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Commentary

COMPUTERS for Research is the subject of a report* presented to Parliament in January. It has been compiled by a Joint Working Group of the Council for Scientific Policy and the University Grants Committee under the Chairmanship of Professor B. H. Flowers. The Working Group was set up in March last and had as its terms of reference 'To assess the probable computer needs during the next five years of users in universities and civil research establishments receiving support from Government funds'. The Working Group is undoubtedly to be congratulated on the thoroughness of its investigations; its recommendations also seem eminently sound. To implement its recommendations would cost, at present day prices, about £30M spread over five years. According to a statement made on 21 December last by the Secretary of State for Education and Science the Government has given general approval to the Flowers Committee's programme but, because of the economic situation and the need to fit the expenditure into the National Plan, it is proposed to implement the programme over six years instead of five. In the first three years' expenditure will be at the rate of about £3M a year; this compares with an average rate of approved orders for universities, over the last three years, of under £0·5M per annum.

Only a few years ago the possession of a computer was a scientific luxury; for modern research it is a necessity and, if this country is to keep pace with other scientifically oriented communities, and particularly the U.S.A., it is essential that workers in universities and other research establishments should have access to adequate computing facilities. This is certainly not the case at present and, in fact, very few universities now possess a computer adequate in size and reliability for their essential needs. From the evidence available it seems that the research demand for computers has a doubling time of between one and two years and there is, as yet, no sign of a slackening in demand. Even in universities which are reasonably well equipped difficulty frequently arises when a piece of work is too large or complex for the local installation to handle; the work then has to be run on a larger installation elsewhere, but this often turns out to be difficult or impossible because the two machines are not compatible, that is, they cannot accept the same computer language, and rewriting a major programme can take weeks or months.

In the Report the Working Group has analysed the needs and made detailed recommendations on a university by university basis. The basic philosophy of these recommendations is that computing facilities should be provided on a regional hierarchial system. It is clear that it would not be economic to provide every university with the largest available computer, each of which might cost £2M. On the other hand any research worker may occasionally need to use a much larger machine than can be provided at his parent institution. Thus it is proposed that universities and Research Council laboratories should be formed into groups, on a regional basis, and that while most of them will be provided with a modest machine of their own they will, at the same time, share a much larger facility provided at one of them. To make such a system workable it is essential that the installations within the hierarchy should be compatible, both in equipment and in programmes, and able to deal with the advanced programming languages required by scientific workers, such as ALGOL and FORTRAN. In this way it is hoped that, at minimum cost, a flexible and efficient arrangement may be achieved whereby any computer may be used as an off line input-output device to any other. Since the demand will expand it is important, where appropriate, to choose machines capable of modular growth.

In the Report considerable attention is given, also, to the need for efficient peripheral equipment in adequate quantity and also the necessity for efficient software and in particular compilers for the major machine languages.

Another matter to which attention is drawn is the need for efficient staffing of the installation and it is stated that already there is great shortage of high quality programmers and that this shortage is likely to become worse as more and larger computing laboratories are set up. A good programmer can, by his more efficient coding of programmes, save tens or even hundreds of hours of computer time each year, and it is clear that the training of such persons must form a vital task for any university computer department.

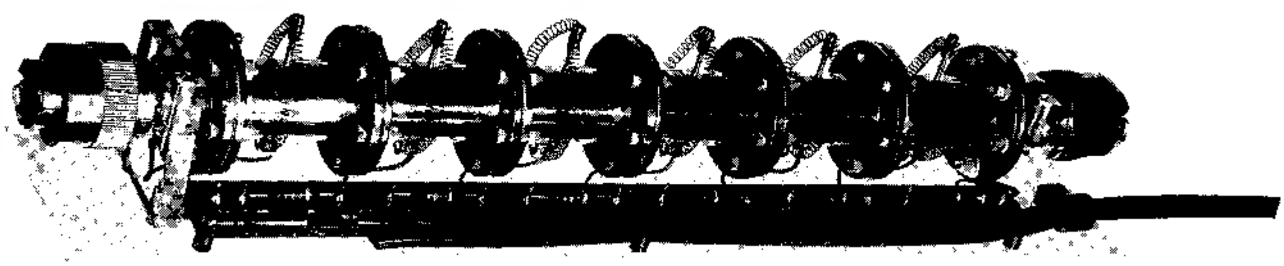
So far as the supply of the equipment necessary to the proposed programme is concerned it seems that British manufacturers can supply all the small and medium sized computers on a reasonable time scale. In some cases this can be done by upgrading existing equipments. In the case of the very large computers it is deemed necessary to acquire American equipment, preferably on a rental basis until such time as suitable British machines become available. To provide such machines would appear to call for a corporate effort on the part of British manufacturers.

If this plan is fully implemented and, unless there is a serious decline in the economic position, it seems that it will be, it will make a very worthy contribution to British research resources and will, at the same time, provide a stimulus for British computer manufacturers and help to make them even more competitive in world markets.

* *A Report of a Joint Working Group on Computers for Research. Cmnd. 2883 (H.M.S.O. January 1966).*

The Design of PIN Diode Coaxial Attenuators and Levellers

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



A wideband pin-diode coaxial attenuator has been designed from computations based on actual diode specifications. Details are given of the design and factors limiting the performance are discussed. The measured performance is in close agreement with that computed and compares very favourably with mechanically tunable wideband devices commercially available.
A simpler version of this attenuator is used to level the output of a swept oscillator. The feedback network and the leveller performance are discussed in detail.

(Voir page 202 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 209)

THE design of wideband coaxial variable attenuators using pin diodes has been investigated[1] and some computed data, covering a variety of cases, have been made available[2] (on a limited scale).

This article is concerned with the practical application of these design data, to produce wideband attenuators and levellers using pin diodes. (A leveller is a device for maintaining constant the output level of a swept oscillator.)

Coaxial transmission lines have been used to accommodate the diodes since the design data are based on this, but it is considered that a well designed strip line mounting would probably give similar results. No theoretical background will be given here as this was dealt with in full in earlier paper[1].

The design of the wideband 'quality' attenuator will be presented first.

Attenuator

Attenuator Design

At the commencement of this work two similar types of A.E.I. stick diodes were available. One was the CV7599 (VX4190) and the other was a prototype with slightly reduced capacitance and a thinner brass mounting shank.

With these at hand and with computed data available[2] for coaxial structure, the actual construction was decided upon. This had to allow for individual biasing of the diodes and has been solved by splitting the outer conductor of the coaxial line into segments separated by very thin insulators as seen in Fig. 1. This line when used without diodes behaved as if it was continuous on account of the large capacitance at the junction flanges. The diodes themselves are closely fitted into the outer conductor so that the diode and its mount present to the line the diode capacitance with resistance across it and in series with fixed inductance mainly due to the diode post (as indicated in Fig. 1).

TABLE 1
(a) Diode Specification

Diodes:	A.E.I. prototype stick diodes
Measured capacitance, C:	0·3pF
Measured diode and post inductance:	1·8nH

(b) Design Data

Diode separation:	$\lambda/4$ at 1·95Gc/s
Conductance ratios:	1, 3·75, 14, 14, 3·75, 1

(c) Computed Results

Insertion loss:	<0·05dB
Attenuation range:	0·05—15dB
V.S.W.R.	<1·33
±0·4dB at	3·5dB
±0·6dB at	15dB

It was decided that the prototype diodes would be used because of smaller capacitance. The larger inductance of the thin shank would tend to compensate for the capacitance effects at higher bias currents. If the diode capacitance had been much smaller this choice would not have been made because the larger inductance would have affected the performance adversely.

The capacitance and total inductance of both types of diodes were evaluated by slotted line measurements. Results are given in Table 1(a).

The photograph above illustrates an attenuator with bias connexions

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Using the measured values of L and C it was now possible to obtain from the computed data, the performance of an attenuator employing such diodes. The recommended method for arriving at a reasonably optimized design is given elsewhere[1] and will not be repeated here.

Six diodes appeared to give reasonable results. If many more diodes had been used then the attenuator performance would have worsened, contrary to the purely resistive case where greater numbers of diodes give better results. This fall off of performance with the diode number is very gradual and so the addition or subtraction of one diode will not produce much noticeable change in behaviour.

The response was further optimized by staggering the diode conductance values in the ratios shown in Table 1(b). There is no unique solution, although the choice of one ratio fixes all the others, but the best values can only be arrived at by inspection of a number of possible ones bearing in mind that the diode resistances are tapered to a lower value towards the middle of the attenuator. As a computer program was at hand this trial and error procedure was carried out on the computer. It should be noted that excessive staggering ratios may cause the lower resistance diodes to absorb a large share of the signal power and thus be overloaded. The inspection of computed results also allows one to choose the best spacing between the diodes. In this case it was decided to space them by $\lambda/4$ at 1·95Gc/s which tends to reduce the effect of C.

At this stage the diode bias resistors could be calculated to permit staggering and also to allow for the spread in diode parameters. At 1mA bias current seven diodes had junction resistances ranging from 29 to 39Ω. (This measurement was performed at 1·5Gc/s using a slotted line and corresponds to the resistance R described in the literature[1].) An almost linear relation between the diode junction conductance and bias current was assumed[1] in which case if the diode currents are adjusted to remove the spread at one particular current then this is true for all other currents. There is also a spread in the diode d.c. characteristic so that each diode must be fed with current from a high enough impedance to remove variations. The biasing network is shown in Fig. 2, which allows for these two types of spread and also produces controlled staggering in the ratios in Table 1.

The capacitance of the mica insulating washers shown in Fig. 1 is also included in the bias network diagram Fig. 2. These mica washers isolate the outer sections to d.c. but cause them to be continuous at the signal frequency. These capacitances have no effect upon the operation of the attenuator for slow variations in bias current.

For routine use two wide-band d.c. isolating sections are attached to both ends of the attenuator to provide the bias return to the centre conductor. These returns have a very low v.s.w.r. and can be designed from the literature[3]. Such isolating sections were built and tried out successfully but were not used in the measurements reported here, instead the bias was supplied through tunable stubs to make sure that no extraneous effects were present.

Measurements on Attenuator

The assembled attenuator and bias network is shown on page 142. A bias of 130V and 5mA was required for an attenuation of 27dB. The v.s.w.r. of the attenuator agreed favourably with the computed values over a wide range of bias, and results for two bias settings are shown in Fig. 3. The attenuator was not intended for use beyond about 15dB but the figure shows that even up to an attenuation level of 27dB, the v.s.w.r. characteristic is quite reasonable.

The attenuation characteristic is shown in Fig. 4 and the flatness of the response is within the predicted computed figure quoted in Table 1(c) (i.e., ±0·4dB at 3·5dB and ±0·6dB at 15dB). Beyond 15dB the response becomes less flat (±2·5dB at 27dB). These attenuator measurements were difficult to perform because the degree of flatness of the response is often within the error of the measurement. Both the modulation method and direct power measuring method were used to obtain these results.

Throughout the measurements the power level was restricted to 50mV and within this range no harmonic generation could be detected.

It is evident from the results that this attenuator compares favourably with the performance of mechanical devices except for power handling capacity. A further substantial improvement could be obtained if the capacitance of the pin diodes could be reduced by a factor of three.

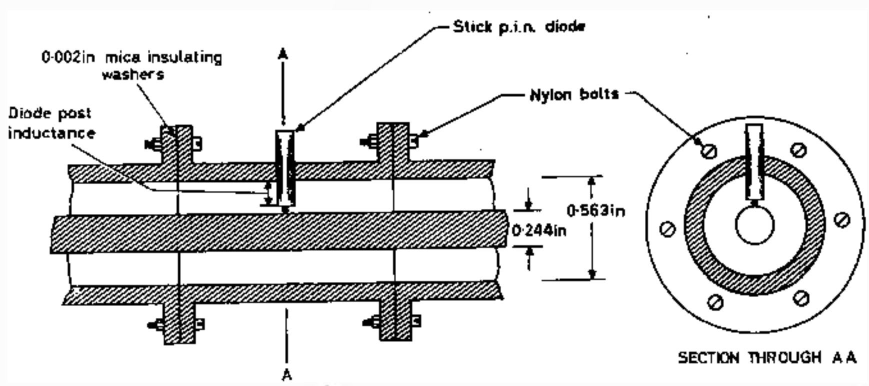
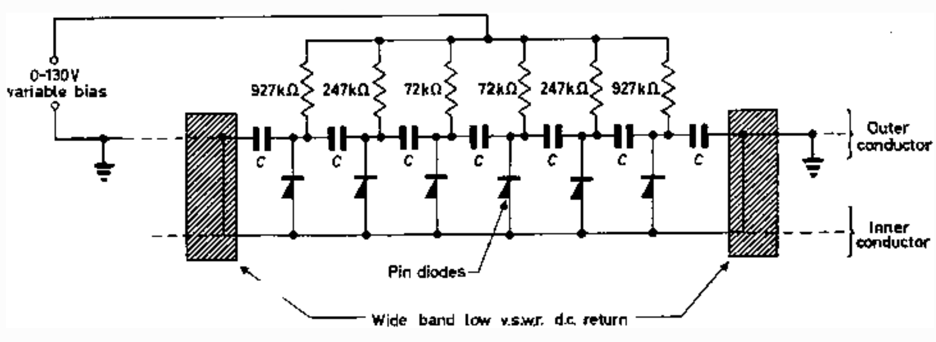


Fig. 1. Insulated section of 50Ω line carrying stick diode

Fig. 2. Circuit diagram showing bias network and decoupling capacitors



$C \simeq$ 150pF (due to mica insulating washers)

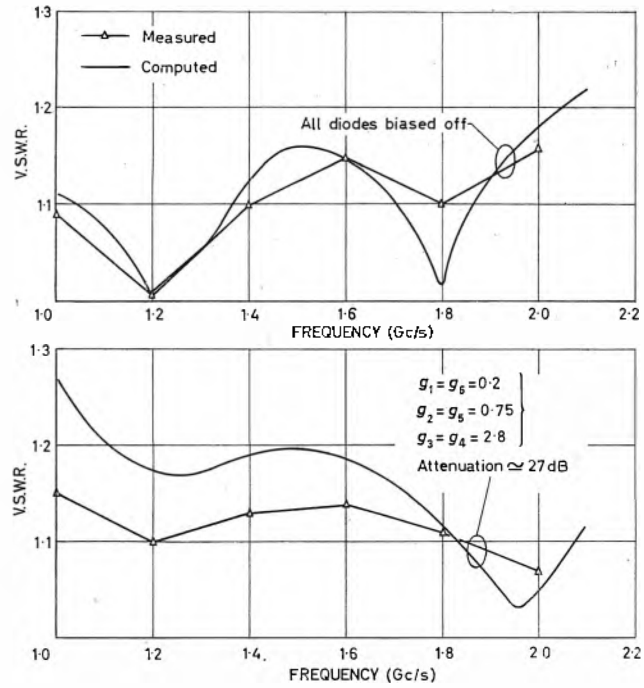


Fig. 3. V.S.W.R. curves for six pin-diode attenuator

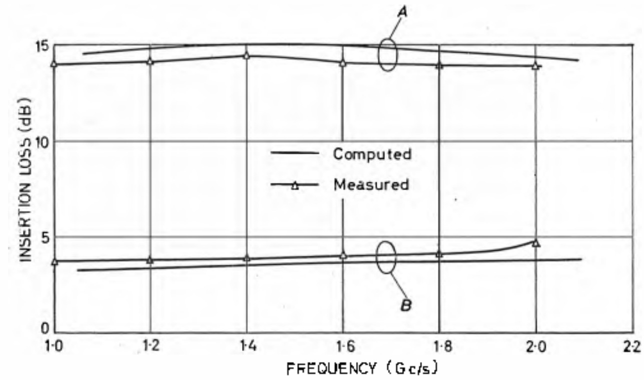


Fig. 4. Attenuator curves for six diode attenuator

A 1 to 2Gc/s Leveller

The basic operation of a microwave pin diode leveller is well known[4] and is illustrated in Fig. 5. The performance of the leveller is very dependent on the flatness of the coupler and detector. For this reason the device to be described operates over a range of 1 to 2Gc/s which is dictated by the coupler. With a suitable coupler, the leveller will operate over the 2 to 4Gc/s range but results on this will not be reported since the design principles are identical to the 1 to 2Gc/s case.

Leveller Design

The pin diode attenuator used in a leveller need not have as flat a response as the attenuator described above and for this reason only three diodes were used. The specification for this attenuator is given in Table 2 and the computed performance in Figs. 6 and 7. The diode spacing was made smaller than need be for 1 to 2Gc/s operation because as mentioned above, with suitable couplers the leveller is to function from 1 to 4Gc/s. Variations in the

TABLE 2
Data for Leveller Attenuator Using Three PIN Diodes

Diodes: CV7599
Measured capacitance, C: 0·35pF
Measured diode and post inductance: 1·1nH
Diode separation: $\lambda/4$ at 3Gc/s
No staggering of diode conductances.

attenuation curve with frequency will cause the control loop gain to vary in the same manner and therefore the attenuator response must not be too peaky. It is clear from Fig. 6 that the attenuation curves vary considerably in the range 3 to 4Gc/s.

Variations in the v.s.w.r. curves do not matter providing the v.s.w.r. is low (say $< 1{\cdot}5$) which is not so for the curves shown in Fig. 7. However this large v.s.w.r. at high attenuation levels is to be expected with such a device and the resulting pulling of the source can be removed by a suitable fixed pad.

The design criterion for a leveller attenuator is therefore clear. If a quality attenuator is used, then the control loop gain is not appreciably influenced by the frequency attenuation response of the attenuator and also the source is not pulled. If only a few diodes are used then the loop gain may be very frequency dependent and also the full

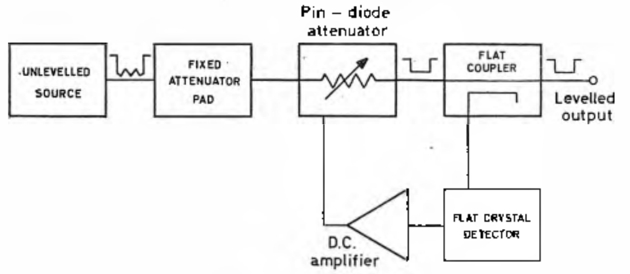
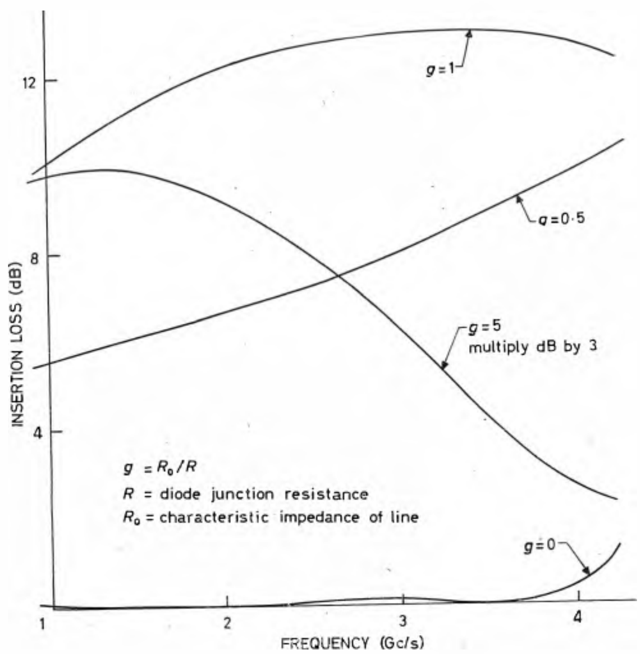


Fig. 5. Arrangement of leveller

Fig. 6. Computed attenuation curves for leveller attenuator



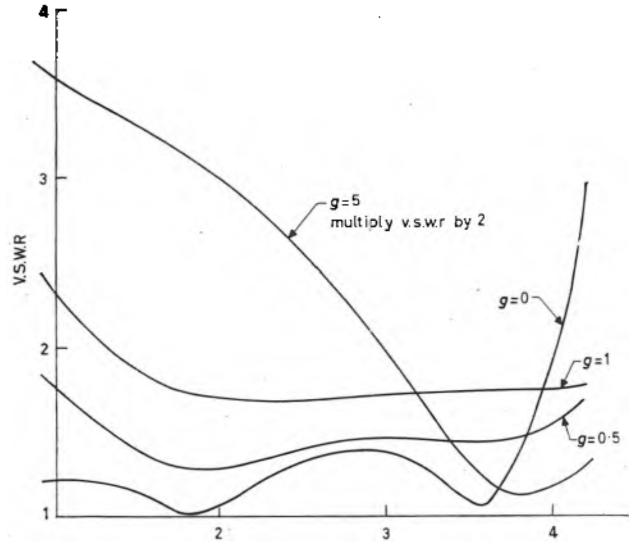


Fig. 7. Computed V.S.W.R. curves for leveller attenuator

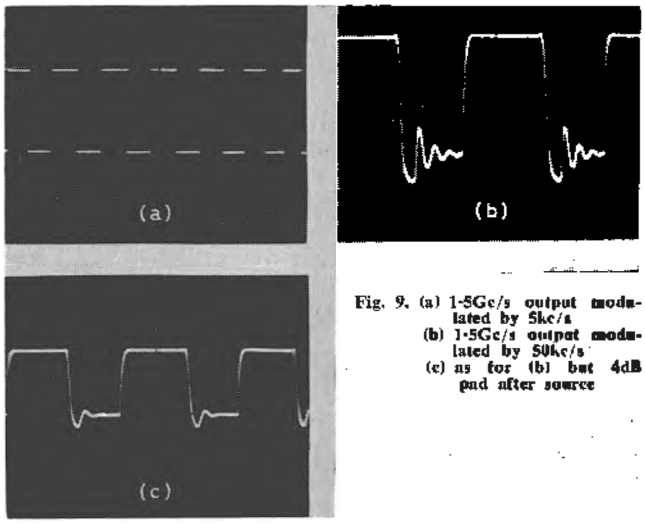


Fig. 9. (a) 1·5Gc/s output modulated by 5kc/s
(b) 1·5Gc/s output modulated by 50kc/s
(c) as for (b) but 4dB pad after source

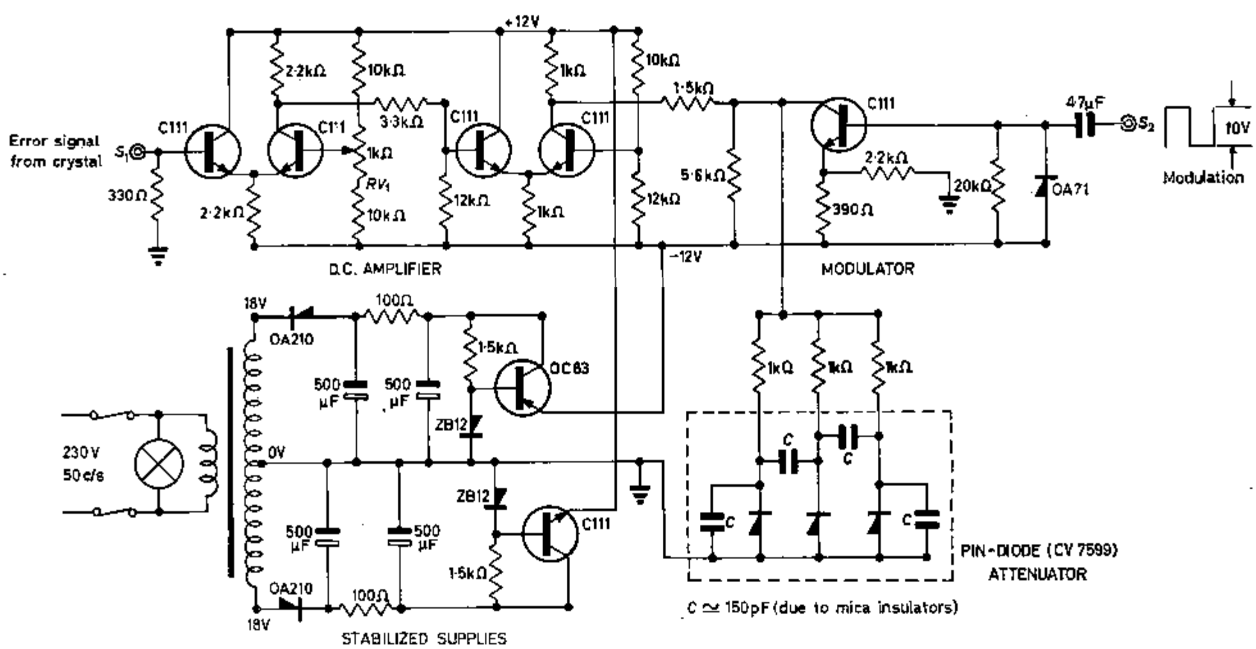
levelled power output of the source cannot be realized because of the fixed pad between source and attenuator. The device to be described falls into the latter category and is of interest because a designer will no doubt be concerned with using the minimum number of diodes.

A circuit diagram of the completed leveller is shown in Fig. 8 and the bias return for the attenuator is taken through the source. If the source had been isolated to d.c. then wideband returns would have been used. Apart from this difference the construction of this three diode attenuator is identical to the previous six diode device, Fig. 1.

The small signal gain of the amplifier is about 50 and since the transistors are of a high frequency type, they introduce very little phase shift over the frequency band of operation. In fact, since the coupler with built-in detector (Hewlett Packard 786D) does not introduce appreciable phase shift the response time of the loop is dictated mainly by the pin diodes and mica coupling capacitors. The transistor amplifier is of the differential type for good temperature stability but no temperature tests on this or the behaviour of the pin diodes at elevated temperature, have been carried out.

It was intended that detailed Nyquist plots of the loop would be performed but the non-linearity of the diodes prevented the use of the usual techniques. As the diodes are biased further into conduction the effect of the coupling capacitors and inherent diode capacitance is lessened so that the response time of the loop becomes a function of attenuation level. In addition to this any pulling of the source can modulate the loop gain and the explanataion of this occurrence is as follows: Since the output level of the microwave source (whether a b.w.o. or klystron, etc.) may change due to mismatching the load, it is possible for the source power to increase as the leveller attenuator is biased further on. This has the effect of reducing the loop gain and the leveller becomes less effective. This may be overcome by a pad at the source as illustrated below. An alternative cure of increasing the feedback amplifier

Fig. 8. Circuit diagram of d.c. amplifier, power supply and diode attenuator



gain as the power output is reduced has not been tried. The performance is described below.

Measurements on Leveller

The leveller can be switched with about 10V modulating signal and the circuit arrangement is shown in Fig. 8. During the 'off' cycle there is no output from the leveller and in the 'on' cycle the output level is determined by the level control. This control is the only variable one on the equipment and is shown in Fig. 9 as RV_1. The performance of the leveller to 5 and 50kc/s modulating signals is shown in Fig. 9. At 50kc/s modulation the response time of the loop is apparent and is about 2μsec. This is probably adequate for most pulse applications and without doubt is very fast acting for levelling, where the sweep rate is often low and the variations in output thus slow. The pulling effects previously mentioned, are apparent at 50kc/s, Fig. 9(b), and cause the loop to perform damped oscillations when the leveller is switched to the 'on' cycle. During these oscillations the frequency of the source (which in this case was a b.w.o.) is no doubt violently pulled. As previously mentioned the pulling is due to the poor attenuator v.s.w.r. in the 'off' cycle and the inclusion of a fixed pad between source and leveller improves matters considerably, Fig. 9(c) but reduces the available output power.

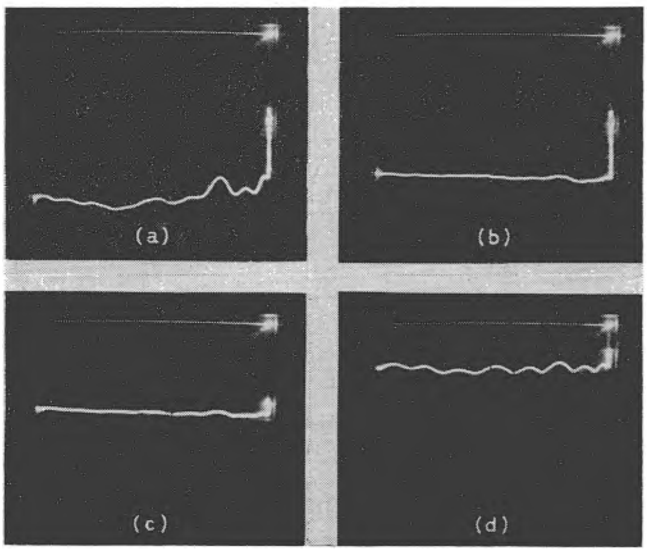


Fig. 10. (a) Unlevelled response
(b) Levelled response
(c) Levelled response with level less than in (b)
(d) Levelled response with level less than in (c)

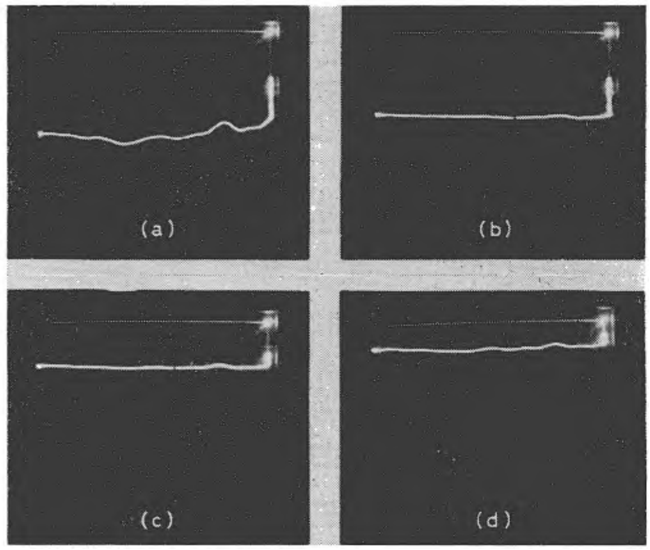


Fig. 11. (a) Unlevelled response with 4dB pad
(b) Levelled response with 4dB pad
(c) Levelled response with 4dB pad and level less than in (b)
(d) Levelled response with 4dB pad and level less than in (c)

The unmodulated performance of the leveller is shown in Fig. 10 and the performance with a 4dB pad included between source and leveller is shown in Fig. 11. In all these figures the frequency sweep is 1 to 2Gc/s from left to right. In Figs. 10(a) and 11(a) the unlevelled output response is shown as displayed on an oscilloscope after detection using a Hewlett Packard 423A detector. The remaining figures show how the levelled output can be controlled using RV_1 and it is apparent that the condition with fixed pad gives better levelling. As already mentioned, this is due to modulation of the loop gain by the mismatching of the load as seen by the source. The effect worsens as the level is reduced and in Fig. 10(d) it is very prominent. The setting for maximum levelled output produces the levelling seen in Figs. 10(b) and 11(b). The use of the 4dB fixed pad causes Fig. 11(b) to be superior and the variations are within ±0·25dB. The response of the coupler and loop detector and also the detector used for displaying the response, contribute to this uneveness because each component has a frequency response of about ±0·2dB over the band.

It was also noticed that the leveller continued to function when presented with very bad mismatches at the leveller output terminal. No measurements of this were recorded because of the difficulty of producing a known mismatch over a wide band.

Conclusions

The results show that computed data can be used satisfactorily to design a wide band attenuator which meets the computed specification.

A six-diode attenuator designed and built to cover the 1 to 2Gc/s range gives negligible insertion loss at zero bias. The variation in attenuation up to 15dB was less than ±0·6dB over the whole frequency range. A substantial improvement could be expected if diodes with lower capacitance were available.

A leveller requires less stringent performance of its attenuator and a three-diode attenuator with a transistor amplifier in the feedback loop has reduced the power variation over the 1 to 2Gc/s range from 3dB to ±0·25dB.

The slow response time of this pin-diode device limits its application as a modulator to frequencies below 100kc/s.

Acknowledgments

The authors wish to thank the Dean, Sir Donald Bailey, for permission to publish this article. Some of the measurements reported have been carried out by Mr. G. Triggs and the attenuator, the leveller and the d.c. returns were made by the Electronics Branch workshop.

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Binary Read-out Circuits Using Transistors and Gas Discharge Numerical Indicator Tubes

By P. R. Adby*, B.Sc.

A read-out for binary counters is described in which the conventional diode decoding matrix is simplified by the combination of base and emitter switching of the output transistors. The resulting matrix contains only nine diodes and offers a considerable reduction in component and production costs. Further simplification is obtained by replacement of the diodes by low tolerance resistors. A memory circuit incorporating bistable, gate, and transfer circuits is also described.

(Voir page 202 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 209)

GAS discharge numerical indicator tubes are used extensively in transistorized counting equipment where visual read-out is required. Counting units normally operate in a system in which four binary circuits are connected together to form a decade. When a display of the count held in the decade is necessary the information in the binaries is used to switch 'on' one of the digits 0 to 9. This is carried out by a decoding system followed by driver amplifiers inserted between the binaries and the indicator. Memory circuits may also be included to retain the read-out while the decade is used to take a further count.

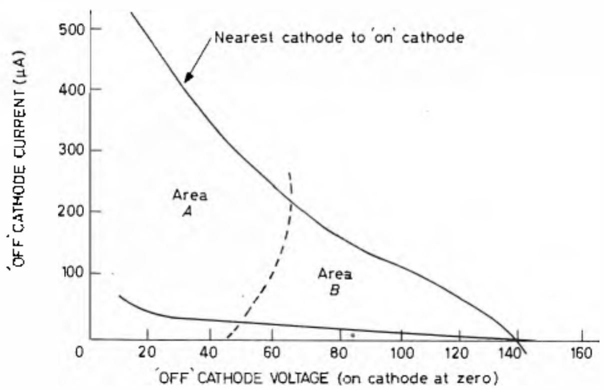

Fig. 1. Typical voltage current characteristics for 'off' cathode of a gas discharge numerical indicator tube

Drive Requirements for Gas Discharge Numerical Indicator Tubes

The indicator tube consists of one anode and ten cathodes mounted in a glass envelope filled with neon at low pressure. The ten cathodes are in the shape of the digits 0 to 9. When a potential of about 200V is applied between the anode and one of the cathodes a discharge takes place. A series resistor sets the discharge current to between 1 and 3mA and the anode potential drops to about 150V. A glow appears at the cathode in the shape of the digit selected and is viewed through the end or side of the glass envelope.

When one cathode is switched 'on' the remaining 9 cathodes will be 'off' and the least amount of background luminosity is obtained if the 'off' cathodes are completely open-circuit. This is possible with relays or other mechanical switching but for considerations of size, power consumption, reliability, or speed, transistor switching is generally preferred.

Ten transistors are required to switch the tube; one connected in series with each cathode. Only one transistor is normally switched 'on' at any time and the remainder are switched 'off'. In the 'off' condition the transistor leakage current can cause background illumination in the indicator and if the currents are large the background may completely swamp the illumination of the 'on' cathode.

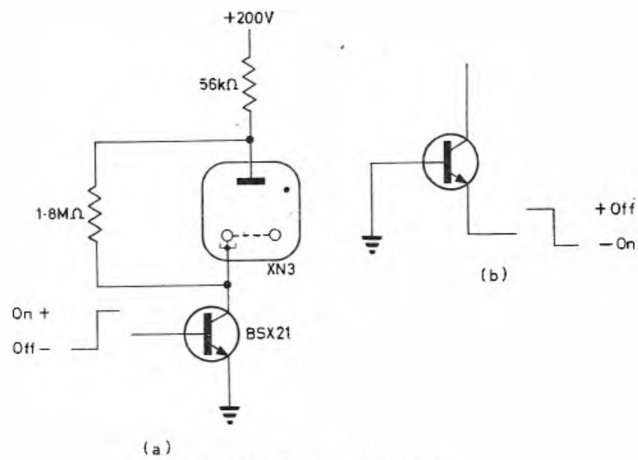

Fig. 2. Indicator drive circuit

Each 'off' cathode has a voltage-current characteristic dependent on the distance between it and the 'on' cathode. Fig. 1 shows typical curves of the voltage-current characteristic of the cathodes nearest to and most remote from the 'on' cathode.

The curves for the remaining seven cathodes lie between the two shown and the area *A* covers the cathode operating points where background luminosity would be considered excessive. In general the 'off' cathode should operate within area *B* of the graph. Transistors with very low leakage currents and high breakdown voltages are therefore necessary.

The circuit shown in Fig. 2 uses the BSX21 transistor and satisfies these requirements. It is possible to omit the 1·8MΩ resistors if the transistor leakage current is very low. The switching current at the base of the transistor must be greater than 50μA in the 'on' condition (assuming $H_{fe} \geqslant 20$ and a collector current of 1mA) and the switching voltage in the 'off' condition must be negative. It is possible to switch the transistor on the emitter as shown in Fig. 2(b) but the full collector current must then be obtained from the switching circuit.

Basic Decoding Circuits

A decoding circuit is inserted between the binary counter and the indicator drive transistors in order to switch 'on' only one digit from the information in the binary. Decod-

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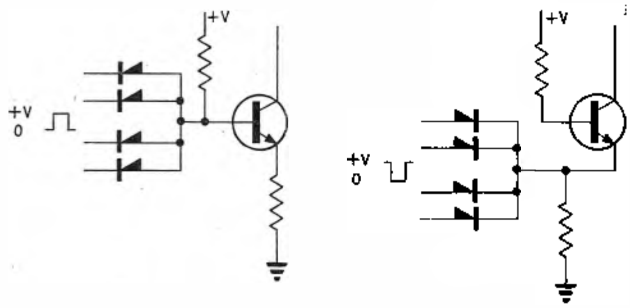
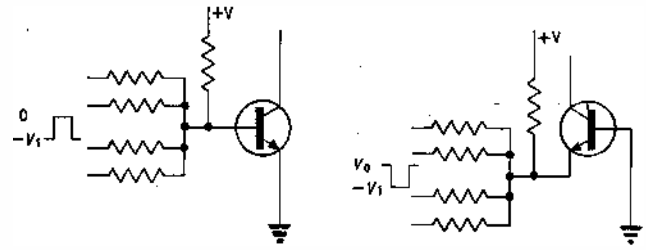


All inputs at +V for "on"
(a)

All inputs at 0 for "on"
(b)

Fig. 3. Basic diode AND circuits
(a) Base switching
(b) Emitter switching



All inputs at 0 for "on"
(a)

All inputs at $-V_1$ for "on"
(b)

Fig. 4. Basic resistor AND circuit
(a) Base switching
(b) Emitter switching

ing is carried out by a separate AND circuit for each digit.

A typical diode logic AND circuit is shown in Fig. 3(a) for base switching and Fig. 3(b) for emitter switching. In the circuit of Fig. 3(a) all diode cathodes are at a positive potential when the transistor is switched 'on'. Similarly in the circuit of Fig. 3(b) all diode anodes are at zero potential when the transistor is switched 'on'.

A resistor logic AND circuit suitable for base switching is shown in Fig. 4(a) and for emitter switching in Fig. 4(b). In the circuit of Fig. 4(a), if any one input falls to $-V$ then the transistor switches 'off'. Similarly when emitter switching is used, if any input rises to V_0 the transistor switches 'off'.

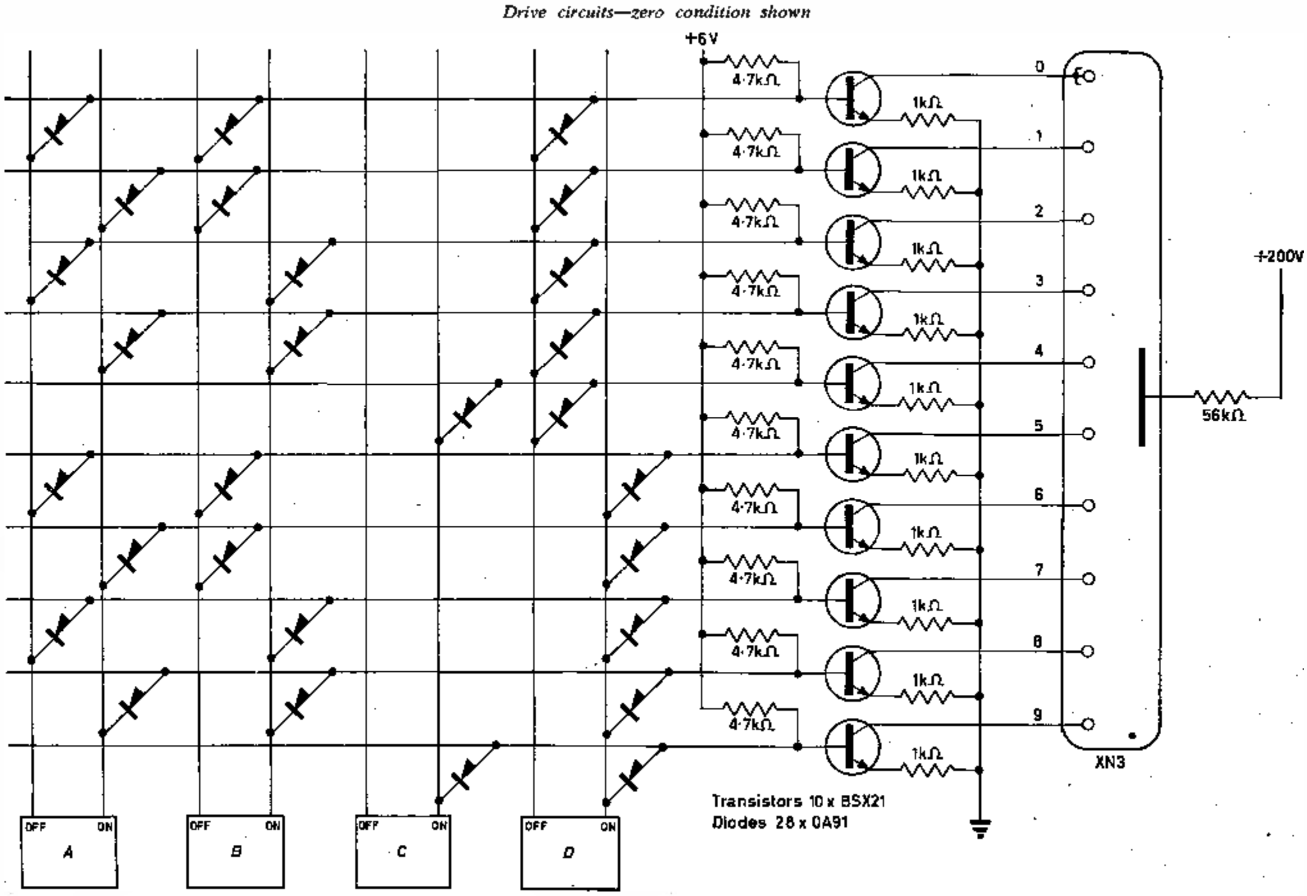
Practical Circuits

A conventional diode matrix for a binary counter utilizing the 2-4-2-1 code is shown in Fig. 5. The input to the matrix comes from drive circuits switching between 0 and +6V and the matrix consists of 28 diodes and 10 resistors. When the cathodes of all the diodes connected to one driving transistor are switched to the +6V the digit comes 'on' but if any one cathode is connected to zero voltage via the binary transistor the digit will be switched 'off'. The decoding table for this circuit is shown in Table 1.

A full matrix corresponding to Table 1 would consist of 40 diodes but twelve of these may be omitted since excess information is available from the binaries. If binary C is at 1 then the state of binary D will decide completely if the output should be 4 or 9. The information in binaries A and B is therefore not required, and the associated diodes are redundant.

It can be seen from the circuit of Fig. 5 that the binary circuit D switches 'on' either the output transistors 0 to

Fig. 5. Diode matrix 2-4-2-1 counter
Drive circuits—zero condition shown



4 or the output transistors 5 to 9. Otherwise 0 and 5, 1 and 6, 2 and 7, 3 and 8, 4 and 9 are five identical pairs. If the circuit D is used to switch the emitters of the output transistors, then the bases of the identical pairs can be connected together and an appreciable reduction in the number of components is possible. A very much smaller matrix, shown in Fig. 6, makes use of the combination of base and emitter switching to drive the five pairs. Nineteen diodes and five resistors have been saved by using this combination.

This simplification can be applied to most binary to decimal read-out matrices in which the decade consists of a divide by two binary and a divide by five consisting of the three other binaries.

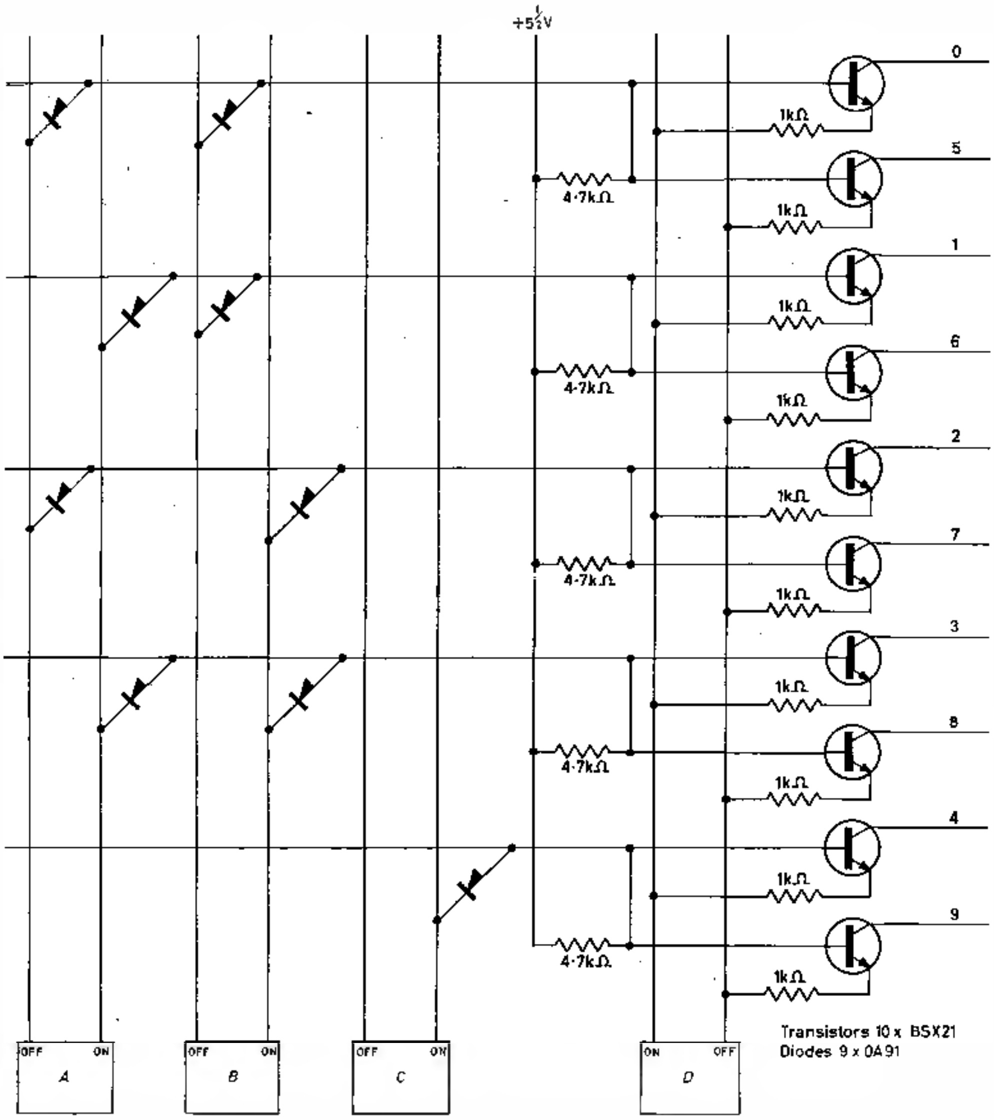
The reduction of the 'fan in' to the base of the output transistor from 5 to 2 by this method makes it easier to design a reliable resistor matrix; lower tolerance resistors can be used and the driver stage current can be more economical[1]. The worst case calculations for a three input resistive AND circuit is made assuming resistor tolerance of ±5 per cent, resistor life shifts of ±5 per cent, and voltage drifts of ±5 per cent. The worst case 'off' condition is shown in Fig. 7(a). Nominal values are given with worst case values shown in brackets.

TABLE 1

Decoding Table for a 2-4-2-1 Counter

	A	B	C	D
0	0	0	0	0
1	1	0	0	0
2	0	1	0	0
3	1	1	0	0
4	1	1	1	0
5	0	0	0	1
6	1	0	0	1
7	0	1	0	1
8	1	1	0	1
9	1	1	1	1

Fig. 6. Diode matrix for 2-4-2-1 counter with emitter switching
Drive circuits—zero condition shown



The equations which apply to this circuit are:

$$\frac{6{\cdot}3 + V}{16{\cdot}2} = i_1 - 2i_2$$

$$\frac{5 - V}{11} = i_1$$

$$V/9 = i_2$$

$\therefore$ $(6{\cdot}3+V)$ $99 = (5 - V)$
$145{\cdot}8 - 356{\cdot}4V$

which gives $V = 0{\cdot}18$V (approximately).

In the worst case 'off' condition therefore the transistor base is 0·18V negative with respect to the emitter.

The worst case 'on' condition is shown similarly in Fig. 8(b).

Base current of the transistor

$$= \frac{5{\cdot}2}{19{\cdot}8} - 0{\cdot}2\text{mA}$$

$= 60\mu$A (approximately)

Assuming $h_{fe(min)} \geqslant 20$

Collector current $\geqslant 1{\cdot}2$mA

The complete decoding circuit for a 2-4-2-1 counter, assuming connexion to a suitable drive circuit, is shown in Fig. 8. As with the diode matrix the emitters are switched by the drive circuits from the binary D.

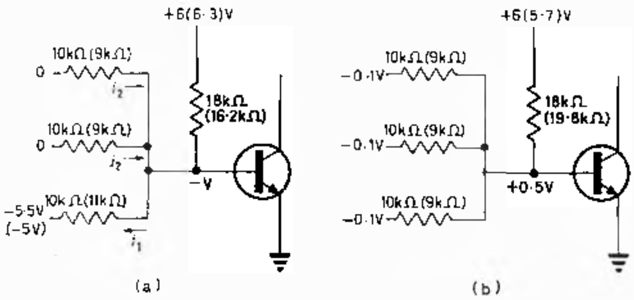


Fig. 7. Resistor AND circuit
(a) Off condition *(b) On condition*

Memory Circuits

It has been assumed that the matrices so far described have been either connected directly to the binary or connected via read-out transistors which switch from 0 to +6V or 0 to −6V. The numerical indicator tubes cannot switch at high speed and are normally read only after a count has been completed. It is usually considered desirable to provide a memory so that the indicator displays the result until a new count is available from the decade.

When using a relay decoding circuit it is relatively simple to provide the relays with double coils and to use one coil to hold the relay during a count and one to read the decade information at the end of the next count. In a transistor system a low speed bistable circuit may be used to perform this function and it is then necessary to have a gate between each counting binary and memory bistable which is only opened after the count is finished. The information is then transferred to the memory circuit and displayed on the numerical indicator tube.

