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Contents

561 Commentary

562 A Transistor Pulse Width Modulator *by R. D. Bell and K. E. Tait*

568 Design of Small Signal Amplifiers using F.E.T. *by W. Gosling*

571 A Cathode-ray Tube Display System

572 The Analysis and Performance of Transistor Choppers *by D. J. Finlay*

579 Q Measurements on Low-Loss Waveguide Cavities *by J. K. Chamberlain*

581 Computers for Air Traffic Control

582 A Waveform Regenerator for Amplitude Sampled Systems *by T. I. Mitchell and V. J. Phillips*

588 Transfer Function Measurement Using Fast Pulses (Part 2) *by R. C. French*

593 Spectrum Analysis of a Train of Modulated Trailing Edge Variable Slope Pulses *by O. E. Kruse and R. W. Montgomery*

x 596 Swept Frequency for General Receiver Tests *by J. F. Golding*

601 A Colour Visual Flight Attachment

602 Some Applications of Double Coil Relays *by H. Biggar*

606 Microminiaturized S.S.R. Equipment

x Extracted.

607 New Books

610 New Equipment

616 Short News Items

FRANCAIS

619 Nouvelles Réalisations

624 Résumés des Principaux Articles

DEUTSCH

626 Neue Ausrustungen

631 Zusammenfassung der wichtigsten Beiträge

Classified Advertisements page 157

Index to Advertisers page 163

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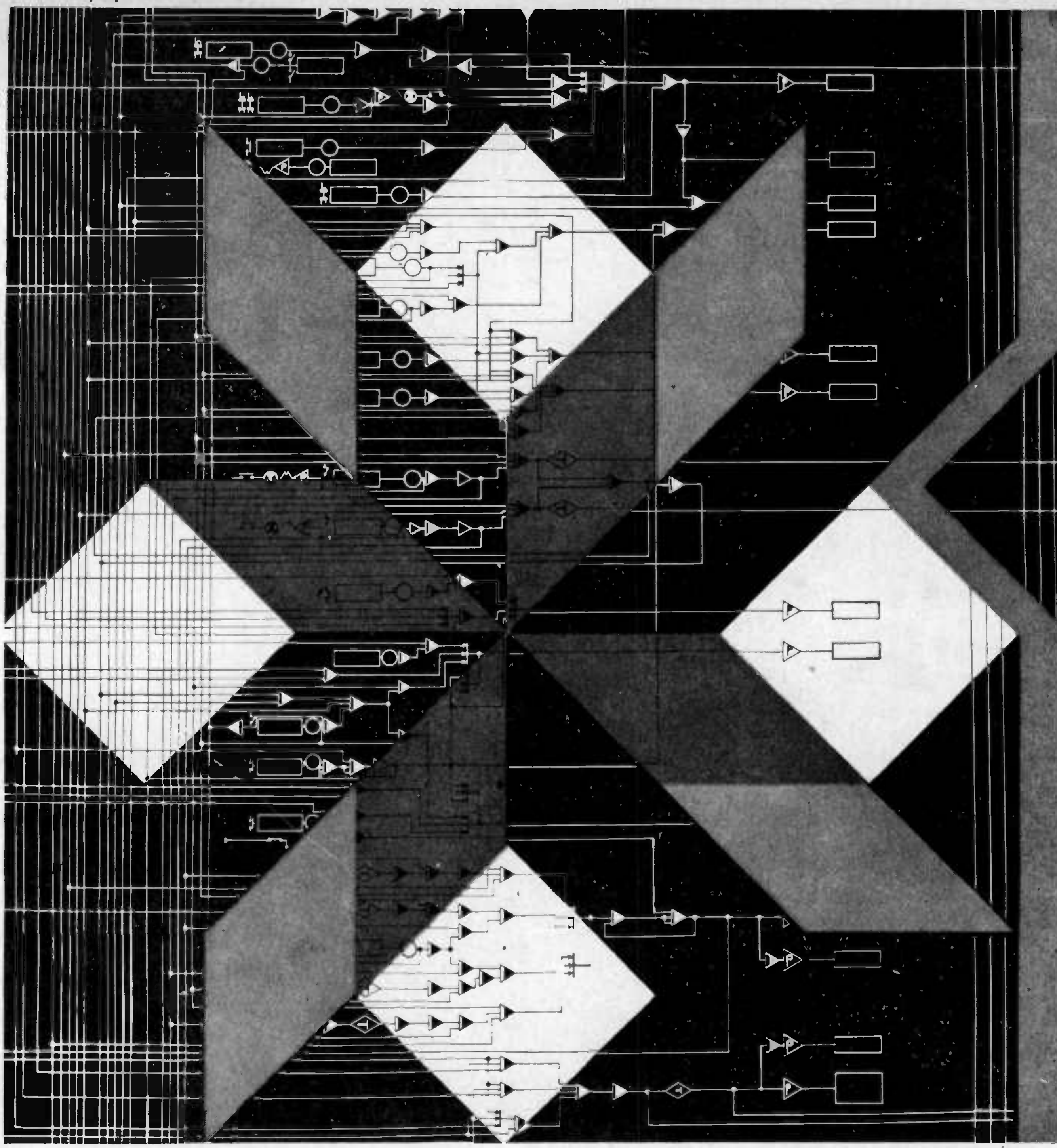

