperelemente gestatten.