The circuit of Fig. 9 shows the combination of a memory bistable, a gate, and a binary to memory transfer circuit. The transfer from the binary to memory circuits is controlled by a pulse occurring at the end of the count but before the decade reset takes place. VT_1 and VT_2 form a conventional bistable circuit which delivers its drive to the matrix from the transistor collectors. VT_3 is held 'off' during a count by VT_5 which is also 'off'. The counting binary is connected to VT_3 by a resistive divider and when

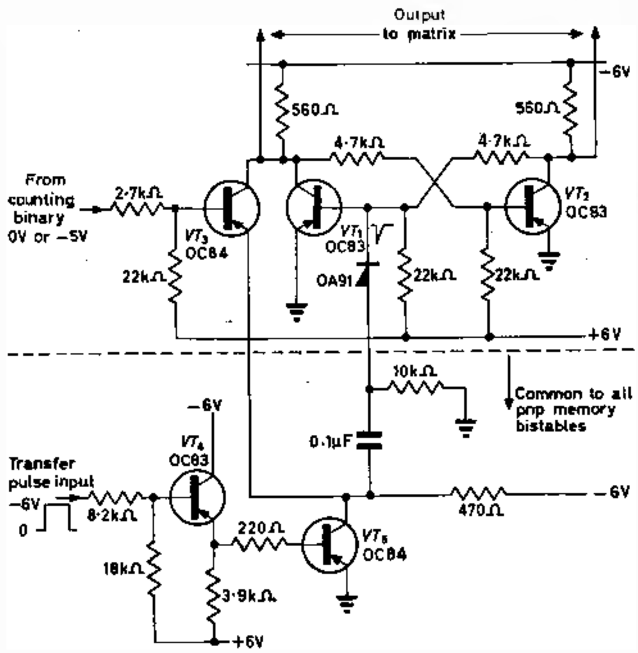
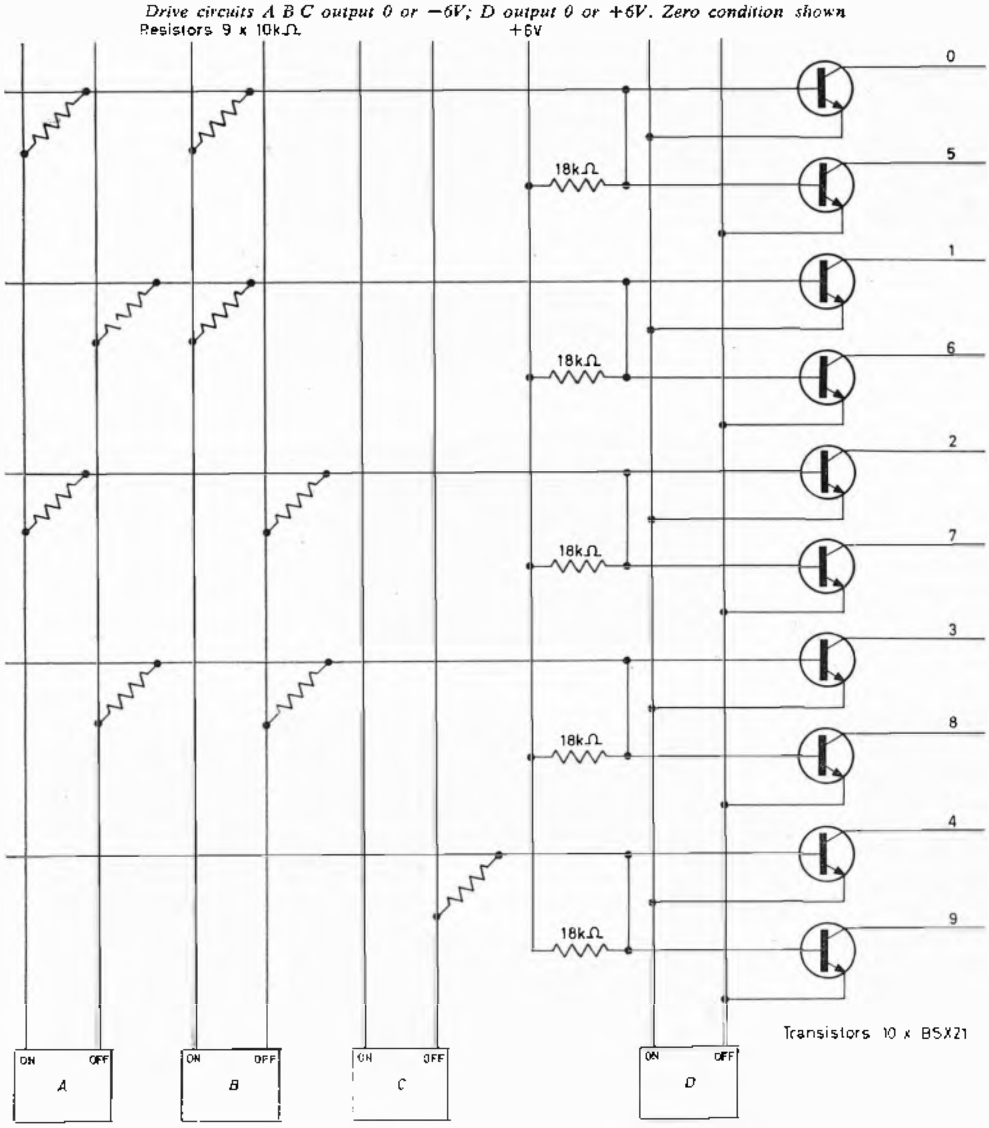


Fig. 9. Memory bistable, gate, and transfer circuits for resistor matrix drive

Fig. 8. Resistor matrix for 2-4-2-1 counter with emitter switching
Drive circuits A B C output 0 or −6V; D output 0 or +6V. Zero condition shown



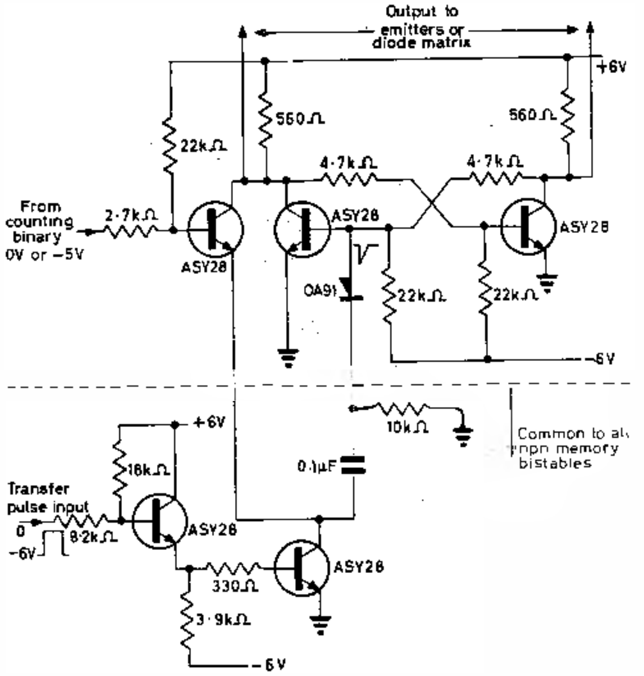


Fig. 10. Memory bistable, gate, and transfer circuits for emitter switching or for diode matrix drive

VT_3 is 'off' the count proceeds without disturbing the memory bistable. At the end of the count a negative pulse turns VT_5 'on', via the emitter-follower VT_4, and the emitter of VT_3 is effectively connected to earth. Also if VT_1 is 'on' a short positive differentiated pulse occurs at the base of VT_1, and switches it 'off'. VT_1 is therefore either 'off', or switched 'off' by the leading edge of the transfer pulse. If at this time VT_3 is switched 'on' by the information from the binary VT_1 is switched 'on' after the differentiated pulse on the base has finished. If, however, VT_3 is switched 'off' then VT_1 remains switched 'off' and is held 'off' by VT_2. The information from the binary counter has therefore been transferred to the memory bistable. The transistors VT_4, VT_5 and the differentiating circuit may be common to all the memory bistables in the counter.

A similar circuit shown in Fig. 10 uses npn transistors and is required to switch the emitters of the output transistors. The transfer in this circuit is controlled by a positive pulse occurring at the end of the count. As before the transfer and differentiating circuits are common to all the npn memory bistables.

The npn memory circuit is used to switch emitters when used with a resistive matrix and only one binary in each decade requires this circuit. The diode matrix of Fig. 6 would however use the npn memory circuits (Fig. 10) with all four binaries in the decade.

The memory bistable must not be free to run with the main counter but when counting over long periods the intermediate totals may be examined periodically by a multivibrator driving the control circuit at say 100c/s. The intermediate totals would then be displayed continuously and corrected every 10msec.

Conclusion

The circuits described were designed to fulfil the need for a small reliable counter read-out system of very low cost. By the use of emitter and base switching of silicon planar transistors driving cold-cathode numerical indicator tubes, an economic design has been achieved using both resistive and diode switching matrices.

Physically smaller read-out systems could be made using integrated circuits or micrologic techniques combined with miniature forms of output transistors. For most applications, however, the cost of these smaller circuits would be high and could rarely be justified in an economic commercial design.

Acknowledgment

The author would like to thank the Directors of Airmec Ltd and the University of Sussex for permission to publish this article.

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A U.H.F. Telemetry System

U.H.F. monitoring of outlying substations is being used by Eastern Electricity in experiments at the Woodland Control centre of the Suffolk Sub-Area at Stowmarket. The objective of these experiments is to improve the security of supply to consumers in rural areas.

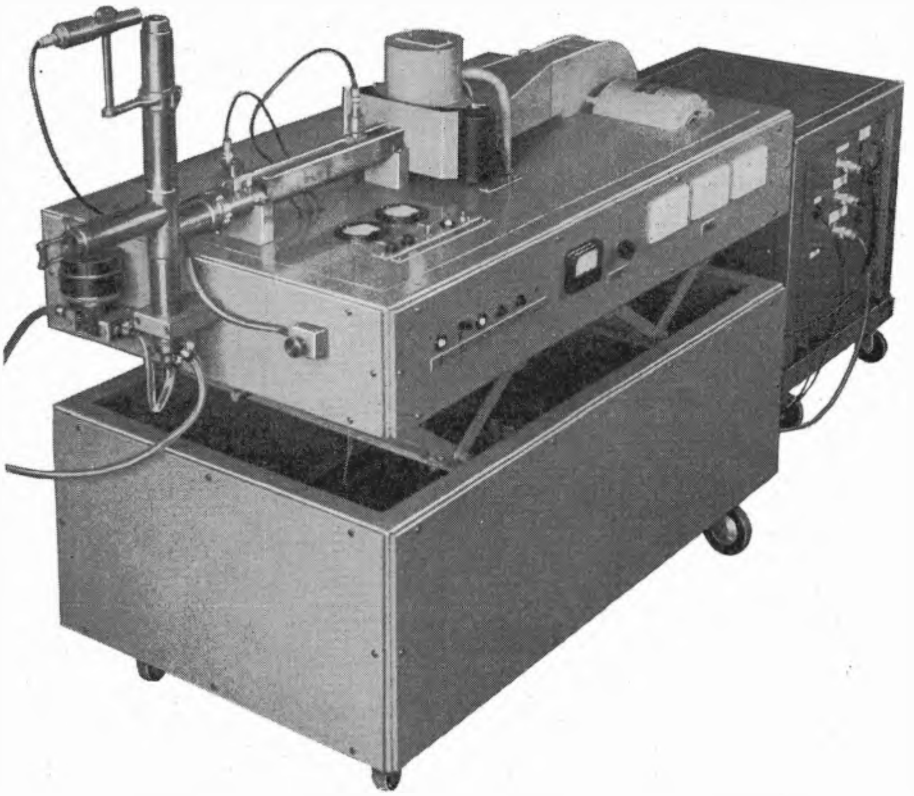
Suffolk's pilot scheme operates from Stowmarket to three outlying substations at Icklingham, Eriswell and Thetford over distances upwards of twenty-five miles. Each of these stations is called in turn by a coded radio signal broadcast in the u.h.f. band from the Control centre; an immediate reply is then transmitted back from the station so addressed indicating whether any incident or change of state affecting the local distribution system has occurred since the previous interrogation was made a few seconds earlier. Should trouble occur the control engineer is alerted by an alarm identifying the station calling. He can then make a detailed investigation over the telemetry system of the individual items of the substations, pinpoint the source of trouble and take the necessary action.

The radio telemetry system installed also includes facilities for checking loading and voltage conditions at the remote substations and is able to carry out selected switching operations which in emergencies can bring in alternative supplies from neighbouring areas.

This initial scheme obtains its outreach into West Suffolk by sharing an existing u.h.f. radio link path used for the engineering mobile radio telephone system. Over this path the system has worked most successfully even during thunderstorms and is capable by simple extension to cover many more of the important high voltage distribution centres in the area.

Prior to the installation of the Pye telemetry equipment, Suffolk had no satisfactory method of monitoring the correct functioning of its automatic substation equipments from the Control Centre, or any means for obtaining loading information or control remotely. With its predominately overhead system of distribution no private telephone type cables are available in Suffolk to link up Control to the rural distribution centres and to rent these from the G.P.O. would be prohibitive in cost even if available. Hence the decision was taken to explore the potential use of u.h.f. radio circuits as channels for two-way transmission between outstation and the Control Centre. By this means a future total of 100 high voltage primary distribution substations required to meet growing loads in Suffolk Sub-Area's two thousand square miles can be effectively linked up.

The time saved by the new system is enormous when compared with the old practices involving the calling out of an engineer, a journey to the substation, a fault finding investigation and a return message; such saving must reflect in a very much improved service to consumers.



A Microwave Plasma Torch

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

A microwave plasma torch is described which provides a simply operated and reliable source of plasma for research purposes. The torch operates at 2·45Gc/s and the power available is 2·5kW.

(Voir page 202 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 209)

THE term plasma is used to describe a gas or other form of matter in which a major portion of the molecules are dissociated into ions and electrons. The ions and electrons are approximately equal in number so that as a whole the system is electrically neutral.

In recent years intensive effort has been devoted to plasma research because plasmas have immense possibilities and are part of many natural phenomena and man-made devices[1,2].

In nature, plasmas occur in the ionosphere, the Van Allen radiation belts, lightning and auroras and the sun, whose inner region is one vast plasma. Geomagnetic storms result from increased plasma discharge during sun-spot activity.

The familiar fluorescent, sodium and mercury discharge lamps are examples of artificially produced plasmas. In electron tubes, plasmas are involved in the operation of industrial rectifiers, thyratrons, plasmatrons, gaseous noise diodes, ignitrons and many other devices.

The purpose of the plasma torch described in this article is to produce a stream of plasma from equipment that is reliable and simple to operate. Examination of the spectral lines from the plasma flame with a spectroscope enables more of its fundamental properties to be established[3] and this particular torch is intended only for research purposes. However the plasma torch is finding increasing application in industry[4]. For example it may be used to cut ¾in thick steel plates in about half the time taken with an equivalent rated oxy-acetylene burner. A plasma jet may be used as a lathe tool to machine refractory materials such as ceramics which cannot normally be cut.

Principle of the Plasma Torch

There are several types of plasma torch in common use. In the induction-coupled torch[5] no electrodes are necessary and power is transferred to the gas fuel by inductive coupling at a frequency of several megacycles per second. Other plasma torches[2] require electrodes to carry energy to the gas fuel and operate at d.c. or a few megacycles per second.

In the microwave torch discussed in this article, a strong microwave field is used to dissociate a gas and on recombination of the ions a plasma flame is produced[6,7]. A

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microwave torch has the advantage that the equipment can be made relatively simple and compact. In addition there are theoretical considerations concerned with the flame equilibrium which suggest the use of a microwave power source[8].

The main components of the present microwave torch are shown in Fig. 1. Basically the torch consists of a magnetron power source, a matching section and the gas fuel. When microwave power is coupled to the matching device an intense electromagnetic field is produced at the open end. As the gas fuel is passed near the open end, dissociation into ions and electrons occurs, the necessary kinetic energy of the particles for the dissociation process being provided by the microwave field. The resultant stored energy of the plasma is then carried by the gas stream to the tip of the torch where it is released as heat energy on the recombination of the charged particles of the dissociated gas.

The maximum power that can be transported by the plasma flame may be estimated approximately as follows. With a gas velocity v cm/sec and a flame cross-sectional area A cm², the number of gas particles entering this area in one second is $n \,.\, A \,.\, v$ where $n \simeq 2{\cdot}7 \times 10^{19}$, Loschmidt's number. If each of these gas particles gives up R electron-volts of energy on recombination, the transported power is given by:

$$\begin{aligned} P &= nAvR \times 1{\cdot}6 \times 10^{-19}\,\mathrm{W} \\ &= AvR \times 2{\cdot}7 \times 10^{19} \times 1{\cdot}6 \times 10^{-19} \\ &= 4{\cdot}32\,AvR \qquad \text{(1)} \end{aligned}$$

For example with a nitrogen flame for which v = 20cm/sec. A = 3·0cm² and R = 9·76eV.

$$P \simeq 2{\cdot}5\mathrm{kW}$$

This simple expression assumes complete dissociation of the gas which certainly does not occur. However, to partly compensate this is the fact that gas is feeding into the flame from the sides as well as from the end to increase the amount of gas available for dissociation. In addition newly associated molecules may be dissociated again and on recombining add to the heating effect. The microwave power source provides the energy for the initial dissociation and in addition some power is radiated and some dissipated in circuit losses.

The flame differs fundamentally from a combustion type flame but they are similar in appearance. The main differences are that no exothermal chemical action takes place and the flame temperature is generally much higher. Because the dissociation of the gas is followed by recombination, the fuel is not consumed in the process and may be recirculated if desired. In addition there are no problems of electrode deterioration.

A pure monatomic gas is not a suitable fuel in the microwave torch. On recombination, monatomic gases release little energy and the flame produced is a very cool one. In addition, the monatomic gas argon has a very low breakdown voltage[9] and an arc can form across the open end of the matching device at a power level as low as 10W. However a diatomic gas or a monatomic one mixed with a small quantity of diatomic gas is capable of producing a stable and extremely hot flame.

To initiate the dissociation of the gas, free electrons must be produced in the vicinity of the intense microwave field at the open end of the matching device. This is done simply by producing a momentary arc discharge near the tip of the torch.

Construction of the Plasma Torch

The main circuit components of the plasma torch are shown in block diagram form in Fig. 2. The magnetron used is the air-cooled Philips 7091 which is capable of delivering 2·5kW and operates at a fixed frequency of 2·45Gc/s. The magnetron is fitted with a thermoswitch which cuts off the mains supply if the temperature of the anode block becomes excessive. The blower unit which cools the magnetron is fitted with an air switch so that in the event of air failure the mains supply is again cut off.

The power supply is a commercial unit (Hirst Electronic Ltd.) and is fitted with the complex control circuits needed for the magnetron. It is usual to operate the anode of the magnetron with unfiltered d.c. However in this application it is vital to use filtered d.c. as the plasma flame would not be sustained if the magnetron operated only for a small portion of a cycle. Another important requirement is that the power output of the magnetron can be increased in steps up to its maximum power. The power supply used had facilities to increase the magnetron power output continuously from 500W to 2·5kW.

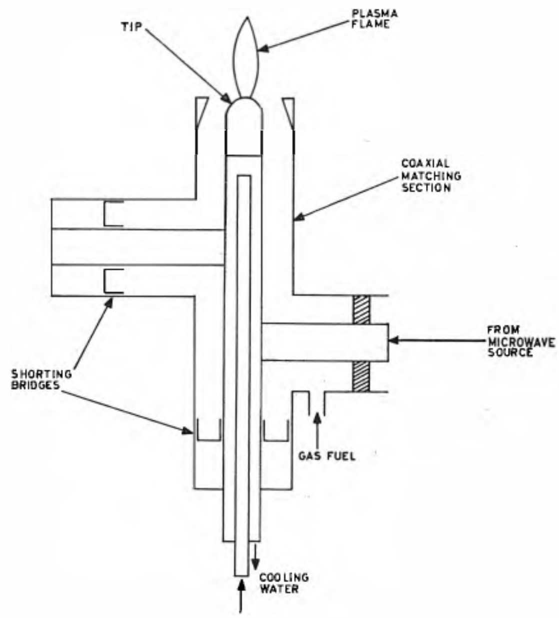

Fig. 1. Basic components of the plasma torch

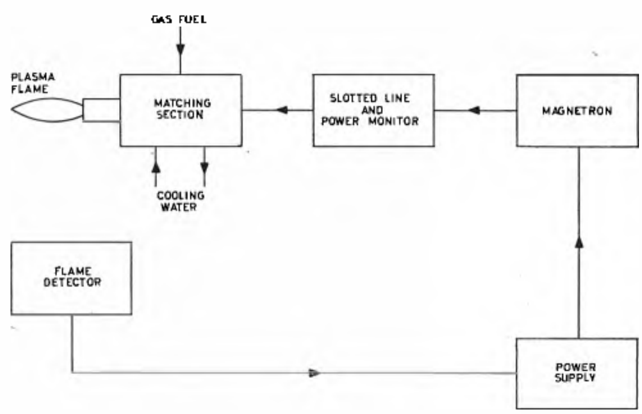

Fig. 2. Circuit components of the plasma torch

The output of the magnetron is fed via a rigid 50Ω coaxial line to the slotted line and power monitor. The slotted line is conventional and enables the impedance match presented by the plasma flame to the 50Ω line to be measured. The magnetron can only be operated safely if the v.s.w.r. on the line is within specified limits. The slotted line may also be used to measure the actual flame impedance at various power levels and with different values of gas flow. Incorporated in the slotted line is a loosely coupled directional coupler which enables the forward power flow to the flame to be monitored.

The matching element (Fig. 1) is connected to the slotted line with rigid 50Ω coaxial line. Impedance matching of the plasma flame to the 50Ω line is accomplished by tuning the bridges in the two stubs placed one quarter-wavelength apart. A v.s.w.r. of 1·2 or better is easily obtained. The tip of the torch is constructed of molybdenum to withstand the heat of the flame and a stream of water runs through the inner conductor of the matching element to cool the tip. The water is fed via a water switch which cuts off the mains supply if the flow is less than 0·25 gallons/minute. The molybdenum tip may be unscrewed and one of a different shape readily inserted. For demonstration purposes air is used as the gas fuel, a small amount being fed from the blower via an adjustable tap to the matching element. Teflon gaskets confine the gas to the correct path at the same time producing negligible discontinuity in the coaxial lines. In practice, gas flows of from 0·05 to 0·25ft^3/min were found suitable. With very low gas flows the flame becomes wide and unstable and if the flow is too great the flame may be extinguished.

A vital part of the protective circuit of the torch is the flame detector. If the torch is operating and the flame is suddenly extinguished, the magnetron is left without a load and may rapidly destroy itself. The flame detector consists of a photoconductive cell in a mount which shields it from extraneous light and protects the cell from microwave radiation. The unit is mounted on the matching element close to the plasma flame. The cell is coupled to a Schmitt trigger circuit which in turn is connected to the magnetron power supply. If the flame goes out while the equipment is operating, the magnetron anode voltage is rapidly switched off. The use of a Schmitt trigger circuit enables the ambient light level to approach that of the flame. Some simpler systems will work only if the ambient light level is very much lower than that of the flame and are usually unsuitable in this application. The present flame detector has adjustable sensitivity to allow for ambient light variations and works effectively at any power level between 500W and 2·5kW. In starting the torch the flame detector circuit is temporarily shorted out.

In operation, the torch is extremely simple to use. Once set, no adjustments to the microwave components are needed and the plasma flame is started or extinguished by only one switch and the starting wand. The starting wand consists of an insulated rod with a pointed tungsten end which is placed on the tip of the torch to produce a spark and initiate the plasma flame. The power control potentiometer incorporates a switch which is connected to the starting button so that the equipment can only be switched on at the minimum power of 500W. Considerable thought was given to the provision of an automatic starter for the plasma flame to operate at any power level and using a corona discharge or radioactive source. However, the present system, although less elegant, proved more reliable.

A photograph of the plasma torch is shown on page 152. The top section of the torch may be moved vertically over a distance of 6in to allow different portions of the flame to be examined with a spectroscope.

With equipment of this type the operator must be carefully protected against two hazards. The plasma flame gives off ultra-violet radiation and special glasses are needed when viewing the flame. These glasses provide protection yet give clear vision, an important point when doing experiments with the plasma flame. A second possible hazard is from the microwave radiation near the equipment[10]. Most authorities consider a power density of above 10mW/cm^2 to be harmful to the eyes and other parts of the body. As the plasma flame is a conductive extension of the inner conductor of the matching element it can act as an aerial and the field near the flame may be quite intense. Tests were carried out with a dipole connected to a low-level power meter and it was found that at a distance as close as 6in to the flame the power density was well below the danger level.

Conclusion

A reliable and simply operated microwave plasma torch has been described which is intended for research purposes. The principle of operation and the general construction have been given and possible hazards mentioned.

Acknowledgment

The author wishes to thank the University of New South Wales for permission to publish this article. Grateful thanks are also extended to Philips who have provided much valuable information on microwave plasma torches built within their organization and to staff member Mr. P. Green for his help in the construction of the torch.

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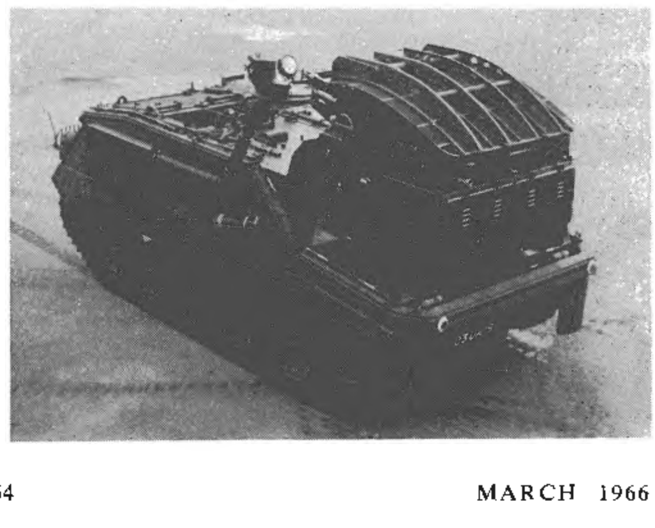
Mortar Locating Radar

EMI Electronics Ltd announce that the West German Government is placing an order worth approximately £3M with the Company for 25 sets of Green Archer mortar locating radar equipment.

The radar equipment pinpoints enemy mortar firing positions by plotting the bombs in flight. It is simple to operate, highly accurate, easy to maintain, and reliable in service. A rugged, all-weather unit, the Green Archer is adaptable to a wide variety of wheeled and tracked vehicles; it can also be transported by aircraft and helicopter. In action, the unit can be quickly dug-in or concealed and can be operated from a remote position.

In addition to locating enemy mortars, the Green Archer can be used for battlefield surveillance, for ranging fire and in various other roles.

The Green Archer equipment mounted on an armoured fighting vehicle



Peripheral Equipment for a Small Digital Computer

By A. D. Booth*, Ph.D., D.Sc., and J. M. S. deVries†, M.Sc.

A small transistorized digital computer has previously been described. In this article a description is given of the peripheral equipment. This consists of a photo-electric paper tape reader, a paper tape punch, a typewriter and the associated control circuits.

(Voir page 203 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 210)

THE general features of the design of the M.3 computer have been described in a previous article[1]. During the period which has elapsed since the completion of the central processor, a new set of peripheral equipment has been designed, constructed, and integrated with the rest of the machine. As far as the computer itself is concerned, the input-output equipment consists of a photo-electric, paper-tape reader, a high-speed paper-tape punch and an electric typewriter. On the original M.3 the typewriter was absent and the paper-tape equipment was of the type originally described[2]. It is the purpose of this article to describe the new equipment and its logical integration with the M.3.

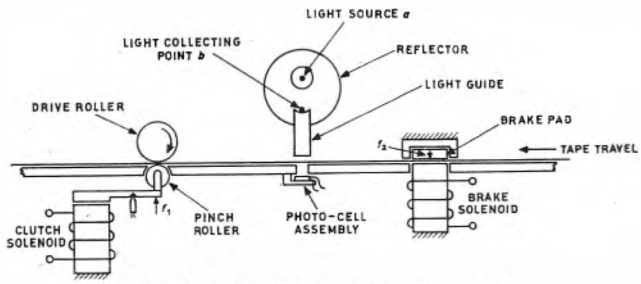


Fig. 1. Basic elements of the tape reader

The Paper-Tape Reader

The original, wire brush, reader[2] was comparatively slow (80-100 characters/sec) and it was decided to replace it by a photo-electric reader whose speed should be 500 characters/sec. This speed was chosen because it represented the greatest speed at which the M.3 could absorb data. In fact the reader which was constructed will operate reliably at 1 000 characters/sec.

The speed of an electromechanical device is limited because of the movements of the parts necessary for the stopping and starting of the tape. To accelerate a mass requires time, which will give rise to delays and limits the speed of operation.

In designing a high-speed mechanical device, careful thought should be given to simplicity of construction and minimizing the mass and travel of the moving parts[3].

The basic elements of the photo-electric reader are shown in Fig. 1. It consists of three assemblies: the clutch, brake and character sensing unit.

The clutch and brake mechanical movements are minimized by having the pinchroller of the clutch assembly and the brake pad in contact with the tape at all times. If the clutch coil is energized, the only movement is that of compressing the tape, which will be a fraction of the tape thickness. The same applies to the brake pad when the brake coil is energized. It should be noted that the brake force (f_2) must be larger than the clutch force (f_1) since, when the tape is travelling at high speed the brake must be able to stop within the diameter of the location

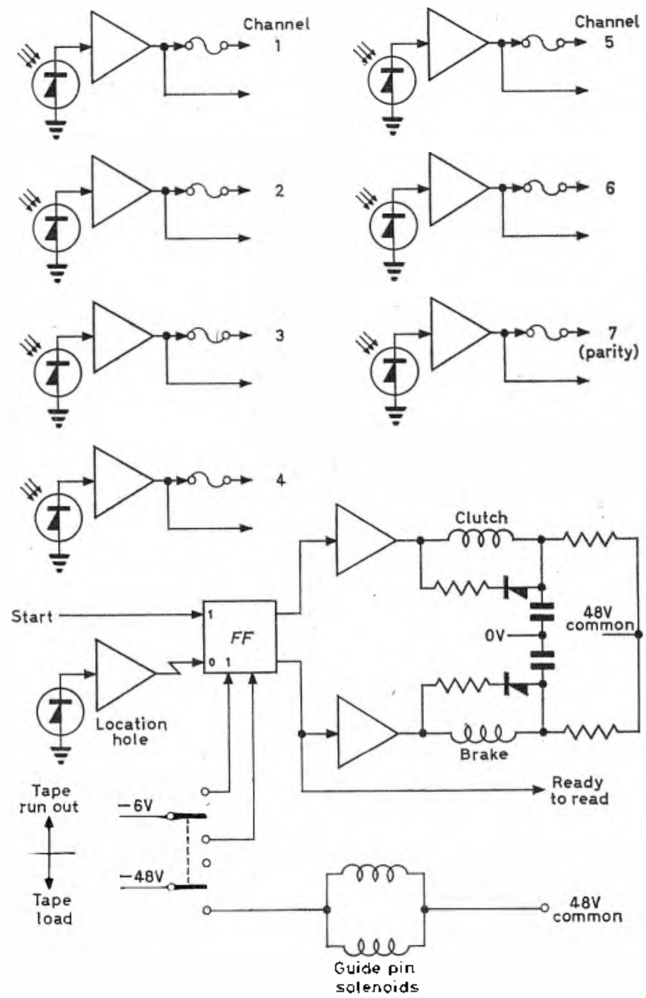


Fig. 2. Tape reader logic

hole (0·045in). In this design f_2 was chosen to be approximately three times f_1 to produce satisfactory results.

The light source and sensing unit is the next assembly of interest, it consists of a 12V, 6W bulb with a straight filament of about 0·5in length, mounted within a reflector as shown in Fig. 1. Direct and reflected light is collected at point *b* and emitted through a Lucite light guide to the photocells. A cylindrical reflector is used for ease of con-

* *University of Saskatchewan, Canada.*
† *Northern Electric Research Labs., formerly University of Saskatchewan.*

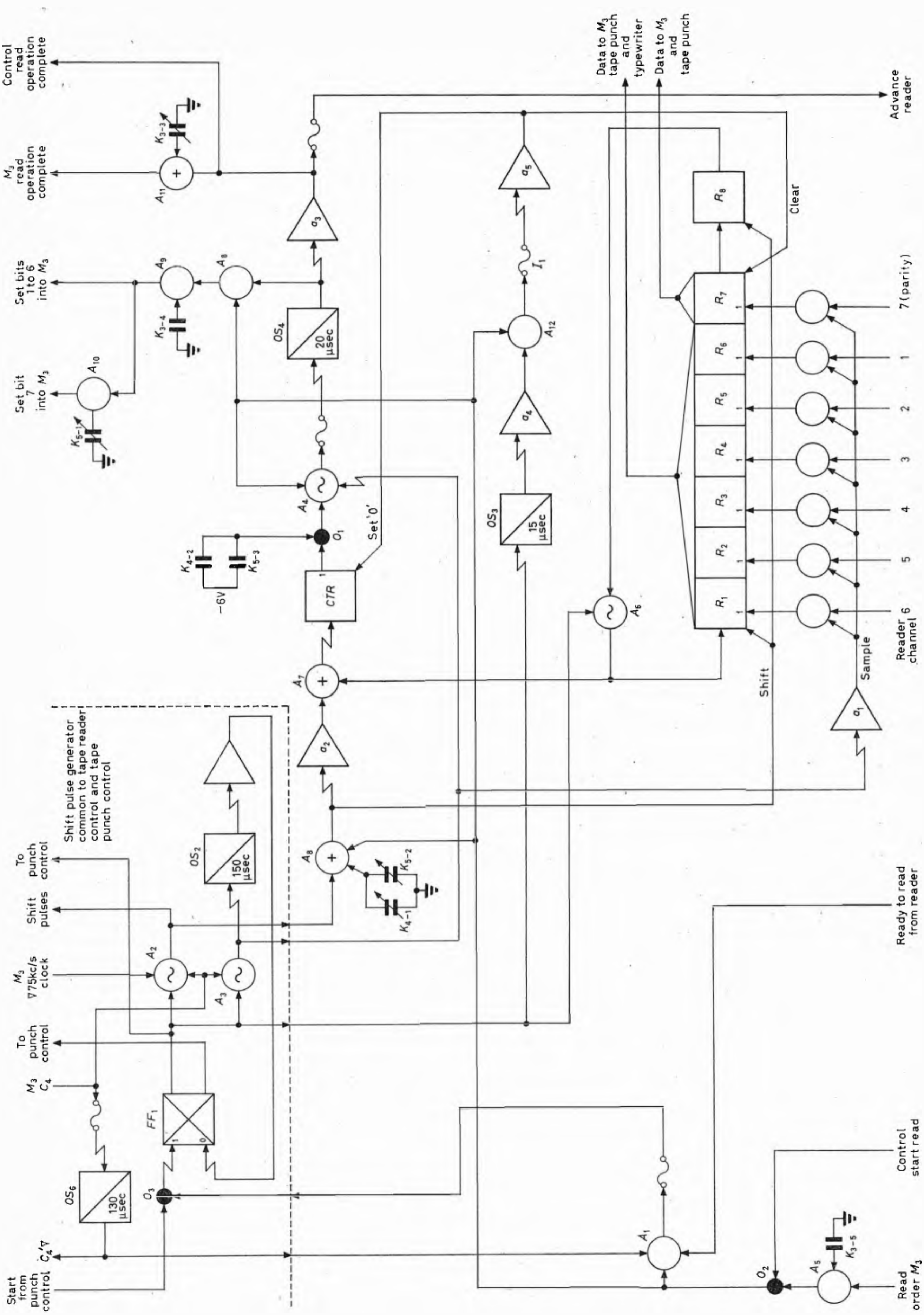


Fig. 3. Tape reader control logic

struction, and the positions of the lamp filament and of the Lucite light guide are chosen to maximize the light presented via the tape to the silicon photocells which form the detecting array.

The tape used has seven data channels and one sprocket channel which is used as a 'clock' to indicate position. The configuration used is: 3 data channels, sprocket, four data channels. Signals from the data channels are amplified and converted to the standard computer levels, −6V for a '1' (hole) or 0V for a '0' (no hole). Inverted outputs are also available to make the reader more flexible. The sprocket hole photocell feeds an amplifier which switches the control flip-flop to the '0' state and energizes the brake. This is shown, schematically, in Fig. 2.

To start the reader a start pulse is required which will set the flip-flop to the '1' state and energize the clutch coil, which causes the advance of the tape to the next character. When the next sprocket hole is detected, the flip-flop is reset and the tape stopped. At the same time a signal is generated to indicate that the character can be read.

The speed is limited to 500 characters/sec. and is determined by the maximum speed of the drive roller (Fig. 1) which has a peripheral velocity of 50in/sec.

To facilitate tape loading, a switch is provided, which when operated, will de-energize both the brake and clutch coils and move the guide pins down so that the tape can be moved into position. A red light indicates when the switch is in the 'tape load' position. Another position on the same switch provides a momentary contact for running out the tape or portion thereof. In the 'run out' position, the clutch is energized and the brake released. Upon releasing the switch, the tape stops. Then the computer will be able to restart the reader as before.

The reader is completely self-contained, needing only a start pulse to advance the tape. A signal is available to indicate when the seven signal lines can be sampled. The output levels of the seven channels and their complements are available at the input-output connector.

The Tape-Reader Control

In the M.3 system 6 of the data channels are used to hold numerical or alpha-numerical information, and the seventh channel is used to hold an odd parity bit. A coupling unit was designed in order to synchronize the reader to the M.3 and to perform the parity check. The logical circuit of this unit is shown in Fig. 3. It will be observed that some of the gates in this figure are controlled by relay contacts (e.g. $K_{3\text{-}5}$). These are shown in the de-energized state, they are operated by a manual selector switch system which enables the whole input-output system of the M.3 to be used for tape preparation, duplication and checking independently of the main machine. The details of this switching system, which is conventional, will be described later. The sequence of operation is as follows.

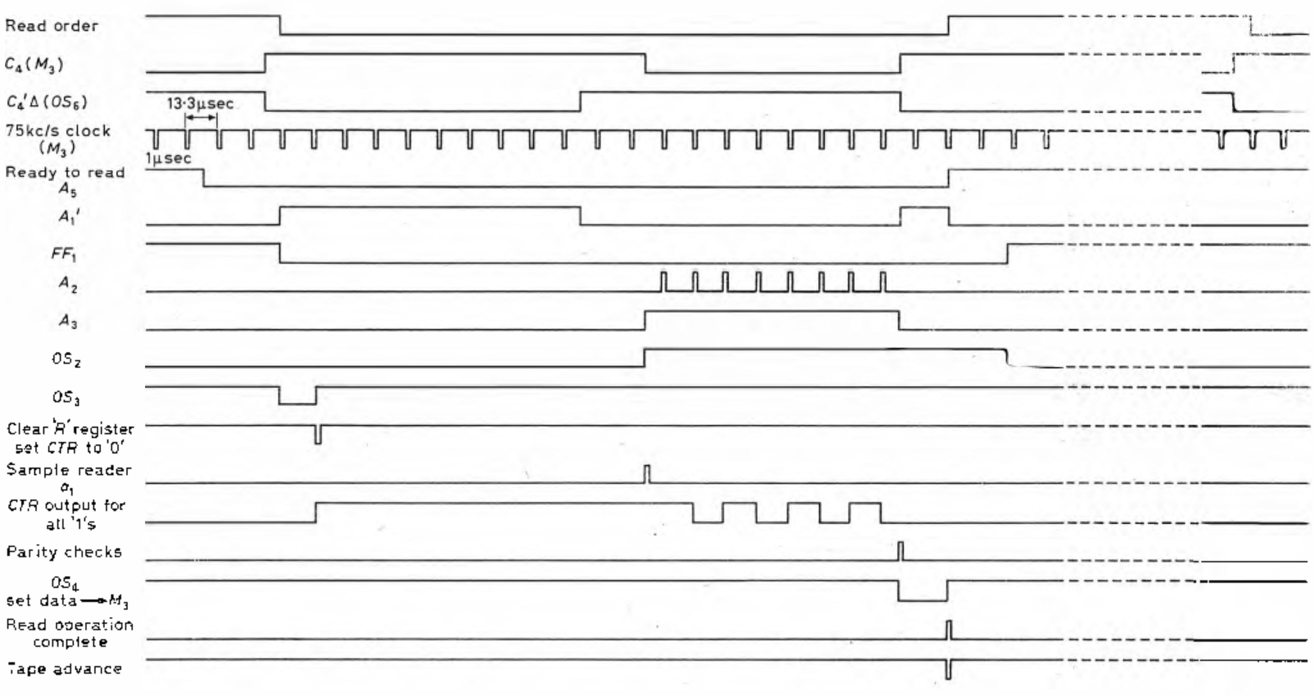
A read order received from the M.3 is fed through the AND gate, A_5, and the OR gate, O_2 to A_1. A_1 is tested and, if the tape reader is ready to be sampled and C_4 delayed is not at −6V, flip-flop 1 is set to the '1' state through O_3.

The waveforms of C_4 and the 75kc/s half clock[1], (Fig. 4) are from the computer and serve as the signals to synchronize the reader cycle with the M.3. C_4 is the fourth stage of the word counter in the M.3 and is negative for the duration of eight pulses at the 75kc/s rate. By singling out one of these groups, eight pulses are available for shift operations. It may be noted that the synchronizing circuit is used for both the reader control and the punch control. This is indicated in Fig. 3. The positive level of C_4 is shortened by the one-shot delay circuit OS_6 in order to avoid starting the read or punch cycle too late, which could result in the loss of the first shift pulse.

The read order from A_1 may now occur at any time during this period (negative level OS_6 output in Fig. 4).

To continue the read cycle after flip-flop 1 is in the '1' state, A_2 is opened and allows eight shift pulses to pass through when the C_4 waveform is negative. At the same time A_3 triggers OS_2 which, after 150μsec resets flip-flop 1.

Fig. 4. Tape reader control timing diagram



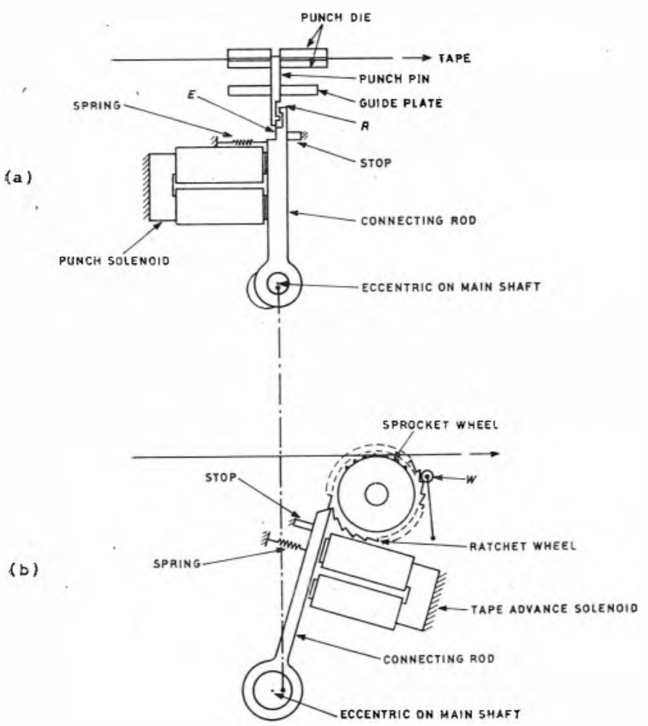


Fig. 5. Basic components of the tape punch
(a) punch mechanism *(b) advance mechanism*

The positive going edge of A_3 also samples the reader and shifts the character into the reader buffer stages 1 to 7 6·6μsec after the sample pulse, the shift pulses start arriving and circulate the buffer contents until the character is back in its original position. Since there are eight shift pulses, a dummy stage R_8 is added. During the shifting operation, R_8 is connected back to R_1 through A_6. The purpose of circulating the character is to facilitate the counting of the number of 'ones' contained in the character. Knowing that this number has to be odd, all that is required is to feed these 'ones' to a preset counter. This is accomplished by feeding the output of A_6 to A_7, which gates out the delayed shift pulses from A_8 and a_2 for 'ones' contained in the output of A_6. The counter CTR, is initially set to '0', thus producing a negative output when the number of pulses from A_7 is odd. This output in turn opens A_4 through O_1 and, if a read order is present at A_8 and if A_9 is open. A_8 and A_9 should have been combined logically, but physically an additional input to A_8 was not possible. At the end of the 20μsec pulse from OS_4 an 'operation complete pulse' is generated in a_3 which is fed through A_{11} to the computer. This pulse also starts the reader, which then advances the tape to the next character.

As mentioned before, CTR was initially zeroed. This is accomplished as follows: 15μsec after FF_1 is set to the '1' state the trailing edge of the pulse from OS_3 through a_4 tests A_{12}. If a read order is present, A_{12} is open and a pulse, generated by invertor I_1 and a_5 sets CTR to '0' and clears the R buffer stages. When the reader has advanced to the next character the line labelled 'ready to read from reader' switches to -6V and the control logic is ready to recycle.

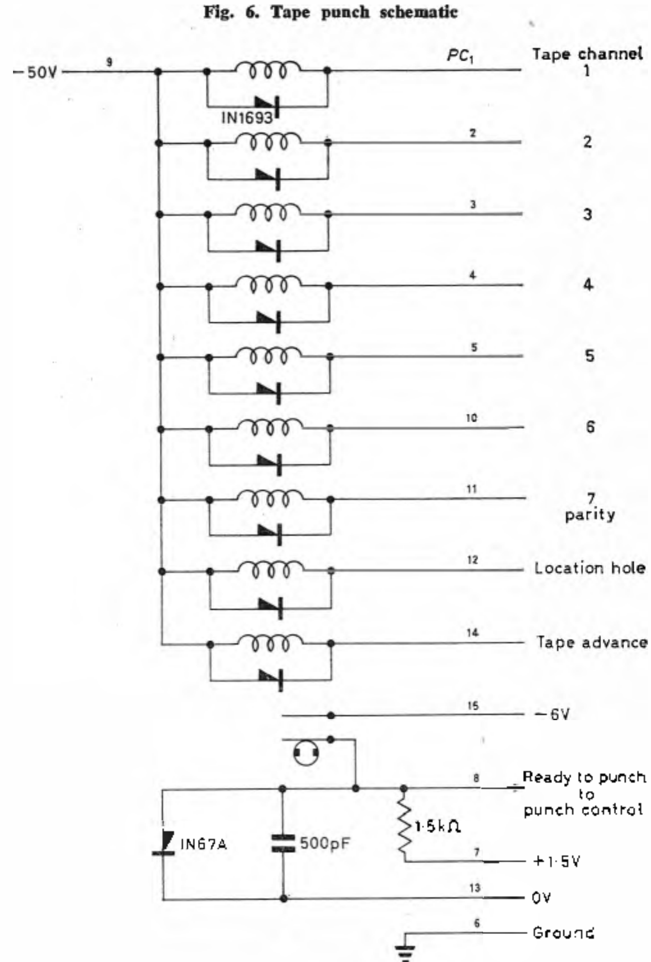
The Paper-Tape Punch

A new tape punch has been developed for the M.3 This has a capacity of 300 characters/sec. when operated in a bath of cooled oil but is derated to 100 characters/sec. and operated in air in the M.3. The basic components are shown in Fig. 5. The punch mechanism is shown in (a) and the tape advance mechanism in (b).

The single main shaft rotates continuously and has two eccentrics, one for the eight punch-connecting rods and the other for the advance-connecting rod. These connecting rods are arranged to reciprocate freely but are held against their stops by springs. To punch a hole, the connecting rod is pulled over by energizing the punch solenoid when the connecting rod is moving up. This causes the edge E to move under the punch pin and push it up, cutting a hole in the tape. As soon as the down stroke begins, the solenoid is de-energized and, on the way down, the step R catches the punch pin and resets it to the normal position as shown. There are eight punch pins of which one, producing the location hole, is always energized regardless of the character to be punched. Besides being used as the index holes for the reader, these location holes are also used to advance the tape by means of the sprocket wheel. Coupled to this sprocket wheel is the ratchet wheel used to translate the linear motion of the advance connecting rod to a rotary motion. This connecting rod also moves with a reciprocating motion missing the ratchet wheel by being held against a stop with the aid of a spring. When the advance solenoid is energized, the connecting rod is moved toward the ratchet wheel causing it to advance sufficiently to be indexed by the spring-loaded wheel W, which corresponds to a tape advance of 0·1in.

The eccentrics on the main shaft are arranged so that the tape can only be advanced after the punches are with-

Fig. 6. Tape punch schematic



drawn from the paper, preventing the tape from being torn.

The punch and advance solenoids are pulsed in succession and controlled by the punch control logic, which is described in the next section.

To indicate when punching can begin, a timing cam mounted on the main shaft, emits pulses to the controlling logic.

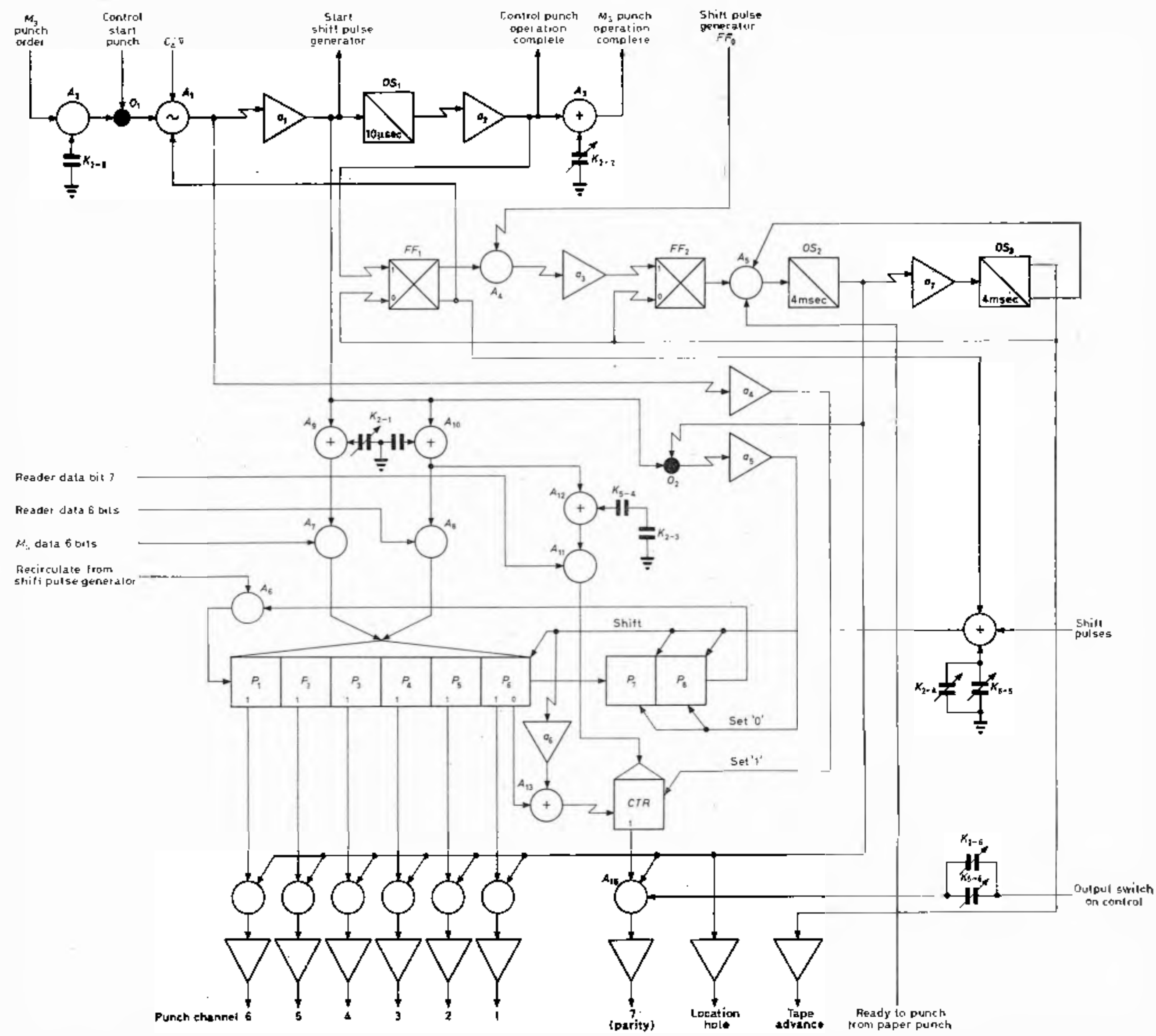
The arrangement of the paper tape punch circuit is shown in Fig. 6. To prevent the high reverse voltage from breaking down the driving transistors, damping diodes are connected across the solenoids but it was not found necessary to use more elaborate methods of damping at the 100 characters/sec rate.

The Tape-Punch Control

The functions of this control unit are: to generate an odd parity bit from the six data bits received from the computer, to add this parity bit to the tape, to synchronize the punch with the computer and disconnect the punch from the M.3 as soon as possible in order to minimize time waste. Fig. 7 shows the tape-punch control logic, while Fig. 8 shows the waveforms at the more important points. It is assumed that the six bits of the character to be punched are in the most significant digits of the shift register of the M.3.