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ELECTRONIC ENGINEERING

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Commentary

UNTIL a month or two ago, the impression was that the colour television situation in this country was clear. It was expected that after the indecision at Vienna last year, the Study Group XI of the C.C.I.R., meeting at Oslo in July, would at last come to agreement and the whole of Western Europe would have a common system. And in this country the Postmaster-General announced in March that the Government had authorized the start of a colour television service towards the end of next year, it seemed that at least we had the signal to go ahead.

However, at the moment things look more confusing than ever, both at home and abroad. The Oslo meeting, like its Vienna predecessor, was one of indecision and the Western European countries are as divided as ever.

A large number of countries (Austria, Denmark, Eire, Finland, Great Britain, Holland, Iceland, Italy, Liechtenstein, Norway, Sweden, Switzerland and West Germany), representing some 41 million television viewers, had voted for the PAL system. Three of these countries, Britain, Germany and Holland, had in fact decided to start broadcasting on the PAL system within the next few months regardless of the Oslo decision.

The rival system—SECAM III—was supported by France, Greece and Monaco, together with the Eastern bloc of countries totalling some 24 million viewers.

During the Oslo conference further support to the PAL system was given by Australia, New Zealand and South Africa, whereas some twenty-two countries outside Europe, including sixteen African countries, seven of whom have no television service at all, voted for SECAM III.

What might have been a way out from the present dilemma was a variation of the SECAM system originally introduced by Russia and known as the SECAM IV or NIR system. It was proposed by France, with Russian support, that the choice should be delayed for a period of at least six months during which time this new SECAM IV system could be further developed, but this was turned down by Britain and West Germany on the grounds that they were too far committed to PAL, so that France had no alternative but to stop development, and it is more than likely now that the various countries will install their own colour television service, based on either PAL or SECAM, regardless of their neighbours, ruling out, so it is alleged, the international exchange of programmes.

But there are some eleven different monochrome standards already in Europe, and this means the same number of standards with a single colour system, and so a European viewer will need an expensive multi-standard television receiver if he wishes to view a programme direct from a neighbouring country.

And how do we stand in this country?

Although we have permission by the Postmaster-General, colour television, using the PAL system, is limited to the BBC2 service on the 625-line standard, so excluding at present the Independent Television Authority who are confined to the 405-line standard and who have not yet been granted a second service.

It was proposed as far back as 1960 by the Television Advisory Committee that all broadcast television should be changed from the 405 to the 625-line standard and subsequently in 1962, the Government of the day, supporting the report of the Pilkington Committee, went a step further and decided that all new programmes should be on 625-line u.h.f. standard including any future ITA second programme.

Unfortunately the prospects of a second ITA television service seem very remote, particularly with the present-day economic crisis, and the only hope the ITA have of introducing their own separate colour programme is to do so on their existing 405-line service.

While the manufacturing industry sympathize with the ITA, they are rigidly opposed to the introduction of colour on what they regard as the outmoded 405-line service on the grounds that the 405-line service is already inadequate for monochrome transmission and the difference between colour transmissions on 405 and 625 lines will be even more marked, and they argue that a dual standard television set to receive colour and monochrome will be a complex and therefore more unreliable affair adding some £25-£30 to the cost of the receiver.

So it would seem that the future rests with colour and monochrome only on the 625-line service.

But is the present BBC2 625-line service all that it is made out to be?

According to a report recently issued by ABC Television Ltd, the BBC2 625-line service on u.h.f. has turned out to be very disappointing, giving a coverage inferior to that predicted, so that extra gap-filling transmitters and a large number of low power boosters will have to be installed if the present coverage of BBC1 and ITA is to be approached, let alone equalled.

What is certainly more surprising in the report is the survey carried out in May this year by the ITA in the u.h.f. areas, where they found that viewers able to receive 625-line programmes stated that these programmes were not superior to those on 405-line despite the additional cost of receiver and aerial installation.

But when we have solved this dual-standard problem, can we still go ahead?