Starting with a punch order from the computer, this signal passes through A_2 and O_1 to A_1. FF_1 is initially in the '0' state, causing A_1 to allow the negative going edge of $C_4'\Delta$ to pass through A_1. Since A_1 is an inverting type AND gate, the positive going output is differentiated by a_1, which starts the shift pulse generator (located in the reader control) and through A_9 and A_7 shifts the data from the M.3 (six bits) into the punch buffer stages P_1 to P_6. At this point the computer is not needed any longer. The output of a_1 also triggers OS_1 and after the 10μsec delay, a_2 produces the 'operation complete' pulse which is fed through A_3 to the computer. FF_1 is set to the '1' state by a_2 and now allows the negative edge of the flip-flop in the shift-pulse generator ('0' side of FF_1 in Fig. 7 which is physically the same as FF_1 in Fig. 3) to set flip-flop 2 to the '1' side through A_4 and a_3. The flip-flop 2 output opens A_5 and waits until the punch is ready to start punching. Before flip-flop 2 is set to '1', the shift pulses

Fig. 7. Tape punch control logic



arrive at A_{14} and, since flip-flop 1 is in the '1' state, these eight pulses shift the contents of the punch buffer, P around the loop through A_6. A_6 is opened by the recirculate pulse also generated in the shift pulse generator (from the '1' side of FF_1, Fig. 7). Since there are six bits in the character and eight shift pulses, two dummy stages are required, P_7 and P_8.

During the recirculating cycle the output of P_6 is gated in A_{13} by delayed shift pulses from a_6, and causes the counter to count. This counter is initially set to '1' and it follows that the output will be '1' for an even number and '0' for an odd number of bits, making the total number of bits to be punched always odd.

As soon as the punch cycle begins, that is, when the output of A_1 goes positive, the counter, CTR, is set to '1' through a_4. The P_7 and P_8 stages of the buffer are set to '0' to ensure that the correct number of bits are counted. P_7 and P_8 are set to '0' when the punch cycle can begin again when it is finished. In this case the output of a_7 and the positive edge of OS_2 are fed to O_2. The output of O_2, through a_5, sets P_7 and P_8 to '0'. The output of OS_2 also resets FF_1 and FF_2. Flip-flop FF_1 is the control flip-flop, while FF_2 is necessary to avoid punching when the shift-pulse generator is used in the read cycle.

The outputs of P_1 to P_6 and the counter, CTR, are fed to the punch solenoids through AND gates which, in turn, are controlled by the output of OS_2. The output of A_5 triggers OS_2 and produces a 4msec pulse which is of sufficient duration to operate the punch solenoids. Fig. 7 shows that regardless of the number of bits in the character, the location hole is punched by feeding the output of OS_2 directly to the location hole solenoid.

Immediately after punching the character the tape must be advanced. This is done by using the trailing edge of OS_2 to trigger OS_3 through a_7. The output of OS_3 now feeds a 4msec pulse to the tape advance solenoid.

The Typewriter

An IBM typewriter is used. This has 48 input lines to energize the solenoids for the alpha-numeric characters, space and punctuation symbols, and functions such as carriage return.

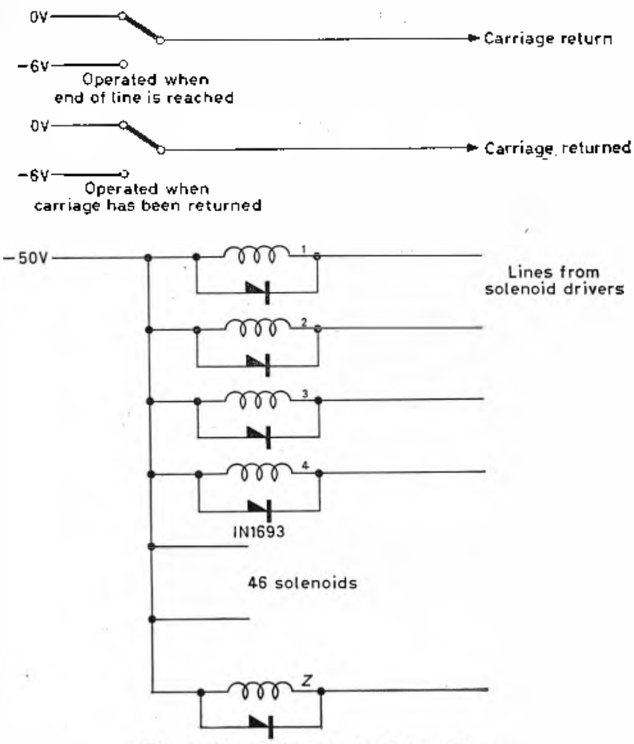
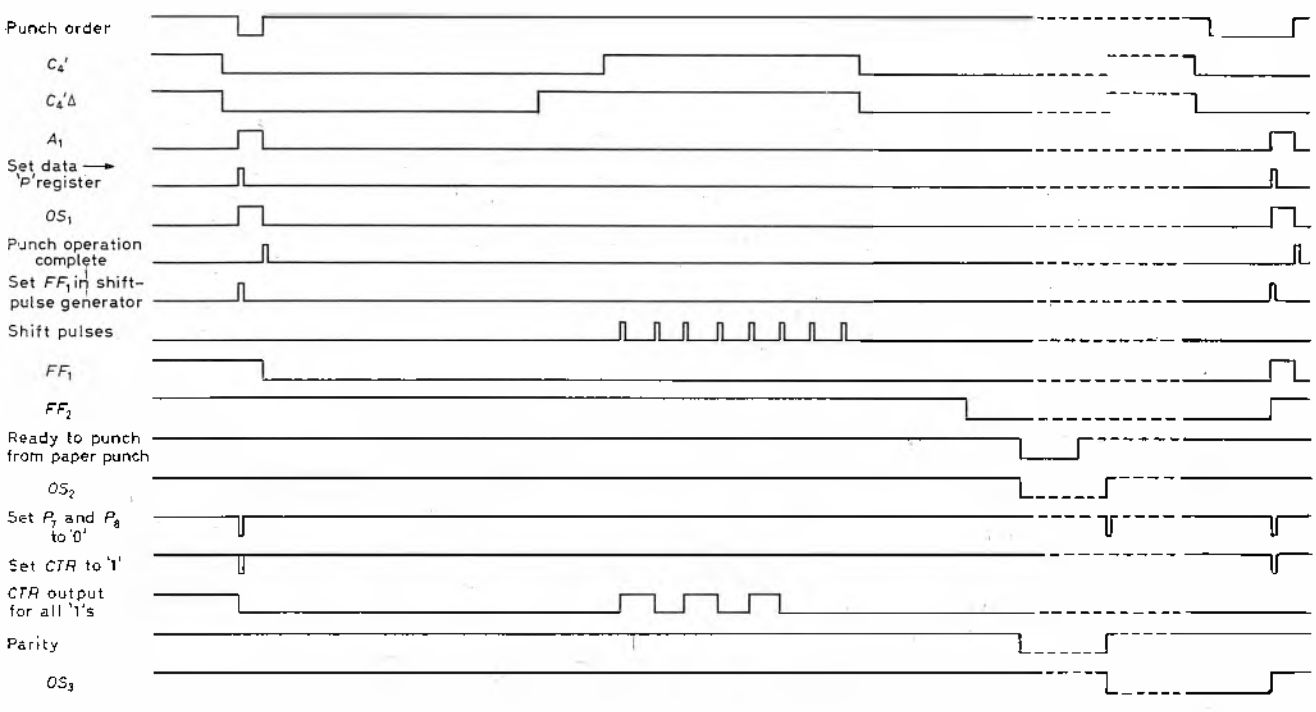


Fig. 9. Electrical elements of typewriter

Fig. 9 shows the essential electrical elements.

Due to the high induced voltage which occurs when a solenoid is switched off, the diodes shown were added to prevent breakdown of the power transistors driving these solenoids. Two microswitches were also added in case the carriage return instruction is forgotten, when it is desirable to be able to return the carriage automatically without losing data. This is accomplished by detecting the end of line with one of the microswitches, thus initiating a $C.R.$ (carriage return) order and also inhibiting the control from feeding data to the typewriter. When the

Fig. 8. Tape punch control timing diagram



carriage has been returned, another microswitch operation signals the control. To ensure proper operation, the signal from this switch is delayed electronically in the control.

The maximum speed of the typewriter is 10 characters/sec.

The Typewriter Control

Before a type order is initiated, it is assumed that the character to be typed is in the most significant end of the shift register in the M.3.

The type order signal is fed through A_6, Fig. 10, and O_4 to A_1, and if the output of O_1 is negative, flip-flop, FF_1 is set to the '1' state. Switching FF_1 on causes a positive pulse from a_1 through A_7 to transfer the data from the M.3 through A_3 to the type buffer T. At this time the computer is no longer needed, and to signify this, the '1' side of FF_1 triggers OS_1. The trailing edge of the 10μsec pulse from OS_1 is shaped in a_2 and fed through A_2 to the M.3, indicating that the operation is complete. The typewriter control is then independent and the following events occur. As soon as the character is fed into the T buffer, the six-bit word is decoded and the appropriate solenoid amplifier energized. The associated solenoid requires a pulse of 25msec duration. This is obtained by setting all the T buffer stages to '1' at the end of the 25msec pulse, generated by OS_2 and a_4. The output from OS_2 resets FF_1 and closes A_1 through O_1. When the OS_2 output pulse ceases, OS_3 takes over to inhibit A_1 in order to prevent the control from accepting the next type order. When the OS_3 output returns to 0V, the control is in the 'ready state' and can start with the next operation.

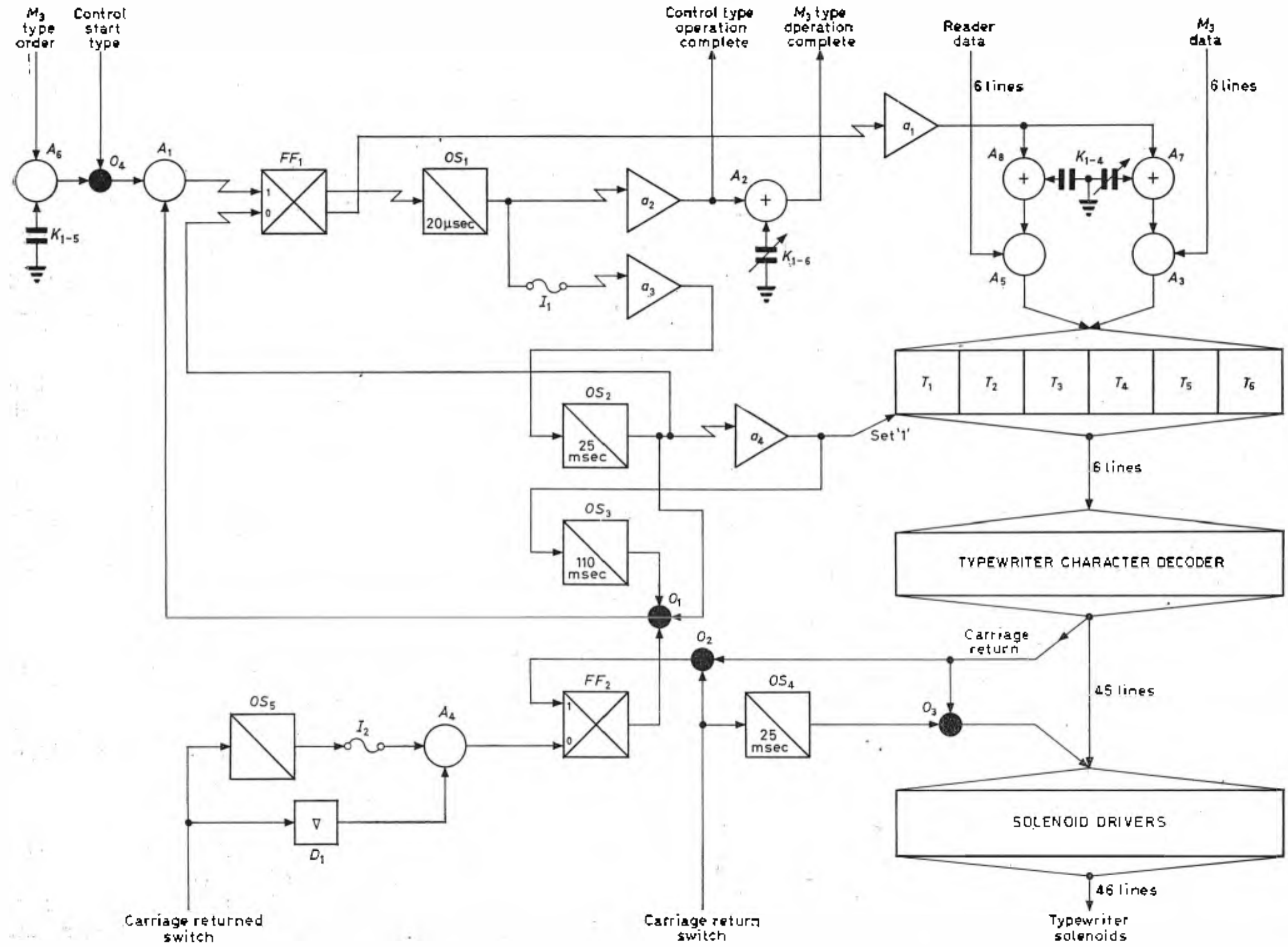
An exception to the above sequence of events occurs in case of a carriage return order or when, in the absence of this order, the end of line switch is sensed.

In the case of the $C.R.$ order, events happen as explained above, but in addition, the following occurs. The 'carriage return' signal from the decoder is fed to the solenoid driver through O_3 and also triggers FF_2 through O_2. FF_2 is now in the '1' state and closes A_1 through O_1, which again inhibits the control. FF_2 remains in the '1' state until the carriage has been returned. When the carriage returns, the 'carriage returned switch' feeds a −6V level to OS_5 and D_1. The combination of OS_5, I_2, A_4 and D_1 delays this level long enough to allow the carriage to settle down. The OS_5 time-constant was chosen for the worst case condition, which occurs when the carriage is in the extreme or end of line station. When released at that time it will accelerate and bounce at the reset position.

The time required to return the carriage from the end of line is a full second; however, by using FF_2 rather than a one-shot of 1sec duration, this time is reduced to a minimum. This saving in time may be significant, especially if rows of numbers are to be typed.

When a $C.R.$ order is forgotten, and if the end of the line is reached, the 'carriage return switch' feeds a −6V level to OS_4 and through O_2 sets FF_2 to the '1' state. OS_4 energizes, through O_3, the $C.R.$ solenoid for 25msec. FF_2

Fig. 10. Typewriter control logic



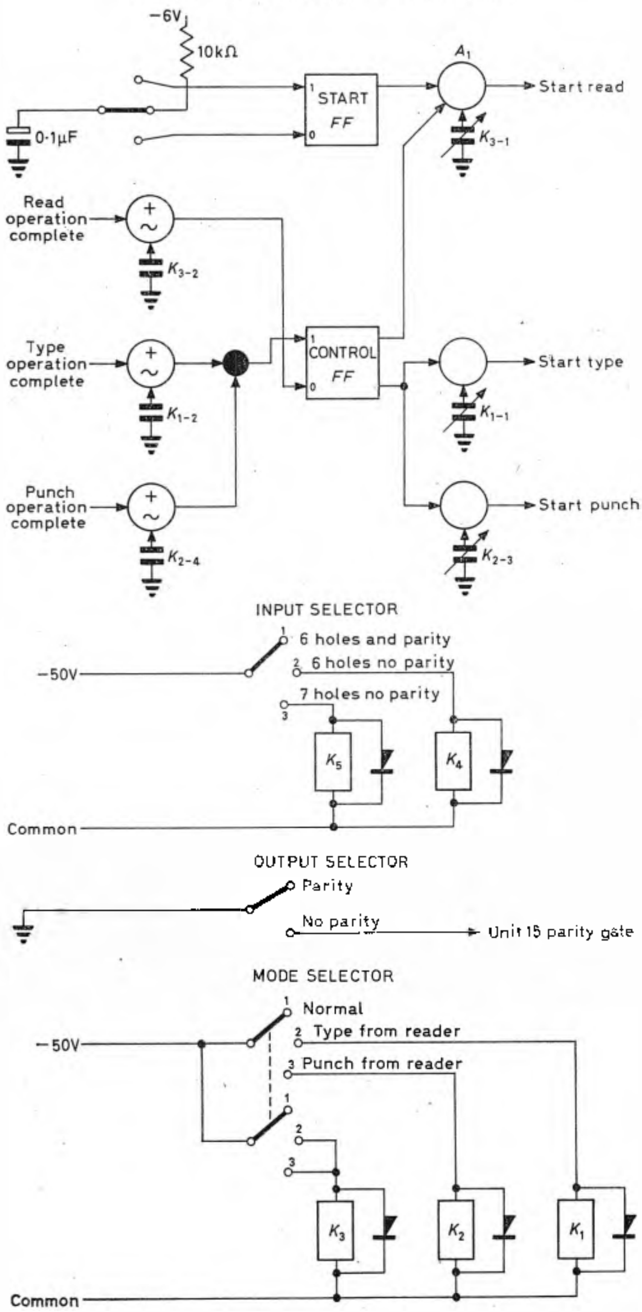
inhibits the control until the carriage has been returned, after which normal operation can resume.

Because of the lack of a + on the typewriter, a space is typed instead. The reason for doing this is that the output routine for the M.3 will instruct the output device to print or punch a + sign. In order to line up a column of both positive and negative numbers, it is necessary to provide a space for a + sign.

Peripheral Equipment Control

For maximum flexibility of the peripheral equipment, it seemed desirable to be able to operate either the tape punch or the typewriter directly from the tape reader. Thus, if one of the frequently used tapes becomes worn, it is possible to copy it by reading it through the reader, which is then directly connected to the tape punch. In this mode of operation it is also possible to read tapes without a parity bit and to produce a similar tape with the parity bit inserted. Alternatively, a tape with a parity bit can be reproduced with the parity bit deleted.

Fig. 11. Peripheral equipment control logic



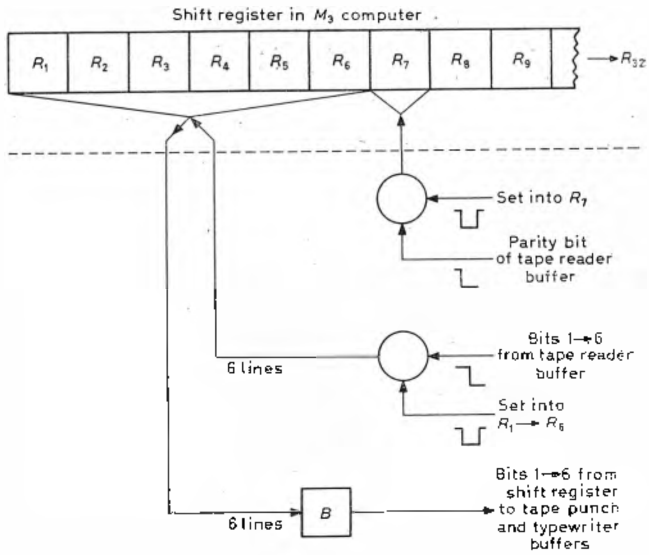


Fig. 12. Peripheral to computer coupling logic

Another advantage of this facility is the gain in speed, since typing occurs at a rate of 10 characters/sec and punching at a rate of 100 characters/sec. This is achieved by using the punch as the output device and then feeding the tape to the reader on mode 2 (type from reader) at any time thereafter. The tape then also serves as a permanent storage.

Different modes of operation can be selected by manipulating the control switches and using the combination that will give the required result. The connexions are shown in Fig. 11. Relays are used since switching will not occur during computation; however, the logic is so designed that if the switch positions are altered, no interference will result.

Fig. 11 shows that the additional logic required consists of a start flip-flop, a control flip-flop and some gates.

For modes 2 and 3 on the 'mode selector', the start-stop switch has to be used. When the computer is switched on, FF_1 (Figs. 3, 7, 8) is set to the '0' and FF_2 to the '1' state. When operating the start switch, FF_1 is set to the '1' state and the reader control logic receives a start pulse through A_1 and O_2 (Fig. 3). A character is then read from the tape, checked for correct parity, if required, and stored in the R buffer. The read 'operation complete pulse is now received and this sets the control flip-flop, FF_2, to the '0' state. This, in turn, starts the punch or type cycle, after which an 'operation complete' pulse is emitted from the appropriate unit, which sets the control flip-flop, FF_2 to the '1' state. The above cycle will be repeated until the start-stop switch is pushed to the stop position.

A final point to note is that when mode 2 (type from reader) has been selected, the punch can still be used as an output for the M.3 and may be operated simultaneously. The same condition applies when mode 3 (punch from reader) has been selected, in which case the typewriter can function as the output device.

Peripheral Equipment to Computer Coupling

The only data link between the M.3 and the peripheral equipment is the shift register of the computer.

Normally, the six most significant stages of the shift register will receive the six-bit code from the tape reader as shown in Fig. 12. The significant bit is fed to R_1 and the least significant to R_6. These six bits provide 64 possible combinations, but in the case of sampling analogue data, an accuracy of 1 per cent or better is required. By deleting the parity bit, it is possible to use seven bits for data in order to obtain 128 combinations and then the analogue signal can be represented by a single character with a 1 per cent accuracy if 100 discrete values of amplitude are used. In this case the seventh bit is then fed to R_7. This mode of operation is selected with the input selector switch.

The outputs of the six shift register stages are continuously available to the peripheral equipment. These signals are buffered with emitter-followers.

In the case of either input or output, it is the responsibility of the programmer to provide the necessary instructions to place the character in the appropriate location of the shift register or to remove the character read into this register.

Conclusion

The complete M.3 system is shown in Fig. 13. The input-output console can be removed from the machine and can be replaced by an analogue-digital, digital-analogue convertor system when it is desired to use the machine in process control applications. The system has now been operating for nearly two years and seems to have a mean-free path between faults of over three months. Such reliability is surprising in a drum-type machine especially as the M.3 drum has been in operation, (partly on an earlier M.2 system) since 1956 and has never been serviced. It is proposed to replace this drum with a core store so constructed as to simulate the serial behaviour and digit rate of the drum but to eliminate the latency of the latter.

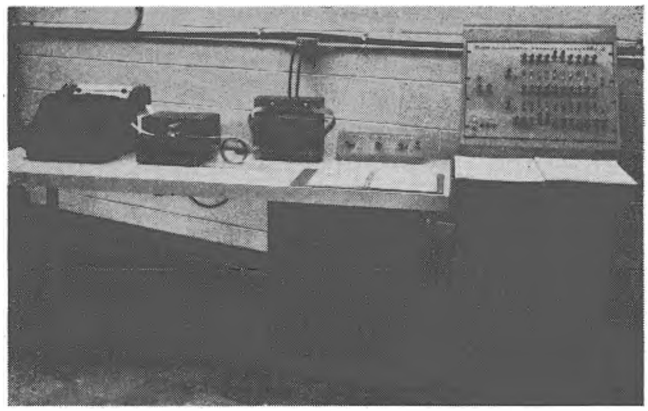


Fig. 13. The complete M.3 system

Acknowledgments

The authors wish to thank Mr. H. Dobbs for his help in constructing the photo-electric reader and to acknowledge, with gratitude the financial support of the National Research Council under Grant N. BT140 and of the Defence Research Board under Grant No. 9931-12.

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2. Booth, A. D. The Physical Realization of an Electronic Digital Computer: Input and Output. *Electronic Engng.* 30, 570 (1958).
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A Suitcase Training Computer

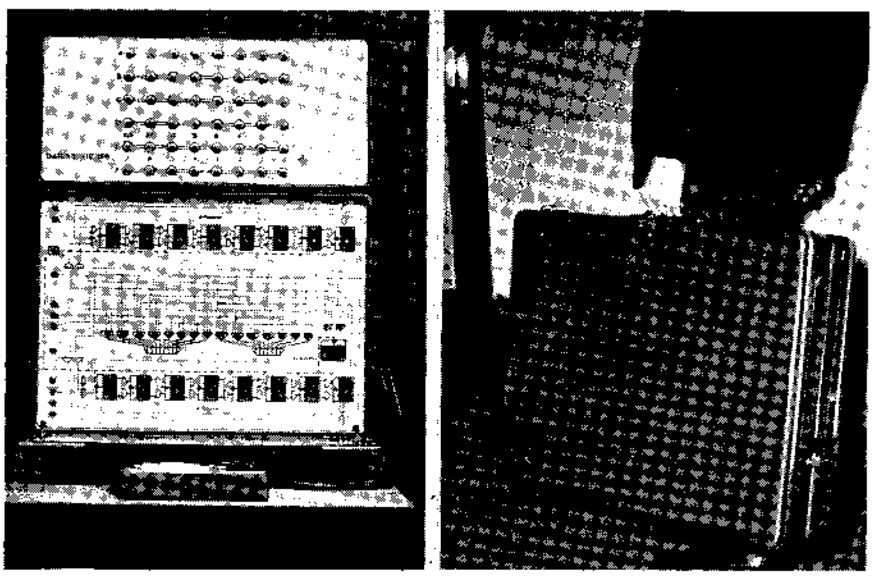
Telefunken A.G. of West Germany is producing a simple digital computer which is intended for teaching purposes.

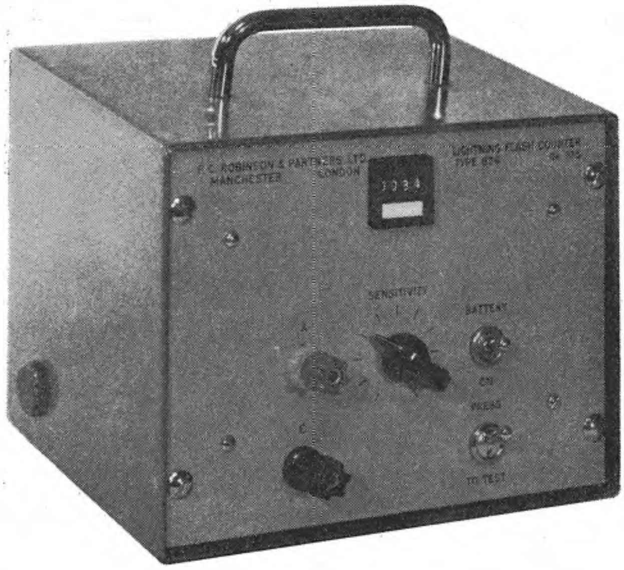
This demonstration unit is built into a three section suitcase which is relatively easy to transport and can be set up anywhere a mains connexion is available. When the case is opened the lower section forms the base containing the control unit which carries the push-buttons and switches required to perform the computing operations. On this base the computer proper stands upright with a clearly drawn circuit diagram and above it, as upper section, the storage in six rows of eight cells. Computer and storage are permanently interconnected by means of special flexible cables in spring-loaded guides, the control unit is connected by means of 'Tuchel' plugs.

By means of pilot lamps which light up each time a signal reaches the clearly marked binary points, storage cells or flip-flop stages, the trainee can trace the various phases of the switching procedure by which the command he has inserted is carried forward by the machine, linked up and stored. The clock frequency generator is operated by means of a slowly running telephone dial in such a manner that the computing process remains perceptible. At the same time it is possible to illustrate each operation separately.

Since the logical connexions and arithmetical operations of the digital computing technique can be traced visually on this demonstration computer, the trainee memorizes very easily the machine sequence of numeric computations in the computer. He sees what the machine does and how. Such a visual demonstration leads to quicker appreciation of the mechanical and electronic procedures in the machine during the processing of commands. In addition programming tasks can be practised on the control unit.

The suitcase computer shown ready for use on the left and packed for transport on the right





A Lightning Flash Counter

By R. H. Golde*, D.Sc., M.I.E.E.

A lightning flash counter is described which has been adopted by a C.I.G.R.E. Study Committee for international use with a view to determining the frequency of occurrence of lightning flashes to earth.

(Voir page 203 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 210)

TO the layman, lightning and thunder constitute either the most majestic natural spectacle or a source of intense anxiety and fear. On the other hand, lightning represents the major source of interference for the telecommunication engineer; it is a source of irritation to the architect who dislikes seeing his buildings festooned with lightning conductors; it causes concern to the aircraft operator and pilot, and in many countries it is responsible for more interruptions of electricity supply than any other cause.

Much is known today about the physics of the lightning discharge and remedial measures can be designed for most purposes. However, if lightning protection is to be arranged on an economical basis, the one factor which assumes predominant importance is the frequency of lightning discharges to earth and it is this factor on which little reliable information is available on a world-wide basis. The frequency of occurrence of earth discharges has been estimated from visual observations, from recordings of certain physical manifestations of the lightning discharge, and from the frequency of lightning strokes to electrical transmission lines. All these methods have their limitations, the most important of which is the short period of observation to which published data are confined.

Having recognized this situation, the E.R.A. approached in 1956 Dr. E. T. Pierce[1] who was then engaged on lightning research at the Cavendish Laboratory, Cambridge, with the request to suggest a device which would be capable of counting the number of earth flashes within a given area and which would be so cheap that it could be installed in large numbers and so simple that it could be operated by non-technical personnel all over the world. In due course, Dr. Pierce produced a simple electronic circuit which fulfilled the last two conditions and which, according to the information then available, was expected to operate on earth flashes only. His original circuit was modified in co-operation with the E.R.A. and in 1958 the first commercially produced lightning flash counter was installed on the Association's premises in Leatherhead.

At about the same time the writer, in his capacity as secretary of C.I.G.R.E. Study Committee No. 8 (Lightning and Surges), informed that Committee of the new development and, after an examination of alternative devices which had been suggested from time to time, the so-called E.R.A. lightning flash counter was adopted by the Study Committee. As a result of this official recognition, counters were installed in fairly large numbers in Scandinavia, Italy, Australia, Rhodesia, Kenya and Japan and later in Germany, Czechoslovakia, South Africa and several other countries. The experience gained with these installations, coupled with research into the relevant characteristics of the lightning discharge, was discussed at annual meetings of a special working group set up by the C.I.G.R.E. Study Committee. As a result of these deliberations, an improved version of the counter[2] was adopted in May 1965 and it is this which will be described in the present article. However, before doing so the physical principles on which the design is based must be briefly reviewed.

A thundercloud[3], which usually extends from one or a few thousand feet to thirty thousand feet or more above ground, contains essentially negative charges in its lower regions and positive charges in its upper part. Some sixty to eighty per cent of all discharges occur between these charge centres and these are called cloud flashes. Most of the remaining discharges occur between lower negative charge centres and earth and these are termed earth flashes. A lightning discharge involves, therefore, a movement of charges and these effects can be registered by oscillographic recording of the resulting field changes on the earth. From such recordings and from photographic studies, it is known that an earth flash consists of a faint leader stroke which travels from the cloud towards earth at a speed of about one thousandth of the speed of light. When the leader stroke contacts the ground, an intense return stroke follows which proceeds from earth to the cloud at a speed which attains almost the velocity of light. This process may be repeated several times, in which case the lightning flash consists of a succession of earth strokes almost all of which follow the same channel so that the whole process of charge movement may last a second or more. A cloud flash also starts with a leader stroke but instead of the severe return stroke in the former case this is followed by a gradual neutralization of the charges involved.

The above photograph shows a version of the lightning flash counter manufactured by F. C. Robinson & Partners Ltd.

* *The Electrical Research Association.*

The field changes caused by the different types of lightning discharge cover a fairly wide frequency spectrum ranging from very low frequencies to frequencies in the kilocycle per second range. With varying distance between the lightning stroke and the place of recording, this spectrum changes appreciably both with respect to its amplitude and wave pattern. However, in the great majority of cases a lightning flash produces a quasi-continuous oscillatory discharge spectrum and it is this which constitutes the main source of interference in high-frequency telecommunication. One of the best known lightning flash counters[4] for use in radio research utilizes this fact and it was therefore given a characteristic which reaches its maximum response at 10kc/s.

A lightning flash counter which is intended to respond to earth strokes only requires different design considerations. As mentioned above, it is the occurrence or absence of a return stroke which affords a definite distinction between earth and cloud flashes and it seems, therefore, obvious to base the design of such a counter on the field change caused by a return stroke.

The amplitude of this field change is a complex function of the intensity of the lightning current and of the distance between the point of recording and the lightning channel. However, if the counter is intended to register earth flashes over a distance not exceeding some tens of miles, the amplitude of the resulting field change can be shown to be proportional, as a first approximation, to the amplitude of the lightning current and inversely proportional to the cube of the distance so that it decreases rapidly with increasing distance. This feature greatly facilitates the determination of the effective distance over which the counter is liable to be actuated by earth flashes, the intensity of which may vary from a few hundred to more than a hundred thousand amperes.

The shape or time variation of the field change within the distances specified is mainly a function of the wave shape of the lightning current and this is well known from many thousands of oscillographic records. On the basis of this information, it was decided to design the counter circuit in such a manner that it would reject the very slow field changes associated with cloud discharges but would respond to field changes with an equivalent frequency of a few hundred cycles per second. A simple electronic circuit was consequently developed which has since been gradually modified in accordance with experience gained over several years and in different countries.

The circuit of the lightning flash counter in its present

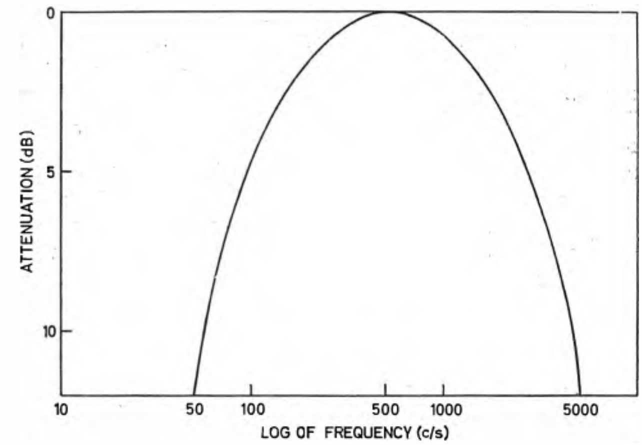


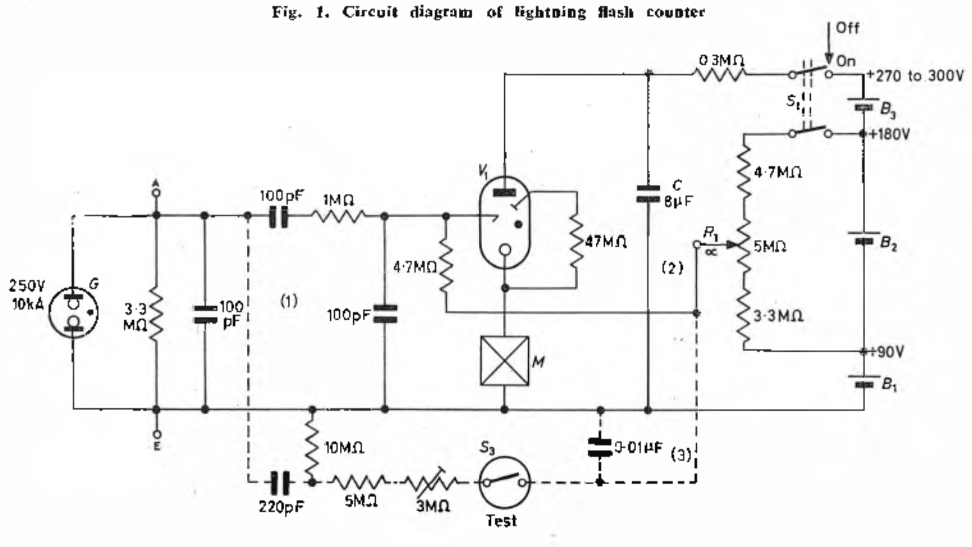
Fig. 2. Response characteristic of filter circuit

state of development is shown in Fig. 1. It consists essentially of three parts, part (1) being the input circuit, part (2) the recording circuit, and part (3) the testing circuit. At point A, the counter is connected to an aerial system of given dimensions, the capacitance to earth of which amounts to about 220pF. A signal received from that aerial is applied at point A to the resistance-capacitance filter circuit (1) which has a maximum transmission at 500c/s while attenuating signals of higher or lower frequencies in accordance with the response characteristics of the filter circuit shown in Fig. 2. The output of this filter circuit, superimposed on a positive bias voltage obtained from the potentiometer across the battery B_2, is applied to the trigger electrode of a cold-cathode triode V_1. The bias voltage is set below the critical voltage of the triode so that in the quiescent state V_1 is not conducting and capacitor C is charged to the full voltage of the battery supply when switch S_1 is closed. The auxiliary electrode in the valve V_1 provides adequate ionization to ensure consistent operation of the valve over long periods of service.

When a signal exceeding 25V (equivalent to a field change of 5V/m) is supplied to point A, the triode V_1 is fired and the capacitor C in the recording circuit (2) is partially discharged through the message register M which records one operation. The resistance in the discharge circuit is so chosen that the triode is deionized after the discharge, and capacitor C is recharged from the battery at a rate which enables the counter to register field changes due to separate lightning flashes in rapid succession while at the same time avoiding operation of the counter by successive discharges in a lightning flash consisting of several component strokes.

Part (3) of the diagram in Fig. 1 incorporates a test circuit which is intended to be used daily to check the correct operation of the device by applying to the grid of the triode a pulse of 25V amplitude and of the correct wave shape. In addition, external test methods using either a signal generator or a

Fig. 1. Circuit diagram of lightning flash counter

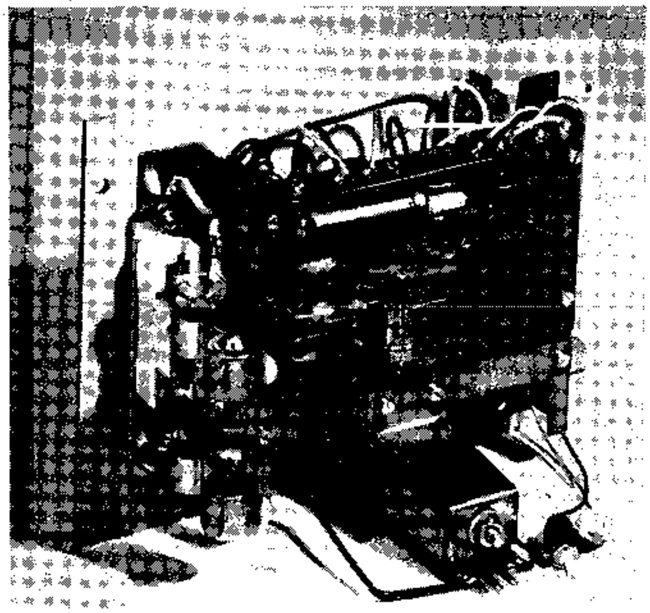


small impulse generator have been devised to check that the frequency response of the counter lies within the following tolerances: 25V (peak) ±1V (peak) at 500c/s and 42·5V (peak) ± 5V (peak) at 100c/s and 2 500c/s. To achieve this standard of reliability, all components have to be selected within specified tolerances and all insulating materials incorporated in the counter have to be of high quality and suitable for operation in tropical conditions. Similarly, the insulation used for the suspension of the aerial and for the lead connecting the aerial to the instrument must be such that the resistance between the aerial and the earth terminals of the counter falls under no conditions below 100MΩ.

Shortly after its initial installation, it was noted by visual observation that the counter was occasionally operated by cloud discharges. This was found to be due to formerly unsuspected field changes which are superimposed on the typical slow field changes produced by cloud discharges and some of which have components of a frequency of a few hundred cycles per second. It must thus be admitted that the lightning flash counter as at present designed will operate on a certain proportion of cloud flashes. Further research is therefore concentrated on two distinct subjects, viz. the development of a counter which will respond only to earth discharges, and to the interpretation of results obtained with the present counter with respect to the frequency of occurrence of earth flashes.

So far as the first problem is concerned, it is clear that a counter which counts only earth flashes requires a much more sophisticated circuit than the present design. Work in this direction is proceeding following a suggestion of Prof. Malan of the University of the Witwatersrand in Johannesburg but much more work will be required before this device is fully developed. In view of its complexity, this counter is unlikely to be used in large numbers but it is hoped to use it as a calibrating device for the simple counter described in this article.

Meanwhile, considerable progress is being made in determining the effective recording range of the existing counter and in estimating the percentage of counts which are due to earth flashes. The first results which have been obtained in this respect have been evaluated in terms of the frequency of occurrence of earth flashes in different parts of the world, and valuable information has already been



Interior view of flash counter

obtained. Thus for temperate regions a figure of about four earth flashes per square mile per ten thunderstorm-days appears to constitute a reasonable estimate. A transistorized version of the present counter is being developed in Australia.

Summarizing the present state of the art, it can be stated with confidence that, while more research work is required, the lightning flash counter described in this article should in due course be capable of providing urgently needed information on the frequency of occurrence of lightning flashes to earth with the ultimate aim of providing a world map showing the geographical distribution of this vital parameter.

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A Semi-Automatic Hardness Testing Machine

Two of the main causes of errors in Rockwell hardness testing machines are:

(a) Lack of precision in timing of the major load application period. This needs to be accurate to within a fraction of a second to ensure good repeatability.

(b) Vibration during the testing period which may cause deeper penetration with a resultant error in reading.

In the latest machine manufactured by Engineering and Scientific Equipment Ltd of Alperton, Middlesex, these difficulties have been overcome by the use of an electronic timer and pneumatic control. The former ensures accurate and precisely repeatable timing while the latter means that no levers have to be operated during the test period which thus obviates one of the main causes of vibration.

The new machine, the type SPRB, has all the features of the standard 'PRB' machine, having 7 loads (30, 60, 100, 125, 150, 187·5, 250kg); all these loads can be operated by presetting of a lever. The machine also incorporates an automatic zero setting mechanism and fast lifting and lowering lever for quick loading and unloading of components in and out of the machine.

The operating cycle is completely automatic and can be restarted in two ways:

By moving the fast lifting lever away from the operator and bringing the component in touch with the diamond indentor, using an internal electrical signal to start the cycle, or by a press-button switch situated in the front panel of the instrument case.

The automatic cycle operates as follows:

(1) The minor load is applied.

(2) The major load is applied for a period which is pre-set by an electronic timer situated on the right-hand side of the instrument case; this electronic timer can be set between 2 and 8sec in steps of 1sec, to conform to B.S.S.

(3) The major load is removed and the indicator light on the front panel of the instrument case comes on giving a signal to the operator to read the 'Rockwell' value on the dial indicator.

The fast lifting and lowering lever mechanism is brought forward by the operator, the component removed, and the machine is then ready to be loaded with another component. The same cycle applies to a machine called the 'SRB'; this machine is identical with the 'SPRB', but has no fast lifting and lowering lever mechanism. The lifting and lowering of the component towards the identor is done only by turning the capstan wheel.

Dynamic Frequency Response Measuring Equipment

By G. R. J. Baldwin*, B.Sc. Grad.Inst.P.

Equipment capable of measuring the phase and amplitude responses of active and passive devices under r.f. pulse conditions with very fast frequency modulation is described.

The accuracy of measurement is $\pm 1°$ in phase (within a 100° range) and $\pm 0{\cdot}1$dB in amplitude (over a 10dB range) for variations in phase and amplitude at rates up to 2·5Mc/s, with f.m. rates up to several Mc/s per microsecond.

The equipment is designed to work in r.f. bands from 400Mc/s to 10 000Mc/s, with interchangeable r.f. heads for the various bands. Pulse width and repetition rate are variable over a wide range.

(Voir page 203 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 210)

THE response of an active or passive electronic device to an input signal whose frequency is swept rapidly between two limits is difficult to predict accurately, but is in general different from the response of the device to a slowly varying frequency sweep.

For instance, the 'dynamic frequency response' of a tuned circuit shows three basic differences from the true response as the sweep rate is increased:

(1) There is an apparent displacement of the resonant frequency, increasing as the sweep speed increases. The direction of displacement depends on the direction of the sweep. A sweep with increasing frequency causes an apparent displacement of the resonance towards a higher frequency and vice versa.

(2) The amplitude of the resonance decreases with increasing sweep rates, giving an apparently increased bandwidth for the circuit (i.e., the Q value of the circuit is apparently reduced).

(3) Ringing occurs on the downward part of the resonance curve. Corresponding changes occur in the phase response of the circuit.

These effects occur because a fast frequency sweep does not allow sufficient time for the resonance to build up to its normal maximum. Fig. 1 shows the effects on the amplitude response qualitatively for three different sweep rates, and a mathematical treatment of the subject is given elsewhere[1].

It is obviously necessary to avoid these effects when the true frequency response of a circuit is to be determined by an instrument which uses the swept frequency principle, and this is easily done by limiting the rate of frequency sweep to a value acceptable for the circuit under test. However, in some modern electronic systems the behaviour of circuits to very fast frequency changes—often during short pulses—is important, and some form of measurement equipment is required which will make tests under these conditions.

The equipment described in this article was designed to display amplitude and phase responses of active and passive devices at frequencies in various bands from 400Mc/s to 10 000Mc/s during single or multiple pulses of variable width and repetition rate while the frequency is swept at rates up to several Mc/s per microsecond.

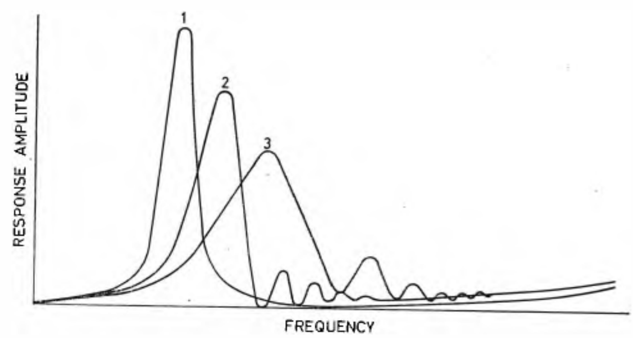


Fig. 1. The effect of three different frequency sweep speeds on the apparent resonance of a tuned circuit. *The sweep speed increases from curve 1 to curve 3.*

The equipment does not employ a sampling technique—every pulse is measured and displayed on oscilloscope display units for pulse rates up to 1kc/s—so that irregular differences between successive pulses can be observed.

The major part of the measurement equipment has a continuous electronic calibration, which eliminates errors in the display units, and can indicate many of the faults which might occur in the equipment.

Principles of the Measurements

The measurement principles are simple, although considerable development has been necessary to produce a practical system which can work with accuracy at the high speeds required.

(Phase and amplitude variations at rates up to 2·5Mc/s can be handled by the equipment during pulse measurement).

Referring to Fig. 2, it can be seen that a radio-frequency head, which is interchangeable for different frequency bands, sends frequency modulated r.f. pulses through the object to be tested, to a matched load. Samples of the input and output of the test object are obtained from directional couplers, and these samples are compared by

* *The Marconi Company Ltd*

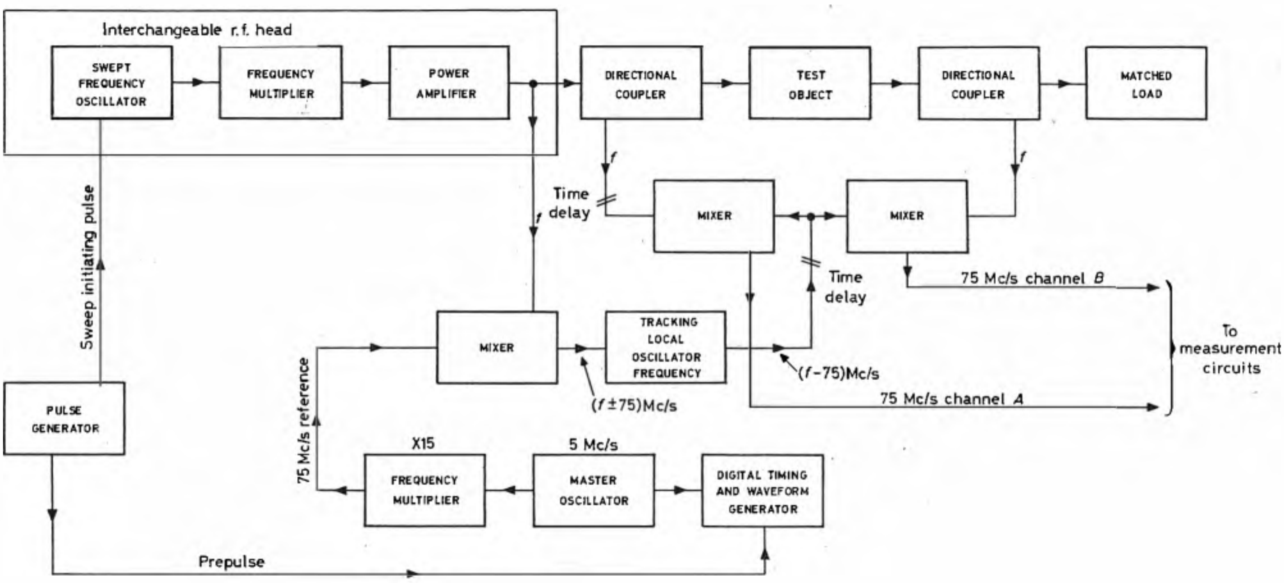


Fig. 2. R.F. part of the equipment

the actual measurement equipment. Pulse duration and repetition rate are controlled by a separate pulse generator, which produces a pre-pulse to switch in the measurement channels before each pulse occurs. Between pulses the equipment has an automatic electronic calibration, which will be described later.

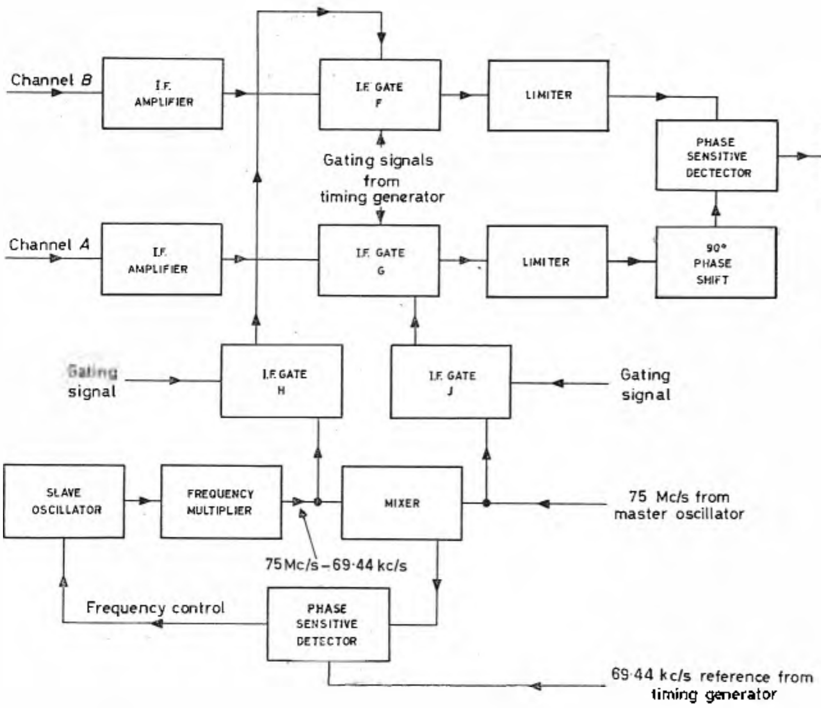
Direct frequency calibration of the display during pulses is not possible, since the fast f.m. sweep does not allow sufficient time for the usual beat frequency calibration methods to give sharp calibration pulses. However, as the law of frequency sweep of the type of r.f. swept oscillator which is used is virtually linear[2], time calibration of the X-axis on the displays, together with a knowledge of the initial and final frequencies in the swept pulse, is sufficient for most purposes.

Pulses derived from a 5Mc/s master oscillator are used to brighten the display screens and give a series of equally spaced dots along the X-axis of the displays. These dots are drawn into vertical lines by applying a slow sawtooth waveform to the display Y-axes during the time calibration periods. The start and finish frequencies of the r.f. pulses can be obtained by a beat frequency method, even though this method is not suitable for continuous frequency calibration.

Conversion to a Fixed Intermediate Frequency

The 5Mc/s crystal controlled master oscillator is used to govern the timing of the calibration equipment by

Fig. 3. Phase measurement and calibration system



digital counting and pulse forming circuits, and also provides a 75Mc/s signal by frequency multiplication. By mixing this 75Mc/s with a sample of the frequency swept r.f. pulse, a signal is obtained whose frequency follows the variation of the pulse frequency, but at a spacing of 75Mc/s below the r.f. This signal is amplified and used as a tracking local oscillator signal to feed two mixers in the main signal paths from the test object. The outputs of these mixers are therefore at a fixed 75Mc/s intermediate frequency, but the signals carry information about the phase and amplitude variations introduced by the test object. Time delay through the test object is compensated by introducing equal delays in the positions shown in Fig. 2.

Conversion to a fixed intermediate frequency facilitates the extraction of the phase and amplitude information, which is faithfully transferred to the intermediate frequency provided that the local oscillator level is much higher than the measurement signal levels.

For convenience, the signal from the input of the test object is labelled 'channel *A*', and that from the output is 'channel *B*'.

Phase Measurement

Fig. 3 shows the system for phase measurement and phase calibration.

Taking the actual measurement first, the signals in channels *A* and *B* are amplified and limited to remove amplitude modulation, after which they are fed into a phase sensitive detector, channel *A* being phase shifted by 90°. The phase sensitive detector is essentially a double balanced ring modulator, which in conjunction with the 90° phase shift gives a cosine law of output for input phase differences. This results in cramping at the extreme edges of the phase scale, but with the calibration system employed it is easily possible to read phase angles to $\pm 1°$ over a 100° range at the scale centre.

The output of the phase detector is amplified by a video amplifier and displayed on an oscilloscope which is modified to allow the very fast bright-up required for the phase calibration.

Calibration signals are switched into the phase measurement system by the two i.f. gates *F* and *G* at all times when the equipment is not actually engaged in measuring a pulse.

Channel *A* receives a sample of the 75Mc/s signal derived from the master oscillator, and channel *B* receives a signal which differs from 75Mc/s by a small amount (actually 69·44kc/s). This signal comes from a slave oscillator, whose frequency is locked by a feedback system and a signal from the timing chain, to give the fixed frequency difference.

The phase detector thus sees the constant difference frequency in the two channels, which amounts to a linear change of phase with time, and its output is a sinewave at the difference frequency. This is fed to the display oscilloscope as usual, but only small sections of it are made visible by brightening the oscilloscope trace for approximately 4nsec at every 5° along the sine wave.

The dots so formed are drawn into horizontal lines by allowing the sine wave to slip across the screen at a rate fast enough for persistence of vision to integrate them. To achieve this a composite trigger waveform is required for the phase display oscilloscope so that the measured pulses are synchronized on the screen, but the calibration wave is not.

This trigger waveform is obtained from the main timing chain, together with the bright-up signal which forms the calibration dots.

Amplitude Measurement

Fig. 4 shows the amplitude measurement and calibration system.

During signal measurement periods, gates *R* and *S* allow the signals from channels *A* and *B* to pass through amplifiers and linear detectors. The detected signals are

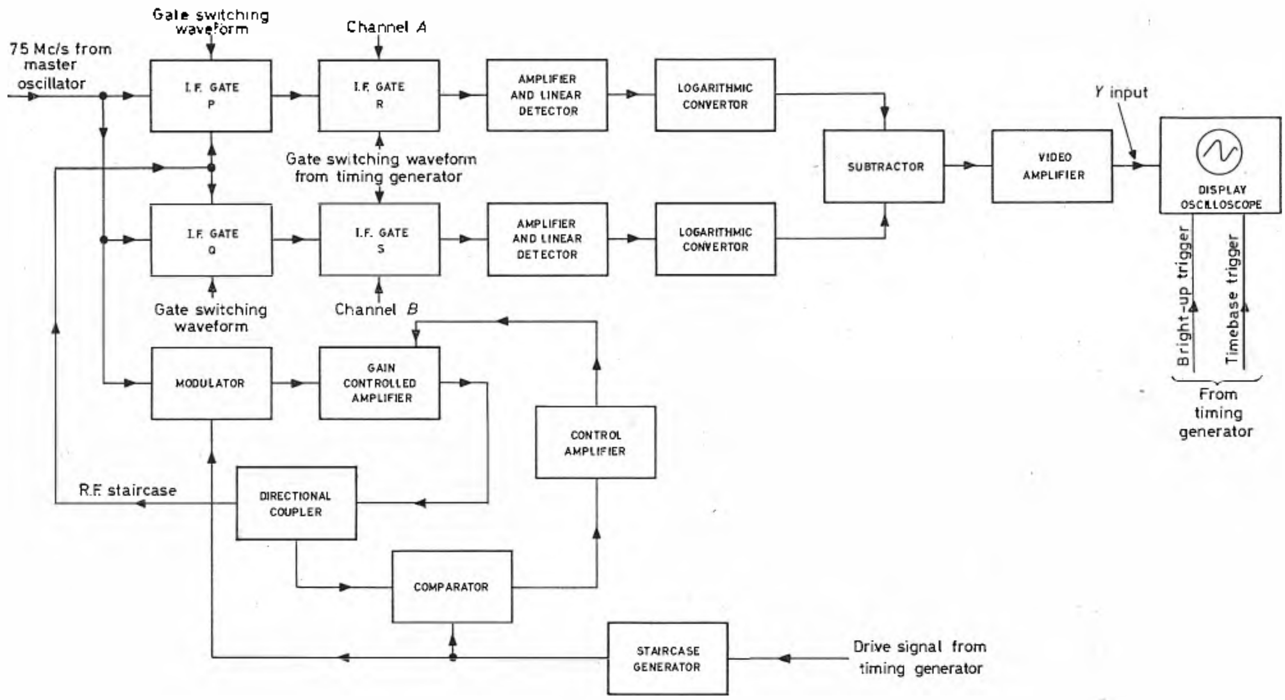


Fig. 4. Amplitude measurement and calibration system

passed through logarithmic convertors, and the difference of the logarithms obtained gives the logarithm of the ratio of the signals. Since this is a logarithm, it may very conveniently be displayed as a measure of the signal ratio directly in decibels, which is done via a video amplifier to improve sensitivity.

Calibration of the amplitude measurement equipment is performed by feeding one measurement channel with a fixed level i.f. reference signal from the master oscillator, while the other channel receives an i.f. signal which is modulated with an accurate anti-logarithmic staircase waveform. At each step of the staircase the amplitude display oscilloscope is triggered, and draws a horizontal line corresponding to the ratio of the fixed reference signal level and the step level concerned. In this way a set of horizontal calibration lines is formed on the display screen. The lines are equally spaced and form a scale of decibels, since the anti-logarithmic staircase calibrates a logarithmic system.

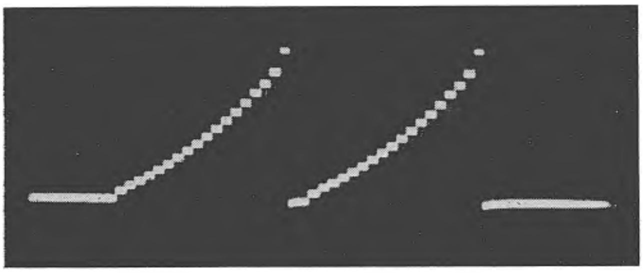


Fig. 5. Photograph of 20 step amplitude calibration staircase

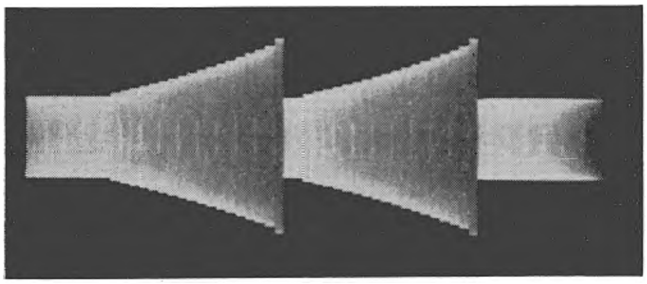


Fig. 6. Photograph of staircase modulated on to 75Mc/s carrier

During a further part of the calibration period, gates *P* and *Q* are switched so that the fixed reference signal goes to the channel which previously received the staircase, and vice versa. This produces a second set of horizontal calibration lines, which should coincide with the first set when the equipment is correctly adjusted.

Faults which may occur in the calibrated part of the equipment will in general be shown up by pairing of the two sets of calibration lines, which therefore also serve to give instant warning of fault conditions.

The staircase waveform is derived from a specially developed staircase generator, whose output is modulated on to a 75Mc/s signal from the master oscillator. A feedback system compares the modulated output with the staircase input, and applies correction by a gain-controlled i.f. amplifier to ensure that the modulated output corresponds very closely to the original staircase levels.

The staircase generator design is very flexible, and any number of steps may be produced per staircase[2]. In the present application up to 32 steps are available, and are used to produce a 15dB range calibrated at intervals of 0·5dB.

For setting up the staircase waveform, the generator can be switched manually from step to step, and a digital voltmeter is used as an indicator while each step is individually adjusted to the required level. Steps can be set to an accuracy of 0·1 per cent by this method.

Fig. 5 shows a photograph of two cycles of a twenty step staircase produced by the generator, and Fig. 6 shows these staircases modulated on to a 75Mc/s carrier wave.

Some Equipment Design Details

Having described the main measurement and calibration principles, it may be of interest to describe briefly some of the individual electronic units which have been developed for this equipment.

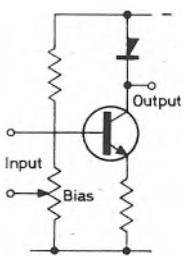
I.F. Amplifiers

Two types of i.f. amplifier are used. The first is a low level amplifier having a bandwidth of 20Mc/s at the −0·1dB points, centred on 75Mc/s, and having a variable gain of 42dB maximum. This type is in common use in radio communication links, and has been adapted for this application.

The second type used is a specially developed high level amplifier, having a bandwidth of 34Mc/s at the −1dB points centred on 75Mc/s and a gain of 32dB. The output level can be up to 10V r.m.s. Printed circuit construction is used to obtain uniform performance from each amplifier, and three variations of the circuit are available. One, is a straightforward amplifier with a 75Ω output impedance at a high level, with provision for gain control if required—this type is used as the i.f. staircase gain controlled amplifier. The other two types have video detectors with either positive or negative going outputs built on the main amplifier board. These detectors when used with a small amount of bias become the linear detectors shown in Fig. 3. They are linear to within 1 per cent over a 20dB signal range.

Logarithmic Convertors

These use the inherent logarithmic forward current-forward voltage characteristic of semiconductor diodes, with transistor constant-current generator drive stages, fed from the linear detectors. The basic convertor circuit is shown in Fig. 7. Since the convertors are highly sensitive to temperature variations, they are enclosed in a Marconi 'change of state' crystal oven, which provides a 250:1 stabilization against ambient temperature changes.



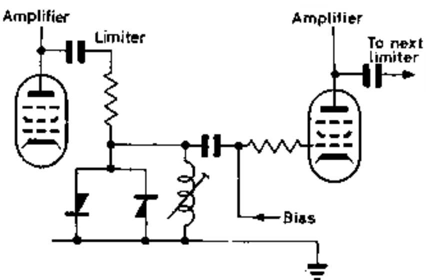


Fig. 7 (left). Basic logarithmic convertor circuit

Fig. 8 (right). Basic limiter circuit

I.F. Signal Limiters

In the phase measurement equipment, the signal limiters are extremely important. They must suppress amplitude variations, which would affect the accuracy of the phase sensitive detector, but phase changes in the input signal must be faithfully transferred to the detector. The basic circuit used is of the well-known crossed diodes type shown in Fig. 8. By using several stages of amplification and limiting, a suppression of 200:1 is obtained for amplitude variations of up to 10dB, while phase changes introduced by the limiters are kept small.

I.F. Gates

These have the circuit shown in Fig. 9 and are essentially single pole double throw electronic switches, using semiconductor diodes as switching elements. In the 'on' position the loss through the gates is approximately 2dB. This figure is maintained constant from gate to gate by adjustments after manufacture to within ±0·1dB, as any variations will affect the calibration and measurement accuracy. The gates are matched to 75Ω impedance in both 'on' and 'off' states, and the isolation in the 'off' state is greater than 60dB.

Main Timing Generator

The timing generator supplies gate switching waveforms, synchronizing and bright-up trigger pulses for the displays, and a signal for locking the slave oscillator to the master oscillator.

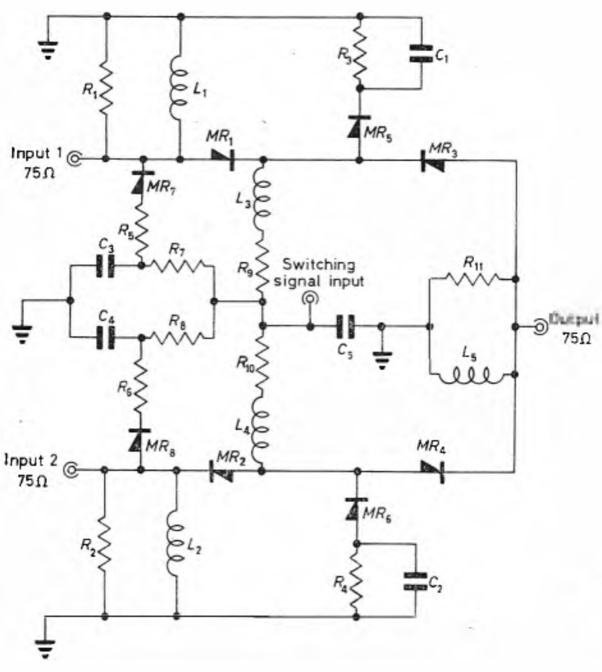
Standard digital techniques are employed for counting down from the master oscillator and pulse forming, the entire equipment being transistorized and of printed circuit construction.

A separate pulse generator supplies pulses to the r.f. head for initiating the r.f. test pulses, and a pre-pulse from this generator switches the equipment from the continuously running calibration condition to the measurement condition before each r.f. pulse is initiated.

Avalanche Pulse Generator

In order to keep the phase and time calibration lines as narrow as possible, very fast bright-up pulses must be applied to the cathodes of the display oscilloscope tubes. By using a transistor in the avalanche mode of operation, with the basic circuit of Fig. 10, pulses of 4nsec duration with a rise time of 1nsec are obtained. The pulse generators are built in to the oscilloscope cases so that very short connexions can be made to the tube cathodes, thus avoiding pulse degradation. Bell[4] gives more general information about avalanche pulse circuits.

Fig. 9. I.F. gate circuit (RD 51619/2)



Conclusion

It has been shown that equipment can be developed to meet the requirements for testgear in modern electronic systems which use high f.m. rates.

Much interesting work remains to be done both in refining the equipment and in applying it to various measuring tasks.

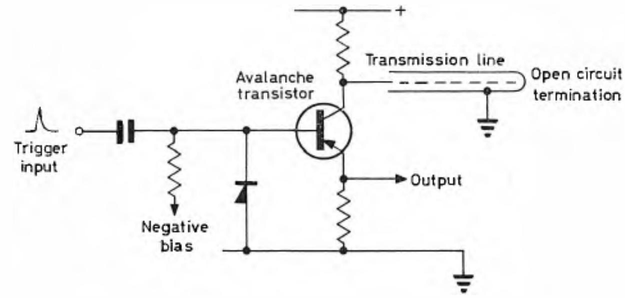


Fig. 10. Basic avalanche pulse generator circuit

The equipment design is quite flexible and will allow pulse widths and repetition rates to be varied to suit the needs of any particular test. The rate and extent of frequency sweep can also be altered as required within a wide range.

One advantage of this equipment over the usual sampling measurement techniques which have been employed for r.f. pulse measurements is the fact that every pulse which the test object receives is measured and displayed. This allows pulse-to-pulse comparison to be done, and variations due to noise, etc., are more readily seen. If desired, a high-speed camera can be used to photograph individual pulses in sequence, and the photographs can later be evaluated with the aid of the superimposed electronic calibration graticules.

Applications of this equipment are expected to include pulse compression radar testing, wideband communication link tests (both telephone and television systems), and in fact any wideband system employing very high data rates where problems with fast f.m. are likely to be encountered.

Acknowledgments

The author wishes to thank his colleagues for their assistance in the preparation of this article, and the Director of Research, The Marconi Co. Ltd, for permission to publish it. Acknowledgement is also made to the Royal Radar Establishment, Ministry of Aviation, for whom this equipment was developed.

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Counting to 1 099 508 482 050 Without Carries

By A. M. Andrew*

The following results are proved for a binary shift register with feedback. (a) For cycle length $(2^n - 1)$, feedback of the modulo-2 sum must be from an even number of elements. (b) The same connexion pattern gives this cycle length whether the feedback is of the modulo-2 sum or its inverse. (c) Cycles of length $(2^n - 2)$ as well as $(2^n - 1)$ are possible. For the case of $n = 20$, feedback connexions are specified for both $(2^n - 1)$ and $(2^n - 2)$ cycle lengths. Two 20-element registers embodying these connexion patterns and fed in parallel with shift pulses have a combined cycle of the length specified in the title.

(Voir page 203 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 210)

THE n elements of a shift register are multistable devices which can be numbered from 1 to n and are so arranged that on the application of a 'shift pulse' the state of the i^{th} element becomes that previously occupied by the $(i-1)^{th}$ element, for $2 \leqslant i \leqslant n$. The information stored in the n^{th} element is not retained by the register. The state of the first element is not defined in the above description, and this element can accept an input signal to determine its new state.

The number of alternative states which each element may occupy is the radix r of the register. In the present discussion only binary registers ($r = 2$) will be considered. The results can certainly be generalized, since the mathematical treatment is based on the work of Elspas[1] who deals with the case of arbitrary radix.

The Shift Register With Feedback

The input signal which determines the state of the first element after application of a shift pulse can be computed from the states of the elements prior to application of the shift pulse. In this case the register may show interesting cyclic behaviour when a succession of shift pulses is applied. Since there are 2^n possible states for a binary register with n elements, the cycle length cannot be greater than this. Very simple feedback arrangements give a cycle of length $(2^n - 1)$, and this is termed a 'maximum-length cycle'.

Where the state of a binary element is represented by 0 or 1, one of the simplest possible forms of feedback is that in which the state of the first element is made to correspond to the modulo-2 sum of the states, prior to the shift, of a subset of the elements. It has been shown[1,2] that maximum-length cycles result from this kind of feedback if the subset of elements is appropriately chosen. It is shown in the Appendix that the number of elements in the subset must be even for a maximum-length cycle.

Applications

The shift register with feedback may be useful wherever a pseudo-random source of digits is required. The fact that two registers can generate identical sequences at remote locations is of obvious value when messages are to be transmitted in such a way that they cannot easily be understood if intercepted.

A register with feedback can also be used as a digital counter if the events to be counted are used to initiate shift pulses. This may allow the construction of physically small counters since there are ways of making very compact shift-registers, e.g. by the use of ferrite cores[3,4]. Another advantage of the feedback shift-register as a counter is its high speed of operation in comparison with a conventional counter using similar components. This is because the shift pulse is applied to all the elements in parallel, with no need for a succession of carry operations following some of the counts.

Compared to a conventional counter the feedback shift-register has the disadvantage that the count is stored in a highly inconvenient form. In some applications this would be of minor importance.

It is interesting to consider the variant of the feedback shift register in which, instead of a string of multistable elements there is a physical medium which can be modified at points along its length. The modification corresponds to setting the state of a multistable element. The shift of the stored pattern along the register corresponds to physical movement of the medium, or, alternatively, to movement of the 'growing point' of the pattern along a stationary medium. The way in which the medium is modified at the 'growing point' is a function of a certain portion of the pattern left behind it. The difference from a feedback shift-register is that the pattern does not disappear at the n^{th} element.

It has been suggested[5,6] that this variant of the feedback shift register may provide a model of the way in which a complex periodic structure in a living organism is determined by a relatively small amount of genetic information. A macromolecule growing in a spiral would be likely to be influenced at its 'growing point' by its existing structure at a number of points some distance behind the 'growing point'.

Feedback of the Inverse Sum

When the feedback signal is the modulo-2 sum, the all-zero state of the register must be a trapping state. Since the maximum-length cycle includes $(2^n - 1)$ of the 2^n states it must always be associated with a trapping state, but since it is natural to think of returning a counter to the all-zero state it may be convenient to arrange that this is not the trapping state. It may be preferable, therefore, to feed back the inverse of the modulo-2 sum. The shift register with direct feedback of the modulo-2 sum will be called a direct feedback shift-register, or d.f.s.r., and one with inverse feedback an i.f.s.r.

When the number of elements from which the feedback is taken is even, it is clear that the all-unity state is a trapping state for the i.f.s.r. This follows whether or not the pattern of connexions gives a maximum-length cycle.

For an even number of feedback elements, the cycle

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structure of an i.f.s.r. and d.f.s.r. with the same connexions must be identical. This follows from the fact that, for any state of the register, the modulo-2 sum of the states of the feedback elements is unchanged when the state of every element is reversed. Any cycle which can be traced by the d.f.s.r. therefore corresponds to a cycle of the i.f.s.r. in which the state of every element is reversed. Table 1 shows a maximum-length cycle of 31 states for a d.f.s.r. with $n = 5$, and Table 2 shows the corresponding cycle for the i.f.s.r. having the same connexions.

Since maximum-length cycles are only produced in a d.f.s.r. by feedback from even numbers of elements, any pattern of connexions which gives a maximum-length cycle in a d.f.s.r. also gives one in an i.f.s.r. This result is derived in a different way in the Appendix.

When the number of feedback elements is odd, the i.f.s.r. may have a cycle of length $(2^n - 2)$, which will be termed a near-maximum-length cycle. In this case the two states not included in the cycle are 010101...... and 101010......, which constitute a cycle of length 2.

Connexion Patterns for Maximum-Length Cycles

From published tables[7] it is possible to deduce connexion patterns to give maximum-length cycles with binary shift-registers having any number of elements up to 19. For $n = 19$, the number of connexion patterns with this property is 27 594.

The properties of feedback shift-registers can be studied in an unsophisticated way by programming a digital computer so that one of its registers behaves as a shift register while cycle lengths are simply counted. The existence of near-maximum cycle lengths was, in fact, discovered in this way. By the use of such a program, one result has been obtained for the case of maximum-length cyclic behaviour, which could not be derived from the tables. For $n = 20$, it was found that the maximum cycle length of 1 048 575 was obtained when the feedback was from the 3rd and 20th elements. Because of a symmetry relationship which holds for all these results, it can be inferred that feedback from the 17th and 20th elements could also be used.

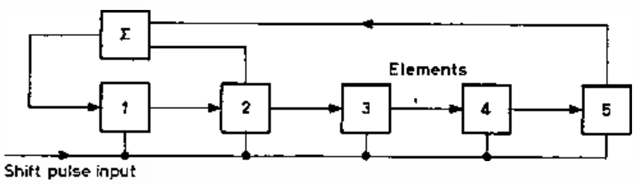

Fig. 1. Shift register having direct feedback from the 2nd and 5th elements

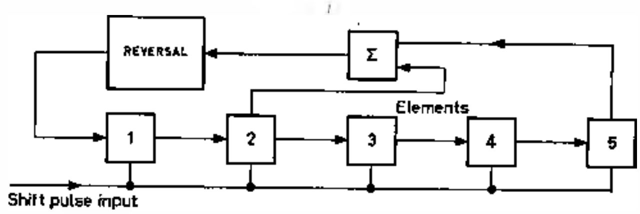

Fig. 2. Shift register having inverse feedback from the 2nd and 5th elements

TABLE 1

Cycle Structure for Shift Register with Feedback Shown in Fig. 1.

	STATES OF ELEMENTS 1 2 3 4 5
CYCLE OF 31 STATES	1 1 1 1 1
	0 1 1 1 1
	0 0 1 1 1
	1 0 0 1 1
	1 1 0 0 1
	0 1 1 0 0
	1 0 1 1 0
	0 1 0 1 1
	0 0 1 0 1
	1 0 0 1 0
	0 1 0 0 1
	0 0 1 0 0
	0 0 0 1 0
	0 0 0 0 1
	1 0 0 0 0
	0 1 0 0 0
	1 0 1 0 0
	0 1 0 1 0
	1 0 1 0 1
	1 1 0 1 0
	1 1 1 0 1
	0 1 1 1 0
	1 0 1 1 1
	1 1 0 1 1
	0 1 1 0 1
	0 0 1 1 0
	0 0 0 1 1
	1 0 0 0 1
	1 1 0 0 0
	1 1 1 0 0
	1 1 1 1 0
TRAPPING STATE	0 0 0 0 0

TABLE 2

Cycle Structure for Shift Register with Feedback Shown in Fig. 2.

	STATES OF ELEMENTS 1 2 3 4 5
CYCLE OF 31 STATES	0 0 0 0 0
	1 0 0 0 0
	1 1 0 0 0
	0 1 1 0 0
	0 0 1 1 0
	1 0 0 1 1
	0 1 0 0 1
	1 0 1 0 0
	1 1 0 1 0
	0 1 1 0 1
	1 0 1 1 0
	1 1 0 1 1
	1 1 1 0 1
	1 1 1 1 0
	0 1 1 1 1
	1 0 1 1 1
	0 1 0 1 1
	1 0 1 0 1
	0 1 0 1 0
	0 0 1 0 1
	0 0 0 1 0
	1 0 0 0 1
	0 1 0 0 0
	0 0 1 0 0
	1 0 0 1 0
	1 1 0 0 1
	1 1 1 0 0
	0 1 1 1 0
	0 0 1 1 1
	0 0 0 1 1
	0 0 0 0 1
TRAPPING STATE	1 1 1 1 1

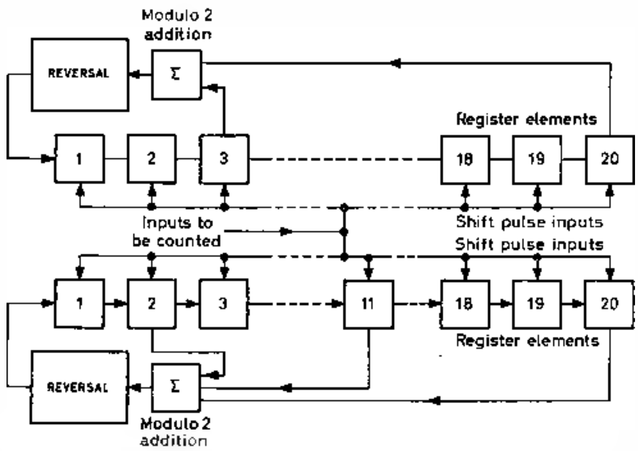

Fig. 3. Combination of shift registers with cycle length approaching 2^{40}

Connexion Patterns for Near-Maximum-Length Cycles

It is shown in the Appendix that, given a connexion pattern giving maximum-length cyclic behaviour in a register with n elements, it is possible to deduce a pattern giving near-maximum cyclic behaviour in a register with $(n + 1)$ elements. Using the above result for $n = 20$, therefore, it is possible to deduce two patterns for a 21-element register giving cycle lengths of $(2^{21} - 2)$. Similarly, from any of the 27 594 patterns for $n = 19$ a pattern can be derived which gives cycle length $(2^{20} - 2)$ for $n = 20$.

A curious property of the i.f.s.r. is that in the trivial cases where $n = 1$ or 2, cycle lengths of 2^n are obtained. For $n = 1$ there is only one element from which to take the feedback, and the states alternate $0, 1, 0, 1 \ldots$ with cycle length 2. When $n = 2$ and the feedback is taken from the second element only, the sequence of states of the register is $00, 10, 11, 01, \ldots$ with cycle length 4.

Operation in Parallel

When the cycle lengths of two feedback shift-registers are relatively prime, operation of the two in parallel gives a combined cycle length equal to the product of the two. The numbers $(2^{20} - 1)$ and $(2^{20} - 2)$ are relatively prime, so parallel operation of registers having these cycle lengths gives a combined cycle length in the region of 2^{40}, the exact figure being given in the title of this article.

Without the discovery of near-maximum cyclic behaviour it would have been possible to obtain a cycle length in the region of 2^{39} by operating in parallel registers having cycle lengths of $(2^{20} - 1)$ and $(2^{19} - 1)$.

One of the possible connexion patterns for the 2^{40} case is shown in Fig. 3. In the case of the upper of the two registers, which gives the maximum-length cycle $(2^{20} - 1)$, the feedback can be of either the modulo-2 sum or its inverse. The inverse is preferred because with it the starting point for the counter can be with all 40 elements set to zero.

Acknowledgments

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Appendix

In refs. 1 and 2 the shift-register with feedback of the modulo-2 sum is treated as follows.

Let the elements $x_1, x_2 \ldots$ of a column matrix represent the states of the elements of a shift register before arrival of the shift pulse. Then the states $y_1, y_2 \ldots$ after application of the pulse can be obtained by multiplying this matrix by a square matrix as follows:

$$\begin{bmatrix} y_1 \\ y_2 \\ y_3 \\ \cdot \\ \cdot \\ y_{n-1} \\ y_n \end{bmatrix} = \begin{bmatrix} c_1 & c_2 & \ldots & c_{n-1} & c_n \\ 1 & 0 & \ldots & 0 & 0 \\ 0 & 1 & \ldots & 0 & 0 \\ \cdot & \cdot & \cdots & \cdot & \cdot \\ \cdot & \cdot & \cdots & \cdot & \cdot \\ 0 & 0 & \ldots & 0 & 0 \\ 0 & 0 & \ldots & 1 & 0 \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \\ x_3 \\ \cdot \\ \cdot \\ x_{n-1} \\ x_n \end{bmatrix} \quad \ldots\ldots\ldots\ldots \text{(1)}$$

The square matrix has a diagonal line of ones just below its principal diagonal. Each first-row entry c_i is equal to unity if the i^{th} element of the shift-register contributes to the feedback, and is zero otherwise. Matrix multiplication is carried out by the usual rows-into-columns rule, with all summations performed modulo-2.

If the square matrix is represented by T, there exists a cycle of length m if, for some X, m is the smallest value of i for which:

$$T^i X = X$$

To transfer from matrix algebra to ordinary algebra, use is made of the Cayley-Hamilton theorem[8] which states: *Every square matrix satisfies its own characteristic equation.*

If the roots of the characteristic equation are such that, for any root r, m is the smallest value of i for which:

$$r^i = 1$$

then the corresponding feedback shift register can traverse a cycle of length m. This is expressed by saying the roots of the equation have period m.

The characteristic equation in λ of the matrix T is obtained by subtracting λ from each element in its principal diagonal and equating the determinant of the resulting matrix to zero. The characteristic equation is:

$$|T - \lambda I| = 0 \quad \ldots\ldots\ldots\ldots\ldots\ldots \text{(2)}$$

where I is the unit matrix.

The characteristic equation for the square matrix in equation (1) can be expanded as:

$$\lambda^n + c_1\lambda^{n-1} + \ldots + c_{n-1}\lambda + c_n = 0 \quad \ldots\ldots\ldots \text{(3)}$$

where each c_i is 0 or 1 and modulo-2 arithmetic is still used.

The condition that a particular feedback shift register has a maximum-length cycle is that the characteristic equation of its matrix be irreducible with roots of periodicity $2^n - 1$.

It is clear from equation (3) that the number of elements from which feedback is taken is one less than the number of non-zero coefficients in the characteristic equation. Since the arithmetic is performed modulo-2, any equation with an even number of non-zero terms has $\lambda = 1$ as a root. Consequently, the only irreducible polynomial with an even number of non-zero terms is $(\lambda - 1)$, or equivalents such as $(\lambda + 1)$. For a maximum-length cycle with $n > 1$, therefore, the equation must have an odd number of non-zero terms, and the number of elements of the register from which feedback is taken must be even.

To treat the case where the feedback is the inverse of the modulo-2 sum, additions can be made to each of the matrices in equation (1), as follows:

$$
\begin{bmatrix} y_1 \\ y_2 \\ y_3 \\ \cdot \\ \cdot \\ y_{n-1} \\ y_n \\ k_y \end{bmatrix}
=
\begin{bmatrix}
c_1 & c_2 & . & . & . & c_{n-1} & c_n & 1 \\
1 & 0 & . & . & . & 0 & 0 & 0 \\
0 & 1 & . & . & . & 0 & 0 & 0 \\
. & . & . & . & . & . & . & . \\
. & . & . & . & . & . & . & . \\
0 & 0 & . & . & . & 0 & 0 & 0 \\
0 & 0 & . & . & . & 1 & 0 & 0 \\
0 & 0 & . & . & . & 0 & 0 & 1
\end{bmatrix}
\begin{bmatrix} x_1 \\ x_2 \\ x_3 \\ \cdot \\ \cdot \\ x_{n-1} \\ x_n \\ k_x \end{bmatrix}
\quad \ldots\ldots\ldots\ldots (4)
$$

It can be seen that $k_y = k_x$, i.e. multiplication by the square matrix leaves the lowest element in the column matrix unchanged. If this element has the initial value zero, the transformation of the upper n elements is exactly the same as is represented by equation (1), and corresponds to the application of the shift pulse in a d.f.s.r. If the lowest element has initial value unity, multiplication by the square matrix in equation (4) corresponds to the application of the shift pulse in an i.f.s.r. The characteristic equation of the square matrix in equation (4) therefore embodies a cycle structure which includes the cycles of both a d.f.s.r. and an i.f.s.r. having feedback from the same subset of elements.

The characteristic equation is easily seen to be:

$$(\lambda - 1) \, . \, P(\lambda) = 0 \quad \ldots\ldots\ldots\ldots\ldots (5)$$

where:

$$P(\lambda) = 0 \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (6)$$

is the characteristic equation of the square matrix shorn of its last row and last column, making it of the same form as the square matrix in equation (1).

Two special cases are of interest:

(a) where $P(\lambda)$ is irreducible with roots of period $(2^n - 1)$, and

(b) where $P(\lambda)$ equals $(\lambda - 1) \, . \, Q(\lambda)$ and $Q(\lambda)$ is irreducible of degree $(n - 1)$, with roots of period $(2^{n-1} - 1)$.

In case (a), a rule given by Elspas for non-repeated factors can be applied. For each of the irreducible factors the cycle structure is expressed in the form:

$$(1 + \textstyle\sum \mu_i(k_i)) \quad \ldots\ldots\ldots\ldots\ldots\ldots (7)$$

where μ_i is the number of cycles of length k_i. In the present case the cycle structure for the $P(\lambda)$ factor is expressed as:

$$(1 + 1 \, . \, (k)) \text{ where } k = 2^n - 1 \quad \ldots\ldots\ldots (8)$$

and for the $(\lambda - 1)$ factor as:

$$(1 + 1 \, . \, (1)) \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (9)$$

Expressions (8) and (9) are 'multiplied' to give the combined cycle structure using the following rule for the multiplication of two terms with respective suffixes i and j:

$$\mu = \mu_i \, \mu_j \text{ g.c.d. } (k_i, k_j)$$
$$k = \text{l.c.m. } (k_i, k_j) \quad \ldots\ldots\ldots\ldots\ldots (10)$$

g.c.d. and l.c.m. being greatest common divisor and lowest common multiple, respectively.

Applying equation (10) to expressions (8) and (9) gives:

$$(1 + 1 \, . \, (k)) \, (1 + 1 \, . \, (1)) = 2 \, . \, (1) + 2 \, . \, (k)$$

showing that the cycle structure consists of two trapping states (cycles of length unity) and two cycles of length $2^n - 1$.

Since one of the trapping states and one of the cycles of length $2^n - 1$ belong to the d.f.s.r., the others must belong to the i.f.s.r. That is to say, the same patterns of connexions give maximum-length cycles with an i.f.s.r. as with a d.f.s.r. This has already been shown in a different way in the text.

In case (b) the equation for the d.f.s.r. is:

$$(\lambda - 1) \, . \, Q(\lambda) = 0 \quad \ldots\ldots\ldots\ldots\ldots (11)$$

and for the complete cycle structure of d.f.s.r. and i.f.s.r. it is:

$$(\lambda - 1)^2 \, . \, Q(\lambda) = 0 \quad \ldots\ldots\ldots\ldots\ldots (12)$$

Equation (11) can be treated by the procedure used in (a) to show that the d.f.s.r. has two trapping states and two cycles of length $(2^{n-1} - 1)$.

When there are repeated factors, as in equation (12), it is not permissible to use equation (10) to combine cycle structures of all the factors taken separately. It is, however, permissible to use it to combine the cycle structure of:

$$(\lambda - 1)^2 = 0 \quad \ldots\ldots\ldots\ldots\ldots\ldots (13)$$

$$\text{and} \quad Q(\lambda) = 0 \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (14)$$

Equation (13) may be rewritten:

$$\lambda^2 + 1 = 0 \quad \ldots\ldots\ldots\ldots\ldots\ldots (15)$$

showing that it corresponds to a shift register with two elements, with feedback taken from the second one only. The cycle structure for this contains two trapping states and a cycle of length 2, and may be represented as:

$$(2 \, . \, (1) + 1 \, . \, (2)) \quad \ldots\ldots\ldots\ldots\ldots (16)$$

The cycle structure of (14) may be represented as:

$$(1 + 1 \, . \, (k')) \quad \ldots\ldots\ldots\ldots\ldots\ldots (17)$$

where $k' = 2^{n-1} - 1$.

Combining (16) and (17) gives:

$$(2 \, . \, (1) + 1 \, . \, (2) + 2 \, . \, (k') + 1 \, . \, (2k'))$$

or 2 trapping states, two cycles of length $(2^{n-1} - 1)$, a cycle of length 2 and a cycle of length $2(2^{n-1} - 1)$, or $2^n - 2$. The trapping states and cycles of length $(2^{n-1} - 1)$ belong to the d.f.s.r., showing that the i.f.s.r. has a cycle of length 2 and one of length $2^n - 2$.

It is now possible to state the rule which allows the design of feedback shift registers with cycle length $2^n - 2$. If $Q(\lambda)$ is an irreducible polynomial of degree $(n - 1)$ with roots of period $(2^{n-1} - 1)$, then: $(\lambda + 1) \, . \, Q(\lambda) = 0$ is the characteristic equation of a d.f.s.r. which has the same connexion pattern as an i.f.s.r. with cycle length $2^n - 2$. (In modulo-2 algebra, $(\lambda + 1)$ and $(\lambda - 1)$ are interchangeable.)

From the tables due to Marsh, two irreducible polynomials, having maximum cycle length, have coefficients as follows:

11 000 000 000 111 111 111 ...... (18)
10 000 000 000 000 111 111 ...... (19)

(By a symmetry relationship, the reversals of these

11 111 111 100 000 000 011
11 111 100 000 000 000 001

could equally well be used.)

Multiplying (18) and (19) by $(\lambda + 1)$ gives:

101 000 000 001 000 000 001 ...... (20)
110 000 000 000 001 000 001 ...... (21)

Comparing (20) with (3), it can be seen that in an i.f.s.r. feedback can be taken from the 2nd, 11th and 20th elements to give cycle length $(2^{20} - 2)$. This result is used in the lower register in Fig. 3.

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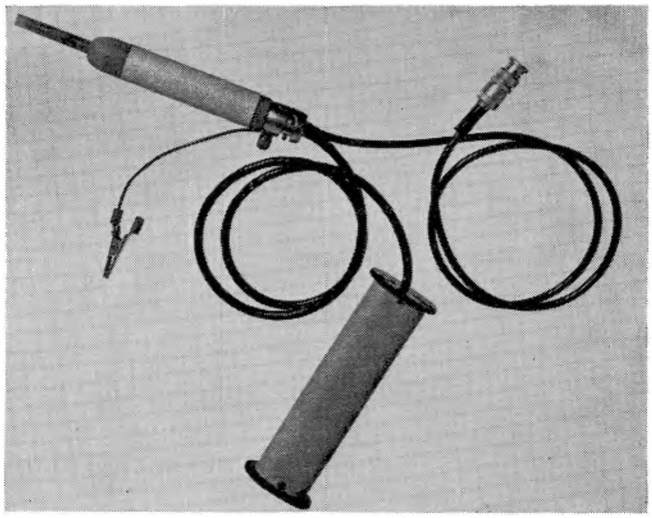
An A.C. Probe with Low Input Capacitance and Unity Gain

By W. H. T. King*



A probe having no attenuation is necessary when low level high frequency signals are to be monitored on the majority of available general purpose oscilloscopes, particularly the smaller types. A simple active probe is described having a gain of 0·98, or better, from 10c/s to 5Mc/s for signals up to 1V r.m.s. and a good response to pulses with fast rise times. The device draws approximately 1·5mA at 18V from batteries only when connected to the circuit under test.

(Voir page 203 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 210)

SEVERAL general purpose oscilloscopes have frequency capabilities of 6Mc/s and greater and are equipped with passive probes. These probes are simple and useful in many applications but for measurements on circuits where signal levels are low, 50mV or less at frequencies above 500kc/s their limitations become apparent. Generally, the input capacitance is about 10pF, but the most serious disadvantage under these signal conditions has been found to be the inherent attenuation factor, usually 10, which cannot be compensated for in the oscilloscope because of bandwidth limitations. To overcome this disadvantage a self-contained active probe was envisaged having a gain of unity with as high a frequency capability as possible, where the advantage of the field-effect transistor could be used to good effect[1,2].

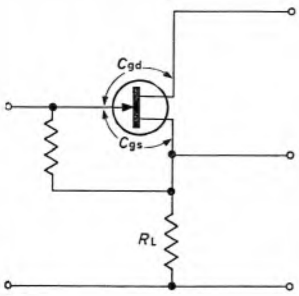


Fig. 1. Simple source-follower

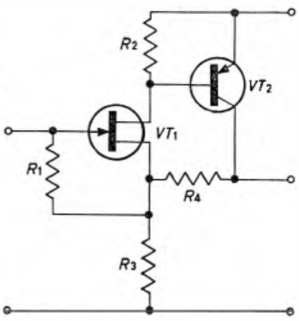


Fig. 2. Source-follower with enhanced gain

Choice of Circuit

It would be desirable to use a single f.e.t. as a source-follower, Fig. 1, where the gain is given by:

$$A_v = \frac{g_m R_L}{1 + g_m R_L} \quad \ldots\ldots\ldots\ldots\ldots\ldots (1)$$

The design aim was to make $A_v = 0{\cdot}98$ or better. In order to achieve this the product $g_m R_L$ must be relatively high at 49, but in the interests of power economy this is not practicable.

A further point of interest is that the input capacitance is reduced by source-follower action:

$$C_{in} = C_{gd} + (1 - A_v) C_{gs} \quad \ldots\ldots\ldots\ldots (2)$$

Consider the circuit of Fig. 2, biasing components are omitted. Provided R_2 is large compared with the input impedance of VT_2 a good approximation to the gain is given by:

$$A_v = \frac{g_m \beta (R_3 + R_4)}{(1 + g_m \beta R_3)} \quad \ldots\ldots\ldots\ldots\ldots (3)$$

if the product $g_m \beta R_3 \gg 1$, then $A_v = (R_3 + R_4)/R_3$. In the application considered the gain is required to be unity and to obtain maximum frequency response it is desirable to make R_4 as small as possible. If R_4 is zero, then:

$$A_v = \frac{g_m \beta R_3}{(1 + g_m \beta R_3)} \quad \ldots\ldots\ldots\ldots\ldots\ldots (4)$$

The circuit behaves as an enhanced source-follower where the effective mutual conductance is now $g_m \beta$ and the criterion for $A_v = 0{\cdot}98$ (i.e. $g_m \beta R_3 = 49$) becomes realizable. It is assumed that the oscilloscope loading is small, and will be capacitive at the higher frequencies.

Design Considerations

The values of R_2 and R_3 are each governed by conflicting requirements. Both are limited by supply voltage restrictions, R_2 must satisfy the conditions of the approximation in equation (3) and R_3 is required to be large enough in attempting to satisfy the criterion for $A_v = 0{\cdot}98$. In practice a resistor R_4 of small value may be included to 'boost' the gain without noticeably influencing the frequency response.

The Ferranti ZFT 14 is a double gate device and has a typical input capacitance of 10pF when gate 2 is connected to source. This connexion gives a lower value of g_m than with both gates connected together. However, the lower capacitance is of greater significance here particularly in view of the fact that this capacitance is dominated by the gate to source capacitance, C_{gs}, which can be effectively reduced, as shown in equation (2).

The choice of device for VT_2 is limited, mainly by the dearth of high frequency pnp devices. Hence the planar V205 is used.

Low capacitance cable is desirable at the output to the oscilloscope so that capacitive loading at the output is as low as possible and the effective value of R_3 is maintained to as high a frequency as possible.

Practical Results

The original circuit Fig. 3, in open layout had a measured input capacitance C_{in} of 1·5pF and the gain was

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unity up to 2Mc/s, falling to 0·707 at 9·5Mc/s from a 75Ω source. Loading of 100pF reduced the 3dB frequency response to 8Mc/s. Maximum input signal amplitude was greater than 3·5V peak-to-peak.

The prototype assembly (shown on page 176) had an input capacitance of 4·6pF with a gain of 0·98 or better to a frequency of 5Mc/s with a 3dB point at about 10Mc/s. At the output a coaxial cable of total capacitance about 20pF was used. The value of input capacitance is the sole limitation to the frequency response at high values of source resistance, and it was considered practicable to reduce this to about 3pF with modified probe layout. The effective shunt input resistance was determined by a comparison method and found to be 125MΩ.

Performance with respect to temperature was satisfactory up to 45°C and is limited by the gate leakage current and the choice of value of the gate bias return resistor R_1.

Fig. 4 gives an indication of the transient response characteristics of the probe, compared with those of a high-quality passive probe. The fall delay time in Fig. 4(a) of approximately 0·1μsec was considered acceptable for the types of oscilloscopes with which the probe is likely to be used.

The probe is supplied by two 9V mercury cells contained

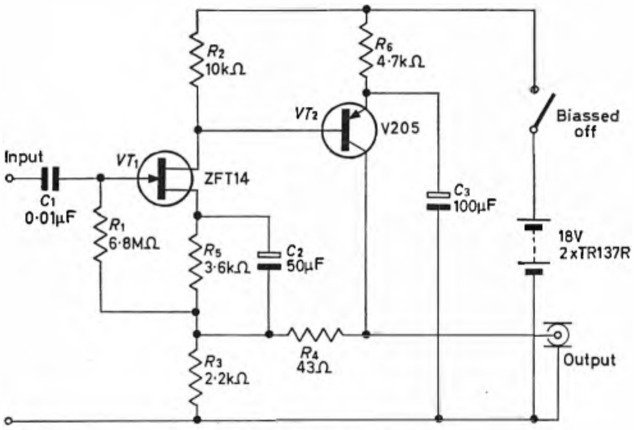

Fig. 3. Complete circuit diagram

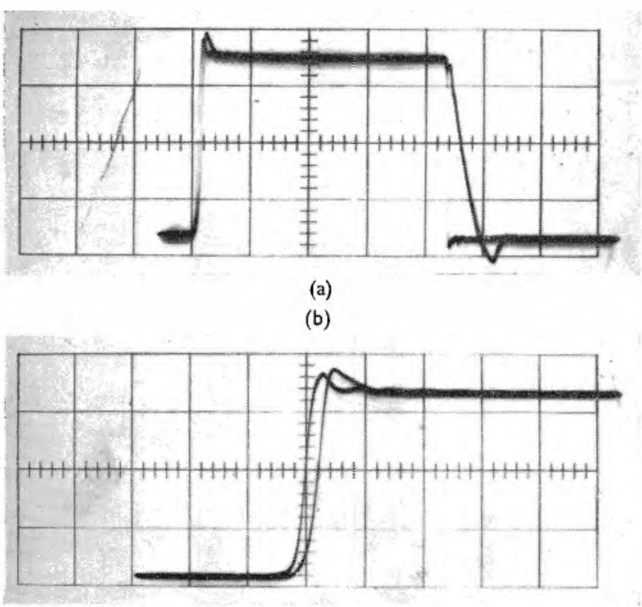

Fig. 4. Transient characteristics compared with high quality passive probe
(a) Scaling: 0·2μsec/cm 1·0V/cm
0·1V/cm (Passive probe)
(b) Scaling: As Fig. 4(a) but time scale 0·05μsec/cm

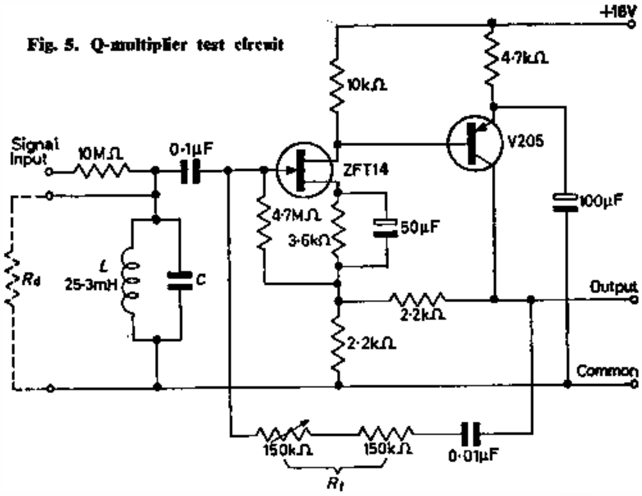

Fig. 5. Q-multiplier test circuit

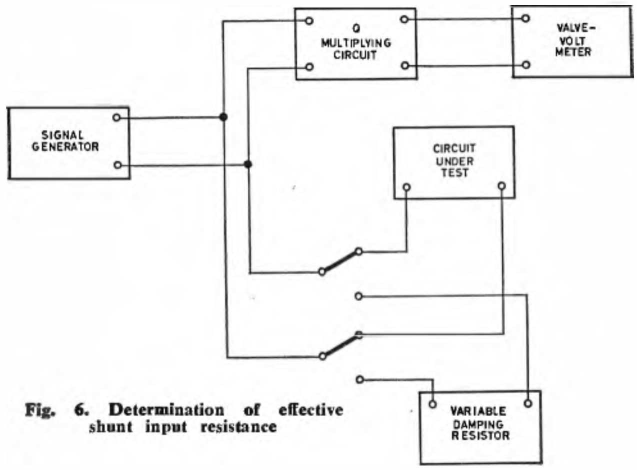

Fig. 6. Determination of effective shunt input resistance

in a separate housing which can be placed conveniently nearby, the supply is switched on only when the probe is connected to the circuit.

Determination of Effective Shunt Input Resistance

The method used involved a Q-multiplying circuit using the basic circuit of the probe as shown. (Fig. 5). Stable Q values of up to 5 000 were obtained by adjustment of the feedback resistor, R_f. The peak output at the tuned frequency was determined for varying values of damping resistance, R_d. From the calibration obtained the effective shunt resistance of the probe was determined. The method is shown in Fig. 6.

Conclusions

The probe overcomes the limitations of passive types giving a useful extended range to the oscilloscopes available at low levels and high frequencies. The use of an enhancement type of m.o.s. transistor will undoubtedly result in a simpler and better probe when these devices eventually become cheaper, more reliable, and more readily available.

Acknowledgments

Thanks are due to the Dean of the College, Sir Donald Bailey, Kt. O.B.E., for permission to publish this article and to Mr. T. Stacey of the Electronics Section of the Departmental workshop for construction of the probe.

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The Design of Thyristor Controlled Current Stabilized D.C. Power Supplies

By A. Davies*, A.M.I.E.E.

The design is discussed of current stabilizers using a bank of transistors for fine control of current together with thyristor circuits for coarse control. An expression is derived for the effective source resistance of such stabilizers which allows the current stability to be calculated for given supply voltage and load resistance changes.

The circuit of a typical stabilizer is described, together with a standard plug-in control circuit which can be used with almost any single phase stabilizer.

(Voir page 203 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 210)

DURING recent years a number of current stabilized d.c. power supplies have been designed to supply powers of up to 3kW at 40A to inductive and non-inductive loads, requiring current stabilities of up to about 0·01 per cent.

In these stabilizers the basic method of stabilization is to connect a bank of parallel transistors between the load and the unstabilized d.c. supply. Feedback from the load current is fed to the transistor bank such that, when the load resistance or supply voltage varies, the voltage dropped across the transistor bank changes to hold the load current constant to within the required limits. In order to reduce the number of parallel transistors required, and to keep the power dissipation of the transistor bank within its rating, the maximum voltage drop across the transistor is limited to a few (typically 5) volts by a secondary coarse stabilizer.

The coarse stabilizer may take a number of forms. The simplest is a manual control where a Variac is operated to keep the voltage drop across the transistor bank within given limits. A development of this which has been used, is a motorized Variac driven by a reversible motor which is switched on by relays when the voltage drop goes outside the limits. In the more recent designs, however, thyristor circuits have been used as a means of coarse control. These have the advantage of having no moving parts and are comparable in cost with Variac controlled circuits. Moreover, because their output voltage is continuously variable, they make a marked improvement in current stability.

* *Atomic Energy Research Establishment.*

The following treatment deals principally with thyristor circuits.

Thyristor Controlled Stabilizers

Fig. 1 shows the block diagram of a typical current stabilized power supply. The mains supply voltage is rectified in the thyristor bridge. The a.c. components in the output voltage are attenuated in the smoothing circuit and the direct current flows via the load R_L, the transistor bank VT_A and the measuring resistor R_M. The voltage across R_M is compared with a voltage developed from a highly stable primary reference and the difference voltage is amplified and fed to the transistor bank, changing the effective resistance of the bank to make the voltage drop across R_M equal to the reference voltage.

Coarse control is achieved by comparing the voltage drop across VT_A and R_M with a secondary reference voltage and using the difference voltage to control the firing angle of the thyristors, and hence the d.c. output, such that the voltage across VT_A and R_M is approximately equal to the secondary reference voltage.

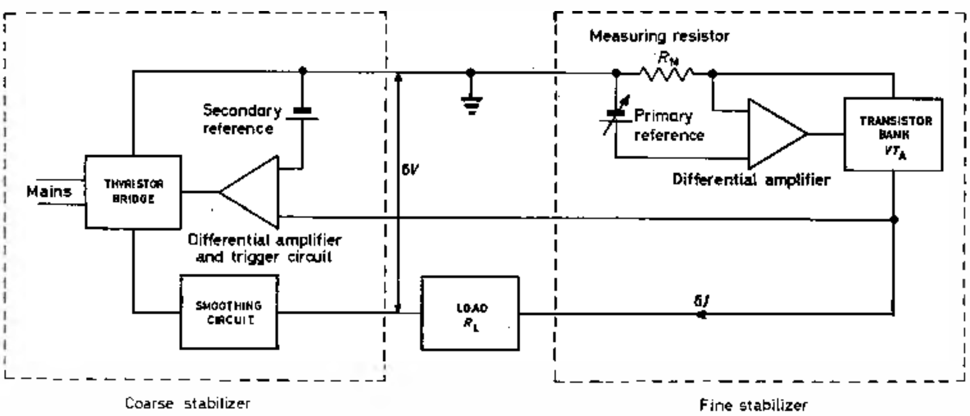

Fig. 1. Typical current stabilized supply

Design Considerations

The object of the design is to make the source resistance of the power supply, seen from the load terminals, look very high compared with changes in load resistance. Then if the load resistance changes, the effect on the current will be swamped by the high incremental resistance of the supply. Similarly this high resistance will reduce the current changes produced by changes in the supply voltage.

In analysing the circuit either supply voltage changes or load resistance changes may be considered. They will lead to the same result since they both show themselves as a change in voltage across the measuring resistor.

For example, consider a load of nominal resistance=5Ω, in which it is desired to hold a current of 20A constant to 1 part in 1 000 (i.e. 20mA) for changes of 20 per cent in either supply voltage or load resistance. Then, with the supply voltage fixed, for a 20 per cent increase in resistance (1Ω) the required source resistance is $1 \times 1\,000 = 1\,000\Omega$ ($R_S \gg R_L$). With a fixed load resistance, for a 20 per cent drop in supply voltage (20V), the required source resistance is $20\text{V}/20\text{mA} = 1\,000\Omega$ again.

To arrive at an expression for the source resistance R_S in terms of the stabilizer parameters, consider the diagram in Fig. 1, where the transistor bank is now considered as a single transistor VT_A with an effective mutual conductance g_m' and an effective collector resistance R_C'. Neglecting for the moment, the coarse stabilizer, the effective source resistance of the fine stabilizer may be calculated by considering the change in current δI caused by a change in voltage δV.

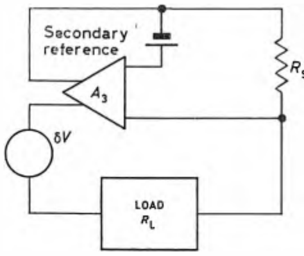

Fig. 2. Simplified arrangement of Fig. 1

i.e. $$R_S' + R_L = \delta V/\delta I = \frac{\delta I(R_M + R_C' + R_L) + \delta I\, R_M A_1 g_m' R_C'}{\delta I}$$

hence

$$R_S' = R_C' + R_M\,(1 + A_1 g_m' R_C')$$

where R_S' = effective resistance of the fine stabilizer
A_1 = voltage gain of the differential amplifier

putting $A_2 = g_m' R_C'$ = voltage gain of VT_A

$$R_S' = R_C' + R_S\,(1 + A_1 A_2)$$
$$\simeq A_1 A_2 R_S$$

Replacing the fine stabilizer with R_S', Fig. 1 can be simplified as shown in Fig. 2, where the thyristor bridge, its firing circuit and differential amplifier are grouped together and shown as an amplifier, effective gain = A_3.

By again considering the effect of a change in supply voltage

$$R_S + R_L = \delta V/\delta I = \frac{\delta I(R_S' + R_L) + \delta I R_S' A_3}{\delta I}$$

hence $R_S = R_S'(1 + A_3)$
$\simeq A_3 R_S'$
$= A_1 A_2 A_3 R_M$

where R_S = effective source resistance of the complete stabilizer.

Because of the slow response of the coarse stabilizer ($\gg 1$ cycle of supply frequency) R_S is only of use in calculating the long term current stability. For short term stability calculations, for example, on the effect of the ripple voltage at the output from the smoothing circuit, the figure for R_S' should be used.

Values of g_m' and R_C'

The transistor bank comprises (Fig. 3) an emitter-follower (VT_B) whose collector current is supplied from the load current, which drives a number of parallel transistors (VT_C). These transistors have load sharing resistors in their emitter circuits to minimize the unequal load sharing due to the spread in the transistor mutual conductance, g_m. The effect of introducing the emitter resistors is to reduce the effective g_m. This is now $\delta I_C/\delta V_1$, and it may be shown that the effective value is:

$$g_m' = \frac{g_m}{1 + g_m R_E}$$

N transistors in parallel may be considered as a single transistor with

$$g_{m(N)}' = N \cdot \frac{g_m}{1 + g_m R_E}$$

and

$$R_{C(N)} = R_C/N$$

The value of R_C is modified by the current flowing through the collector resistance of VT_B and by considering an increment in collector voltage and the corresponding change in load current it may be shown that the effective collector resistance of the transistor bank is

$$R_C' \simeq \frac{(R_{C1}/N) \cdot R_{C2}}{(\beta R_{C1}/N) + R_{C2}}$$

where R_{C1} = collector resistance of VT_C
β = current gain of VT_C
R_{C2} = collector resistance of VT_B

For a typical power transistor R_C may be 240Ω, g_m, 10A/V ± 50 per cent and $\beta = 50$. Considering a specific case where there are three such transistors in parallel with a similar transistor used as an emitter-follower, as shown in Fig. 3, then for the spread in current to be reduced from ±50 per cent to ±10 per cent, the emitter resistors must be about 0·5Ω, and g_m' is then 1·67A/V. R_C' in this circuit is 4·5Ω.

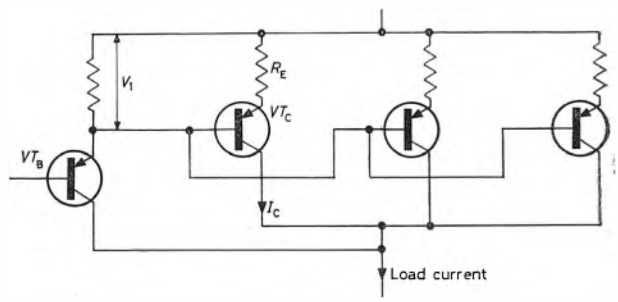

Fig. 3. Transistor bank

Since the source resistance is directly proportional to the product $g_m' \cdot R_C'$ these figures serve to show the considerable degeneration in performance that can occur. To improve matters the emitter-follower VT_B should be chosen to have a high collector resistance, and if only a few parallel transistors are used, it is worth going to some lengths to replace them with a single transistor so that R_E may be eliminated.

Value of A_3

For a single phase thyristor bridge circuit with commutating diode[1] the effective gain of the thyristors and their firing circuit is derived from the relationship

$$V_{dc} = V_{pk}/\pi\,(1 + \cos\alpha)$$

where V_{dc} = d.c. component of output from the thyristor bridge
V_{pk} = peak a.c. voltage applied to the bridge
α = thyristor firing angle.

α is directly proportional to the input voltage (V_{in}) to the firing circuit, hence

$$V_{dc} = V_{pk}/\pi\,(1 + \cos kV_{in})$$

where k is typically 150°/V

$$A_3 = \left|\frac{dV_{dc}}{dV_{in}}\right| = k\,(V_{pk}/\pi)\sin kV_{in} \quad (k \text{ in radians/volt})$$

As a rough approximation V_{dc} may be considered to vary linearly between the maximum output voltage

$2V_{pk}/\pi$ and 0 as the firing angle is varied between 30° and 150°. If V_2 is the change in input voltage required to shift the firing angle by this amount, then

$$A_2 \simeq 2V_{pk}/\pi V_2$$

In practice V_2 is made about 1V and A_2 is generally between 25 and 250.

Voltage Drifts

In addition to changes in load current brought about by changes in supply voltage and load resistance, internal voltage drifts will appear to the differential amplifier as changes in load current, and cause the load current to change. The principal sources of drift are the primary reference voltage, the input stage of the differential amplifier and the measuring resistor R_M.

The reference voltage is derived from a low temperature coefficient Zener diode. If the current in this diode is held constant the voltage stability will be as good as the temperature coefficient, which can be as low as 0·001 per cent/°C.

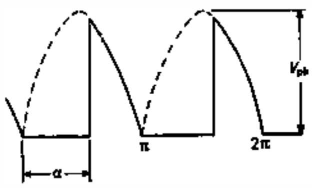

Fig. 4. Thyristor bridge output waveform

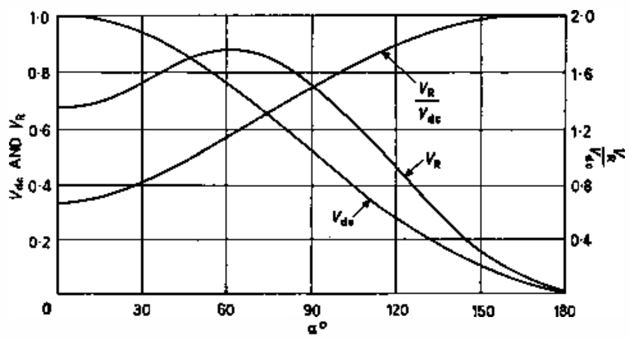

Fig. 5. Variation of V_{dc} and V_R with firing angle

By using a long-tailed pair, the amplifier voltage drift referred to the input can be made about 10μV/°C. The maximum input voltage from the measuring resistor is generally about 1V, so that, at maximum load current, this drift is about 0·001 per cent/°C.

Minalpha wire is generally used for the measuring resistor. Its temperature coefficient has a point of inflection at 26°C, and over a range of 20°C between 17° and 37°C, the change in resistance is 0·005 per cent.

If the Zener diode and amplifier input stage are placed in an oven which stabilizes the temperature to 1°C, it is quite feasible to produce a stabilizer with a long term current stability of better than 0·01 per cent.

Smoothing Circuit

In order to keep the power dissipation in the transistor bank as low as possible it is desirable to be able to predict with confidence the ripple voltage at the output from the smoothing circuit, so that the mean voltage across the transistor bank can be made just high enough to prevent the negative peaks on the ripple voltage from bottoming the transistors.

The design of the smoothing circuits for use with uncontrolled rectifiers is well established[2], and there has recently been published[3] a harmonic analysis of the waveforms occurring at the outputs of single-phase and three-phase thyristor circuits, so that, knowing the harmonic content of the output waveform, the smoothing circuit can be designed according to the practice employed with uncontrolled rectifiers.

In the case considered of a single phase bridge with commutating diode, feeding into an inductive load, the output waveform is as shown in Fig. 4.

Harmonic analysis of this waveform produces the following expression for the d.c. component and second harmonic.

$$F(t) = V_{pk}/\pi\ [(1 + \cos\alpha) - a_2 \cos 2\omega t - b_2 \sin 2\omega t]$$

$$\text{where } a_2 = \tfrac{2}{3} + \cos\alpha - \frac{\cos 3\alpha}{3}$$

$$b_2 = \sin\alpha - \frac{\sin 3\alpha}{3}$$

and hence the peak second harmonic ripple voltage is $V_{pk}/\pi\ \sqrt{(a^2 + b^2)}$.

Knowing the amplitude of the second harmonic ripple, the smoothing circuit can then be designed to give the required ripple attenuation. In practice the smoothing circuit is designed to give 1-2V ripple at its output.

Harmonics higher than the second are neglected since the series for $F(t)$ converges fairly quickly and the output is fed into a smoothing circuit which has an attenuation of 12dB/octave.

Critical Inductance

It is desirable that the thyristor output be fed into a choke input filter since this limits the peak current supplied by the thyristors and diodes in the bridge and gives a smoother control of the d.c. output voltage.

For the circuit to act as a choke input filter, the choke current must never fall to zero at any part of the cycle, i.e. the amplitude of the peak negative ripple current must always be less than the d.c. component of the current. The critical value of inductance (L_{crit}) is that value for which they are equal, i.e.

$$V_{dc}/R_L = V_R/\omega L_{crit}$$

where V_R = peak second harmonic ripple voltage

or

$$L_{crit} = R_L/\omega\ .\ V_R/V_{dc}$$

Fig. 5 shows the variation of d.c. output voltage, second harmonic ripple voltage and V_R/V_{dc} with firing angle. There are two main points of interest

(a) The largest ripple (at $\alpha = 60°$) is only 30 per cent greater than the ripple from uncontrolled rectifiers.

(b) As the firing angle approaches 180° the value of V_R/V_{dc} becomes three times the value when $\alpha = 0°$.

Therefore if the thyristors are to operate over the full range of firing angle, the critical inductance is three times that required with uncontrolled rectifiers, i.e. if the mains frequency is 50c/s $L_{crit} = R_L/100\pi$.

Output Ripple

The inherent ripple in the output current is obtained by dividing the ripple voltage at the output from the smoothing circuit by the resistance (R_S') of the fine stabilizer. Additional ripple is introduced by extraneous ripple voltages coupling with the input to the differential amplifier. To limit these voltages it has been found desirable always to screen the smoothing choke, and to pay careful attention to the wiring of the stabilizers, in particular, to prevent the earth return currents from the power supplies and trigger circuit from interacting with the amplifier earth returns.

6A 24V Current Stabilizer

To illustrate the design, the circuit of a recently designed electromagnet current stabilizer is shown in Fig. 6. This is a relatively low power stabilizer, and the required current stability was 0·5 per cent of maximum current for load resistance changes of 25 per cent and mains changes of ± 10 per cent. In view of the stability required, no attempt has been made to control the temperature of the temperature sensitive components.

In this circuit $A_1 = 50$

$$A_2 = g_m' R_c' = 20$$

$$A_3 = 20$$

$$R_M = 0{\cdot}18\Omega$$

and hence

$$R_S = 3\,600\Omega$$

The value of R_S will give a stability much better than that called for, and in this case the limiting factor is the internal drift.

The control circuit which comprises the voltage references, differential amplifier and thyristor trigger circuits has been built on a printed plug in board (shown inside the dotted rectangle in Fig. 6). This board is capable of controlling almost any single phase stabilizer and is designed for use with a measuring resistor developing 1V at full current. It will supply a drive current of up to 40mA to the transistor bank and has two isolated outputs for triggering thyristors.

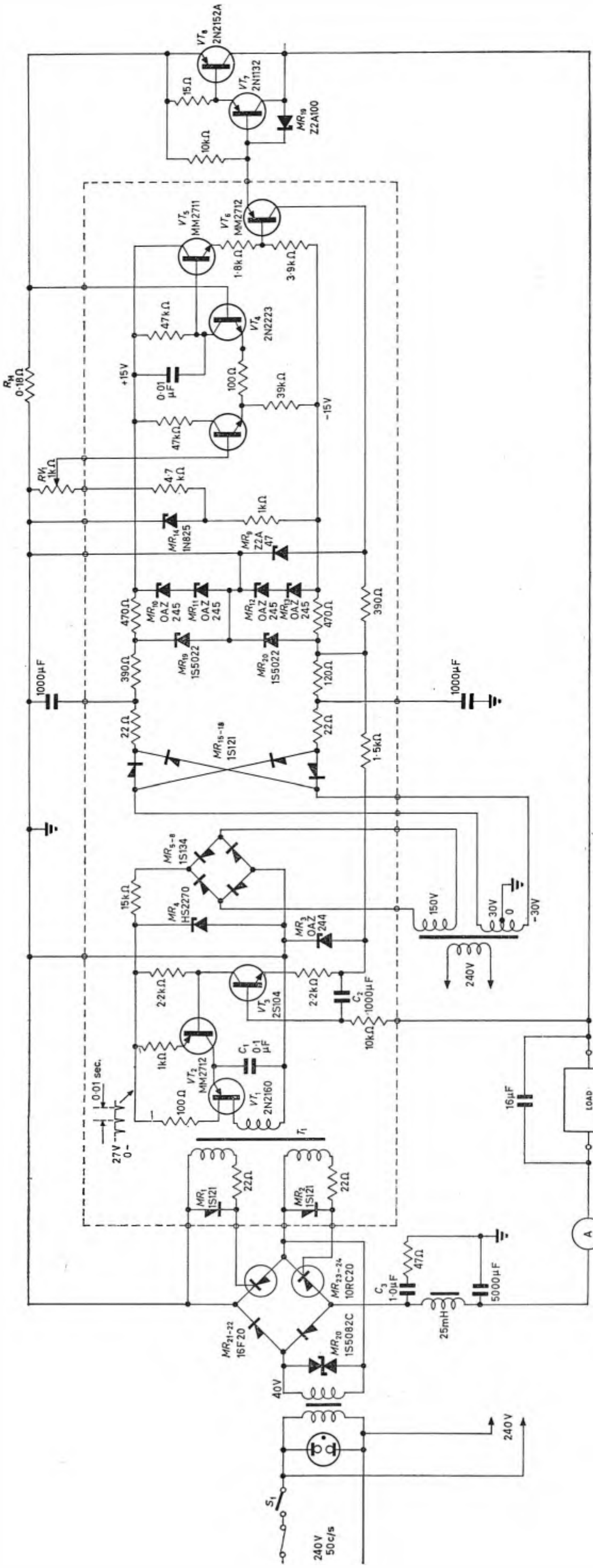

Fig. 6. A 6A, 24V supply

Circuit Description

The primary reference voltage is stabilized in three stages using Zener diode MR_{20} for the first stage of stabilization, low Zener resistance diodes MR_{12} and MR_{13} to provide high stability against mains variations, and the final low temperature coefficient diode MR_{14}. The voltage from MR_{14} is tapped down to give 1V across RV_1 and the voltage on the slider of RV_1 is compared in the dual transistor VT_4 with the voltage developed across R_M by the load current. The 100Ω resistor in the VT_4 emitter circuit is to nullify the offset voltage between the two bases which might otherwise prevent the load current from being reduced to zero.

The output from VT_4 is fed via emitter-followers VT_5, VT_6 and VT_7 to provide the drive to VT_8 which carries the bulk of the load current, only one transistor being used in this case. VT_8 is mounted on a convection cooled heat sink and is rated to dissipate 20W at an ambient temperature of 70°C. The voltage at VT_8 collector is compared in VT_3 with the secondary reference voltage derived from MR_3. The voltage at VT_3 collector is used to drive a conventional[4] unijunction trigger circuit. VT_2 is a constant current source which linearly charges C_1 until its potential reaches the breakdown potential of VT_1. C_1 discharges into T_1, which has three similar windings, and fires whichever thyristor has its anode positive at the time. C_1 then commences to charge again. The voltage on C_1 is reset to zero every half cycle by its discharging through VT_1 when its positive supply voltage falls to zero. By these means the output voltage from the thyristor bridge is controlled to maintain the collector voltage of VT_8 some 1 to 2V positive (dependent on load current) with respect to the voltage on the anode MR_3.

C_2 and the associated 10kΩ resistor are used to attenuate the ripple from VT_8 collector and prevent low frequency oscillation in the coarse stabilizer. The value of C_2 is determined empirically and is proportional to the effective gain of the thyristors and firing circuit. For a gain of up to 50, 1 000μF is adequate.

The 16μF capacitor across the load terminals is used to bypass the transient produced in an inductive load which is caused by the load current pulse due to pick-up at the amplifier input from the thyristor switching transient.

MR_{20} protects the thyristor bridge against voltage transients which may occur when S_1 is opened.

C_3 provides a low impedance circuit for the thyristors to fire into, thereby allowing the thyristor current to rise above the holding current before the trigger pulse decays.

Results

With an initial load resistance of 3Ω the measured stability at 5A was 0·02 per cent for 25 per cent increase in resistance and 0·02 per cent for 10 per cent change in mains voltage.

The long term stability measured over 4 hours was better than ±0·03 per cent.

At maximum load current, the main component of ripple was a short 5mA pulse (≃0·1 per cent of load current) coincident with the firing of the thyristors. The amplitude of the remaining ripple was negligible compared with this.

The load current was continuously variable down to 5mA, this being the leakage current in VT_8.

Conclusions

The measured stability against load variations compares very well with the calculated figure. For variations in supply voltage, in the case of the particular stabilizer considered, the stability is half that expected from considerations of the source resistance alone. This indicates that changes in mains voltage are causing small changes in the control circuit supply voltages which in turn cause the output from the differential amplifier, and hence the load current, to change. This has not been investigated further since the figure given in the results is well within the required stability.

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An Electronic Filming Aid

A compact electronic 'conversion pack' that fits on the side of a standard 35mm film camera to provide the cameraman with a television picture instead of the conventional optical viewfinder, has been developed by a team of British engineers.

Known as Add-a-Vision, the system also enables production executives to watch action as filming progresses on television monitors placed anywhere in the studio.

The equipment has been designed by the Communications Division of the Livingston Group at Watford, to work in conjunction with the Mitchell B.N.C. camera.

Add-a-Vision can help film makers keep pace with the growing worldwide demand for programme material by enabling production techniques to be speeded-up. The director of the picture, studio executives and floor staff can monitor action during rehearsals to decide when a scene is ready to be shot. No film need be exposed until the scene is ready to be taken, and then the monitors relay exactly what the cameraman is shooting to as many remote points as required.

The electronic viewfinder used by the camera crew displays a picture of photographic quality which enables them to keep a continual check on focus and framing much more easily than is possible with the standard optical viewfinder.

If required production executives can have the television output from the camera—or cameras—connected directly to a video tape recorder to provide instant playback. The need for the film studio ritual of waiting for daily 'rushes' to be processed before the director knows if he has a good 'take' in the 'can' is therefore eliminated.

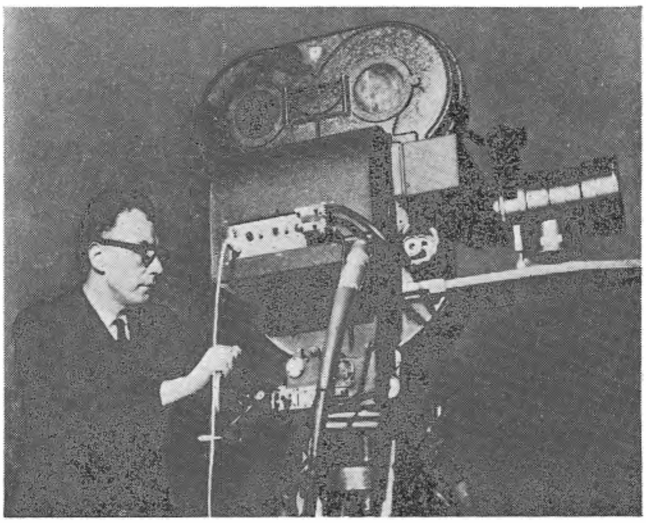
Greatly improved quality of the television image has been made possible by the use of the Philips plumbicon pick-up tube. The camera viewfinder uses a 7in flat faced tube producing a picture of such high quality that it ensures focus accuracy at all times.

Another design feature contributing to the high picture quality produced by the Livingston system is that the optics for both video and film are common.

The lens assembly consists of a specially designed Angenieux 24mm to 240mm zoom lens with two optical outlets—one of which forms an image on the film and the other on the face of the plumbicon tube. Flicker is totally absent since the video image is not taken from a revolving mirror shutter.

No significant increase in set lighting is required when monochrome or colour film stock is used and all the original B.N.C. camera features have been retained including complete freedom of access for film loading and rapid checking of the film gate. The system uses solid-state techniques and once set-up rarely needs adjustment. Plug-in printed circuit boards are used to enable maintenance to be carried out easily.

The Mitchell B.N.C. camera with the Add-a-Vision attachments



SHORT NEWS ITEMS

The Third Colloquium on Microwave Communication, organized by the Hungarian Academy of Sciences and the Scientific Society of Telecommunication, is to be held at Budapest on 19 to 22 April this year, and will deal with the following subjects:

Communication Systems Theory
Network Theory
Electromagnetic Theory
Microwave Theory and Techniques
Microwave Electronics
Systems and Microwave Measurements

Further details are obtainable from the Administrative Secretary, Mrs. A. Valko, Technika Haza, Budapest, V., Szabadsag ter 17. Hungary.

CSF United Kingdom Ltd has received a contract to supply more than three million gold-bonded diodes, over a period of 12 to 15 months, for I.C.T.'s new 1900 Series computers.

CSF United Kingdom Ltd is handling this contract on behalf of COSEM, an associate company of the CSF group.

These diodes will be produced at the automated purpose-built COSEM factory at Saint-Egreve, near Grenoble, France, which is currently turning out 100 000 000 germanium and silicon diodes per year.

The Ministry of Aviation has awarded a contract worth £250 000 to the Marconi Co. Ltd for the provision of a tropospheric scatter radio system in Malaysia. This new system will provide the Royal Air Force with 24 voice channels linking Singapore with Penang. The system will be suitable for use with either telephone or telegraph traffic.

Operating in the 900Mc/s frequency band with 10kW transmitters, this tropospheric scatter link will cover the 370 miles of different country without any repeater stations. The tropospheric scatter terminals themselves, will be supplied by Radio Engineering Laboratories (REL), a Division of the Dynamics Corporation of America. Sixty foot dish aerials will be provided by Marconi for both the Penang and Singapore stations. At Penang, where the terminal is situated on a steep hillside, with the access road passing under the aerials, a huge steel gantry has to be built, to carry the two 27 ton aerials.

The Plessey Company Ltd is installing its type 424 radars for final approach control at the Polish airports of Cracow and Poznan.

The Polish contract provides two transmitter/receivers and two displays for each airfield, and duplicate v.h.f. equipment. Duplication will ensure uninterrupted service and materially assist the Polish Civil Aviation authorities in their policy of improving the safety factor for internal civil air traffic. All-weather operation will be secured by variable polarization.

Installation at Poznan is being completed by Polish engineers under Plessey supervision, to be followed later by that at Cracow. The Polish air traffic controllers will be trained on site by Plessey Radar controllers, who will also carry out commissioning and calibration.

The Plessey Company Ltd has also received an order from the Government of Zambia for a type 42A weather radar and an AR-1 air surveillance radar for installation at Lusaka airport.

The type 42A X band system has been installed by the U.K. Meteorological Office at the London Post Office Tower.

Smiths Industries Ltd is the new title of the parent company previously known as S. Smith & Sons Ltd (England) Ltd.

Tape Recording of Biological Signals is the subject of a one-day symposium being organized by the Scottish Section of the Institution of Electronic and Radio Engineers and the Scottish Electronics and Control Section of the Institution of Electrical Engineers. It will be held in the Department of Medical Physics, The Royal Infirmary, Lauriston Place, Edinburgh 3, on 29 March this year.

Further information together with registration forms may be obtained from the Honorary Secretary of the Joint I.E.R.E.-I.E.E. Scottish Medical Electronics Committee: P. M. Elliott, 21 Craigmount Loan. Corstorphine, Edinburgh 12.

E.M.I. Ltd has put forward a new solution to the problem of testing microcircuits to provide records suitable for quality control, reliability prediction and design evaluation.

Previously, to enable tests to be carried out without disturbing the equipment, infra-red radiometer techniques had been proposed. Although adequate sensitivity was obtainable with these techniques, there was considerable practical difficulty in obtaining an absolute record due to the dynamic range of the emissivity indices of the various components and surfaces and the dynamic range of the temperatures to be considered.

The new solution proposed by EMI consists of converting the measurement to a null indication rather than one of absolute meaning by means of a suitable infra-red variable emissivity test pattern interposed between the equipment under test and the radiometer. Such a filter could conveniently be produced at the same time as the high formulized artwork techniques already current for this type of work, and has the effect of normalizing the radiations such that all points of importance are nominally at the same temperature. The radiometer can now work over a restricted dynamic range, conveniently on a GO—NO GO basis and ambient measurements are obviously much less critical.

The British Radio Valve Manufacturers' Association (BVA) and Electronic Valve and Semi-Conductor Manufacturers' Association (VASCA) state that the total sterling value of their members' sales of valves, tubes and semiconductor devices during the three months ended 30 September 1965 was up by £1M as compared with the same period in 1964.

Sales during the third quarter of 1965 were as follows:

Valves and tubes	£11·0M
Semi-conductor devices	£6·6M
Total	£17·6M

This brings the total value of sales for the first nine months of the year to £53M. The figure for the whole of 1964 was £72M.

The Office for Scientific and Technical Information has made a grant of £47 114 to the University of Cambridge for the support of crystallographic data activities. In addition, it is transferring and extending a grant of £5 420 made originally to Birkbeck College, London. For the next three years the combined activity will be situated at the University Chemical Laboratory.

Briefly the work covered by these grants is aimed at producing the highest quality results from X-ray crystallographic measurements of single crystals and powders. New methods of data reduction using computers are being developed which can be and are used to recalculate earlier measurements, resolving some paradoxes and resulting in more accurate values of molecular dimensions and structures.

This work is part of a co-operative international effort. Data produced in several countries, including Britain, are sent to the National Bureau of Standards in Washington where a book of crystal data is shortly to be published. The printing of the book of crystal data will be computer controlled and from later this year the computer tapes will be held in Cambridge, as well as in the U.S.A., and will be used for data retrieval in Britain. The workers at Cambridge will be responsible for refining and extending the data stored on the tapes.

The work is thus intended to be the first stage in the setting up of an X-ray Crystallographic Data Centre in Britain.

The British Scientific Instrument Research Association (SIRA), in association with the Scientific Instrument Manufacturers' Association (SIMA), is to hold a conference on 'Exploiting Instrument Development' at the Grand Hotel, Eastbourne, on 22 to 23 March this year.

It is the object of the conference to give British instrument manufacturers and users an opportunity of discussing ways in which the successful exploitation of instrument development is achieved by the instrument industry and how the industry's performance in this respect can be improved.

On the first day of the conference an attempt will be made to uncover the various factors which have been found by experience to lead to success or failure in designing, developing and securing the widespread application of new instruments.

The second day will be divided into two sessions entitled 'Technical stimulation of demand' and 'Speeding the pay-off from research'.

Further information may be obtained from the Head of Information Group (Miss A. E. S. Mills), SIRA, South Hill, Chislehurst, Kent.

The two British electronics companies, the Marconi Co. Ltd and Ferranti Ltd, have announced that a licence agreement has been signed in a bid to win a major share of the growing world market for microelectronics. Under the terms of this new agreement, the Marconi Company will manufacture and sell the advanced range of silicon microcircuits designed and sold by Ferranti under the name of Micronor II.

Extensive new production facilities for Micronor II are being commissioned by the Marconi Company at Witham, near Chelmsford, and by Ferranti at their Manchester headquarters. The two companies already manufacture more than 70 per cent of the silicon integrated circuits in the United Kingdom, and this new agreement will provide them with a greatly increased sales potential in world markets.

Micronor II is a new range of ultra fast silicon integrated circuits designed for use in computer and other logic applications. They are available in many different circuits including multiple gates, power stages, J-K flip-flops, etc., all of which have been subject to 12 months' intensive systems proving trials. The full range is already in quantity production in Manchester and it is anticipated that Government qualification approval will be obtained during 1966.

An international conference on the automatic operation and control of broadcasting equipment sponsored jointly by the television groups of the IEE Electronics Division and the Institution of Electronic and Radio Engineers, the Television Society and the Institute of Electrical and Electronics Engineers Region 8, will be held at the Institution of Electrical Engineers, London, from 9 to 11 November.

The conference will cover all aspects of the automatic operation and control of broadcasting equipment including:

Programme distribution systems; picture and sound generation equipment (including automatic lighting control); switching and distribution equipment; communication equipment; primary and relay transmitting equipment; monitoring equipment (reference to domestic receivers will not be included).

Provisional titles and 250-word synopses of contributions should be sent before 31 March to the Conference Secretariat, Institution of Electrical Engineers, Savoy Place, London, W.C.2, from whom further details of the conference will be available.

The colour subcarrier frequency and associated transmission characteristics of the BBC experimental colour television transmissions from Crystal Palace Channel 33 on the PAL system was modified at the end of last month in accordance with a recommendation recently agreed at a meeting of representatives of the European Broadcasting Union.

The characteristics of PAL colour television system used for test transmissions are now as follows:

Line standard: 625-line standard 'I'.
Luminance bandwidth: 5·5Mc/s.
Subcarrier frequency: 4·43361875Mc/s.
Line period: 64·00μsec.
R-Y bandwidth: 1·6Mc/s.
B-Y bandwidth: 1·6Mc/s.
Burst amplitude: 0·3V peak-to-peak.
Burst phase: 45° clockwise from the −(B-Y) axis for +(R-Y): 45° anti-clockwise from the −(B-Y) axis for −(R-Y).
Burst position: as for NTSC (back porch).
Field frequency: 50c/s.
Switched phase line-by-line: R-Y axis.

The electronics division of Finnish Cable Works Ltd, Helsinki, has designed an electronic recording system for saw-mill sorting plants. For each board package the system produces printed information on the running number of package, board grade (up to three grades), kind of wood, board thickness and width, total number of running feet, number of boards of each length (up to 16 lengths) and total number of boards. The system is fully transistorized and contains a ferrite core store for data storage. The output printer starts automatically when a preset number of boards of one grade have passed the sorting plant. Forty-one of these systems will be delivered this year to the Soviet Union, together with sorting plants manufactured by Lahden Rautateollisuus Oy, Lahti.

The Microelectronics Division of the Marconi Co. Ltd announce an advanced new microcircuit which will allow simpler design and more reliable high stability v.h.f. wideband amplifiers. It is a silicon planar integrated v.h.f. cascode circuit and is the first of its type available in this country which will operate at frequencies as high as 200Mc/s.

With this circuit it is now possible to eliminate entirely the neutralization techniques that are necessary with existing v.h.f. wideband amplifier circuits. This is achieved by a unique circuit configuration which provides extremely low internal feedback characteristics.

The circuit performance figures are as follows:

Frequency range: d.c. to 200Mc/s.
Power gain at 100Mc/s: 17dB.
Power gain at 150Mc/s: 14dB.
Noise figure at 100Mc/s: 5dB.
Maximum line voltage: +25V.
Operational temperature range: −55°C to +125°C.

This circuit, type 316-04, is packaged in eight pin, TO-5 header. It is now in production at the Division's factory at Witham, Essex, and is available in production quantities.

'Energy Beams and their Uses' is the title of a conference being organized by the Electronics Group of the Institute of Physics and the Physical Society. It will be held at the University of York from 28 to 30 September next.

The Conference will deal with methods of transferring energy by means of collimated and controllable beams, and the scope will include the industrial and scientific uses of such beams, e.g. in polymerization, sputtering, ion implatation, welding, etching, sensitizing, etc. The main areas of interest will be concerned with the science and technology of electron, photon, and ion beams and some consideration will be given to neutron sources. Modern techniques for the production, collimation, and control of energy beams will be treated. A principal aim of the conference will be to obtain a comparison of competing methods of achieving the same purpose, e.g. between laser and electron beams for welding operations.

Review lectures have been invited to deal with the main fields of interest and additional contributions describing recent work are welcomed. Intending contributors are invited to submit three copies of the title and synopsis of their paper

(200-300 words preferably typed on quarto paper in double line spacing) to the Conference Technical Secretary, Mr. G. Siddall, Electrical Research Association, Leatherhead, Surrey, as soon as possible but not later than 15 June, 1966.

Residential accommodation will be available and advance registration will be necessary. Further details and application forms will be available in May from the Meetings Officer, The Institute of Physics and the Physical Society, 47 Belgrave Square, London, S.W.1.

Estimated figures issued by the Economic and Statistical Division of The British Radio Equipment Manufacturers' Association (BREMA) for the first 11 months of 1965 show that home disposals of radio receivers have been adversely affected by the imports from low-cost countries. At 1 594 000 for the period there is a 401 000 drop over the same months of 1964, and 849 000 less when compared with 1963.

Television receivers delivered to the trade at 1 508 000 continue to show an increase over 1963, but are still lower than the 1 737 000 for the 11 months of 1964.

Deliveries of radiograms were also lower at 203 000 compared with 250 000 for 1964 and 240 000 for 1963.

The Radiation and Nuclear Engineering Division of Vickers Ltd Engineering Group has received an order from the National Research Council in Ottawa, for an electron linear accelerator.

Valued at approximately Canadian $500 000 and won against strong United States competition, the order covers the design, manufacture and commissioning of the accelerator—more usually known as a 'linac', and follows the recent delivery of another large machine to Toronto University.

The linac for Ottawa, although designed for use in a wide variety of research applications is particularly required for X-ray and electron dosimetry, while the Toronto linac will be used primarily for nuclear physics work.

To meet the exacting requirements called for by dosimetry calibration, the linac for Ottawa is designed to operate over a wide energy range, 3MeV to 35MeV, at the same time maintaining good energy homogeneity. The low energy characteristics of the machine are improved by having a separate low energy output beam, which can be used up to 10MeV. Although powered by only one 20MW klystron, the acceleration length is 8m and, at low beam current, the maximum energy is 58MeV. Under short pulse conditions, using the 'field collapse' technique, a maximum beam current of not less than 5A is provided.

A high beam current of 600mA, under steady rate conditions, is provided by the Toronto linac. Powered by two 20MW klystrons, and having a 4m acceleration length, the energy at full power is 35MeV but this can be increased to 45MeV using a low beam current, or under short pulse conditions.

Orders for relays and relay piece parts totalling £500 000 spread over the next four years have recently been placed with Keyswitch Relays Ltd by Middle East countries.

Well over 50 per cent of the orders, which in the first year will total nearly £100 000, are from countries such as Israel with high import tariffs on completed components; in these countries the importers have agreed to buy part-manufactured relays exclusively from Keyswitch, in exchange for manufacturing assistance and technical advice.

Keyswitch, a private company, has for several years maintained an increase in turnover and profits of 50 per cent per annum, a growth rate which is among the highest in the electronics and electrical industries.

The British Standards Institution now comes under the sponsorship of the Ministry of Technology which has assumed responsibility for its Government grant-in-aid.

The Board of Trade will continue to deal direct with the B.S.I. on matters relating to industries for which they are the sponsoring Department, and on matters primarily affecting the consumer. Other Departments also will continue direct relationships on matters within their respective responsibilities.

The first analogue computer to be installed in Sardinia will be a Solartron TY 1451, ordered by the Cagliari University. The order comes two years after Solartron supplied the first analogue computer to be used on another Mediterranean island, Sicily. This was ordered by the University of Palermo.

The TY 1451 will be used in the Institute of Electrical Engineering and Electronics at the University. The Institute head, Prof. Gallerio, will use it for teaching purposes and research into control systems.

The order came after Prof. Gallerio made a special flight to Turin where the TY 1451 was being demonstrated to the Alfa Romeo, Lancia and Fiat motor companies.

It is a flexible 24-amplifier computer with a low cost removeable problem board. Six non-linear modules can all be interchanged or converted to linear use if required. It was one of the first medium sized computers fitted with independent control on its integrators, allowing iterative and hybrid problems to be solved.

Hadley Telephone & Sound Systems Ltd has been acquired by Shipton Automation Ltd and a programme of integration between the Hadley organization and Modern Telephones Ltd (another Shipton Group company) is being carried out.

The combined resources of Hadley and Modern Telephones will be devoted to providing a comprehensive range of communication equipment for industry and commerce. This range includes the latest type of pocket paging system (to be used in the London Stock Exchange) which provides speech communication between pocket receivers and telephones with electronic push-button dialling, with optional loud speech.

James Scott (Electronic Engineering) Ltd has recently been appointed as the U.K. agent for Trak Microwave Corporation of Tampa, Florida.

Trak specializes in microwave oscillators and has built up a comprehensive catalogue of these devices.

The frequency coverage for the oscillators is 400Mc/s to 9·4Gc/s including units for c.w. and pulsed applications. There is also a range of solid state oscillators with outputs between 50Mc/s and 3·4Gc/s, amplifiers in the range 0·96Gc/s to 2·3Gc/s, harmonic generators with outputs between 0·6Gc/s and 16·75Gc/s and also a new range of laboratory receiver heads in L, S, C, or X-band.

English Electric Leo Marconi Computers Ltd announce that an order has been placed by Great Universal Stores Ltd on behalf of its Manchester Mail Order Division for a System 4-50 computer. It will be delivered in the third quarter of 1967.

System 4 is the latest British computer and the first series in the world to use microintegrated electronic circuits, originally developed for military and space use. These give three main advantages—less cost, greater reliability and higher operating speeds—over computers using transistor circuits.

System 4-50 is a multi-programming machine which can run up to 15 jobs at once. This facility will be extensively used to cope with the current daily work load and the increase of work which will take place in the future at the company's Manchester Mail Order Division.

This is the second Manchester order for the new System 4, together worth £1 000 000: on 23 November last the North West Electricity Board stated that it was to order a System 4-70, the largest of the new range.

EMI Electronics Ltd has received an order worth approximately £20 000 to supply television broadcast equipment for the Swiss Television service.

Included in the order are four image orthicon cameras, camera control and remote control units, special lenses and other accessories.

The order is being handled by EMI Electronics' Swiss agents, G. Glatz and Co.

LETTERS TO THE EDITOR

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

A Simple Approximation for Calculating the Phase Shift of Two-Loop CR Networks

DEAR SIR,—Phase advance and phase retard circuits built up of capacitors and resistors are widely used in control and communication systems. The phase

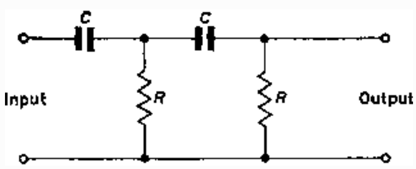


Fig. 1. Two CR sections cascaded

advance angle for a simple single loop *CR* network is just

$$\phi = \tan^{-1}(1/\omega CR) \quad \ldots\ldots\ldots (1)$$

and the magnitude of the output voltage is given by

$$|e_{out}| = \cos(\phi) \,.\, |e_{in}|$$

Evidently the network can only be used for $0 < \phi < 90°$, and for values of ϕ approaching $90° |e_{out}|$ is very small. For $\phi > 90°$, or to achieve shifts of phase of less than 90° with increased $|e_{out}|$, two *CR* sections may be cascaded (Fig. 1). The phase shift (assuming two identical sections) is not twice that of a single section, since the first section is loaded by the second. The exact value of phase shift of the first section is very easily calculated, and is just

$$\phi_1 = \tan^{-1}\left\{\frac{1 + 2\omega^2C^2R^2}{2\omega CR + \omega^3C^3R^3}\right\} \quad \ldots\ldots\ldots\ldots (2)$$

from which the total phase shift can be calculated. However, it is often of more interest to know the angle by which the phase shift of the first section is changed, due to loading effects. Calling this 'error' angle α, and writing ϕ_2 for the phase shift of the second (unloaded) section, which may be obtained from equation (1),

$$\phi_1 = \alpha + \phi_2 = \alpha + \tan^{-1}(1/\omega CR)$$

Hence, from equation (2), α can be calculated as

$$\alpha = \tan^{-1}\left\{\frac{\tan\phi_2\,(1 - \tan^2\phi_2)}{1 + 4\tan^2\phi_2 + \tan^4\phi_2}\right\}$$

by substitution in the formula for the tangent of the sum of two angles. Re-expressing this relationship in terms of trigonometric functions of multiples of ϕ_2, it can be shown that

$$\alpha = \tan^{-1}\left\{\frac{\sin 4\phi_2}{5 - \cos 4\phi_2}\right\} \quad \ldots\ldots\ldots\ldots (3)$$

Hence

$$\tan(\alpha) = (1/5)\sin 4\phi_2 + (1/50)\sin 8\phi_2 - (1/250)\sin 8\phi_2 \,.\, \cos 4\phi_2 + (1/1\,250)\sin 8\phi_2 \,.\, \cos^2 4\phi_2, \text{ etc.} \quad \ldots\ldots\ldots\ldots\ldots (4)$$

The amplitude of the first term of the series is only about 11½ degrees, thus the tangent and radian measure of the angle coincide to within 1 per cent, hence to a close approximation the total phase shift of the two section network is θ where

$$\theta = 2\phi_2 + (1/5)\sin(4\phi_2) \text{ in radians}$$

or

$$= \phi_2 + 11{\cdot}5\sin(4\phi_2) \text{ in degrees} \quad \ldots\ldots\ldots\ldots (5)$$

The error in this expression (as a percentage) is (as shown in Fig. 2) an oscillating function, decreasing in magnitude with increasing θ, and less than 2 per cent for $\theta > 35°$.

This calculation has been carried out for a phase advance network, but it can easily be shown that the expression also holds for a lag network, except that both θ and ϕ_2 are now negative, but the correction term remains positive.

More complex approximations, involving trigonometric functions of $8\phi_2$, are also, obviously, possible and will reduce the error resulting, but will rarely

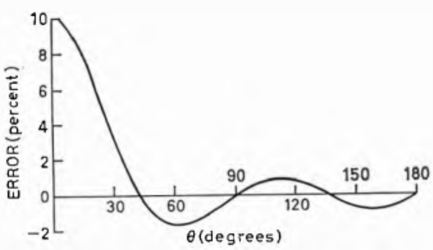


Fig. 2. Error shown as percentage

be justified in view of the errors from component inaccuracies and stray reactances.

Yours faithfully,

W. GOSLING.

University College of Swansea.

A Bridged-T Network for A.C. Magnetic Testing

DEAR SIR.—I have read in the October issue of *Electronic Engineering*, the article by S. N. Bhattacharyya and J. K. Choudhury.

This arrangement of this bridged-T network is a modification of the configuration used by J. K. Choudhury and P. C. Sen [*J. Sci. Instrum.*, *35*, 145 (1958) and *Electronic & Radio Engr.*, *36*, 422 (1959)]. These authors are using a symmetrical *CRC* T-network which is **shunted with a magnetic coil.**

In the present work the authors are using a symmetrical *CLC* T-network which is shunted with a resistor *R*.

I quoted these two configurations in the *Bulletin de la Société Scientifique de Bretagne* some years ago.

(a) *CLC* T-bridged with a resistor[1] *R*

(b) CRC T-bridged with a magnetic coil[2] *L*.

The first arrangement (a) I have given[1] is better because the value of the resistor *R* is lower than in the second arrangement (b). This is also quoted by the authors.

I have pointed out[3] that care must be taken when the magnetic circuit of the coil is open. When a permanent magnet is approached near the core, the bridge being balanced, an unexpected deviation is observed on the detector.

Yours faithfully,

LÉON GRILLET.

Maître de Conférences Honoraire,
Faculté des Sciences,
Rennes, France.

REFERENCES

1 GRILLETT, L., *Bull. Soc. sci. Bretagne*, *33*, 9 (1958).

2 GRILLETT, L., *Bull. Soc. sci. Bretagne*, *35*, 49 (1960).

3 GRILLETT, L., *Bull. Soc. sci. Bretagne*, *35*, 63 (1960).

Distortion of a Saturated Transistor

DEAR SIR,—In the article 'A Constant Volume Amplifier Covering Wide Dynamic Range' by A. G. Morris, published in the August 1965 issue, a Fourier development for the voltage drop of a diode is given. The coefficients, calculated by Professor A. Stone of Rochester University as mentioned by the author, appear to be wrong. In fact rewriting the original expression (8):

$$f(\Theta) = \log(1 + K\sin\Theta\}$$

$$= \log[1+m^2]/m + 2\left[m\sin\Theta + \frac{m^2}{2}\cos 2\Theta - \frac{m^3}{3}\sin 3\Theta - \frac{m^4}{4}\cos 4\Theta + \ldots\right]$$

where $m = (1/K)\,[1-K^2]^{\frac{1}{2}}$

it can easily be seen that when $K \to 0$, all coefficients become ∞ which is not correct.

The function $f(\Theta)$ fulfils the Dirichelet conditions and can be represented by a Fourier series only for $-1 \leqslant K \leqslant +1$ which is also the useful range of the saturated transistor. I believe that probably a misprint might have occurred, because the harmonic coefficients should read slightly differently, as given in equation (1) of this letter and the d.c. term should be changed to $-\log(1+m^2)$.

I can add that using $\cos\Theta$ instead of $\sin\Theta$, all the coefficients are the same, but as the function is even, no sinus terms are required. The correct series, required by the author for calculating the distortion of the saturated transistor is the following:

$$\log(1 + K\cos\Theta) = -\log(1+m^2) + 2\left[m\cos\Theta - \frac{m^2}{2}\cos 2\Theta + \frac{m^3}{3}\cos 3\Theta - \ldots\right]$$

where $m = 1/K\,[1-(1-K^2)^{\frac{1}{2}}]$

$$\text{for } -1 \leqslant K \leqslant +1 \qquad (1)$$

Here all the terms tend to zero as $K \to 0$ and the d.c. term is always negative as it must be.

This Fourier development is important for electronic applications as it gives the voltage distortion of a forward biased junction diode when driven by a superimposed sinusoïdal current.

I hope this letter will contribute to clarify the results of Professor Stone's derivation.

Yours faithfully,

ALBERTO BILOTTI,

University of Buenos Aires, Argentine.

The Author replies:

DEAR SIR,—I am most grateful to Mr. Bilotti for pointing out the errors in equation (8), which were unfortunately introduced during re-typing of the manuscript. It should, of course, be:—

$$\log(1 + K\sin\Theta) = -\log(1+m^2) + 2\left[m\sin\Theta + \frac{m^2}{2}\cos 2\Theta - \frac{m^3}{3}\sin 3\Theta - \frac{m^4}{4}\cos 4\Theta + \ldots\right] \quad \ldots\ldots\ldots \text{(i)}$$

$$\text{where } m = \frac{1}{K}[1-(1-K^2)^{\frac{1}{2}}] \quad \ldots\ldots \text{(ii)}$$

Once these series are known, they can be verified by considering the real part of the series expansion of $\log[1 - jm\exp(j\Theta)]$ in the sine case and $\log[1 + m\exp(j\Theta)]$ in the cosine case. The series is convergent for $1 > K > -1$ and not $1 \geqslant K \geqslant -1$ as suggested by Mr. Bilotti. This can be seen quite clearly by substituting $K = \pm 1$ in the left-hand side of equation (i). Expansion of equation (ii) is interesting as it dictates the behaviour of the distortion curves. Hence:

$$m = \frac{K}{2} + \frac{K^3}{8} + \frac{K^5}{16} + \ldots \qquad \text{(iii)}$$

On differentiating this expression, it can be seen that for small values of K the curves have slopes of 1, 2 and 3 for the second, third and fourth harmonic components, when plotted logarithmically. As the signal approaches 26mV, K approaches unity and the distortion increases very rapidly (not shown in Fig. 3).

Since publication of the article, it has become apparent that the OC201 is not the most suitable transistor for use as a bottomed attenuator, since there is no guaranteed figure for saturation resistance. If the bottomed resistance does not cover a suitable range, the dynamic range of the amplifier could be severely restricted. This parameter is quoted for chopper transistors, although far lower saturation resistances will be found among the transistors designed as high current switches. These transistors have low ohmic resistances and due to their low bottoming voltages at high currents, have inherently high inverse current gains. These facts can be deduced from equations (5) and (6). A further deduction from equation (5) is that the lowest bottomed impedance will be obtained in the normal mode, although the induced collector voltage may cause trouble. Suitable transistors are the BFY50 and 2N2904, both of which have saturation resistances of the order of 2Ω.

Yours faithfully,

A. G. MORRIS,
Datran Ltd,
Hitchin, Hertfordshire.

The Determination of the Impedance Matrix of a Complicated Linear Network

DEAR SIR,—In the November 1965 issue, W. A. Crabbe suggested the Kron method for relatively complicated networks.

On page 753, the impedance matrix Z is correct but the second and third columns of the rearranged matrix must be exchanged.

Thus:

$$\begin{bmatrix} E_1 \\ E_3 \\ E_2 = 0 \\ E_4 = 0 \end{bmatrix} = \begin{bmatrix} Z_{i1} + Z_1 & 0 & Z_{r1} + Z_1 & -Z_1 \\ 0 & Z_{o2} & -Z_{f2} & Z_{o2} \\ Z_{f1} + Z_1 & -Z_{r2} & Z_{o1} + Z_1 + Z_{i2} & -Z_1 - Z_{i2} \\ -Z_1 & Z_{o2} & -Z_1 - Z_{f2} & Z_1 + Z_2 + Z_{o2} \end{bmatrix} \begin{bmatrix} I_1 \\ I_3 \\ I_2 \\ I_4 \end{bmatrix}$$

and C and $Z_{11}, Z_{13}, \ldots$ are not simple.

The Kron method can be used to solve complicated networks but it is advised to take terminal 1 for input and terminal 2 for output when input and output have a common terminal.

Yours faithfully,

P. MONFORT,
Lycée technique de Puteaux, France.

The Author replies:

DEAR SIR,—My thanks to P. Monfort for pointing out an error in the rearranged impedance matrix appearing at the top of the second column, page 753, Nov. 1965 issue. It should read:

$$\begin{bmatrix} E_1 \\ E_3 \\ E_2 = 0 \\ E_4 = 0 \end{bmatrix} = \begin{bmatrix} Z_{i1} + Z_1 & 0 & Z_{r1} + Z_1 & -Z_1 \\ 0 & Z_{o2} & -Z_{f2} & Z_{o2} \\ Z_{f1} + Z_1 & -Z_{r2} & Z_{o1} + Z_1 + Z_{i2} & -Z_1 - Z_{i2} \\ -Z_1 & Z_{o2} & -Z_1 - Z_{f2} & Z_1 + Z_2 + Z_{o2} \end{bmatrix} \begin{bmatrix} I_1 \\ I_3 \\ I_2 \\ I_4 \end{bmatrix}$$

Then it follows that:

$$\begin{bmatrix} E_1 \\ E_3 \end{bmatrix} = \frac{1}{|C|}\begin{bmatrix} Z_{11} & Z_{13} \\ Z_{31} & Z_{33} \end{bmatrix}\begin{bmatrix} I_1 \\ I_3 \end{bmatrix}$$

where $C = \begin{bmatrix} Z_{o1} + Z_1 + Z_{i2} & -Z_1 - Z_{r2} \\ -Z_1 - Z_{f2} & Z_1 + Z_2 + Z_{o2} \end{bmatrix}$

and

$$Z_{11} = (Z_1 + Z_2 + Z_{o2})\{(Z_{i1} + Z_1)(Z_{o1} + Z_1 + Z_{i2}) - (Z_{r1} + Z_1)(Z_{f1} + Z_1)\}$$
$$+ (Z_1 + Z_{r2})\{Z_1(Z_{r1} + Z_1) - (Z_{i1} + Z_1)(Z_1 + Z_{f2})\}$$
$$+ Z_1\{(Z_1 + Z_{f2})(Z_{f1} + Z_1) - Z_1(Z_{o1} + Z_1 + Z_{i2})\}$$
$$Z_{13} = Z_{r1}Z_{r2}(Z_1 + Z_2) + Z_1\{Z_2Z_{r2} - Z_{r1}Z_{o2} + Z_{o2}Z_{i2} + Z_{o1}Z_{o2} - Z_{r2}Z_{f2}\}$$
$$Z_{31} = Z_{f1}Z_{f2}(Z_1 + Z_2) + Z_1Z_{f2}(Z_2 - Z_{r2}) + Z_1Z_{o2}(Z_{o1} + Z_{i2} - Z_{f1})$$
$$Z_{33} = Z_1Z_{o2}(Z_{o1} + Z_{i2} + Z_2) + Z_2Z_{o2}(Z_{o1} + Z_{i2}) - Z_{f2}Z_{r2}(Z_1 + Z_2)$$

Yours faithfully,

W. A. CRABBE,
Royal Military College of Canada.

NEW BOOKS

The Dynamics of Linear and Non-Linear Systems

By P. Naslin. 586 pp. Med. 8vo. Blackie. 1965. Price 105s.

COLONEL NASLIN, who is Chief Engineer in the French Corps of Armaments, has himself translated this book from his native French into excellent English. This by itself is an outstanding achievement. The book is very clear, and there are frequent specific numerical examples in which full detail is given, by means of graphs or tables or both, to enable the reader to learn how to 'worry it out' for himself, even if he starts with no previous knowledge of the particular point or technique under discussion. Where approximate graphical or numerical methods are discussed, comparison with an exact solution, when available, is helpfully included.

Block diagrams and signal-flow graphs are introduced in the first chapter and applied in the second, to transients in linear systems. The reviewer can only record that he personally, being by training a mathematician, finds these confusing and unhelpful, and prefers to work in terms of matrices and the impedance concept, and a calculus which is more completely operational than that used by the author. To many potential users of the book, however, signal flow graphs will doubtless aid understanding, but they do tend to increase the length and price of the book. In Chapter 3, the delay operator E and associated transforms are introduced as a tool of the general theory of linear systems; the discussion of sampled-data systems is postponed to Chapter 7. The intervening chapters deal with the sinusoidal steady state, the relation between transient and frequency response, and stability and damping.

Chapter 8, entitled, "Filtered Non-Linear Systems," in which the harmonics generated by the non-linearities may be neglected, discusses the complex equivalent gain or describing function appropriate to such systems, and methods (notably those of Tsypkin and of Douce) of calculating this. This chapter is applicable to the majority of non-linear systems of order greater than two, whereas second order systems are better analysed in the phase plane (Chapter 9) in which the co-ordinates are displacement and velocity. Chapter 10 deals with the author's "method of numerical simulation" and the "slop-line" method of integration of de Juhasz which are applicable to any system, linear or non-linear. Appendices deal with Graeffe's method of solving algebraic equations by root squaring, harmonic analysis, and a comparison between the operational calculus used by the author and Laplace-transform calculus.

Because of its comprehensive cover and clarity in detail, this book should be easily accessible in the libraries of organizations concerned with control engineering and related fields.

J. W. HEAD

Traffic Control—Theory and Instrumentation

Edited by T. R. Horton. 218 pp. Med. 8vo. Consultants Bureau Enterprises. 1965. Price $12.50

THIS is an interesting and informative book on a subject, of great public interest, which is beginning to take on a more scientific flavour. In the past, traffic engineering has tended to be an art and, what is more, an art which could vary from one borough, municipality or corporation to another. Now, at last, reference books on the subject are becoming available which will help to create a more uniform approach and this book is a notable contribution.

"Traffic Control" is a collection of papers reviewing the state of the art, with a foreword by Henry A. Barnes, the Commissioner for the New York City Department of Traffic. This book is, of course, largely concerned with American practice and, for this reason, may be of more limited value internationally.

The book is in three parts. The first provides a general background; the second describes traffic instrumentation and control technique and the third gives five case studies. On the fly-leaf it says that the book is intended for physicists, psychologists, mathematicians and communications engineers. Unfortunately it has been written by traffic experts who, like everyone else, have their esoteric language. For such a wide class of readers it would have been beneficial had it contained a layman's guide to terminology.

From the British point of view it is extremely encouraging to read remarks concerning the work of the Road Research Laboratory and the Buchanan report. It is a little disappointing to find only two references in the bibliography to British papers. But then, as has been the case for many years, a schism has existed between British and American practice. One looks forward to a British book complementing this one.

In conclusion, a quotation from Evan Herbert's paper gives food for thought.

"If drivers cannot be encouraged or made to conform to a control systems imposed upon them—and, from a systems point of view, there is no agreement on the best control pattern to impose—then it might be better to let the control system adapt to the traffic, or to put automatically controlled vehicles on the highway.

By detecting stationary or moving vehicles by pneumatic tubes, ultrasonics, radar, strain gauges, magnets or magnetometers, traffic can actuate signals. For years, England has employed traffic actuated signals, but individual signals controlled by a highly localized pattern of traffic simply don't work when traffic is heavy, Traffic flow is an area problem."

Does this mean, perhaps, that the Americans are beginning to realize that the natural evolution is from fixed time, programmed control to vehicle actuated traffic lights and from these to wide area control?

B. A. HUNN

Random Processes in Non-Linear Control Systems

By A. A. Pervozanskii. 341 pp. Med. 8vo. Academic Press. 1965. Price 100s.

THE combination of non-linear systems with random excitation would naturally be expected to give rise to some rather stringent theory in order to describe their dynamic behaviour. The reader will not be disappointed in this respect. Combine this with, here and there, a rather unfamiliar turn of phrase and a somewhat humourless text and one is faced with a rather heavy going piece of technical literature. Furthermore, it soon becomes apparent that considerable demands are to be made on the reader's acquaintance with a large body of linear and non-linear control system theory, and statistical concepts. Granted this pre-requisite, which is briefly covered in the fifteen-page Introduction, the rest of the book provides a valuable review of approximate and exact methods of analysis and synthesis in this difficult field of realistic systems with realistic inputs.

The main text is divided into four large chapters. Chapter 1 opens with a discussion of lag-free non-linearities. Examples are analysed for a variety of well-defined non-linearities such as symmetrical limits, relay characteristics, asymmetrical limits and other more complicated arrangements including multivalued functions. Non-linear transformations which introduce lags are dealt with next. An ideal relay followed by a linear transfer function is taken as a basic example to demonstrate a method of calculation in terms of correlation functions and the spectral density at the output of the system. However, more complex situations are discussed and analytical solutions are given for the special cases when the input signal is amenable. Synthesis of optimal systems which transform input signals into some desirable form is the next topic. The emphasis is naturally on non-linear filter synthesis, linear aspects being relegated almost only to a study of references. The first chapter concludes with methods to determine the best statistical description of a non-linear system in terms of a linear scheme, or statistical linearization.

All the preceding discussion is essentially for open-loop systems. Chapter 2 reviews some six possible classes of analytical methods for dealing with closed-loop systems including the application of Markov processes, the interlacing of piecewise linear functions for regions of system response, and various other approximation methods. The application of statistical linearization is strongly featured, and also frequency distortion effects are calculated in some detail. Synthesis of linear compensation networks for systems having pronounced non-linearities also form an important section of this chapter. In passing, readers may care to note an amusing (but quite untypical) misprint on page 124 where it is suggested that the word 'not' should be replaced by 'now' in relation to equation 2.88.

By now the groundwork is laid for dealing with non-stationary situations, and the complexities of combined slowly-varying signals with high frequency random disturbances in non-linear systems is analysed at some length in the third chapter.

Chapter 4 provides the logical culmination to the preceding theory with a theoretical discussion on the behaviour and design of extremal systems composed of combinations of linear and non-linear elements working with noise as a more or less useful component of the signal. Both proportional and discrete systems are considered.

Five appendices provide relevant analytical data such as special describing functions for common non-linearities, an integral transformation theorem, statistical linearization coefficients, and a description of Markov processes.

The inevitably difficult topic of this book still comes across as a rather piecemeal and formidable collection of mathematical devices, although a serious attempt has been made to form a constructive treatment, and considerable inroads into solving many hitherto seeming intractable situations are clearly indicated. However the 'great interpretation' of this topic is still yet to come.

K. C. Garner

Understanding Lasers and Masers

By S. Leinwoll. 88 pp. Demy 8vo. Iliffe Books Ltd. 1965. Price 13s. 6d.

Originally published in America, this book is an introductory guide for the scientist and technologist requiring a general survey of the subject. It opens with a review of the first maser produced by Prof. C. H. Townes and then deals with the solid state ruby laser and the gas and injection lasers.

A chapter is devoted to the potential use of the laser in the field of communications.

Trunk Waveguide Communication

By A. E. Karbowiak. 208 pp. Crown 8vo. Chapman & Hall Ltd. 1965. Price 30s.

This small book serves as an introduction to trunk waveguide communication and gives a full account of the manufacture, installation and testing of waveguide equipment. It contains a list of no less than 730 references to papers dealing with all aspects of the subject.

Electrical Phenomena in Gases

By R. Papoular. 198 pp. Demy 8vo. Iliffe Books Ltd. 1965. Price 45s.

This book, originally published in France in 1963 is intended for those engaged in research in basic plasma physics and gaseous discharges and for students specializing in physics, chemistry and electrical and electronic engineering.

Some knowledge of mathematics is assumed but the main emphasis is on the physical principles of the processes involved. Recent experimental data has been included.

Network Analysis for Telecommunications and Electronics

By R. A. Lampitt. 269 pp. Demy 8vo. Iliffe Books Ltd. 1965. Price 63s.

The first of the seven chapters of this book serves as a revision of the fundamentals of this subject such as the application of the rotating vector and complex functions to circuit analysis. The remaining chapters deal with basic networks, four terminal and tuned networks, wavefilters and transmission lines ending with an introduction to the Fourier analysis of waveforms.

Electromechanical Energy Conversion

By A. J. Ellison. 200 pp. Demy 8vo. George G. Harrap & Co. Ltd. 1965. Price 27s. 6d.

This is an introductory textbook dealing with electrical machines including transformers, contactors and instruments, describing the physical principles of the functioning of these machines in terms of their parameters.

It is based on a series of lectures given by the author at Queen Mary College, London.

The Construction of High-field Electromagnets

By D. De Klerk. 122 pp. Demy 8vo. Newport Instruments Ltd. 1965. Price 25s.

This book is based on four articles published by the author in *Netherlands Tijdschrift voor Natuurkunde* in 1960 and includes iron and coil magnets, superconducting coil magnets and pulsed fields up to 10^6 oersted with pulse times of 10^{-2} to 10^{-6} seconds.

The Mathematical Theory of Linear Systems

By B. M. Brown. 279 pp. Demy 8vo. Chapman and Hall. 1965. Price 25s.

The second edition of this book, originally published in 1961, contains a new chapter on multi-variable systems including an account of the relevant parts of the theory of matrices, a subject which was dealt with very briefly in the first edition.

Principles of Electronics in Medical Research

By D. W. Hill. 288 pp. Demy 8vo. Butterworth & Co. 1965. Price 67s. 6d.

This book is based on the series "Some Fundamentals of Medical Electronics" which appeared in the British Journal of Anaesthesia and on the author's experience in teaching scientists and doctors in the Laboratory Course in Medical Electronics held at the Royal College of Surgeons of England. It provides a general description of the more commonly used terms, components and circuit techniques and gives, in many cases, references for further reading so that workers in a particular specialty can find additional information on techniques which will be of use to them.

Widerstände, Kondensatoren, Spulen und ihre Werkstoffe (Resistors, Capacitors, Inductances and their Materials)

By O. Zinke. 232 pp. Med. 8vo. Springer Verlag. 1965. Price DM.28.50

This book has mainly been written for students of electrical engineering courses. It deals with all passive components and the materials required for their production. Extensive bibliographies are given at the end of each chapter.

Automat und Mensch Kybernetische Tatsachen und Hypothesen (Automatic Machines and Man Cybernetic Facts and Hypotheses)

By K. Steinbuch. 454 pp. Med. 8vo. Springer Verlag. 1965. Price DM.36.-

This is the third enlarged edition within four years of its first publication and shows that the author has hit the right note in the presentation of this difficult subject to a growing circle of readers.

Technische Elektronik 1. Band: Grundlagen und Vakuumtechnik (Technical Electronics Vol. 1: Fundamentals and Vacuum Technology)

By M. Knoll and J. Eichmeier. 390 pp. Royal 8vo. Springer Verlag, Berlin. 1965. Price DM.42.-

Technical electronics is defined by the authors as the principles of construction and operation of discharge devices in the wider sense, i.e. including ion discharges. The first volume deals with the physical and technological basis of modern high vacuum, gas and solid state discharge devices. The presentation chosen is designed to inform the reader as quickly as possible and without extensive reference to mathematics.

Handbook of Electronic Circuits

Edited by R. Feinberg. 195 pp. Crown 4to. Chapman and Hall Ltd. 1966. Price 50s.

About 75 electronic circuits each accompanied by diagrams, their uses, a list of components and a brief description of operation, are contained in this book.

Each circuit, designed and tested by an authority in the particular field, is fitted into a numbering system to accommodate additional related circuits and to allow replacement should the circuit become outmoded by later development.

This book is designed for experimental and laboratory workers who are not familiar with electronic circuits and who wish to make an electronic circuit to perform a particular function.

Correction

It is regretted that in the review of 'Fundamental Analogue Techniques', by R. J. A. Paul, on page 118 of the February issue, the publisher was incorrectly given as Souvenir Press. In fact, this book is published by Messrs. Blackie & Sons Ltd.

NEW EQUIPMENT

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

(Voir page 197 pour la traduction en français; Deutsche Übersetzung Seite 204)

ULTRASONIC EQUIPMENT

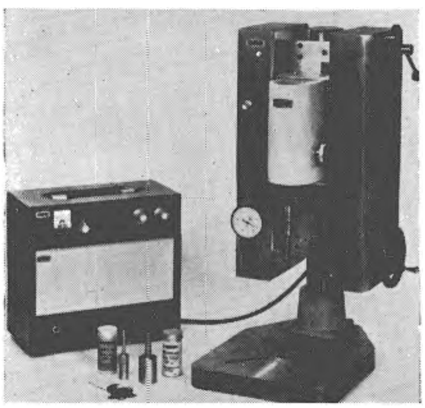
The M.E.L. Equipment Co. Ltd, Manor Royal, Crawley, Sussex

(Illustrated below)

The first items in a new range of ultrasonic equipment have been introduced by the M.E.L. Equipment Co. Ltd.

The basic unit is an amplifier (type L.665) having an output of 100W. The frequency of operation is controlled automatically by feedback from the transducer. The amplifier may be operated continuously and conforms to U.K. and European electrical safety and radio interference standards.

The amplifier can be used to drive



either a 2 litre ultrasonic cleaning bath (type L.673), or a 4 litre bath (type L.674). Both baths are of stainless steel and are fitted with ferrite transducer elements. A changeover switch (type COS.10) is also being manufactured which allows alternative operation of any combination of two of the baths from one amplifier.

A control unit (type L.666) has been introduced which in conjunction with the new amplifier provides a versatile 100W source of ultrasonic power. It incorporates a phase control for use when the units are operated as a feedback amplifier, and a drive control for manual adjustment of oscillator output when they are operated as a generator.

An important innovation is an amplitude monitor which indicates in microns the amplitude of the actual physical movement of the probe.

A new probe transducer (type L.667) is announced which can be employed in a wide variety of applications including drilling, plastics welding, tinning, emulsification and bacteria disruption. It is designed to be driven by the new amplifier (type L.665) and control unit (type L.666). The L.667 comprises a nickel magnetostrictive transducer permanently attached to a half-wavelength aluminium bronze velocity transformer. Polarization is provided by permanent magnets, and two lead zirconate pick-ups are fitted to provide signals for automatic frequency control and the amplitude monitor.

The transducer assembly is partly enclosed in a cylindrical oil-filled cooling jacket which is mounted in a casting. The casting acts as a heat sink and is cooled by a fan.

By mounting the transducer (type L. 667) on a new drill stand (type L.668), an ulltrasonic drill of advanced design is obtained. M.E.L. state that whereas an earlier 60W drill had a maximum cutting tool cross-section of ¾in, the new 100W equipment will cope with tools of at least 1¼in cross-section with increased speed, depth and accuracy.

The complete drilling equipment incorporates the new amplifier and control unit and, consequently, the control and amplitude monitoring facilities which they provide.

The drill stand itself has a number of features that recommend it for industrial applications. The column and base are of high linear accuracy and stability, while the transducer mounting position may be accurately aligned. A rack and pinion drive allows the transducer assembly to be moved smoothly up and down the main column to accommodate the workpiece. The transducer slides on precision-ground roller bearings between the 'out-of-work' and cutting positions, and there is a quick release button and dash-pot for bringing it into the cutting position.

The illustration shows the ultrasonic machining equipment comprising L.665 100W amplifier with L.666 control unit mounted above and L.667 probe transducer mounted on L.668 drill stand.

EE 91 751 *for further details*

TACHOMETER GENERATOR

Muirhead & Co. Ltd, Beckenham, Kent

(Illustrated above right)

Muirhead & Co. Ltd now has in production a size 11 servomotor tachometer generator which is especially designed for use in precision integrating and velocity servos.

By careful design and manufacture the linearity of the output voltage with speed has been kept as low as 0·07 per cent and the maximum variation of the output voltage with temperature is only 0·25 per cent over a temperature range of

−15°C to +75°C. By the adoption of the 'inverted' tachometer design the output voltage is increased to 2·75V per 1 000 rev/min. This is nearly five times higher than that of the earlier types of size 11 tachometers.

Many alternative motor windings are available to suit customers' requirements.

EE 91 752 *for further details*

DETECTOR-FILTER UNIT

Associated Engineering Ltd, 60 Kenilworth Road, Leamington Spa, Warwickshire

(Illustrated below)

A new detector-filter unit—model DF/2—is now available from Associated Engineering Ltd. The unit has been designed for use with matched pairs or double versions of A.E. inductive displacement transducers for the non-contact measurement of displacement and vibration under difficult environmental conditions.

The equipment enables accurate measurement to be made over a wide range of applications which include the movement of bearing journals, oil film thicknesses and bearing wear rates; complex vibration effects in delicate structures where the measuring device must not load the body being measured; and measurements to metrological accuracies on machine tools and machining jigs.

The new DF/2 unit will measure displacement or vibration at frequencies from zero to 10kc/s over a distance of 0·015in with a resolution of up to 10μin depending on the material under measurement.

The temperature stability of the system

including the transducers is better than $\pm 2\mu in/°C$ when the transducers are temperature cycled over the range from ambient to 150°C, and the long-term system stability for static displacement measurement over an eight-hour period is better than $\pm 100\mu in$ for ferrous materials.

EE 91 753 *for further details*

RECTANGULAR POTENTIOMETERS

Reliance Controls Ltd, Sutherland Road, London, E.17

(Illustrated below)

Reliance Controls Ltd now has a new range of rectangular trimmer potentiometers in full production; it is known as the CW.90 series. Good reliability is achieved with diallyphthalate body, ceramic former, stainless steel lead screw and welded end terminations. The units are mechanically interchangeable with existing Reliance and other models. Terminals are on a 0·1in grid, the resistance range is 10Ω to 50kΩ and the temperature range −40°C to +100°C. The standard units are wound with Karma

wire giving temperature coefficients of 20×10^{-6} over the majority of the temperature range.

A fully sealed version of this unit will be available shortly.

EE 91 754 *for further details*

COAXIAL AERIAL CHANGEOVER RELAY

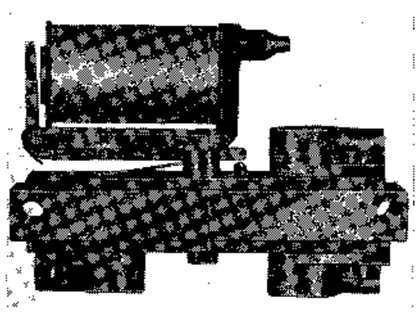
Magnetic Devices Ltd, Exning Road, Newmarket, Suffolk

(Illustrated above right)

A new coaxial aerial changeover relay, designed to eliminate coaxial plugs and sockets in the relay body, thus reducing size and cost, is now available from Magnetic Devices Ltd. The d.c. operated series 951 relay is for aerial switching of frequencies in the order of 450Mc/s, and may also be used for applications requiring exceptionally low inter-contact capacitance.

A small fast operating relay with a single changeover contact set enclosed in a brass housing, it is designed for use with U.R.43 cables which are easily solder connected without disturbing the adjustment of the relay, and firmly secured in position by brass clamps. The clamps each provide continuity of screening and cable outlets in any of three directions.

The series 951 relay has a maximum working voltage of 100V d.c., maximum coil wattage of 1·5W, and maximum coil resistance of 9 350Ω. Tested to withstand 1kV r.m.s. between all terminals



and frame and coil to frame, nominal frequency is 450Mc/s, voltage standing wave ratio approximately 1·1 : 1, and crosstalk 39dB.

Operate time is 20msec, release 8msec, with contact rating 1A or 30W maximum, and nominal impedance 50Ω. The series 951 can be offered to cover the temperature range −40°C to +100°C if specified, has a silver plated finish, and weighs 3½oz.

EE 91 755 *for further details*

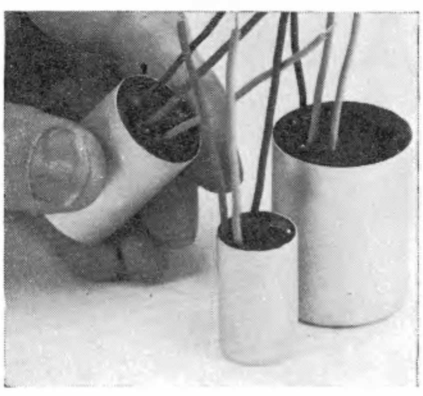
MINIATURE BATTERY CHARGERS

Kynmore Engineering Co. Ltd, 19 Buckingham Street, London, W.C.2

(Illustrated below)

A new range of miniature constant-current charger modules for use with nickel-cadmium rechargeable batteries is available from Kynmore Engineering Co. Ltd. The new charger modules are no larger than the batteries themselves and have so little reverse (leakage) current that they may be left permanently connected to their batteries.

The module consists of a tubular aluminium can containing a capacitor, resistors and a semiconductor-type rectifying arrangement encapsulated in Araldite, from which the three 2½in colour-coded connecting leads protrude. The gold-bonded diodes rectify the supply current and ensure that reverse current is practically zero. In operation, twice the r.m.s. equivalent of the charging current is drawn from the a.c. supply, the major part of the impedance to which occurs across the capacitor. Thus the impedance, and therefore the charging rate, remain substantially constant irrespective of the state of charge of the accumulator. This makes the modules suitable for charging even completely exhausted cells, and no external smoothing is necessary.



Six different sizes are available in the standard range, with capacities ranging from 2 to 45mA assuming a 230V a.c. supply. All are designed for a maximum of ten cells. Corresponding dimensions range from 0·8in long by 0·6in diameter to 1·25in diameter.

EE 91 756 *for further details*

ENVIRONMENTAL TEST CHAMBER

Medical Electronics Ltd, Colford Works, King Henry's Walk, London, N.1

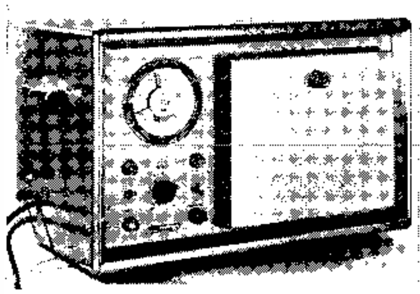
(Illustrated below)

This portable environmental test chamber, type ETC 505, is suitable for temperature testing diodes, transformers and other small components.

Temperatures can be set or automatically cycled between preset limits at preselected rates.

The working chamber is of stainless steel and electrical lead-outs can be provided on the door to permit dynamic testing of equipment.

Heating is by forced air circulation via sheathed heaters. For cooling liquid



CO_2 is dispersed in the forced air, where it is rapidly mixed, eliminating frost build-up.

Full adjustment of heating and cooling rates can be made from a single front panel control. This interval timer can also be adjusted for closer temperature control, the temperatures can be selected on a clearly visible temperature indicator controller which features two independent control points.

The working volume of the chamber is 9in by 7in by 8in. The temperature range is −70°C to +200°C; the cooling rate is 5°C/min and the heating rate 10°C/min.

EE 91 757 *for further details*

DIGITAL COMPUTER

Hawker Siddeley Dynamics Ltd, Whitley Works, Nr. Coventry, Warwickshire

(Illustrated on page 192)

The IAD digital computer DCC.1 is designed to meet the need for an inexpensive, flexible machine which can perform arithmetic and logical operations, and is capable of being incorporated into recording and data handling systems. In this context, extensive use has been made of micro-programming techniques to reduce the amount of hardware and facilitate programming.

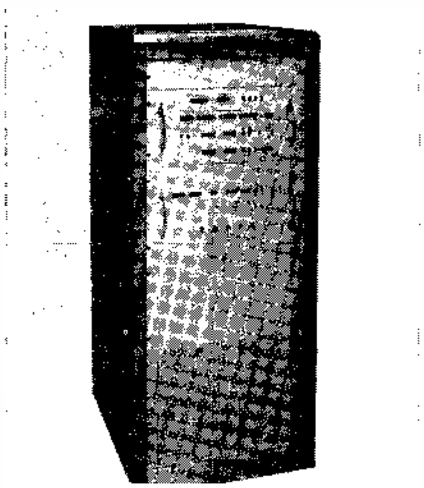
The computer will accept information of Boolean states in binary form, or arithmetic information in binary-coded decimal form, from an eight-hole punched tape; and for on-line applications, information from analogue-to-digital convertors or digital output transducers.

When incorporated in the IAD data logging/recording system, it will be instructed by the programme control unit to perform calculations on data supplied to it by the rest of the system, either in accordance with a predetermined programme or on demand as required.

Information is stored in a ferrite-core matrix of up to 1 024 locations with six bits per location. Each location can hold either a signed decimal digit and Boolean state, or a programme character.

The computer output can feed print-out devices, digital-to-analogue convertors or transducers capable of effecting plant control.

The equipment is mounted on plug-in boards, and circuits are mainly 2Mc/s logic.



The computer comprises three principal sections:

(1) Registers and adder.
(2) Micro-programme.
(3) Digit store.

The registers provide temporary storage of information for processing by the micro-programme.

The micro-programme, which is controlled by a local order counter, directs information and orders between the registers, adder and digit store. It consists of a diode matrix which may readily be changed to give the computer order code most suited to the system in which it is used.

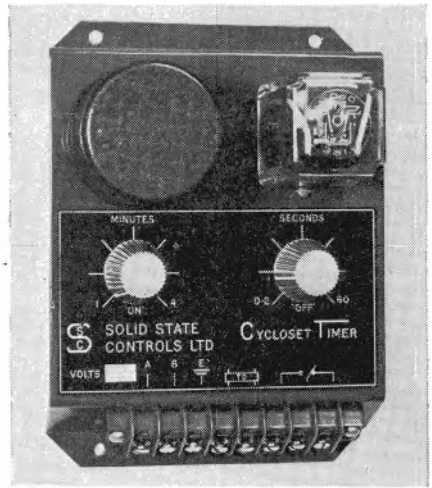
The digit store is a random access, ferrite-core store having 6μsec cycle time and incorporating a nest. Information is stored in the nest on the basis of last in, first out.

EE 91 758 *for further details*

TIMER

Solid State Controls Ltd, 30-40 Dalling Road, London, W.6

(Illustrated above right)



The Cyclo-Set comprises a standard, Cyclo-Set Mini-Module, which is also available as a separate unit, together with a 12V d.c., 8 pin, plug-in relay on a robust, pressed steel chassis. Another version of the timer is available in a totally enclosed die cast box. The Mini-Module has a life of up to 100 million operations, is fully transistorized, with no moving parts and requires no adjustment throughout its life. The sealed relay provides a single pole, changeover output contact with a rating of either 5A or 10A at 250V a.c. as required, and is suitable for both resistive and inductive loads.

The twin stage timing unit has two fully adjustable 'on' and 'off' timing periods; from 0·4sec to 4min as standard, and to 20min to meet special requirements.

This timer is available in six operating voltages from 12V to 415V a.c. and in two d.c. voltages, 12V and 24V d.c. Additionally the unit may be operated both from a normally closed as well as a normally open contact. The Cyclo-Set is also virtually unaffected by fluctuations in operating voltage of ±10 per cent and ambient temperatures within the range −20°C to +55°C.

EE 91 759 *for further details*

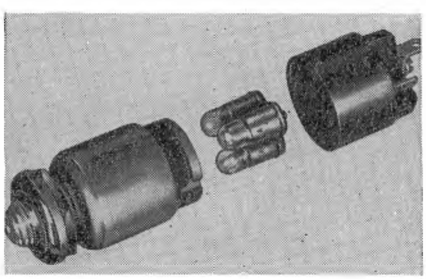
THREE COLOUR INDICATOR LAMPS

Thorn Special Products Ltd, Great Cambridge Road, Enfield, Middlesex

(Illustrated below)

The Thorn three colour indicator is now being produced with new internal filters of glass projecting the selected colours through the single front glass lens.

2W 28V lamps may be used continuously, thereby substantially increasing the light output. In addition to this feature the use of glass for the lenses



allows the colour co-ordinates to be more closely controlled and there is little tendency for colour fading. Standard colours of signal red, signal green and signal yellow to BS. 1376 are now used.

The indicator has an overall length of 2·23in and a diameter of 0·812in. The unit can be fitted in panels up to 0·125in thick and requires a hole fitting size of $^{19}/_{32}$in.

The three colour indicator uses the minimum of panel space and can often replace three separate signal lights. Easily mounted by a locknut and locking ring, the indicator is readily accessible for servicing without removal.

The unit can be supplied complete with three Atlas Midget Panel Lamps with ratings of 6, 12 or 28V as required.

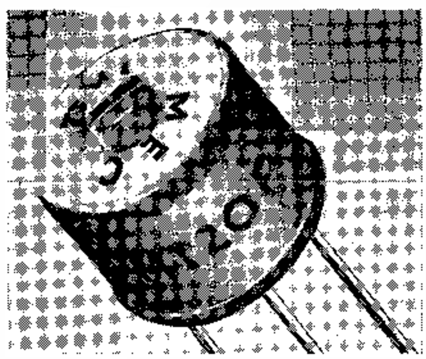
EE 91 760 *for further details*

MINIATURE POTENTIOMETERS

Miniature Electronic Components Ltd. Copse Road, St. Johns, Woking, Surrey

(Illustrated below)

The latest addition to the range of wire-wound trimming potentiometers available from Miniature Electronic



Components Ltd is the MT20. In a TO-5 transistor size case this unit is suitable for printed circuit mounting and is available in standard resistances of 50, 100, 200, 500, 1k, 2k, 5k and 10kΩ. It is specifically designed as a simple, reliable device for use in commercial and industrial electronic applications where high quality at reasonable cost is the criterion. Temperature range −20°C to 125°C; one watt at 70°C rating; single-turn adjustment, positive and stop, humidity and dust proof are some of the salient features.

EE 91 761 *for further details*

BRIDGE OSCILLATOR-DETECTOR

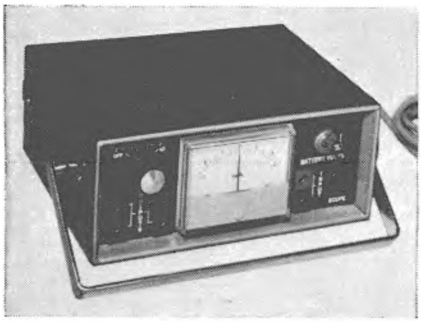
H. Tinsley & Co. Ltd, Werndee Hall, South Norwood, London, S.E.25

(Illustrated on page 193)

The Tinsley type 5720 bridge unit consists of a square wave generator giving two isolated and balanced outputs of 2·5V each. The detector is synchronized with the generator.

It is suitable for strain gauge bridges or resistance thermometer bridges where a combined oscillator and detector is required.

An attenuator controls the sensitivity of the detector whose output is read on a 3in meter. The sensitivity can be con-



trolled to make each division of the meter correspond to an exact amount of strain or temperature according to the way the unit is used.

The unit contains a chargeable battery which may be recharged from the mains.

EE 91 762 *for further details*

MILLIWATT TEST SETS

Standard Telephones & Cables Ltd, Testing Apparatus Division, Corporation Road, Newport, Monmouthshire

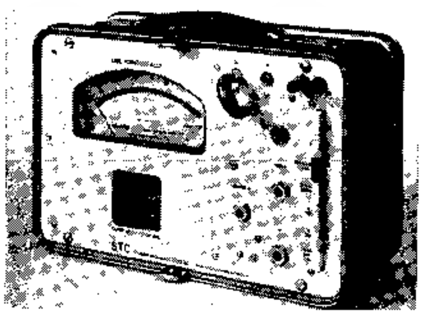
(Illustrated below)

Standard Telephones & Cables Ltd has now introduced redesigned versions of the 74166-A milliwatt test set, providing the same facilities in a restyled metal case occupying only half the volume of the 74166-A. In addition to this greater portability, push-button switches replace the previous rotary switch, and the meter is located behind the front panel to give it more protection and present a neater appearance.

The two models (H and J) are identical apart from their input connectors, the 74166-H being fitted with P.O. No. 1 coaxial plugs, and the 74166-J having BNC connectors (Amphenol 31-221). They will make accurate terminated level measurements of +1dB to −1dB referred to 1mW on 75Ω unbalanced circuits at frequencies up to 30Mc/s, and they will also measure heater voltages of 5·6V to 7·0V.

A wired-in thermocouple is used with heater impedance shunted down to 75Ω, and therefore the instruments cannot be used for measuring 'through levels'. High measuring accuracy is ensured by a built-in miniature Weston standard cell, against which the circuit is standardized by a series of simple switching and adjusting operations. A further feature is that the standardizing circuit can be used to send a d.c. power of 1mW into external 75Ω circuits for calibrating other apparatus.

Both models operate from three 1·5V



dry cells housed in the case, which measures 12¾in by 8in by 7¼in with the lid in position.

EE 91 763 *for further details*

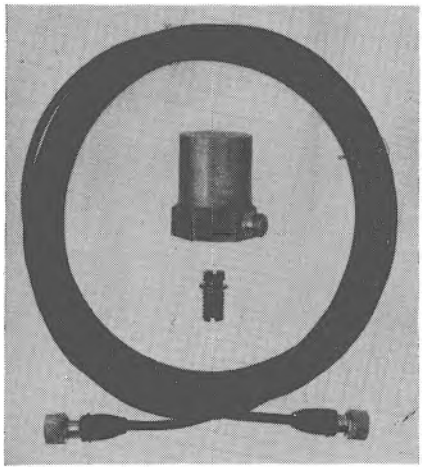
ACCELEROMETERS

Environmental Equipments Ltd, Station Yard, Wokingham, Berkshire

(Illustrated below)

The AQ series accelerometers are constructed from stainless steel throughout the mechanical assembly. Adhesives have been avoided in the crystal assembly to obviate problems at high temperatures resulting from a loss of bond strength and changing stiffness.

The maximum temperature rating is 260°C, and the charge and voltage sensitivity deviation is less than 5 per cent down to −65°C and less than 15 per cent to +160°C. The maximum transverse sensitivity is 5 per cent. The internal capacitance is 1 200pF and that



of the 300cm low noise cable assembly is 300pF.

Four models are available and these are known as the AQ5, AQ10, AQ20 and AQ40. The model numbers indicate the nominal charge sensitivity in pC (peak)/g (peak).

EE 91 764 *for further details*

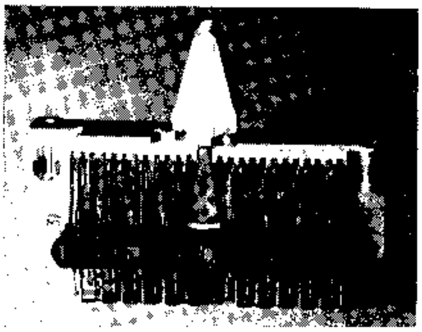
MINIATURE LEVER SWITCHES

Keyswitch Relays Ltd, Cricklewood Lane, London, N.W.2

(Illustrated above right)

Now available from Keyswitch Relays Ltd for light current switching applications is a range of P.O. 1000 type miniature double throw lever keys with a mechanical life in excess of 750 000 operations per switching cycle. The contact action provided is changeover, and is available in combinations of locking, non-locking and stop action, with a maximum of 16 changeovers (eight per action). Silver twin contacts are standard, and the range supplements the eight-changeover lever keys introduced by Keyswitch in 1964.

Contact ratings are 100V at 300mA d.c., 12V at 1A d.c., 100V at 1A a.c., with contact pressures initially 15g minimum (10g minimum after 750 000 operations).



Each key is supplied with two fixing screws, and handles are available in red, green, yellow, grey, ivory or black to blend with various panel facia designs.

Overall length is 2·78in, depth including handle 2·19in, and width 0·62in.

EE 91 765 *for further details*

PORTABLE THICKNESS GAUGES

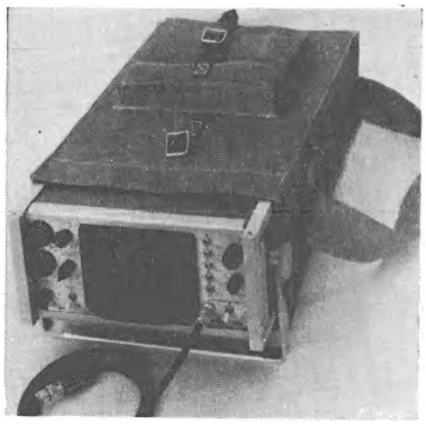
Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustrated below)

Two new versions of the type 1805 Sonoray portable ultrasonic thickness gauging equipment are now available from Dawe Instruments Ltd. Designated the types 1805B and 1805C, they replace the earlier type 1805A, having greatly improved specifications.

The type 1805B, the simpler of the two new models, is similar in appearance, operation and facilities to the old type 1805A with the following improvements: as the instrument is designed primarily for battery operation in the field, power consumption has been reduced to give over five hours of continuous use from one charge. The instrument can, however, also be operated from the mains, and has accuracy and resolution equal to or better than those of most production line gauging units.

Size (15in by 9½in by 4½in), weight (16 lb complete) and general layout are the same as those of the type 1805A, but silicon transistors are used in place of germanium. The result is increased testing depth (to 450in of steel), frequency range 1Mc/s to 8Mc/s with further usable performance up to 15Mc/s) and temperature range (−10°F to +150°F). Pulse repetition rate is 250c/s or 1 000c/s, and pulse echoes are



again recorded on a square-section cathode-ray tube, which has an effective screen area (5cm by 5cm graticule) equivalent to those of much larger circular tubes. As before, the scale can be expanded until the full width represents $\frac{1}{2}$in thickness of steel, and gauging accuracy is again $\pm 0{\cdot}005$in thickness in the range 0·1 to 6in.

The new type 1805C is basically the same as the type 1805B with a number of facilities added to make it suitable for use in automatic flaw detection, thickness measuring, recording and similar applications which have previously been the province of larger units.

The basic additions in the 1805C include a built-in attenuator, variable in 2dB steps of 0dB to 62dB, a built-in flaw alarm and provision for connecting ancillary equipment. Modular construction enables the user to convert his type 1805B into a type 1805C after purchase, and then to select the particular ancillary units required. The present range of ancillaries (not yet completed) includes a digital micrometer, time analogues and universal flaw monitors, all in the form of standard solid-state modules. Any two of these plug into an ancillary housing which couples to the Sonorary itself.

In all cases transducers must be matched to the frequency in use; they are normally of the high-performance piezoelectric Z type. A 6ft transducer coupling cable is normally supplied, but cables up to 200ft in length may be employed without need for modifications. For on-site applications carrying frames, canvas packs and brown leather cases are available, and are interchangeable for both models.

EE 91 766 *for further details*

OSCILLOSCOPE

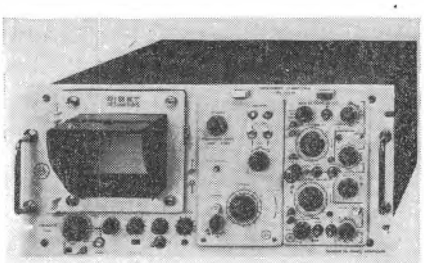
Ribet-Desjardins, 13/17 Rue Perier, Montrouge (Seine), France

(Illustrated below)

This new oscilloscope is completely transistorized and is available in two versions: the type 340A is a portable unit for bench use while the type 340AR is for rack mounting.

Plug-in vertical amplifiers are employed providing either one or two channels. Using the widest bandwidth of d.c. to 100Mc/s the deflexion sensitivity is 50mV/cm with a rise time of 3·5nsec. Using a d.c. to 60Mc/s bandwidth the sensitivity is 5mV/sec.

Either single or dual time-base units are available providing from 5sec/cm to 100µsec/cm with an expansion of ×10. Synchronization is possible up to 200Mc/s.



The horizontal amplifier for use with the dual time-base has a bandwidth of d.c. to 2Mc/s with a maximum sensitivity of 0·1V/cm; with the single time-base the bandwidth is d.c. to 3Mc/s with a sensitivity of 20mV/cm.

EE 91 767 *for further details*

SEMICONDUCTOR TESTER

Rohde & Schwarz, 15 Mühldorfstrasse, 8000 Munich 8, Germany

(Illustrated above)

The Semiconductor tester type BMA is intended for the rapid measurement and checking of the principal parameters of semiconductor components. It can be programmed by type cards or punched tapes and augmented by means of a modular system. The maximum sampling rate is 0·1sec per parameter. The data, which the tester delivers as go/no go decisions or in digital form, can be processed and statistically evaluated by means of the data acquisition and processing equipment of the type series UC. The new semiconductor tester is particularly suitable for production tests, incoming inspection and reliability tests carried out by the customer. The new UC system for data acquisition and processing, which is designed according to the building-block principle and can be adapted to meet various needs, evaluates series of measurements according to mathematical and technical statistics (statistical quality control). It classifies the digital data and stores the absolute frequencies, which can be printed. A screen displays the distribution curve which an *XY* plotter records—either undistorted or distorted for a probability nomograph.

EE 91 768 *for further details*

X-Y READER

P.C.D. Southern Ltd, 4 Blackwater Station Estate, Camberley, Surrey

The new '*X-Y*' reader type ZAE-1 offers facilities for data chart analysis which until now have only been available in far more complex and costly instruments. Among the features of this reader is the accuracy which the semi-automatic *X* axis incremental shift facility, combined with the easily operated sight, allows the tabulating of *X-Y* ordinates, instantaneous slope of the line and area under the line, with a considerable reduction in the degree of the operator skill required. Both linear and non-linear functions may be processed, the cursor sight simply being set on the trace line and the read lever pressed to initiate print-out.

Overall size of the reader is 42in by 24in by 9$\frac{1}{2}$in. Total reading area is 12in in the *Y* axis and 19$\frac{1}{2}$in in the *X* axis, the chart rolls being stored in easily accessible reels operated by a simple manual control, unprocessed data normally being stored on the right-hand reel and 'reeled on' to the left-hand reel in 50cm units of length. Records up to approximately 200ft can be accommodated.

For reading single sheets or graphs, thin magnetic 'hold down' strips are provided and these can also be used to restrain edge curling of long records. When used for multi-channel records the channel number indicator (1-12 inclusive) may be connected to external contacts which close to indicate the channel in use. The 28V supply for the indicator is provided in the reader. The reader is designed so that an optical projection unit may be fitted to enable film records to be projected on to a screen which replaces the standard reading table.

Normally the reader accepts the reference voltage supply for the *X* and *Y* potentiometers from an external source. This voltage must not exceed 50V, nor should the potentiometer wiper currents exceed 50µA if the accuracy is to be maintained.

The reader may be used with any analogue-to-digital convertor and may be directly connected to Southern Instruments Ltd chart digitizing and programming units.

EE 91 769 *for further details*

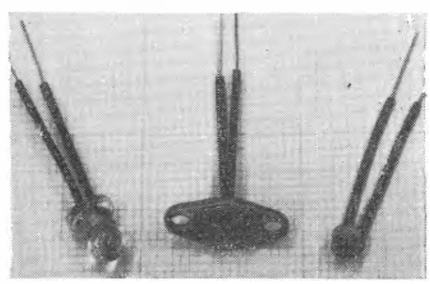
GALLIUM ARSENIDE LAMPS

The Plessey Co. Ltd, Dielectric and Magnetic Division, Towcester, Northamptonshire

(Illustrated below)

A new range of gallium arsenide infra-red lamps is entering quantity production at the Towcester factory of the Dielectric and Magnetic Division, Plessey Components Group.

These subminiature lamps are pn junction devices, emitting a cone of light of relatively narrow spectral width in the infra-red. Three models are available; a resin encapsulated emitter with no lens, a similar emitter fitted with a moulded resin lens $\frac{1}{8}$in diameter, and a unit mounted in a TO37 header, forming a heat sink.



The new emitters used in conjunction with a silicon photovoltaic cell can replace miniature tungsten filament lamps in a very wide range of equipment, and combine reduced physical size and power consumption with greatly improved reliability. Applications include punched tape read-out, vending, weighing and ticket issuing machines, and encoding devices. Very fast response time enables the light output to be modulated at up to 100Mc/s, enabling the emitters to be used in line of sight communication equipment.

EE 91 770 *for further details*

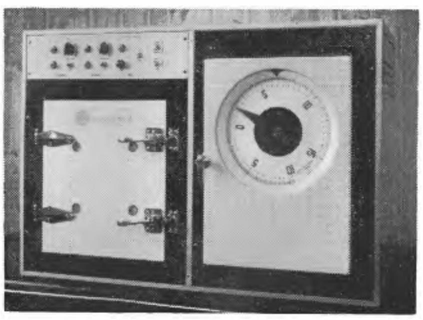
ENVIRONMENTAL TEST CHAMBERS

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

(Illustrated below)

This range of temperature test chambers is now available with an improved specification. A typical specification of the new standard range is:

Temperature span: −185°C to +250°C; or −75°C to +250°C.



Temperature control sensitivity: ±0·1°C.

Temperature gradient at any temperature: ±0·5°C.

Thermal shock facilities: +100°C to −100°C within three minutes.

Chambers are manufactured to BS-DEF specifications, and are made from the finest materials, many extras are available—special doors, racks, supporting jigs, automatic programme controls and multi-point pen recorders.

A feature of all chambers is the module system, which enables special requirements to be met with comparative ease at moderate cost.

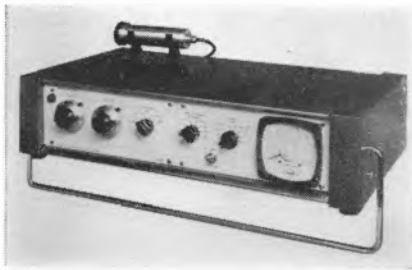
EE 91 771 *for further details*

RATEMETER

Panax Equipment Ltd, Holmethorpe Industrial Estate, Redhill, Surrey

(Illustrated above right)

A new transistorized wide-range accurate ratemeter for scientific and other laboratory applications has been added to the Panax 'Redhill' range of non-modular nucleonic instruments. Suitable for general purpose Geiger-Müller and scintillation counting, in laboratories, hospitals and clinics, the P.7900 ratemeter is a compact mains-powered unit with a built-in stabilized e.h.t. supply, an amplifier and discriminator having a wide bias range. It is a lighter and more



versatile replacement for the Panax type 5054 ratemeter, which is no longer in production.

There are nine separate count ranges and seven preset counting times. Count-rate is read on a large 240-degree moving-coil meter, and there is also a small speaker to give an audible indication when required.

A 1mA output is available at a co-axial socket on the instrument for operating an external meter or a pen recorder, such as the Panax SREC.2 flat-bed recorder or REC.3 PX module.

Provision can be made for the rate-meter to be powered from a 12V d.c. supply, such as an automobile battery, if no mains supply is available. This makes the instrument particularly suitable for accurate measurements which have to be made in the field.

The 'Redhill' series of complete counting instruments has been developed in parallel with the PX modular instrumentation produced by Panax, and is intended primarily for users requiring low-cost accurate equipment without the infinite extension capability offered by the modular systems.

EE 91 772 *for further details*

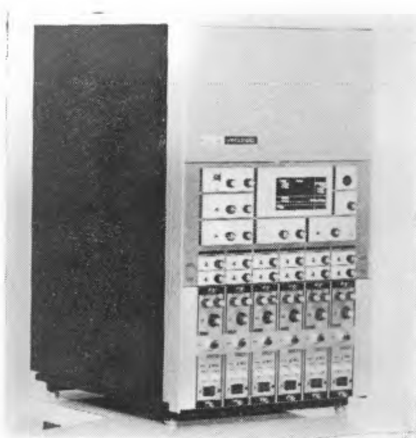
PROGRAMMABLE PULSE GENERATOR

Digital Equipment Corporation (UK) Ltd, 11 Castle Street, Reading, Berkshire

(Illustrated below)

This programmable pulse generator introduced by Digital Equipment Corporation (U.K.) Ltd is designed for use in laboratory and production line testing of ferrite memory components.

The PPG 2120 is a table-top system offering 4-, 8- or 16-step testing programmes for such devices as cores and films operating at speeds between 10kc/s and 5Mc/s.



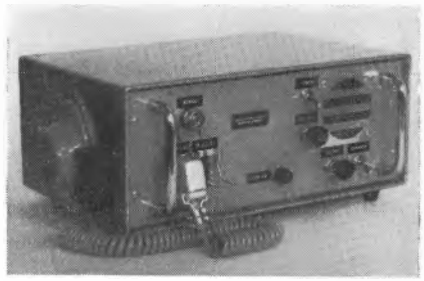
The 2120 provides six output channels fed by high-speed current drivers, with individual adjustments for pulse width, shape and position. Pulse width can be varied from zero to 10μsec, and each pulse can be delayed after the start of a program step by up to 10μsec.

The new system has two independent repeat channels, permitting the user to select step repeats at any program steps and pair repeats for any even-odd pair. Repetition intervals are continuously variable from 200nsec to 3msec.

Synchronizing pulses to trigger an oscilloscope can be selected at any steps. These pulses can also be delayed up to 10μsec.

Operating controls and switches are mounted on the front, and the cabinet swings open for access from the front to internal components.

EE 91 773 *for further details*



RADIO-TELEPHONE

Radio Communications Co., 16 Abbey Street, Crewkerne, Somerset

(Illustrated above)

This instrument has been specially developed to fulfil the need for an equipment of low current consumption for use as a mobile station or fixed station in isolated surroundings where a mains power supply is not available.

It is designed for operation from a 12V battery and uses transistors throughout the transmitter and receiver sections. The standby current is only 26mA, this figure rising to 1¾A while transmitting. A power output of approximately 5W is produced. The power output of the receiver is 1W.

The TRT/6 is intended for a.m. speech operation on 1 to 12 closely spaced channels in the 40Mc/s to 100Mc/s band for 25kc/s or 50kc/s channel spacing.

When used with a battery that is charged with solar cells or a wind-charger, many months of unattended base station operation can be achieved.

The TRT/6 is normally supplied complete with a plastic fist microphone or lightweight handset and operates from an external 75Ω aerial system which may be of the directional type to give greater range of communication.

For repeater operation, a modified version of the TRT/6 can be supplied, namely TRT/6R.

A transportable version, type TRT/4, is also available complete with telescopic aerial and internal 28V sealed rechargeable battery.

EE 91 774 *for further details*

Meetings THIS Month

THE INSTITUTION OF ELECTRICAL ENGINEERS

All meetings will be held at Savoy Place, commencing at 5.30 p.m. unless otherwise stated.

Ordinary Meetings

Date: 3 March.
Lecture: The Generation of Cloud Electricity.
By: B. J. Mason.
Date: 24 March.
Lecture: The Place of the Engineer in Society.
By: Lord Snow.
(Joint meeting with the Institution of Civil and of Mechanical Engineers.)

I.E.E./R.Ae.S. London Joint Group on the Applications of Electricity and Electronics in Aircraft

Date: 22 March. Time: 6 p.m.
Discussion: Solid State Switching—for Better or for Worse?
Opened by: H. Latham.

Control and Automation

Date: 1 March.
Lecture: The Analysis and Design of Model Reference Adaptive Control Systems.
By: A. J. White.
Date: 8 March.
Lecture: A New Approach to Measurement and Data Transmission for Process Control Systems.
By: I. R. Young, A. T. Keefe and G. Moss.
(Joint meeting with the Society of Instrument Technology.)
Date: 14 March.
Lecture: The Explanation of Some Fundamental Phenomena of Modern Physics using a Ballistic Theory of Light.
By: R. A. Waldron.
Date: 23 March.
Discussion: Technical Information—a New Teaching Subject.
Opened by: G. H. Wright.
Date: 25 March.
Discussion: Semiconduction in Organic and other Weakly Conducting Materials.
Opened by: D. C. Northrop.
Date: 28 March. Time: 10 a.m.
Colloquium: Making Plant Models.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Date: 29 March. Time: 10.30 a.m. and 2.30 p.m.
Colloquium: The Engineering of Computer Control Systems in Industry.
(Joint meeting with I.E.R.E. Computer Group. Tickets are available from the Secretary I.E.R.E.)

Electronics

Date: 1 March. Time: 2 p.m.
Discussion: Gamma and Infra-red Camera Techniques.
(Joint meeting with the I.E.R.E. Medical and Biological Group.)
Date: 2 March. Time: 9.30 a.m. and 2.15 p.m.
Colloquium. Microphones.
(All wishing to attend must apply to the Secretary for a ticket. No fee for I.E.E. members; non-members £1 1s.)
Date: 9 March. Time: 6 p.m.
Lecture: The Scanning Electron Microscope and other Electron Probe Instruments.
By: C. W. Oatley.
Date: 9 March.
Lecture: Applications of Electron-Statistics.
By: A. W. Bright and P. L. Secker.
Date: 10 March.
Discussion: The Multi-carrier Performance of Communications Satellite Repeaters.
Opened by: W. L. Wright, K. U. Bolwell and P. N. Sargeaunt.
Date: 14 March.
Lecture. The Ten Element Band Pass Filter Section and its Applications.
By: G. C. S. Bown.
Date: 21 March. Time: 6 p.m.
Lecture: Electron Beam Welding and Machining.
By: H. N. G. King.
Date: 25 March.
Discussion: U.H.F. and S.F. Propagation.
Date: 28 March.
Lecture: Semiconductors in Television Receivers.
By: R. Bridgen, K. E. Martin and P. L. Mothersole.
(Joint meeting with I.E.R.E., Television Group and the Television Society.)

Date: 29 March.
Discussion: Gunn Effect Devices.
Opened by: J. E. Carroll, J. S. Heeks and A. K. Jonscher.

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Joint I.E.E.-I.E.R.E. Medical Electronic Group

Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Date: 1 March. Time: 2 p.m.
Discussion: Infra-red Camera Techniques.

Electro-Acoustics Group

Held at: I.E.R.E. Lecture Room, 9 Bedford Square, London, W.C.1.
Date: 23 March. Time: 6 p.m.
Lecture: The Propagation of Sound Through Liquids.
By: R. W. B. Stephens.

Joint I.E.E.-I.E.R.E. Television Group and Television Society

Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Date: 28 March. Time: 5.30 p.m.
Lecture: Semiconductors in Television Receivers.
By: P. L. Mothersole, R. Bridgen and K. E. Martin.

Joint I.E.R.E.-I.E.E. Computer Group

Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Date: 29 March. Time: 2.30 p.m. and 5.30 p.m.
Symposium: Computer Control in Industry: Equipment Design and Application Engineering.
(Registration necessary.)

Radar Group

Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Date: 31 March. Time: 2.30 p.m. and 5.30 p.m.
Symposium: Monitoring of Ground and Airborne I.L.S. Equipment for Automatic Landing.
(Registration necessary.)

Thames Valley Section

Held at: School of Electronic Engineering, R.E.M.E., Arborfield, Berks.
Date: 10 March. Time: 7.30 p.m.
Lecture: Random Signal Testing.
By: P. Atkinson.

South Western Section

Held at: City of Plymouth College of Technology.
Date: 22 March. Time: 7 p.m.
Lecture: Electronic Circuits.
By: G. King.
Held at: University of Bristol Engineering Laboratories, University Walk, Clifton, Bristol, 8.
Date: 23 March. Time: 7 p.m.
Lecture: High Speed Magnetic Thin Film Memories.
By: A. T. Gibson.
(Joint meeting with the Bristol Branch of the British Computer Society.)

South Wales Section

Held at: South Wales Institute of Engineers, Cardiff.
Date: 7 March. Time: 6 p.m.
Lecture: Telemetry—The Present Position and Future Trends.
By: R. E. Young.
(Joint meeting with the Institution of Electrical Engineers.)

East Anglian Section

Held at: Chelmsford Technical High School, Patching Hall Lane, Broomfield, Chelmsford.
Date: 22 March. Time: 6.30 p.m.
Lecture: Various Aspects of Microwave Ferrites.
By: J. A. Penney and others.
Held at: Harlow Technical College, The High, Harlow.
Date: 8 March. Time: 7 p.m.
Lecture: Automatic Tracking from Surveillance Radars.
By: T. Buckley.

West Midland Section

Held at: Lanchester College of Technology, Priory Street, Coventry.
Date: 14 March. Time: 7 p.m.
Lecture: Radio Astronomy.
By: H. Gent.

Scottish Section

Held at: Department of Natural Philosophy, The University, Drummond Street, Edinburgh.
Date: 9 March. Time: 7 p.m.
Lecture: Laser Range Finders—Systems Analysis and Electronic Circuits.
By: G. Hamilton and A. Fowler.
Held at: Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow, C.2.
Date: 10 March. Time: 7 p.m.
Lecture: Laser Ranger Finders—Systems Analysis and Electronic Circuits.
By: G. Hamilton and A. Fowler.
(Joint meeting with the Institution of Electrical Engineers.)
Held at: The Royal Infirmary, Lauriston Place, Edinburgh, 3.
Date: 29 March.
Symposium: Tape Recording of Biological Signals.
(Registration necessary. Details from P. M. Elliott, 21 Craigmount Lane, Corstorphine, Edinburgh, 12.)

South Midland Section

Held at: B.B.C. Club, High Street, Evesham.
Date: 29 March. Time: 7 p.m.
Lecture: U.H.F. Tuners.
By: R. Bridgen.

Southern Section

Held at: Highbury Technical College, Cosham, Portsmouth.
Date: 2 March. Time: 6.30 p.m.
Lecture: Proximity Sensing by Magnetic Induction Applications and Techniques.
By: D. Barnard.
Held at: Lanchester Theatre, University of Southampton.
Date: 8 March. Time: 6.30 p.m.
Lecture: Electron Microscopy and Microelectronics.
By: W. C. Nixon.
(Joint meeting with the Institution of Electrical Engineers.)
Held at: Farnborough Technical College.
Date: 31 March. Time: 7 p.m.
Lecture: Random Access Communication Systems.
By: L. C. Walters.

Yorkshire Section

Held at: Huddersfield College of Technology, Department of Electrical Engineering.
Date: 3 March. Time: 7 p.m.
Lecture: Automatic Tracking from Surveillance Radars.
Held at: University of Leicester.
By: T. Buckley.

East Midland Section

Date: 15 March. Time: 6.30 p.m.
Lecture: Laser and Direct Energy Weapons.
By: R. C. Smith.

Merseyside Section

Held at: Walker Art Gallery, William Brown Street, Liverpool, 3.
Date: 16 March. Time: 6.30 p.m.
Lecture: Radiophony—The Synthesis of Sound Effects for Radio and Television.
By: F. C. Brooker.

North Eastern Section

Held at: Institute of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle.
Date: 9 March. Time: 6 p.m.
Lecture: The Application of Analogue Computers.
By: C. Cox.

THE TELEVISION SOCIETY

Meetings are held in the Conference Suite, I.T.A., 70 Brompton Road, London, S.W.3, and commence at 7 p.m.

Date: 10 March.
Lecture: Colour Film for Colour Television.
By: C. B. C. Wood.
Date: 24 March.
Discussion: The Impact of Semiconductors on Television Receiver Design.

NOUVELLES Réalisations

Traduction des pages 190 à 195

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

EQUIPEMENT À ULTRASONS

The M.E.L. Equipment Co. Ltd, Manor Royal, Crawley, Sussex

(Illustration à la page 190)

Les premiers éléments d'une nouvelle gamme d'instruments à ultrasons ont été mis au point par la société M.E.L. Equipment Co. Ltd.

L'élément de base est un amplificateur (type L.673) d'une sortie de 100 W. La fréquence de fonctionnement est contrôlée automatiquement par réaction du transducteur. L'amplificateur peut être actionné de manière continue et il est conforme aux normes britanniques et continentales en matière de sécurité électrique et d'interférences radiophoniques.

L'amplificateur peut être utilisé pour actionner un bain de nettoyage à ultrasons de 2 litres (type L.673) ou un bain de 4 litres (type L.674). Les deux bains sont en acier inoxydable et munis d'éléments transducteurs en ferrite. Un commutateur de permutation (type COS.10) permet, en variante, l'utilisation de n'importe quelle combination de deux bains à partir d'un seul amplificateur.

Un élément de commande (type L.666) a également été réalisé; ce dernier, en liaison avec le nouvel amplificateur, constitue une source de puissance à ultrasons de 100 W d'une grande souplesse d'emploi. Il comporte une commande de phase pouvant être utilisée lorsque les appareils sont employés comme amplificateurs à réaction, ainsi qu'une commande d'entraînement pour le réglage manuel de la sortie de l'oscillateur, lorsqu'ils sont employés comme générateurs.

Une innovation importante a été réalisée par le contrôleur d'amplitude qui indique en microns l'amplitude du mouvement physique effectif de la sonde.

Un nouveau transducteur à sonde (type L.667) a été également réalisé. Il peut être utilisé dans une gamme étendue d'applications comprenant le forage, le soudage de matières plastiques, l'étamage, l'émulsification et le contrôle de bactéries. Il est prévu pour être actionné par le nouvel amplificateur (type L.665) et l'élément de contrôle (type L.666). Le L.667 comprend un transducteur magnétostrictif en nickel, fixé de manière permanente à un transformateur de vitesse bronze/aluminium d'une demi-longueur d'onde. La polarisation est assurée par des aimants permanents, et deux capteurs en zirconium de plomb sont prévus pour fournir des signaux pour la commande automatique de fréquence et le contrôleur d'amplitude.

L'assemblage à transducteurs est enfermé en partie dans une chemise de refroidissement cylindrique remplie d'huile et montée dans un moulage. Le moulage fait fonction de source froide et il est refroidi par ventilateur.

En montant le transducteur (type L.667) sur un nouveau support de forage (type L.668) on obtient une foreuse à ultrasons d'un modèle perfectionné. Selon les constructeurs une ancienne foreuse de 60 W avait un maximum de coupe transversale de 19 mm, tandis que le nouvel instrument de 100 W peut convenir à des outils d'une coupe transversale de 38 mm au minimum avec une vitesse, une profondeur et une précision accrues.

L'équipement de forage complet comprend le nouvel appareil de contrôle et d'amplification et, offre par conséquent, des possibilités de contrôle d'amplitude.

Le support de forage lui même comprend un certain nombre d'avantages propres aux applications industrielles. La base et la colonne sont d'une plus grande précision linéaire et d'une stabilité accrue, tandis que la position de montage du transducteur peut être alignée avec précision. Un dispositif l'entraînement permet de déplacer l'assemblage de transducteurs de bas en haut et inversement le long de la colonne principale pour recevoir la pièce sur laquelle on travaille. Le transducteur glisse sur des roulement à billes de précision entre les positions de coupage et d'arrêt et il est muni d'un bouton à déclenchement rapide pour le mettre dans la position de coupage.

Notre gravure montre l'ensemble du matériel à ultrasons comprenant l'amplificateur de 100 W L.665, le contrôleur L.666 monté au-dessus et le transducteur à sonde L.667 monté sur le support de forage L.668.

EE 91 751 *pour plus amples renseignements*

GÉNÉRATEUR DE TACHYMÈTRE

Muirhead & Co. Ltd, Beckenham, Kent

(Illustration à la page 190)

La société Muirhead & Co. Ltd produit maintenant un générateur à tachymètre de moteur d'asservissement, format 11, qui a été spécialement conçu pour les servomécanismes intégrateurs et à grande vitesse de précision.

Grâce aux soins tout particuliers apportés à l'étude et à la construction de ce générateur, la linéarité de la tension de sortie en fonction de la vitesse est maintenue à un minimum de 0,07% tandis que la variation maxima de la tension de sortie en fonction de la température n'est que de 0,25% dans une gamme de température de −15° C à +75° C. Par l'emploi du dessin tachymétrique " inversé," on a pu porter la tension de sortie à 2,75 V par 1000 tours/min. Cette valeur est presque cinq fois plus élevée que celle des tachymètres format 11 des modèles précédents.

Les constructeurs ont prévu, en outre, différents types d'enroulements de moteur répondant aux besoins particuliers de chaque client.

EE 91 752 *pour plus amples renseignements*

FILTRE DÉTECTEUR

Associated Engineering Ltd, 60 Kenilworth Road, Leamington Spa, Warwickshire

(Illustration à la page 190)

La société Associated Engineering Ltd fournit maintenant un nouveau filtre détecteur, modèle DF/2, pouvant être utilisé avec des paires assorties ou des versions doubles de transducteurs de déplacement inductif AE pour la mesure sans contact du déplacement et des vibrations dans des conditions d'environnement défavorables.

L'appareil permet d'effectuer des

travaux de mesure précis dans une gamme étendue d'applications comprenant le mouvement de paliers, l'épaisseur de films d'huile et le taux d'usure de coussinets, les effets complexes de vibrations dans les structures délicates où le dispositif de mesure ne doit pas charger le corps soumis à la mesure et, enfin, la mesure de précision métrologique sur machines-outils et gabarits d'usinage.

Le nouveau filtre détecteur DF/2 mesure le déplacement ou les vibrations à des fréquences allant de zéro à 10 kHz sur une distance de 0,015 pouce avec une résolution maxima de 10 micro-pouces selon le matériau soumis à la mesure.

La stabilité du système par rapport au changement de température, de même que la stabilité des transducteurs, est supérieure à ± 2 pouce/° C lorsque les transducteurs sont dans un cycle de température allant de la température ambiante à 150° C, et la stabilité à long terme du système pour la mesure de déplacement statique pendant une période de huit heures est supérieure à ± 100 μ-pouce pour les matériaux ferreux.

EE 91 753 *pour plus amples renseignements*

POTENTIOMÈTRES RECTANGULAIRES

Reliance Controls Ltd, Sutherland Road, London, E.17

(Illustration à la page 191)

La société Reliance Controls Ltd produit maintenant une nouvelle gamme de potentiomètres d'équilibrage rectangulaires; cette série porte la référence de type CW.90. La grande fiabilité de ces potentiomètres a pu être obtenue grâce à l'emploi d'un corps au phthalate diallyl, d'un cadre en céramique, de terminaisons à extrémités soudées et à vis en acier inoxydable. Les éléments sont mécaniquement interchangeables avec les modèles Reliance ainsi qu'avec d'autres modèles. Les bornes sont sur un réseau de $\times 0{,}1$ pouce, la gamme de résistance va de 10 Ω à 50 kΩ et la gamme de température s'étend de $-40°$ C à $+100°$ C. Les éléments standard sont à bobinage en fil Karma donnant des coefficients de température de 20×10^{-6} sur la plus grande partie de la gamme de température.

Une version entièrement scellée de ce potentiomètre sera disponible prochainement.

EE 91 754 *pour plus amples renseignements*

RELAIS DE COMMUTATION D'ANTENNE COAXIALE

Magnetic Devices Ltd, Exning Road, Newmarket, Suffolk

(Illustration à la page 191)

Un nouveau relais de commutation d'antenne coaxiale, conçu pour éliminer les fiches et douilles coaxiales dans le corps du relais et de réduire ainsi les dimensions et le coût, est fourni maintenant par la société Magnetic Devices Ltd. Le relais à tension continue, série 951, a été étudié pour la commutation par antenne de fréquences de l'ordre de 450 MHz, et il peut être utilisé également pour les applications exigeant une capacité d'inter-contact exceptionnellement faible.

Un petit relais à action rapide avec un seul contact de commutation est logé dans un coffret en laiton. Ce relais a été étudié pour l'emploi avec des câbles U.R.43 pouvant être facilement reliés par soudage sans déranger le réglage du relais et pouvant être, en outre, fermement fixé par des pinces en laiton. Ces pinces assurent la continuité du blindage et des sorties de câble dans trois directions au choix.

Le relais série 951 a une tension de régime maxima de 100 V c.c., une tension maxima de bobine de 1,5 W et une résistance de bobine maxima de 9350 Ω. Il a été contrôle pour résister à 1 kV de tension efficace entre toutes les bornes ainsi qu'entre le cadre et la bobine; la fréquence nominale est de 450 MHz, le taux d'ondes stationnaires est d'environ 1,1:1, et la diaphonie est de 39 dB.

La durée de fonctionnement est de 20 msec, la durée de déclenchement de 8 msec, avec une puissance nominale de contact de 1 A ou de 30 W au maximum, l'impédance nominale étant de 50 Ω. Le relais série 951 peut être prévu pour couvrir la gamme de température de $-40°$ C à $+100°$ C si nécessaire. Il est à revêtement argenté et son poids est de 108,25 gr.

EE 91 755 *pour plus amples renseignements*

CHARGEURS DE BATTERIES MINIATURE

Kynmore Engineering Co. Ltd, 19 Buckingham Street, London, W.C.2

(Illustration à la page 191)

La société Kynmore Engineering Co. Ltd produit une nouvelle gamme de chargeurs miniature à courant courant constant pour batteries rechargeables en nickel-cadmium. Les nouveaux chargeurs ne sont pas plus grands que les batteries elle-mêmes et leur courant de fuite est si faible qu'on peut les laisser reliés aux batteries de manière permanente.

Le chargeur consiste en un réservoir tubulaire en aluminium contenant un condensateur, des résistances et un dispositif redresseur du type à semi-conducteurs encapsulé dans de l'Araldite, dont émergent les trois câbles de connexions à codage de couleurs de 6,25 cm. Les diodes à agglomérant d'or redressent le courant d'alimentation et maintiennent le courant de fuite à virtuellement zéro. Durant le fonctionnement, on obtient de la tension alternative deux fois l'équivalent efficace du courant de charge; la majeure partie de l'impédance vers la tension alternative se produit aux bornes du condensateur. L'impédance et, par conséquent le taux de charge demeurent substantiellement constants, quel que soit l'état de charge de l'accumulateur. Les chargeurs peuvent ainsi charger, le cas échéant, des piles entièrement épuisées, nulle approximation extérieure n'étant nécessaire.

Dans la gamme standard, six formats différents sont prévus dont les capacités vont de 2 à 45 mA pour une alimentation en courant alternatif de 230 V. Tous les chargeurs sont étudiés pour un maxima de 10 piles. Les dimensions varient de 0,8 pouce de long $\times$ 0,6 pouce de diamètre à 1,25 pouce de diamètre.

EE 91 756 *pour plus amples renseignements*

CHAMBRE D'ESSAI D'ENVIRONNEMENT

Medical Electronics Ltd, Colford Works, King Henry's Walk, London, N.1

(Illustration à la page 191)

Cette chambre d'essai d'environnement portative, type ETC 505 a été conçue pour le contrôle de température de diodes, de transformateurs et d'autres petits composants.

Les températures peuvent être réglées ou automatiquement mises en cycle entre des limites préréglées et à des taux présélectionnés.

La chambre de contrôle est en acier inoxydable et des câbles électriques peuvent être fixés à la porte pour permettre le contrôle dynamique d'appareils.

Le chauffage se fait par circulation d'air forcé au moyen de dispositifs de chauffage blindé. Pour le refroidissement de liquides, on répand du CO_2 dans l'air forcé dans lequel il est rapidement mélangé, éliminant ainsi toute congélation.

Le réglage du chauffage et du refroidissement s'effectue à partir d'une seule commande de panneau frontal. Cette commande peut être utilisée également pour le réglage de températures précises; ces dernières sont choisies sur un indicateur de température à lecture facile qui comprend deux commandes indépendantes.

Le volume de régime de la chambre est de 22,96 cm $\times$ 17,78 cm $\times$ 20,33 cm. La gamme de température s'étend de $-70°$ C à $+200°$ C; le taux de refroidissement est de 5° C/min et le taux de chauffage est de 10° C/min.

EE 91 757 *pour plus amples renseignements*

CALCULATRICE NUMÉRIQUE

Hawker Siddeley Dynamics Ltd, Whitley Works, Nr. Coventry, Warwickshire

(Illustration à la page 192)

La calculatrice numérique IAD. DCC.1, répond à la demande d'un appareil peu coûteux et d'une grande souplesse d'emploi pouvant effectuer des opérations logiques et mathématiques et pouvant être logé dans des systèmes de traitement et d'enregistrement de données. A cette fin, les constructeurs on fait un usage étendu des méthodes de micro-programmation afin de réduire le nombre de pièces et de faciliter la programmation.

La calculatrice peut recevoir des données de Boole sous forme binaire, ou des données arithmétiques sous forme décimale à code binaire, d'une bande perforée à huit trous; elle peut recevoir également pour les applications en ligne, des données de convertisseurs analogiques/numériques ou de transducteurs à sortie numérique.

Lorsque la calculatrice est incorporée au système d'enregistrement et de concentration de données IAD, l'élément de contrôle de programme lui donne les ordres nécessaires pour effectuer des calculs sur des données qui lui sont fournies par le reste du système, soit suivant un programme déterminé soit sur demande.

L'information est emmagasinée dans une matrice à noyau de ferrite à 1024 emplacements avec six chiffres binaires par emplacement. Chaque emplacement peut recevoir soit un chiffre décimal et une formule de Boole soit un caractère de programme.

La sortie de calculatrice peut alimenter des dispositifs imprimeurs, des convertisseurs numériques/analogiques ou des transducteurs pouvant effectuer le contrôle de matériel.

L'appareil est monté sur plaquette à fiches et ses circuits logiques sont principalement de 2 MHz.

La calculatrice comprend trois sections principales:
(1) les registres et l'additionneur
(2) le micro-programme
(3) le magasin à chiffres.

Les registres assurent l'emmagasinage temporaire d'informations devant être traitées par le micro-programme.

Le micro-programme, qui est contrôlé par un compteur local d'ordres, dirige l'information et les ordres entre les registres, l'additionneur et le magasin à chiffres. Il se compose d'une matrice à diodes qui peut être facilement changée pour fournir à la calculatrice le code d'ordre le plus approprié au système dans lequel elle est utilisée.

Le magasin à chiffres est un réservoir à noyau de ferrite et à temps de cycle de 6 μsec; il comprend une niche dans laquelle l'information est emmagasinée sur une base de "première reçue, première sortie".

EE 91 758 *pour plus amples renseignements*

MINUTERIE

Solid State Controls Ltd, 30-40 Dalling Road, London, W.6

(Illustration à la page 192)

Le "Cyclo-Set" comprend un minimodule standard Cyclo-Set, pouvant être fourni également comme élément à part, ainsi qu'un relais à fiches et à huit broches de 12 V c.c., monté sur un robuste châssis en acier. Une autre version de la minuterie est fournie dans un boîtier coulé, entièrement enfermé. Le minimodule a une durée de vie allant jusqu'à 100 millions d'opérations; il est entièrement transistorisé, ne comporte pas de partie mobile et ne nécessite aucun réglage. Le relais scellé fournit un contact de sortie de commutation unipolaire d'une puissance nominale de 5 ou de 10 A à 250 V c.a., convenant tant aux charges résistives qu'aux charges inductives.

La minuterie à deux étages comprend deux périodes de minutage, "arrêt" et "marche," entièrement réglables, de 0,4 sec à 4 min pour l'appareil standard, et de 0,4 sec à 20 min pour des besoins spéciaux.

La minuterie est prévue pour six tensions alternatives de fonctionnement allant de 12 à 415 V, et pour deux tensions continues, de 12 et de 24 V. De plus, la minuterie peut être mise en oeuvre à partir d'un contact normalement fermé aussi bien qu'à partir d'un contact normalement ouvert. La minuterie Cyclo-Set est pratiquement insensible aux fluctuations de tension de fonctionnement de ± 10 % et aux températures ambiantes dans la gamme de $-20°$ C à $+55°$ C.

EE 91 759 *pour plus amples renseignements*

LAMPES INDICATRICES À TROIS COULEURS

Thorn Special Products Ltd, Great Cambridge Road, Enfield, Middlesex

(Illustration à la page 192)

L'indicateur à trois couleurs Thorn est construit maintenant avec de nouveaux filtres intérieurs de verre projetant les couleurs choisies à travers la lentille de verre à face unique. Des lampes de 2 W, 28 V, peuvent être utilisées de façon continue et augmentent ainsi sensiblement la sortie de lumière. En plus de cet avantage, l'emploi du verre pour les lentilles permet de contrôler plus rigoureusement les coordonnées de couleur et de réduire la tendance à l'altération des couleurs. Les couleurs normalement employées actuellement sont le rouge "signal", le vert "signal" et le jaune "signal", selon la norme britannique No. 1376.

La longueur hors-tout de l'indicateur est de 5,71 cm et son diamètre est de 1,98 cm. Il peut être monté dans des panneaux de 0,31 cm d'épaisseur et la dimension des ouvertures requises est de 1,5 cm.

L'indicateur à trois couleurs utilise le minimum d'espace de panneau et il peut souvent remplacer trois lampes de signalisation séparées. Facilement monté au moyen d'un anneau de verrouillage, l'indicateur est d'un accès aisé pour l'entretien: il n'est donc pas nécessaire de le déplacer.

Il peut être fourni avec trois lampes de panneau "Atlas Midget", d'une puissance nominale de 6, de 12 ou de 28 V.

EE 91 760 *pour plus amples renseignements*

POTENTIOMÈTRES MINIATURE

Miniature Electronic Components Ltd, Copse Road, St. Johns, Woking, Surrey

(Illustration à la page 192)

La société Miniature Electronic Components Ltd vient d'ajouter le potentiomètre MT20 à sa série de potentiomètres d'équilibrage bobinés. Enfermé dans un boîtier à transistors du format TO-5, ce potentiomètre peut être monté sur circuit imprimé et il est prévu pour des résistances normales de 50, 100, 200, 500, 1 k, 2 k, 5 k et 10 kΩ. De conception simple et sûre, il est particulièrement indiqué pour les applications de l'électronique commerciale et industrielle qui exige des pièces de haute qualité à un prix raisonable. Ses principales caractéristiques sont les suivantes: gamme de température de $-20°$ C à 125° C; 1 watt à une température nominale de 70°; réglage par un seul tour; à l'épreuve de l'humidité et de la poussière.

EE 91 761 *pour plus amples renseignements*

OSCILLATEUR-DÉTECTEUR EN PONT

H. Tinsley & Co. Ltd, Werndee Hall, South Norwood, London, S.E.25

(Illustration à la page 193)

L'oscillateur-détecteur en pont Tinsley type 5720 se compose d'un générateur d'ondes carrées donnant deux sorties isolées et équilibrées de 2,5 V chacune. Le détecteur est synchronisé avec le générateur.

Il a été prévu pour les ponts extensométriques ou les ponts à thermomètre de résistance qui exigent un oscillateur et un détecteur combinés.

Un atténuateur contrôle la sensibilité du détecteur dont la sortie est indiquée sur un instrument de mesure de 7,5 cm. La sensibilité peut être contrôlée de manière à faire correspondre chacune des divisions de l'instrument de mesure à une quantité précise de contrainte ou de température, suivant la façon dont l'élément est utilisé.

Il comprend une batterie chargeable pouvant être rechargée par le courant secteur.

EE 91 762 *pour plus amples renseignements*

APPAREIL DE CONTRÔLE À MILLIWATTS

Standard Telephones & Cables Ltd, Testing Apparatus Division, Corporation Road, Newport, Monmouthshire

(Illustration à la page 193)

La société Standard Telephones & Cables Ltd vient de créer une nouvelle version de l'appareil de contrôle à milliwatts 74166-A, assurant les mêmes services que ce dernier. Le nouvel appareil est enfermé dans un nouveau coffret qui n'occupe que la moitié du volume du 74166-A. En plus de cette portabilité accrue, des interrupteurs à bouton-poussoir remplacent l'ancien interrupteur rotatif et l'instrument de mesure est logé derrière le panneau frontal afin de mieux le protéger et de lui donner une meilleure apparence.

Les deux modèles (H et J) sont identiques, sauf en ce qui concerne leur connecteur d'entrée, le 74166-H étant muni de fiches coaxiales No. 1 et le 74166-J ayant des connecteurs BNC (Amphenol 31-221). Les nouveaux appareils assurent la mesure précise de niveaux allant de $+1$ dB à -1 dB par rapport à 1 mW sur circuits déséquilibrés de 75 Ω à des fréquences atteignant 30 MHz. Ils peuvent également mesurer des voltages de chauffe de 5,6 à 7,0 V.

Un thermocouple bobiné est utilisé avec une impédance de chauffe shuntée à 75 Ω, les instruments ne pouvant donc pas être utilisés pour mesurer les niveaux de transit. Une précision de mesure élevée est assurée par une cellule standard miniature incorporée Weston, en fonction de laquelle le circuit est normalisé par une série d'opérations simples de commutation et de réglage. Le circuit de normalisation peut être utilisé en outre pour émettre une tension continue de 1 mV dans des circuits externes de 75 Ω pour l'étalonnage d'autres appareils.

Les deux modèles fonctionnent sur trois piles sèches de 1,5 V logées dans un coffret mesurant 311 mm × 203 mm × 281 mm avec son couvercle.

EE 91 763 *pour plus amples renseignements*

ACCÉLÉROMÈTRES

Environmental Equipments Ltd. Station Yard, Wokingham, Berkshire

(*Illustration à la page 193*)

Les accéléromètres de la série AQ sont construits en acier inoxydable (assemblage mécanique). L'assemblage piézoélectrique est effectué sans adhérents afin d'éviter les problèmes se produisant aux températures élevées par suite de pertes de force d'agglomération et de changement de rigidité.

La température nominale maxima est de 260° C; la déviation de sensibilité de charge et de tension est inférieure à 5 % jusqu'à −65° C et inférieure à 15 % jusqu'à 160° C. La sensibilité transversale maxima est de 5 %. La capacité interne est de 1200 pF, celle de l'assemblage de câble à faible bruit de 300 cm étant de 300 pF.

Les quatre modèles prévus portent les référence AQ5, AQ10, AQ20 et AQ40. Les numéros des modèles indiquent la sensibilité de charge normale en p^C (pointe)/g (pointe).

EE 91 764 *pour plus amples renseignements*

INTERRUPTEUR À LEVIER MINIATURE

Keyswitch Relays Ltd. Cricklewood Lane, London, N.W.2

(*Illustration à la page 193*)

La société Keyswitch Relays Ltd a réalisé une gamme d'interrupteurs miniature à levier à deux directions du type postal 1 000 dont la durée de vie mécanique dépasse 750.000 opérations par cycle de commutation. Ces interrupteurs à deux directions sont prévus pour la commutation de faibles courants. Le contact s'effectue par commutation et on a prévu des combinaisons de contacts de verrouillage, de non-verrouillage et d'arrêt avec un maximum de 16 commutations (huit par contact). Les contacts jumelés en argent sont de type standard et la nouvelle gamme complète les interrupteurs à huit leviers de commutation crées par la société Keyswitch en 1964.

Les puissances nominales de contact sont de 100 V à 300 mA c.c., 12 V à 1A c.c., 100 V à 1 A c.a., avec des pressions de contact initiales de 15 g au minimum (et de 10 g au minimum après 750.000 opérations).

Chaque interrupteur est fourni avec deux vis de fixation et les leviers comportent un choix de 6 couleurs (rouge, vert, jaune, gris, ivoire ou noir) pour pouvoir les assortir à divers dessins de panneaux.

La longueur totale de l'interrupteur est de 7,06 cm, la profondeur, poignée y comprise, est de 5,56 cm et la largeur de 1,57 cm.

EE 91 765 *pour plus amples renseignements*

CALIBRES D'ÉPAISSEUR PORTATIFS

Dawe Instruments Ltd. Western Avenue, Acton, London, W.3

(*Illustration à la page 193*)

Deux nouvelles versions du matériel de mesure d'épaisseur à ultrasons portatif, Sonaray type 1805, ont été réalisées par la société Dawe Instruments Ltd. Ces nouveaux calibres, types 1805B et 1805C, remplacent le calibre précédent, type 1805A dont ils constituent des versions nettement plus perfectionnées.

Le calibre type 1805B, qui est en fait le plus simple des deux nouveaux modèles, est semblable en apparence et quant au fonctionnement au modèle ancien, type 1805A, mais comporte néanmoins les perfectionnement suivants: la consommation électrique a été réduite pour assurer cinq heures de fonctionnement continu à partir d'une seule charge, l'instrument étant principalement prévu pour l'utilisation sur batterie sur le terrain. Il peut, cependant, être employé également sur courant secteur, sa précision et sa résolution étant égales ou supérieures à celles de la plupart des calibres d'épaisseur ordinaire.

Les dimensions (38,1 cm × 30,13 cm × 11,43 cm), le poids (7,25 kg complet) et la disposition générale sont les mêmes que ceux du type 1805A, mais des transistors au silicium sont utilisés à la place du germanium. Il en résulte une profondeur de contrôle accrue (jusqu'à 24,3 cm d'acier), une gamme de fréquence plus étendue (de 1 à 8 MHz avec une performance utilisable allant jusqu'à 15 MHz) et une gamme de température de −10° F à +150° F. Le taux de répétition des impulsions est de 250 Hz ou de 1000 Hz et les échos d'impulsion sont enregistrés sur un tube cathodique à section carrée dont la surface d'écran effective (5 × 5 cm) est équivalente à la surface de tubes beaucoup plus grands. Comme dans le modèle précédent, l'échelle peut être étendue jusqu'à ce que la largeur totale représente 1,25 cm d'épaisseur d'acier et la précision du calibrage est, une fois de plus, de ±0,005 pouce d'épaisseur dans la gamme de 0,1 à 6 pouces.

Le nouveau type 1805C est essentiellement pareil au type 1805B mais il peut être employé également pour la détection automatique de pailles, la mesure d'épaisseur, l'enregistrement ou d'autres applications qui étaient jusqu'ici le propre d'appareils plus grands (bien que n'étant pas nécessairement plus précis).

Les perfectionnements de base apportés au type 1805C comprennent un atténuateur incorporé, variable en plots de 2 dB de 0 à 62 dB, une alarme incorporée de pailles et la possibilité de brancher des appareils auxiliaires. La construction modulaire du 1805C permet à l'utilisateur de convertir son type 1805B en un type 1805C, puis de choisir les éléments auxiliaires dont il a besoin. La gamme actuelle d'éléments auxiliaires (bien qu'incomplète) comprend un micromètre numérique, des instruments analogiques de temps et des contrôleurs de pailles universels, tous sous forme de modules standard constitués de corps solides. Deux de ces éléments, au choix, peuvent être branchés dans un logement auxiliaire accouplé au Sonoray lui-même.

Dans tous les cas, les transducteurs doivent être adaptés à la fréquence utilisée; il sont normalement du type piézoélectrique Z à performance élevée. Un câble de couplage de transducteur de 1,82 m est normalement fourni mais des câbles pouvant aller jusqu'à 61 mètres de long peuvent être utilisés sans modification aucune. Pour les applications sur le terrain, des cadres de transport, des enveloppes en grosse toile et des sacoches en cuir marron sont livrables. Ces accessoires de transport sont interchangeables pour les deux modèles.

EE 91 766 *pour plus amples renseignements*

OSCILLOSCOPE

Ribet-Desjardins, 13/17 Rue Perier, Montrouce (Seine), France

(*Illustration à la page 194*)

Ce nouvel oscilloscope est entièrement transistorisé et prévu en deux versions: le type 340A est un appareil portatif pour banc d'essai tandis que le type 340AR est un appareil pour montage sur bâti.

Les deux modèles sont munis d'amplificateurs verticaux à fiches fournissant un ou deux canaux. Lorsqu'on utilise la largeur de bande la plus étendue, soit 0 à 100 MHz, la sensibilité de déviation est de 50 mV/cm avec un temps de montée de 3,5 nsec. Lorsqu'on utilise une largeur de bande de 0 à 60 MHz, la sensibilité est de 5 mV/sec.

Les appareils peuvent être munis de bases de temps simples ou doubles fournissant de 5 sec/cm à 100 μsec/cm avec une extension de X10. La synchronisation peut s'effectuer jusqu'à 200 MHz.

L'amplificateur horizontal utilisé avec la base de temps double a une largeur de bande de 0 à 2 MHz, avec une sensibilité maxima de 0,1 V/cm; dans le cas de la base de temps simple, la largeur de bande est de 0 à 3 MHz, avec une sensibilité de 20 mV/cm.

EE 91 767 *pour plus amples renseignements*

CONTRÔLEUR DE SEMI-CONDUCTEURS

Rohde & Schwarz, 15 Mühldorfstrasse, 8000 Munich 8, Germany

(*Illustration à la page 194*)

Le contrôleur de semiconducteurs, type BMA, a été étudié pour la mesure rapide et le contrôle des principaux paramètres de semiconducteurs. Sa programmation s'effectue par cartes de bandes ou par bandes perforées et elle est augmentée au moyen d'un système modulaire. La vitesse d'échantillonnage maxima est de 0,1 sec par paramètre. Les données que fournit le contrôleur, sous forme de décisions ou sous forme numérique, peuvent être traitées et évaluées statistiquement au moyen du matériel d'acquisition et de traitement de données du type série UC. Le nouveau contrôleur de semiconducteurs est particulièrement indiqué pour les essais de production, la vérification d'entrée et les essais de fiabilité exécutés par le client. Le nouveau système UC pour l'acquisition et le traitement de données a été conçu suivant le principe des blocs de construction et il peut être adapté pour répondre à différents besoins. Il évalue des séries de mesure suivant des statistiques mathématiques et techniques (contrôle de qualité statistique), il classe les données numériques et emmagasine les fréquences absolues qui peuvent être imprimées. Un écran affiche la courbe de distribution qui est enregistrée par un appareil de report XY. Cet enregistrement est déformé ou non-déformé pour un nomographe de probabilités.

EE 91 768 *pour plus amples renseignements*

LECTEUR X-Y

P.C.D. Southern Ltd, 4 Blackwater Station Estate, Camberley, Surrey

Le nouveau lecteur "X-Y", type ZAE-1, permet l'analyse de diagrammes de données d'une manière qui n'a été possible jusqu'ici que sur des instruments beaucoup plus complexes et coûteux. Ce lecteur se caractérise par la précision du dispositif semi-automatique de décalage différentiel de l'axe "X" lequel, allié au système de pointage d'un fonctionnement aisé, permet de tabuler les ordonnées "X-Y", la pente instantanée de la ligne et de la surface sous la ligne avec une réduction considérable de l'habileté requise de l'opérateur. Des fonctions linéaires et non-linéaires peuvent être traitées en fixant simplement le viseur du curseur sur la ligne de trace et en pressant le levier de lecture pour déclencher l'impression.

Les dimensions hors-tout du lecteur sont de 107 cm × 61 cm × 24 cm. La surface de lecture totale est de 30 cm dans l'axe "Y" et de 50 cm dans l'axe "X", les rouleaux de diagrammes étant emmagasinés dans des bobines facilement accessibles et actionnés par simple commande manuelle, les données non-traitées étant normalement emmagasinées dans la bobine de droite et enroulées dans la bobine de gauche par unités de 50 cm de longueur. Des enregistrements d'une longueur approximative maxima de 60 mètres peuvent être emmagasinés.

Pour la lecture de feuilles simples ou de graphiques, les bandes magnétiques minces de fixation sont prévues. Ces bandes peuvent être utilisées également pour empêcher les enregistrements de s'écorner. Lorsque l'appareil est utilisé pour des enregistrements mutlivoies, l'indicateur du numéro de la voie (1 à 12 inclusivement) peut être branché aux contacts externes qui se ferment pour indiquer que la voie est utilisée. L'alimentation de 28 V pour l'indicateur est prévue dans le lecteur. Ce dernier a été conçu de façon à ce qu'un élément de projection optique puisse être fixé pour la projection d'enregistrements de film sur un écran qui remplace la table de lecture standard.

Normalement, le lecteur reçoit la tension de référence pour les potentiomètres "X" et "Y" d'une source extérieure. Cette tension ne doit pas dépasser 50 V; les courants de balayage du potentiomètre ne doivent également pas dépasser 50 micro-ampères si l'on veut maintenir la précision.

Le lecteur peut être utilisé avec n'importe quel convertisseur analogique/numérique et il peut être relié directement aux appareils de programmation et de chiffrage de diagrammes de la Southern Instruments Ltd.

EE 91 769 *pour plus amples renseignements*

LAMPES À L'ARSENURE DE GALLIUM

The Plessey Co. Ltd. Dielectric and Magnetic Division, Towcester, Northamptonshire

(*Illustration à la page 194*)

Une nouvelle gamme de lampes infrarouges à l'arsenure de gallium est maintenant produite en série à l'usine de Towcester de la Division Diélectrique et Magnétique du Groupe des Composants Plessey.

Ces lampes subminiature sont à jonction pn et émettent un cône lumineux, de largeur spectrale relativement étroite, dans la région de l'infrarouge. Trois modèles sont prévus: un émetteur encapsulé dans la résine et sans lentille, un émetteur analogue muni d'une lentille en résine moulée de 0,63 cm de diamètre et un élément monté dans un dispositif TO37 formant une source froide.

Les nouveaux émetteurs, utilisés en liaison avec une cellule photovoltaique au silicium, peuvent remplacer des lampes miniature à filament de tungstène dans une gamme très étendue d'appareils; ils allient des dimensions physiques réduites et une faible consommation de courant à une fiabilité nettement améliorée. Leurs applications comprennent la lecture sur bande perforée, les machines de pesage à émission de tickets et les dispositifs d'encodage. Le temps de réponse très rapide permet de moduler la sortie de lumière jusqu'à 100 MHz au maximum et, par conséquent, d'utiliser ces émetteurs dans le matériel de communication à portée optique.

EE 91 770 *pour plus amples renseignements*

CHAMBRES D'ESSAI D'ENVIRONNEMENT

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

(*Illustration à la page 195*)

Cette gamme de chambres d'essai de température comporte maintenant des améliorations dans sa spécification, à savoir:

Gamme de température
$-185°C$ à $+250°C$:
ou
$-75°C$ à $+250°C$:
Sensibilité de contrôle de température
$\pm 0,1°C$:
Gradient de température à n'importe quelle température
$\pm 0,5°C$:
Gamme de chocs thermiques
$+100°C$ à $-100°C$
en trois minutes.

Les chambres sont construites selon les spécifications BS-DEF et comprennent les meilleurs matériaux. De nombreux accessoires sont prévus: portes spéciales, châssis, bâtis de support, commandes de programme automatique et enregistreur à plume à plusieurs prises.

Toutes les chambres sont du système modulaire qui permet de faire face aisément à des travaux particuliers et à un prix modéré.

EE 91 771 *pour plus amples renseignements*

COMPTEUR

Panax Equipment Ltd, Holmethorpe Industrial Estate, Redhill, Surrey

(*Illustration à la page 195*)

Un nouveau compteur transistorisé de précision et à gamme étendue, pour les applications scientifiques et de laboratoire, a été ajouté à la série des instruments nucléaires non-modulaires Panax "Redhill". Le compteur P.7900 a été conçu pour le comptage universel Geiger-Müller et de scintillations dans les laboratoires, hôpitaux et cliniques. C'est un appareil compact et alimenté par courant secteur comprenant un bloc d'alimentation stabilisée incorporé, un amplificateur et un discriminateur à gamme de polarisation étendue. C'est une version plus légère et d'une plus grande souplesse d'emploi que le compteur Panax type 5054 dont la production a cessé.

Neuf gammes de comptage séparées et sept temps de comptage préréglés sont prévus. La vitesse de comptage est indiquée sur un grand instrument de mesure à cadre mobile de 240° et un petit haut-parleur donne, en outre, une indication audible, si nécessaire.

Une sortie de 1 mA est fournie sur une douille coaxiale de l'instrument pour actionner un instrument de mesure extérieure ou un enregistreur à plume, comme l'enregistreur Panax SREC.2 ou le module REC.3 PX.

Le compteur peut être alimenté par un courant continu de 12 V, comme celui fourni par une batterie de voiture automobile, au cas où le courant secteur ferait défaut. Cet avantage permet d'utiliser l'instrument pour des travaux de mesure précis sur le terrain.

La série Redhill d'instruments de comptage complets à été réalisée parallèlement aux instruments modulaires PX produits par Panax et elle a été conçue principalement à l'intention des utilisateurs ayant besoin de matériel précis à bas prix, sans les possibilités infinies d'extension qu'offrent les systèmes modulaires.

EE 91 772 *pour plus amples renseignements*

GÉNÉRATEUR D'IMPULSIONS PROGRAMMABLE

Digital Equipment Corporation (UK) Ltd, 11 Castle Street, Reading, Berkshire

(Illustration à la page 195)

Ce générateur d'impulsions programmable, présenté par la Digital Equipment Corporation (UK) Ltd, a été conçu pour le contrôle de composants de mémoire de ferrite en laboratoire et sur chaîne de montage.

Le PPG 2120 est un système pour montage sur table assurant des programmes de contrôle de 4, 8 ou 16 plots pour noyau et film fonctionnant à des vitesses entre 10 kHz et 5 MHz.

Le 2120 comprend six canaux de sortie alimentés par excitateurs de courant à grande vitesse, avec réglages individuels de largeur, de forme et de position d'impulsions. On peut faire varier la largeur d'impulsions de 0 à 10 μsec, et chaque impulsion peut être retardée après le début d'un programme de 10 μsec au maximum.

Le nouveau système comprend deux canaux de répétition indépendants, permettant à l'utilisateur de choisir des répétitions de plots sur n'importe quel programme et des répétitions de paires pour n'importe quelle paire. Les intervalles de répétition sont à variation continue de 200 nsec à 300 msec.

Des impulsions de synchronisation pour déclencher un oscilloscope peuvent être obtenues à n'importe quel plot. Ces impulsions peuvent être retardées de 10 μsec au maximum.

Les commandes de fonctionnement et les commutateurs sont montés sur le panneau frontal, et le coffret peut être ouvert et donner accès par l'avant au composants internes.

EE 91 773 *pour plus amples renseignements*

RADIOTÉLÉPHONE

Radio Communications Co., 16 Abbey Street, Crewkerne, Somerset

(Illustration à la page 195)

Cet appareil a été spécialement conçu pour répondre à la demande d'un instrument à faible consommation de courant pouvant être utilisé comme station mobile ou station fixe dans des lieux isolés et dépourvus d'une alimentation secteur.

Il a été étudié pour fonctionner sur batterie de 12 V et ses sections émetteur et récepteur sont entièrement munies de transistors. Le courant d'attente n'est que de 26 mA, cette valeur montant à 1½ A durant l'émission. La sortie de puissance est d'environ 5 W. La sortie de puissance du récepteur est de 1 W. Le TRT/6 est prévu pour une à 12 voies de communication radiotéléphonique très rapprochées, dans la bande de 40 à 100 MHz, pour un espacement de voie de 25 kHz à 50 kHz.

Lorsqu'il est utilisé avec une batterie chargée par cellule solaire ou par chargeur à vent, l'appareil peut être employé pendant de longs mois à une station de base sans surveillance aucune.

Le TRT/6 est normalement fourni avec un microphone à poignée en matière plastique ou un combiné de poids léger et il fonctionne sur un système à antenne externe de 75 Ω qui peut être du type directionnel pour fournir une portée de communication plus étendue.

Pour le fonctionnement à répétition, une version modifiée du TRT/6 peut être fournie, à savoir le TRT/6R.

Il existe également une version portative, type TRT/4, qui comprend une antenne télescopique et une batterie rechargeable intérieure scellée de 28 V.

EE 91 774 *pour plus amples renseignements*

Résumés des Principaux Articles

La réalisation d'atténuateurs et de niveleurs coaxiaux à diodes P-I-N pour la gamme de 1-2GHz par J. R. James et M.H.N. Potok

Résumé de l'article aux pages 142 à 146

On a pu réaliser un atténuateur coaxial à diodes P-I-N à large bande au moyen de calculs basés sur une spécification effective de diodes. L'auteur donne des détails de la réalisation et indique les facteurs limitant le comportement. Le comportement mesuré est en étroit accord avec le comportement calculé et se compare très favorablement au dispositifs commerciaux à large bande et à accord mécanique.

Une version simplifiée de cet atténuateur a été utilisée pour niveler la sortie d'un oscillateur de balayage. Le réseau à réaction et le comportement du niveleur sont analysés en détail.

Circuits de lecture binaires à tubes inducteurs numériques à transistors et à décharge gazeuse par P. R. Adby

Résumé de l'article aux pages 147 à 151

Cet article décrit le dispositif de lecture de compteurs binaires dont la matrice de décodage à diodes classiques a été simplifiée par la commutation combinée de la base et de l'émetteur des transistors de sortie. La matrice qui en résulte ne comprend que neuf diodes et permet une réduction considérable des frais de composants et de production. Cette simplification peut être accrue en remplaçant les diodes par des résistances à faible tolérance. Un circuit à mémoire comprenant des circuits bistables de porte et de transfert est également décrit.

Une torche à plasma microondes de 2,5kW par J. Swift

Résumé de l'article aux pages 152 à 154

Il s'agit ici d'une torche à microondes d'un fonctionnement simple constituant une source sûre de plasma pour la recherche. La torche fonctionne à 2,45GHz et le courant prévu est de 2,5kW.

Equipement périphérique pour une petite calculatrice numérique par A. D. Booth et J. M. S. de Vries

Résumé de l'article aux pages 155 à 163

Dans un article précédent, les auteurs ont décrit une petite calculatrice numérique transistorisée. Ils décrivent maintenant l'équipement périphérique de cette calculatrice. Cet appareillage se compose d'un lecteur à bande de papier photoélectrique, d'une perforatrice de bande de papier, d'une machine à écrire et des circuits de commande connexes.

Un compteur d'éclairs de coups de foudre par R. H. Golde

Résumé de l'article aux pages 164 à 166

Cet article décrit un compteur d'éclairs de coups de foudre adopté par le comité d'études du C.I.G.R.E. pour l'usage international en vue de déterminer la fréquence des éclairs de foudre sur la terre.

Equipement de mesure à réponse de fréquence dynamique par G. R. J. Baldwin

Résumé de l'article aux pages 167 à 171

Cet article décrit un équipement pouvant mesurer les réponses de phase et d'amplitude de dispositifs actifs et passifs dans des conditions d'impulsions H.F. avec modulation de fréquence très rapide.

La précision de mesure est de $\pm 1°$ en phase (dans une gamme de 100°) et de $\pm 0,1$dB en amplitude (dans une gamme de 10dB) pour des variations de phase et d'amplitude à des taux allant jusqu'à 2,5MHz, avec des taux de modulation de fréquence allant jusqu'à plusieurs MHz par microseconde.

L'équipement a été conçu pour l'utilisation dans les bandes H.F.de 400MHz à 10.000MHz, avec têtes H.F. interchangeables pour les différentes bandes. La largeur d'impulsion et le taux de répétition sont variables dans une gamme étendue.

Le comptage sans retenue jusqu'à 1.099.508.482.050 par A. M. Andrews

Résumé de l'article aux pages 172 à 175

Les résultats suivants ont été contrôlés sur un registre à décalage binaire à réaction. (a) Pour une longueur de cycle de (2^n-1), la réaction de la somme du module-2 doit provenir d'un nombre pair d'éléments. (b) Le même dessin de connexions donne cette longueur de cycle, que la réaction soit de la somme du module-2 ou qu'elle soit l'inverse. (c) On peut obtenir des cycles de la longueur (2^n-2) aussi bien que des cycles de la longueur (2^n-1). Dans le cas de $n=20$, les connexions de réaction sont spécifiées tant pour les longueurs de cycle (2^n-1) que pour les longeuurs (2^n-2). Deux registres à 20 éléments comportant ces dessins de connexions et alimentés en parallèle par impulsions de décalage ont un cycle combiné de la longueur spécifiée dans le titre.

Une sonde à courant alternatif à faible capacité d'entrée et à gain unitaire par W. H. T. King

Résumé de l'article aux pages 176 à 177

Une sonde sans atténuation est nécessaire pour contrôler des signaux haute fréquence de faible niveau dans la plupart des oscilloscopes universels, en particulier les petits modèles. L'auteur décrit une sonde active simple dont le gain est de 0,98, ou supérieur à cette valeur, de 10Hz à 5MHz pour des signaux allant jusqu'à 1V efficace et dont la réponse à des impulsions à temps de montée rapide est excellente. La consommation électrique de la sonde est d'environ 1,5mA à 18V sur batterie seulement lorsqu'elle est reliée au circuit soumis au contrôle.

L'étude de blocs d'alimentation en tension continue stabilisée commandés par thyristors par A. Davies

Résumé de l'article aux pages 178 à 182

L'auteur décrit des stabilisateurs de courant utilisant un banc de transistors pour la commande précise de courant ainsi que des circuits à thyristors pour la commande approximative. Une formule dérivant de la résistance de source effective de ces stabilisateurs permet de calculer la stabilité du courant en fonction de changements de tension d'alimentation et de résistance de charge.

L'auteur décrit le circuit d'un stabilisateur-type, ainsi qu'un circuit de commande classique à fiches pouvant être utilisé avec n'importe quel stabilisateur monophasé.

NEUE AUSRÜSTUNGEN

Übersetzung der Seiten 190 bis 195

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Ultraschall-Geräte

The M.E.L. Equipment Co. Ltd. Manor Royal, Crawley, Sussex

(Abbildung Seite 190)

Die ersten Geräte eines neuen Fertigungsprogramms für Ultraschall-Geräte wurden von M.E.L. Equipment Co Ltd herausgebracht.

Das Grundgerät ist der Verstärker L.665 mit 100 W Ausgangsleitung. Die Betriebsfrequenz wird automatisch durch Rückkopplung vom Schwinger geregelt. Der Verstärker ist für Dauerbetrieb bemessen und entspricht den britischen und europäischen elektrischen Schutz- und Funkstörungsvorschriften.

Der Verstärker kann als Treiber einer 2 - Liter - Ultraschall - Reinigungswanne L.673 oder einer 4-Liter-Wanne L.674 eingesetzt werden. Beide Wannen werden aus rostfreiem Stahl hergestellt und mit Ferrit-Schwingern ausgerüstet. In diesem Programm gibt es auch einen Umschalter COS.10, der umschichtiges Treiben von zwei dieser Wannen in beliebiger Kombination durch einen Verstärker gestattet.

Das Steuergerät L.666 bildet zusammen mit dem neuen Verstärker eine vielseitige 100-W-Quelle für Ultraschallenergie. Es enthält einen Phasenregler, der bei Einsatz der Geräte als rückgekoppelte Verstärker Verwendung findet, und einen Regler für manuelles Einstellen der Oszillatorleistung, wenn die Einheiten als Generator betrieben werden.

Sehr nützlich ist der neue Amplitudenmonitor, der die tatsächliche mechanische Bewegung der Sonde in Mikrons anzeigt.

Ein angekündigter neuer Sondenschwinger L.667 hat vielseitige Anwendungsmöglichkeiten, u.a. beim Bohren, Schweissen von Kunststoffen, Verzinnen, Emulgieren und Bakterienspalten. Der neue Schwinger ist für Treiben mittels des neuen Verstärkers L.665 und Steuergerätes L.666 konstruiert. Der L.667 besteht aus einem magnetostriktiven Nickelschwinger, der fest mit einem Aluminiumbronze-Halbwellen-Geschwindigkeitstransformator verbunden ist. Die Polarisation erfolgt durch Dauermagnete, und zwei Bleizirkonat-Abtaster geben die Signale für die automatische Frequenzregelung und den Amplitudenmonitor ab.

Der Schwinger-Baustein ist teilweise von einem ölgefüllten Kühlmantel umgeben, der auf einem Gussstück sitzt, das die Wärmeabfuhr unterstützt, und wird durch einen Ventilator gekühlt.

Durch Montage des Schwingers L.667 in einen Bohrständer L.668 entsteht ein Ultraschall-Bohrer modernster Konstruktion. Angaben von M.E.L. ist zu entnehmen, dass die maximale lineare Abmessung des Querschnittes eines Schneidwerkzeugs für den älteren 60-W-Bohrer 19 mm war, die neue 100-W-Ausrüstung bei höherer Geschwindigkeit, Tiefe und Genauigkeit jedoch mindestens 38 mm bewältigen kann.

Zur kompletten Bohrmaschine gehören noch der neue Verstärker und das Steuergerät und damit die Steuer- und Amplitudenmonitoreinrichtungen.

Der Bohrständer hat eine Anzahl für industrielle Anwendung besonders sinnvolle Konstruktionsmerkmale. Säule und Grundplatte haben hohe lineare Genauigkeit und Stabilität, und die Halterung für den Schwinger lässt sich genau ausrichten. Der Schwinger-Baustein kann für Aufnahme des Werkstückes mittels eines Zahntriebes stossfrei entlang der Hauptsäule auf und ab bewegt werden. Zwischen der Schneid- und Bereitschaftsposition gleitet der Schwinger auf Präzisions-Rollenlagern; er kann über eine Schnellvorschubtaste und Stossdämpfer in die Schneidposition gebracht werden.

Die Abbildung zeigt eine Ultraschall-Maschine, die aus dem auf den 100-W-Verstärker L.665 montierten Steuergerät L.666 sowie dem Sondenschwinger L.667 auf dem Bohrständer L.668 besteht.

EE 91 751 *für weitere Einzelheiten*

Tachogenerator

Muirhead & Co. Ltd. Beckenham, Kent

(Abbildung Seite 190)

Bei Muirhead & Co. Ltd ist die Fertigung eines Stellmotor-Tachogenerators Grösse 11 angelaufen, der besonders für Verwendung in integrierenden und Geschwindigkeits-Präzisionsregelungen konstruiert ist.

Durch sorgfältige Konstruktion und Fertigung wird die Linearität der Ausgangsspannung mit der Geschwindigkeit innerhalb einer Unsicherheit von nur 0,07 Prozent gehalten; im Temperaturbereich $-15°C \ldots +75°C$ ist die maximale Änderung der Ausgangsspannung mit Temperatur nur 0,25 Prozent. Durch Anwendung der "umgekehrten" Tachometerkonstruktion wurde die Ausgangsspannung auf 2,75 V je 1 000 UPM erhöht, d.h. etwa fünfmal höher als die der früheren Tachometer der Grösse 11.

Viele alternative Motorwicklungen sind lieferbar, um Kundenanforderungen zu genügen.

EE 91 752 *für weitere Einzelheiten*

Such-Filtergerät

Associated Engineering Ltd, 60 Kenilworth Road, Leamington Spa, Warwickshire

(Abbildung Seite 190)

Associated Engineering Ltd kann jetzt das neue Such-Filtergerät DF/2 liefern, das für kontaktloses Messen von Wegen oder Schwingungen mittels induktiver A.E.-Weggeber in gepaarter oder Doppelausführung unter schwierigen Umgebungsbedingungen ausgelegt ist.

Das Gerät gestattet genaues Messen in einem breiten Anwendungsgebiet, einschliesslich der Bewegung von Lagerlaufflächen, Ölfilmdicke und Lagerabnutzungsraten, von komplexen Schwingungseffekten in empfindlichen Strukturen, deren Körper nicht durch Messeinrichtungen belastet werden darf, sowie Messungen an Werkzeugmaschinen und Vorrichtungen mit metrologischer Genauigkeit.

Mit dem neuen DF/2 lassen sich Wege und Schwingungen bis zu 0,38 mm bei Frequenzen zwischen 0 und 10 kHz und —je nach dem zu messenden Werkstoff —mit Auflösung bis zu 0,25 µm messen.

Die Temperaturkonstanz des Systems

einschliesslich Messwertwandler ist besser als $\pm 0{,}05$ μm/°C bei periodischen Schwankungen von Umgebungstemperaturen bis zu 150° C. Für statische Wegmessungen ist die Langzeitkonstanz des Systems für Eisenmetalle über eine 8stündige Periode besser als $\pm 2{,}5$ μm.

EE 91 753 *für weitere Einzelheiten*

Rechteckpotentiometer

Reliance Controls Ltd, Sutherland Road, London, E.17

(Abbildung Seite 191)

Die laufende Fertigung des neuen Rechteck-Trimmerpotentiometers Baumuster CW.90 wurde bei Reliance Controls Ltd aufgenommen. Durch einen Diallylphthalat-Pressling, keramischen Wickelkörper, eine Spindel aus rostfreiem Stahl und geschweisste Anschlüsse wird hohe Zuverlässigkeit erzielt. Diese Bauelemente sind abmessungsmässig gegen seitherige Reliance und andere Modelle austauschbar. Die Anschlüsse sind für ein Raster von 2,54 mm ausgelegt, der Widerstandsbereich ist 10 Ω bis 50 kΩ und der Temperaturbereich −40° C... +100° C. Die Standardelemente sind mit Karma-Draht gewickelt, was über den grössten Teil des Temperaturbereiches einen Beiwert von 20 × 10^{-6} ergibt.

Eine hermetisch dichte Ausführung geht in Kürze in die Fertigung.

EE 91 754 *für weitere Einzelheiten*

Koaxialantennen-Umschaltrelais

Magnetic Devices Ltd, Exning Road, Newmarket, Suffolk

(Abbildung Seite 191)

Magnetic Devices Ltd liefert nunmehr ein neues Koaxialantennen-Umschaltrelais, dessen Konstruktion Koaxialstecker und -buchsen im Relaiskörper überflüssig macht und somit Abmessungen und Kosten reduziert. Das gleichstrombetätigte Relais Baumuster 951 ist für Antennenumschalten bei Frequenzen von etwa 450 MHz ausgelegt und hat ausserdem überall dort Anwendungsmöglichkeiten, wo ungewöhnlich niedrige Kapazität zwischen den Kontakten erforderlich ist.

Das mit einem Umschaltkontaktsatz bestückte schnelle Relais ist in ein Messinggehäuse eingebaut und für Verwendung mit Kabel Typ UR 43 konstruiert, das leicht eingelötet werden kann, ohne die Relaisjustierung zu stören, und mit einer Messingschelle festgeklemmt wird. Jede der Schellen stellt ausserdem die Verbindung zwischen Abschirmung und Kabelausgang in jeder beliebigen von drei Richtungen her.

Relais des Baumusters 951 haben eine maximale Betriebsspannung von 100 V−, maximale Spulenbelastung von 1½ W und einen maximalen Spulenwiderstand von 9350 Ω. Sie werden zwischen allen Klemmen und Masse sowie zwischen Spule und Masse mit 1 kV_{eff} getestet; die Nennfrequenz ist 450 MHz, das Stehwellenverhältnis 1,1:1 und Nebensprechen 39 dB.

Die Ansprechzeit ist 20 ms, die Abfallzeit 8 ms: die Kontakte sind für 1 A oder 30 W max. bemessen, und die Nennimpedanz ist 50 Ω. Das Baumuster 951 ist auf Wunsch für den Temperaturbereich −40° C...+100° C lieferbar, versilbert und wiegt etwa 100 g.

EE 91 755 *für weitere Einzelheiten*

Kleinladegerät

Kynmore Engineering Co. Ltd. 19 Buckingham Street, London, W.C.2

(Abbildung Seite 191)

Kleinladegeräte für das Aufladen von Nickel-Kadmium-Akkumulatoren mit gleichbleibendem Strom hat die Kynmore Engineering Co Ltd in einem neuen Sortiment herausgebracht. Die neuen Ladermoduln sind nicht grösser als die Akkumulatoren und haben einen so niedrigen Sperrstrom, dass sie dauernd an ihre Akkus angeschlossen bleiben können.

Der Modul besteht aus einem röhrenförmigen Aluminiumgehäuse, das einen Kondensator, Widerstände und eine Halbleiter-Gleichrichteranordnung in Araldite verkapselt enthält, und aus dem drei farbgekennzeichnete, 64 mm lange Anschlussdrähte hervorragen. Golddraht-Dioden richten den Speisestrom gleich und gewährleisten, dass der Sperrstrom praktisch Null ist. In Betrieb wird dem Wechselstromnetz zweimal das Äquivalent des Effektivladestroms entnommen, und der grösste Teil der Impedanz tritt dabei am Kondensator auf. Daher bleibt die Impedanz —und damit auch der Ladestrom—im wesentlichen unabhängig vom Ladezustand des Akkus und konstant. Deshalb sind die Moduln auch für das Laden selbst völlig entladener Akkus geeignet, und externes Glätten ist überflüssig.

Im Standardprogramm gibt es sechs verschiedene Grössen, die bei Anschluss an ein 230V∼ -Netz für 2 bis 45 mA bemessen sind. Alle Modelle können bis zu 10 Zellen gleichzeitig laden. Die Abmessungen der Moduln schwanken bei 20,3 mm Länge zwischen 15,2 mm und 32 mm Durchmesser.

EE 91 756 *für weitere Einzelheiten*

Umgebungstestkammer

Medical Electronics Ltd, Culford Works, King Henry's Walk, London, N.1

(Abbildung Seite 191)

Die tragbare Umgebungstestkammer ETC 505 ist für das Temperaturtesten von Dioden, Transformatoren und anderen kleinen Bauelementen geeignet.

Temperaturen können entweder eingestellt oder innerhalb vorgewählter Grenzen mit vorgewählter Geschwindigkeit automatisch wiederholt werden.

Die Arbeitskammer ist aus rostfreiem Stahl gefertigt, und an der Tür können elektrische Zuleitungen vorgesehen werden, die dynamisches Testen von Ausrüstungen ermöglichen.

Beheizung erfolgt mittels Druckluft über ummantelte Heizer. Die Kühlung erfolgt durch Dispersion von flüssigem CO_2 in die Druckluft, mit der es sich schnell mischt, wodurch zunehmender Frost vermieden wird.

Einstellung der Heiz- und Kühlraten erfolgt über ein Bedienelement auf der Frontplatte. Der Intervallschalter lässt sich auch auf Feinregelung der Temperatur einstellen: die Temperatur kann auf einem eindeutig ablesbaren Temperaturanzeiger und -regler mit zwei unabhängigen Regelpunkten eingestellt werden.

Die Innenabmessungen der Kammer sind 229 × 178 × 203 mm. Der Temperaturbereich ist −70° C...+200° C, die Kühlrate 5° C/min und die Heizrate 10° C/min.

EE 91 757 *für weitere Einzelheiten*

Digitalrechner

Hawker Siddeley Dynamics Ltd, Whitley Works, Nr. Coventry, Warwickshire

(Abbildung Seite 192)

Der IAD-Digitalrechner DCC.1 ist zur Deckung des Bedarfs an einer preiswerten, anpassungsfähigen Anlage ausgelegt, die arithmetische und logische Operationen durchführen kann und sich zum Einbau in Datenerfassungs- und Datenverarbeitungssysteme eignet. Weitgehende Anwendung der Mikroprogrammtechnik reduziert die Anzahl der erforderlichen Bausteine und erleichtert das Programmieren.

Boolesche Zustände in Binärform oder arithmetische Information in binärcodierter Dezimalschreibweise können mittels 8-Kanal-Lochstreifen eingegeben werden; im prozessgekoppelten Betrieb kann man die Information über Analog-Digitalwandler oder direkt von Messwertgebern mit Digitalausgang zuführen.

In einem IAD-Datenerfassungs- und Registriersystem führt die Anlage unter Kontrolle des Programmiergerätes die Verarbeitung der vom Rest des Systems zugeführten Daten durch, und zwar entweder wie vorprogrammiert oder auf Verlangen.

Information wird in einer Ferritkernmatrix mit bis zu 1024 Zellen gespeichert, die sechs Bits aufnehmen. Jede Zelle kann entweder eine Dezimalziffer mit Vorzeichen und Booleschem Zustand oder ein Programmsymbol speichern.

Der Rechnerausgang kann in Druckwerke, Digital-Analogwandler oder andere Wandler, die Steuerung einer Betriebseinrichtung bewirken können, gespeist werden.

Die Anlage ist in Steckkartentechnik gebaut, und die Schaltungen sind grösstenteils 2-MHz-Logikschaltungen.

Der Rechner besteht aus drei Hauptteilen:

(1) Register und Addierwerk

(2) Mikroprogramm

(3) Ziffernspeicher

Die Register dienen als Zwischenspeicher für Information, deren Verarbeitung durch das Mikroprogramm geregelt wird.

Das durch einen örtlichen Befehlszähler gesteuerte Mikroprogramm regelt die Übertragung von Information und Befehlen zwischen Register, Addierwerk und Ziffernspeicher. Das Mikroprogramm besteht aus einer Diodenmatrix, die sich ohne Schwierigkeiten zur Erstellung des für das benutzte System am besten geeigneten Rechnerbefehlscodes auswechseln lässt.

Der Ziffernspeicher ist ein Ferritkernspeicher mit direktem Zugriff, 6 μs Zykluszeit und eingebautem Nest. Information wird im Nest auf der Basis "letzte ein—erste aus" gespeichert.

EE 91 758 *für weitere Einzelheiten*

Zeitschalter

Solid State Controls Ltd, 30-40 Dalling Road, London, W.6

(*Abbildung Seite 192*)

Der Cyclo-Set besteht aus einem Normal—dem auch getrennt lieferbaren Cyclo-Set-Kleinstmodul— mit einem 8poligen Einsteckrelais für 12 V— und robustem, gepressten Stahlchassis. In einer anderen Ausführung ist der Zeitschalter völlig umkapselt in einem Spritzgusskästchen lieferbar. Der Kleinstmodul hat eine erwartete Lebensdauer von bis zu 100 Millionen Vorgängen, ist volltransistorisiert, ohne bewegliche Teile gebaut und braucht während der ganzen Lebensdauer keinerlei Justierung. Das abgeschlossene Relais hat einen einpoligen Umschalterausgang, der entweder für 5 oder 10 A bei 250 V~ bemessen werden kann und für rein ohmsche sowie induktive Lasten geeignet ist.

Der zweistufige Zeitgeber hat zwei voll regelbare "Ein-" und "Aus-"Zeitperioden: 0,4 s...4 min als Standard und bis 20 min für Spezialaufgaben.

Der Zeitschalter ist für sechs Wechselspannungen von 12...415 V~ und die beiden Gleichspannungen 12 und 24 V— lieferbar. Ausserdem ist der Baustein sowohl über Ruhe- wie auch Arbeitskontakte steuerbar. Der Cyclo-Set ist von Netzschwankungen bis zu ±10 Prozent und Änderungen der Umgebungstemperatur zwischen −20° und +55° C praktisch unbeeinflusst.

EE 91 759 *für weitere Einzelheiten*

Dreifarben-Signallampen

Thorn Special Products Ltd, Great Cambridge Road, Enfield, Middlesex

(*Abbildung Seite 192*)

Die Thorn-Dreifarbensignale werden nunmehr mit neuen internen Glasfiltern hergestellt, die die gewählte Farbe durch die stirnseitige Einzelglaslinse projizieren.

2 W, 28 V können laufend benutzt werden, was die Lichtausbeute wesentlich erhöht. Ausserdem gestattet die Verwendung von Glas für Linsen engere Kontrolle der Farbkoordinaten, so dass wenig Gefahr besteht, dass Farben verblassen. Standardfarben sind Signalrot, Signalgrün und Signalgelb nach der britischen Norm BS.1376.

Die Gesamtlänge der Signallampe ist etwa 57 mm, ihr Durchmesser 20,6 mm, und sie ist für Montage auf bis zu 3,2 mm starke Frontplatten geeignet, die ein 15-mm-Loch für Montage haben müssen.

Die Dreifarben-Signallampe nimmt nur sehr wenig Frontplattenfläche ein und ersetzt oft drei getrennte Signallampen. Sie wird mit Klemmutter und Klemmring montiert und ist ohne Ausbauen für Service zugänglich.

Der Baustein kann auf Wunsch komplett mit drei Atlas-Zwerglampen für 6, 12 oder 28 V Nennspannung geliefert werden.

EE 91 760 *für weitere Einzelheiten*

Miniaturpotentiometer

Miniature Electronic Components Ltd, Copse Road, St. Johns, Woking, Surrey

(*Abbildung Seite 192*)

Miniature Electronic Components Ltd hat ihr Programm von Draht-Trimmerpotentiometern durch den Typ MT20 ergänzt. Das in ein Transistorgehäuse TO-5 eingebaute Bauelement eignet sich für Montage auf gedruckte Schaltungen und ist in Standardwerten von 50, 100, 200, 500, 1000, 2000, 5000 und 10 000 Ω lieferbar. Die Entwicklung wurde speziell auf ein einfaches, zuverlässiges Bauelement für die kommerzielle und industrielle Elektronik sowie Anwendungsgebiete, in denen hohe Qualität und mässige Kosten massgebend sind, abgestellt. Ein Temperaturbereich von −20° C...+125° C, Belastbarkeit 1 W bei 70° C, Einregelung über eine Umdrehung mit festem Endanschlag, sowie feuchtigkeits- und staubfeste Ausführung sind die wichtigeren Eigenschaften.

EE 91 761 *für weitere Einzelheiten*

Brücken-Oszillator-Demodulator

H. Tinsley & Co. Ltd, Werndee Hall, South Norwood, London, S.E.25

(*Abbildung Seite 193*)

Das Tinsley-Brückengerät 5720 besteht aus einem Rechteckwellengenerator mit zwei isolierten und symmetrischen Ausgängen von je 2,5 V und einem mit dem Generator synchronisierten Demodulator.

Es ist für Dehnstreifen-Messbrücken und Widerstandsthermometer-Messbrücken, die eine Generator-Demodulatorkombination erfordern, geeignet.

Die Empfindlichkeit des Demodulators, dessen Ausgang ein 76-mm-Anzeigegerät ansteuert, wird mittels eines Abschwächers geregelt. Die Empfindlichkeit kann so eingeregelt werden, dass jede Teilung des Messgerätes je nach Einsatz einem bestimmten Dehnungswert oder einer bestimmten Temperatur entspricht.

Das Gerät enthält eine vom Netz aufladbare Batterie.

EE 91 762 *für weitere Einzelheiten*

Milliwattmeter

Standard Telephones & Cables Ltd, Testing Apparatus Division, Corporation Road, Newport, Monmouthshire

(*Abbildung Seite 193*)

Neukonstruierte Ausführungen des Milliwatt-Testgerätes 74166-A wurden von Standard Telephones & Cables Ltd bekanntgegeben. Die gleichen Testeinrichtungen nehmen in einem neugestalteten Metallgehäuse nur das halbe Volumen des 74166-A ein. Der bisherige Drehschalter wurde durch Drucktastenschalter ersetzt und das Messgerät hinter der Frontplatte positioniert, um es besser zu schützen und dem Gerät ausser der grösseren Tragbarkeit eine geschmackvollere Gestaltung zu geben.

Mit Ausnahme der Eingangssteckverbindungen sind die beiden Modelle (H und J) identisch: Modell 74166-H ist mit Koaxialsteckern No. 1 der britischen Post, Modell 74166-J mit BNC-Steckverbindungen (Amphenol 31-221) ausgerüstet. Sie sind für genaue, abgeschlossene Pegelmessungen von +1 dB bis zu −1 dB bezogen auf 1 mW an unsymmetrischen 75-Ω-Schaltungen bei Frequenzen bis zu 30 MHz ausgelegt und können ausserdem Heizerspannungen von 5,6...7,0 V messen.

Ein eingebautes Thermoelement wird mit einer durch Nebenschluss auf 75 Ω reduzierten Heizerimpedanz benutzt, und die Messgeräte können daher nicht zum Messen von "Durchgangspegeln" eingesetzt werden. Die eingebaute Weston-Miniatur-Normalzelle, gegen die die Schaltung durch eine Serie einfacher Schalt- und Abgleichoperationen geeicht wird, gewährleistet hohe Messgenauigkeit. Ausserdem kann die Eichschaltung eine Gleichstromleistung von 1 mW zur Eichung anderer Apparate in eine externe 75-Ω-Schaltung speisen.

Beide Modelle werden aus drei, im Gehäuse untergebrachten 1,5-V-Trockenbatterien betrieben. Die Aussenabmessungen des Gerätes sind mit Deckel 311 × 203 × 181 mm.

EE 91 763 *für weitere Einzelheiten*

Beschleunigungsgeber

Environmental Equipments Ltd, Station Yard, Wokingham, Berkshire

(Abbildung Seite 193)

Die mechanischen Teile des Beschleunigungsgebers Baumuster AQ sind durchweg aus rostfreiem Stahl gebaut. Für den Quarz wurden Klebstoffe vermieden, um die durch Nachlassen der Haftfähigkeit und Änderungen der Steife bei hohen Temperaturen auftretenden Probleme zu vermeiden.

Die maximale Nenntemperatur ist 260° C, und Abweichungen der Ladungs- und Spannungsempfindlichkeit liegen bis zu −65° C unter 5 Prozent und bis zu +160° C unter 15 Prozent. Die maximale Querempfindlichkeit ist 5 Prozent. Die interne Kapazität beträgt 1200 pF, die des 3m langen, rauscharmen Kabels 300 pF.

Lieferbar sind vier Modelle, die mit AQ5, AQ10, AQ20 und AQ40 bezeichnet werden. Die Modellnummern geben gleichzeitig die Nennladungsempfindlichkeit in pC_s/g_s.

EE 91 764 *für weitere Einzelheiten*

Miniatur-Hebelschalter

Keyswitch Relays Ltd, Cricklewood Lane, London, N.W.2

(Abbildung Seite 193)

Für Anwendung in der Schwachstromtechnik liefert Keyswitch Relays Ltd eine Auswahl von Miniatur-Kippumschaltern Typ P.O.1000 mit einer mechanischen Lebensdauer von 750 000 Betätigungen je Schaltzyklus. Die Kontaktbetätigung bewirkt eine Umschaltung und ist in Rast-, Tast- und verriegelbarer Ausführung lieferbar, bestückt mit maximal sechzehn Umschaltern (acht per Hebelstellung). Silber-Doppelkontakte sind Normalausführung; die Baureihe ergänzt die 1964 von Keyswitch eingeführten Kippschalter für acht Umschaltungen.

Kontakte sind für 100 V bei 300 mA−, 12 V bei 1 A− und 100 V bei 1 A∼ mit Anfangskontaktdruck von mindestens 15 g (mindestens 10 g nach 750 000 Betätigungen) lieferbar.

Jeder Schalter wird mit zwei Befestigungsschrauben geliefert, und Griffe kommen zur Anpassung an die Bedienfeldgestaltung in rot, grün, gelb, grau, elfenbein und schwarz.

Die Aussenabmessungen sind 70,6 mm lang, 55,6 mm tief einschliesslich Griff und 15,7 mm breit.

EE 91 765 *für weitere Einzelheiten*

Tragbare Dickenmesser

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Abbildung Seite 193)

Dawe Instruments Ltd kann nunmehr zwei neue Ausführungen des tragbaren Ultraschall-Dickenmessers Sonaray 1805 liefern, die mit 1805B und 1805C bezeichnet werden. Sie lösen den seitherigen Typ 1805A ab und genügen einem bedeutend verbesserten Pflichtenblatt.

Typ 1805B, das einfachere der beiden neuen Modelle, ist in Aussehen, Arbeitsweise und Einrichtungen dem alten Typ 1805A ähnlich, jedoch mit folgenden Verbesserungen: Da das Instrument hauptsächlich für den Aussendienst bestimmt ist, wurde die Energieaufnahme so reduziert, dass eine Ladung für über fünf Stunden Dauerbetrieb ausreicht. Das Instrument kann aber auch ans Netz angeschlossen werden. Genauigkeit und Auflösungsvermögen sind mindestens genau so gut wie die der meisten in der Fertigung eingesetzten Dickenmesser.

Abmessungen (381 × 241 × 114 mm) und Gewicht (7,25 kg) sowie die allgemeine Anordnung sind dieselben wie die des Types 1805A, doch ist das neue Modell mit Silizium-Transistoren statt der bisherigen Germanium-Typen bestückt. Dadurch wurden erhöhte Testtiefe (bis zu 11,43 m Stahl), erhöhter Frequenzbereich (1...8 MHz mit weiterer nutzbarer Leistung bis zu 15 MHz) und erhöhter Temperaturbereich (−23° C...+65° C) erzielt. Die Impulsfrequenz ist 250 Hz oder 1000 Hz, und die Echos werden weiterhin auf dem Rechteckschirm einer Oszillografenröhre mit einer Nutzfläche (Raster 5 × 5 cm), die dem viel grösserer, runder Röhren entspricht, aufgezeichnet. Wie seither kann die Skala gedehnt werden, bis die volle Breite 12,7 mm dicken Stahl darstellt. Die Messunsicherheit im Bereich von 2,5...152,5 mm Dicke ist wieder ±0,127 mm.

Grundsätzlich ist der neue Typ 1805C derselbe wie Typ 1805B, ist aber mit einer Reihe zusätzlicher Einrichtungen ausgerüstet, die Einsatz in der automatischen Rissprüfung, Dickenmessung, Registrierung und ähnlichen Anwendungsgebieten gestatten, die bisher von grösseren, aber nicht immer genaueren Geräten beherrscht wurden.

Die im 1805C vorhandenen zusätzlichen Einrichtungen sind ein eingebauter Abschwächer von 0 bis 62 dB in 2-dB-Schritten, ein eingebauter Rissalarm und die Möglichkeit, weitere Hilfseinrichtungen anzuschliessen. Die modulare Bauweise gestattet Nachrüsten des Types 1805B auf den Typ 1805C sowie Wahl besonders geeigneter Hilfsausrüstungen. Zum noch nicht kompletten Sortiment der Hilfsausrüstungen gehören ein Digitalmikrometer, Zeitanaloge und Universal-Rissmonitoren, alle als Standardmoduln in Festkörpertechnik lieferbar. Jeweils zwei beliebige dieser Hilfsausrüstungen können in ein mit dem Sonoray gekoppeltes Hilfsgehäuse eingeschoben werden.

In allen Fällen müssen die Wandler der benutzten Frequenz angepasst werden; sie gehören üblicherweise zum piezoelektrischen Hochleistungstyp Z. Ein 1,80 m langes Kupplungskabel für den Wandler ist Standard, man kann jedoch Kabel bis zu 61 m Länge ohne Änderung der Instrumente verwenden. Für den Aussendienst sind für beide Modelle geeignete Tragrahmen, Segeltuchtaschen und Ledertaschen lieferbar.

EE 91 766 *für weitere Einzelheiten*

Oszillograf

Ribet-Desjardins, 13/17 Rue Perier, Montrouce (Seine), Frankreich

(Abbildung Seite 194)

Der neue Oszillograf ist volltransistorisiert und in zwei Ausführungen lieferbar: Typ 340A ist ein tragbares Gerät für Tischanwendung, Typ 340AR ist für Gestelleinbau.

Vertikalverstärker in Einschubtechnik gibt es für ein oder zwei Kanäle. Bei Anwendung der grössten Bandbreite (0... 100 MHz) ist der Ablenkfaktor 50 mV/cm mit 3,5 ns Anstiegszeit. Bei 0...60 MHz ist der Ablenkfaktor 5 mV/cm.

Einzel- oder Doppelablenkgeräte sind mit Geschwindigkeiten von 5 s/cm bis zu 100 μs/cm und 10x Dehnung lieferbar. Synchronisierung ist bis zu 200 MHz möglich.

Der Horizontalverstärker für Einsatz mit dem Doppelzeitablenkgerät hat eine Bandbreite von 0...2 MHz mit 0,1 V/cm Höchstempfindlichkeit; mit dem Einzelzeitablenkgerät ist die Bandbreite 0...3 MHz mit einem Ablenkfaktor von 20 mV/cm.

EE 91 767 *für weitere Einzelheiten*

Halbleitertester

Rohde & Schwarz, 15 Mühldorferstrasse, 8000 München 8, Deutschland

(Abbildung Seite 194)

Der Halbleiter BMA wurde für schnelle Messung und Kontrolle der wichtigsten Parameter von Halbleitern entwickelt. Er kann mittels Typenkarten oder Lochstreifen programmiert und mit Hilfe eines Modulsystems ausgebaut werden. Die maximale Messfolge ist 0,1 s/Parameter. Die vom Tester in Form von "Gut-Schlecht"-Entscheidungen oder als Digitalinformation abgegebenen Daten werden in den Datenerfassungs- und Verarbeitungsgeräten der Serie UC ausgewertet. Der neue Halbleitertester ist besonders für die Fertigungsprüfung, Wareneingangskontrolle und Zuverlässigkeitsprüfungen beim Kunden geeignet. Das neue System UC für Datenerfassung und Verarbeitung ist nach dem Baukastenprinzip konstruiert und lässt sich den verschiedensten Anforderungen anpassen. Es wertet die Messreihen im Sinne der mathematischen und technischen Statistik (statistische Qualitätskontrolle) aus, klassiert digitale Daten und speichert die absoluten Frequenzen, die ausgedruckt werden können. Ein Bildschirm zeigt die Verteilungskurve—entweder unverzerrt oder für das Wahrscheinlichkeitsnomogramm verzerrt—die ein X-Y-Schreiber aufzeichnet.

EE 91 768 *für weitere Einzelheiten*

X-Y-Leser

P.C.D. Southern Ltd. 4 Blackwater Station Estate. Camberley, Surrey

Der neue X-Y-Leser Typ ZAE-1 bietet Anwendungsmöglichkeiten in der Kurvenblattanalyse, die bisher nur in viel komplizierteren und teureren Anlagen zur Verfügung standen. Bemerkenswert ist die Genauigkeit der halbautomatischen inkrementalen Verschiebung auf der X-Achse, die zusammen mit dem einfach zu betätigenden Visier das Tabellarisieren von X-Y-Ordinaten, momentaner Steigung der Kurve und Fläche unter der Kurve mit wesentlich geringerer Schulung der Bedienenden als bisher gestattet. Sowohl lineare wie nichtlineare Funktionen können verarbeitet werden: der Läufer wird einfach über die Kurve gestellt und der Hebel "Lesen" zur Einleitung des Ausdruckens betätigt.

Die Aussenabmessungen des Lesers sind 107 × 61 × 24 cm, die Lesefläche ist in der X-Achse 30 cm, in der Y-Achse 50 cm lang. Die Kurvenbahn ist in leicht zugänglichen Rollen untergebracht, die durch einfache manuelle Bedienelemente betätigt werden. Üblicherweise weise sind die unverarbeiteten Daten in der rechten Rolle aufgewickelt und werden in Abschnitten von 50 cm nach der linken Rolle vorgeschoben. Auf diese Weise lassen sich Aufzeichnungen bis zu 60 m Länge verarbeiten.

Kurven und grafische Darstellungen auf Einzelblättern werden durch dünne magnetische Streifen festgehalten, die auch zum Flachhalten der Kanten langer Bahnen Verwendung finden können. Bei mehrkanaligen Aufzeichnungen kann man den Kanalnummeranzeiger (1...12 einschliesslich) mit externen Kontakten verbinden, die anzeigen, welcher Kanal analysiert wird. Der Leser enthält die 28-V-Stromversorgung für den Anzeiger und ist so ausgelegt, dass ein optischer Projektor nachgerüstet werden kann, mit dem Filmaufzeichnungen auf einen Schirm projiziert werden, der den Standard-Lesetisch ersetzt.

Die Bezugsspannungen für die X- und Y-Spannungsteiler werden dem Leser üblicherweise von aussen zugeführt. Diese Spannung soll 50 V nicht überschreiten. Der durch den Schleifer des Spannungsteilers fliessende Strom darf 50 μA nicht überschreiten, wenn auf Aufrechterhalten der Genauigkeit Wert gelegt wird.

Der Leser kann mit jedem Analog-Digitalwandler eingesetzt und direkt an die Kurvenverarbeiter- und Programmiergeräte der Southern Instruments Ltd angeschlossen werden.

EE 91 769 *für weitere Einzelheiten*

Galliumarsenidlampen

The Plessey Co. Ltd. Dielectric and Magnetic Division. Towcester. Northamptonshire

(*Abbildung Seite 194*)

Die Fertigung einer Reihe neuer Galliumarsenid-Infrarotlampen läuft im Werk Towcester der Dielectric and Magnetic Division der Plessey-Bauelementegruppe an.

Diese Kleinstlampen sind PN-Übergangselemente, die einen Lichtkegel verhältnismässig schmaler Spektralbreite im infraroten Gebiet emittieren. Sie kommen in drei Modellen: ein harzverkapselter Strahler ohne Linse, ein ähnlicher Strahler mit gepresster Harzlinse von 6,35 mm Durchmesser und ein Baustein in einem Gehäuse TO37 als Kühlkörper.

Die neuen Strahler können zusammen mit einem Silizium-Fotoelement Miniatur-Wolframlampen in vielen Ausrüstungen ersetzen; sie kombinieren kleinere Abmessungen und niedrigen Stromverbrauch mit wesentlich verbesserter Zuverlässigkeit. Anwendungsmöglichkeiten bestehen u.a. in Lochstreifenlesern, Warenautomaten, Wiege- und Fahrkartenausgabemaschinen sowie Verschlüsselungseinrichtungen. Die sehr schnelle Ansprechzeit gestattet Modulation des emittierten Lichtes mit bis zu 100 MHz, so dass diese Strahler über Sichtlinien als Nachrichtenverkehrsmittel eingesetzt werden können.

EE 91 770 *für weitere Einzelheiten*

Umgebungstestkammern

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

(*Abbildung Seite 195*)

Die Temperaturtestkammern dieses Programms werden jetzt nach verbesserten Pflichtenblättern hergestellt. Typische Kenndaten der neuen Baureihe sind:

Temperaturbereich
$-185°$ C...$+250°$ C
oder
$-75°$ C...$+250°$ C
Regelempfindlichkeit
$\pm 0{,}1°$ C
Temperaturgefälle bei beliebiger Temperatur
$\pm 0{,}5°$ C
Wärmestosseinrichtung
$+100°$ C auf $-100°$ C innerhalb 3 Minuten

Die Kammern werden nach britischen Normen und DEF-Pflichtenblättern und aus den feinsten Werkstoffen hergestellt. Gegen Aufschlag sind u.a. lieferbar: Spezialtüren, Gestelle, Stützvorrichtungen, automatische Programmsteuerungen und Mehrkanalschreiber.

Alle Kammern sind nach dem Modulsystem konstruiert, das verhältnismässig einfach Sonderwünsche bei mässigen Kosten zu berücksichtigen gestattet.

EE 91 771 *für weitere Einzelheiten*

Ratemeter

Panax Equipment Ltd. Holmethorpe Industrial Estate. Redhill, Surrey

(*Abbildung Seite 195*)

Ein neues, transistorisiertes Präzisions-Grossbereich-Ratemeter für wissenschaftliche Zwecke und Anwendung im Labor wird von Panax als Ergänzung des Redhill-Programms von nichtmodularen kerntechnischen Instrumenten angekündigt. Das kompakte, netzbetriebene Ratemeter P.7900 ist für universelles Zählen mit G-M-Rohr und Szintillationsdetektor in Laboratorien, Krankenhäusern und Kliniken geeignet. Es hat eine eingebaute Konstanthochspannungsversorgung, einen Verstärker und Diskriminator mit besonders breitem Vorspannungsbereich. Dieses Modell ist leichter und vielseitiger als das Panax-Ratemeter 5054, dessen Fertigung eingestellt wurde.

Das Gerät hat neun getrennte Zählbereiche und sieben voreingestellte Zeitkonstante. Die Zählrate wird auf einem grossen 240°-Drehspulinstrument angezeigt, und ein kleiner Lautsprecher gibt auf Wunsch eine akustische Anzeige.

Einer koaxialen Buchse kann 1 mA für Betrieb eines externen Anzeigegerätes oder Schreibers, z.B. des Panax-Flachschreibers SREC.2 oder PX-Modul REC.3, entnommen werden.

Das Ratemeter lässt sich auf Wunsch auch aus einer 12-V-Stromversorgung—z.B. einer Fahrzeugbatterie—speisen, wenn kein Netzanschluss vorhanden ist. Dadurch eignet sich das Instrument vor allem für genaue Messungen im Aussendienst.

Die Redhill-Serie kompletter Zählgeräte wurde parallel mit der von Panax hergestellten Modul-Instrumentenserie PX entwickelt und ist hauptsächlich für Anwender gedacht, die preisgünstige, genaue Geräte ohne die in der Modul-Bauweise gebotene unendliche Ausbaufähigkeit benötigen.

EE 91 772 *für weitere Einzelheiten*

Programmierbarer Impulsgeber

Digital Equipment Corporation (UK) Ltd. 11 Castle Street, Reading, Berkshire

(*Abbildung Seite 195*)

Dieser von Digital Equipment Corporation (UK) Ltd eingeführte programmierbare Impulsgeber wurde für das Testen von Ferritspeicherelementen in Labor und Fertigung entwickelt.

Der PPG 2120 ist ein Tischgerät und bietet Programme zum Testen von Kernen und Filmen mit Arbeitsgeschwindigkeiten zwischen 10 kHz und 5 MHz in 4, 8 oder 16 Schritten.

Der 2120 hat sechs Ausgangskanäle, die durch schnelle Stromtreiber mit getrennt regelbarer Impulsdauer, -form und -position gespeist werden. Die Impulsdauer ist zwischen 0 und 10 μs regelbar, und jeder Impuls kann nach Start des Programmschrittes bis zu 10 μs verzögert werden.

Das neue System hat zwei getrennte Wiederholungskanäle, die dem Anwender gestatten, jeden beliebigen Programmschritt oder für beliebige gerade-ungerade Paare die Paare zu wiederholen. Wiederholungsintervalle sind zwischen 200 ns und 3 ms kontinuierlich regelbar.

In jedem Schritt können Synchronisierimpulse zum Triggern von Oszillografen eingestellt und bis zu 10 μs verzögert werden.

Bedienelemente und Schalter sind auf der Frontplatte angeordnet; das Gehäuse lässt sich so öffnen, dass interne Bauelemente von vorn zugänglich sind.

EE 91 773 *für weitere Einzelheiten*

Funkfernsprecher

Radio Communications Co., 16 Abbey Street, Crewkerne, Somerset

(*Abbildung Seite 195*)

Dieses Gerät wurde speziell zur Deckung eines Bedarfs an einer Anlage mit niedrigem Stromverbrauch für Einsatz als bewegliche oder feste Station in isolierter Umgebung ohne öffentliche Stromversorgung entwickelt.

Es ist für Betrieb aus einer 12-V-Batterie ausgelegt; sowohl Sender- wie Empfängerteil sind durchweg mit Transistoren bestückt. Der Bereitschaftsstrom ist nur 26 mA und steigt während des Sendens auf 1½ A; es wird eine Ausgangsleistung von etwa 5 W erzeugt. Der Empfänger hat eine Ausgangsleistung von 1 W.

Modell TRT/6 ist für AM-Sprechbetrieb mit 1 bis 12 Kanälen in geringem Abstand voneinander im 40..100-MHz-Band mit 25 oder 50 kHz Raster bestimmt.

Bei Einsatz mit durch Sonnenzellen oder Windgenerator geladenem Sammler ist unbemannter Betrieb einer Feststation für viele Monate möglich.

Modell TRT/6 wird üblicherweise mit Handmikrofon in plastischem Gehäuse oder leichtem Fernsprechhörer geliefert und mit einem externen 75-Ω-Antennensystem betrieben. Zur Erzielung einer grösseren Verkehrsreichweite kann man eine Richtantenne verwenden.

Für Relaisbetrieb gibt es den Typ TRT/6R, eine abgewandelte Ausführung des TRT/6.

Eine transportable Ausführung komplett mit teleskopischer Antenne und dichter, aufladbarer interner 28-V-Batterie ist unter der Bezeichnung TRT/4 lieferbar.

EE 91 774 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Kristalloszillator mit integrierter Verstärkerschaltung von A. H. James

Zusammenfassung des Beitrages auf Seite 142-146

Der Beitrag beschreibt einen Oszillator, in dem ein Quarzkristall als Serienresonanzelement in Verbindung mit einem Festkörperverstärker verwendet wird. Die so entstehende Schaltung ist kompakt und einfach, hat eine Frequenzkonstanz von 60×10^{-6} zwischen $-50°C$ und $+100°C$ sowie $0{,}5 \times 10^{-6}$ je Volt Hochspannungsänderung.

Binäranzeigeschaltungen mit Transistoren und Gasentladungs-Ziffernröhren von P. R. Adby

Zusammenfassung des Beitrages auf Seite 147-151

Es wird eine Anzeige für Binärzähler beschrieben, in der die herkömmliche Dioden-Decodiermatrix durch Kombination von Basis— und Emitterschalten des Endtransistors vereinfacht wird. Die entstehende Matrix enthält nur neun Dioden, benötigt wesentlich weniger Bauelemente und ist billiger herzustellen. Eine weitere Vereinfachung wird durch Ersatz der Dioden durch Widerstände mit engen Toleranzen erreicht. Eine Speicherschaltung mit bistabilen, Tor- und Übertragungsschaltungen wird auch beschrieben.

Eine 2,5-kW-Mikrowellen-Plasmafackel von J. Swift

Zusammenfassung des Beitrages auf Seite 152-154

Eine Mikrowellen-Plasmafackel, die eine einfache und zuverlässige Plasmaquelle für die Forschung bildet, wird beschrieben. Sie arbeitet bei 2,45 GHz und gibt eine Leistung von 2,5 kW ab.

Periphere Geräte für kleine Digitalrechner von A. D. Booth und J. M. S. de Vries

Zusammenfassung des Beitrages auf Seite 155-163

Ein kleiner transistorisierter Digitalrechner ist bereits früher beschrieben worden. In diesem Beitrag werden die peripheren Geräte beschrieben, und zwar ein fotoelektrischer Lochstreifenleser, ein Streifenlocher, eine Schreibmaschine, sowie die zugehörigen Steuerschaltungen.

Ein Blitzstrahlzähler von R. H. Golde

Zusammenfassung des Beitrages auf Seite 164-166

Ein beschriebener Blitzstrahlzähler wurde von einem Forschungsausschuss der Internationalen Hochspannungskonferenz (C.I.G.R.E.) für internationalen Einsatz zur Bestimmung der Frequenz des Auftretens von Blitzstrahlen zur Erde übernommen.

Gerät zur Messung des dynamischen Frequenzgangs von G. R. J. Baldwin

Zusammenfassung des Beitrages auf Seite 167-171

Ein Gerät, das den Phasen- und Amplitudengang aktiver und passiver Elemente unter HF-Impulsbedingungen mit sehr schneller Frequenzmodulation messen kann, wird beschrieben.

Die Messunsicherheit ist $\pm 1°$ Phase (innerhalb eines 100°-Bereiches) und $\pm 0,1$ dB Amplitude (über einen 10-dB-Bereich) bei Phasen- und Amplitudenänderungsraten bis zu 2,5 MHz, mit FM-Raten bis zu mehreren MHz per Mikrosekunde.

Das Gerät ist für Betrieb in den HF-Bändern von 400 MHz bis zu 10 GHz mit auswechselbaren HF-Köpfen für die verschiedenen Bänder konstruiert. Impulsdauer und -frequenz sind über einen breiten Bereich regelbar.

Bis zu 1 099 508 482 050 zählen, ohne zu übertragen von A. M. Andrews

Zusammenfassung des Beitrages auf Seite 172-175

Die folgenden Resultate wurden für ein binäres Schieberegister mit Rückführung bewiesen: (a) Für die Zykluslänge (2^n-1) muss die Rückführung der Modulo-2-Summe von einer geradzahligen Anzahl von Elementen erfolgen; (b) Dieselbe Verknüpfungskonfiguration gibt diese Zykluslänge, einerlei ob die Rückführung der Modulo-2-Summe oder ihrer Umkehrung erfolgt; (c) Zyklen der Länge (2^n-2) sind genauso möglich wie die der Länge (2^n-1). Für den Fall $n=20$ werden Zykluslängen für (2^n-1) und (2^n-2) angegeben. Zwei 20-Element-Register, die diese Verknüpfungskonfiguration verkörpern und parallel mit Schiebeimpulsen gespeist werden, haben einen kombinierten Zyklus der im Titel angegebenen Länge.

Ein Wechselstromtaster mit niedriger Eingangskapazität und Verstärkung von 1 von W. H. T. King

Zusammenfassung des Beitrages auf Seite 176-177

Für Überwachung schwacher Hochfrequenzsignale mit den meisten der greifbaren Mehrzweckoszillografen braucht man—besonders für die kleineren Typen—einen Tastkopf, der keine Abschwächung hat. Es wird ein einfacher aktiver Taster beschrieben, der zwischen 10 Hz und 5 MHz für Signale bis zu 1 V_{eff} eine Verstärkung von 0,98 oder besser hat und auf Impulse mit schneller Anstiegszeit gut anspricht. Der Taster entnimmt Batterien etwa 1,5 mA bei 18 V, jedoch nur, wenn er mit der zu testenden Schaltung verbunden ist.

Entwurf einer thyristorgeregelten Gleichstromversorgung mit konstantem Strom von A. Davies

Zusammenfassung des Beitrages auf Seite 178-182

Der Entwurf von Stromkonstanthaltern mit Transistoren für die Feinregelung zusammen mit Thyristorschaltungen für die Grobregelung wird besprochen. Ein für den effektiven Quellenwiderstand solcher Konstanthalter abgeleiteter Ausdruck gestattet Berechnung der Stromkonstanz für eine gegebene Speisespannung sowie Widerstandsänderungen der Last.

Die Schaltung eines typischen Konstanthalters wird zusammen mit einer Standard-Einsteckregelschaltung, die mit fast jedem einphasigen Konstanthalter Verwendung finden kann, beschrieben.