On economic grounds it seems unlikely.

The original estimate for transmitting stations alone put forward by the Television Advisory Committee in 1960 was given as £25 million and it is certain that at today's prices this figure is quite unrealistic, and the viewers will find it all very expensive—at least £200-£250 for a receiver or 35s. a week on a rental basis, together with an increase in the licence fee.

The Design of a Transistor Pulse Width Modulator Suitable for Control Applications

By R. D. Bell*, B.E. (Hons), and K. E. Tait†, B.E. (Hons), B.Sc.

The design of a transistor pulse width modulator suitable for use in analogue computer studies of pulse width modulated control systems is presented. A theoretical prediction of the limitations and accuracy of the designed unit is given and these limitations are critically compared with practical results.

(Voir page 624 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 631)

A NUMBER of devices and circuits have been described which produce a pulse duration determined by an analogue input signal[1,2,3,4,5]. Although valves tend to have more advantages in this application than transistors, it was desirable in this design to use the latter and consequently, circuits similar to the Phantastron or Sanatron valve circuits could not be used.

Transistor circuits used in this application are quite often of the 'Bootstrap integrator' type but unfortunately these circuits suffer from a relatively long recovery time compared to the duration of the maximum output pulse and the dependence of the trailing edge of the pulse on the actual input voltage at that time. An additional disadvantage is that for sudden changes in the input signal the circuit could be triggered resulting in an output pulse starting before a sampling interval.

To overcome these difficulties a different approach was necessary and the block diagram of the system adopted is shown in Fig. 1. Although the linearity would not be as good as the Bootstrap circuit unless a more complicated discharge circuit was used, it was considered that the removal of other disadvantages would outweigh the small loss in linearity.

Detailed Description of the Circuit

The circuit designed is shown in the block diagram of Fig. 2. The multivibrator is the normal free running type and can be synchronized to a desired input signal if required in the usual way. This multivibrator produces a suitable pulse which cannot exceed 10 per cent of the sampling interval and is used to control the sampling of the input voltage. The input voltage is initially taken through an impedance convertor to reduce excessive load on the voltage source and to produce a sufficiently low output impedance so that the timing capacitor can be charged in the required sampling time. The sampler is used to charge the timing capacitor which then discharges through a constant current circuit to produce a sawtooth, the amplitude of which is dependent on the input voltage. It was found necessary to include two samplers and two *RC* timing networks so that for small input voltages maximum sensitivity could be obtained.

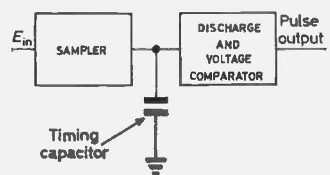

Fig. 1. Simplified block diagram

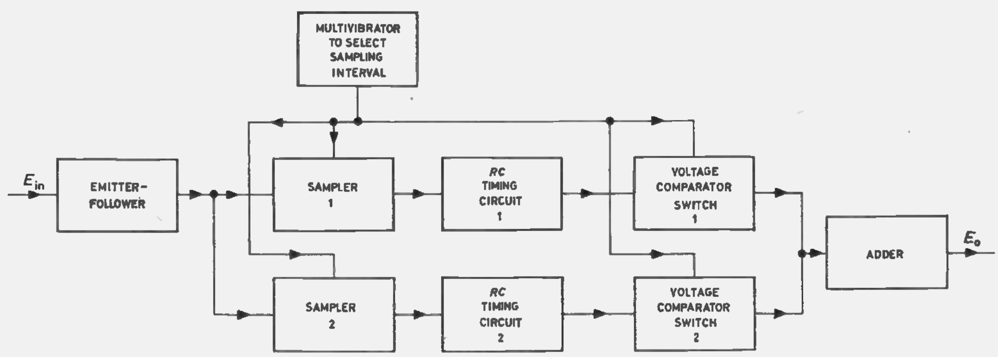

Fig. 2. Arrangement of the modulator

The sawtooth produced by this *RC* timing circuit is then applied to a sensitive voltage comparator to obtain the necessary output pulses. Pulses produced by each channel are then summed by a simple resistance adder to produce the desired output pulse train.

Sampling Interval Multivibrator

The complete circuit of the multivibrator is shown in

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† *University of New South Wales, Australia.*

Fig. 3. This is the normal circuit of a free running multivibrator except for two modifications.

(1) Transistor VT_1 which charges capacitors C_4 or C_5 (depending on the switch position) when VT_2 is cut off and VT_3 is saturated. These capacitors would normally charge through R_3, but since a small charge time was required it was necessary to include VT_1. The resistor R_2 in the collector of VT_1 is to limit the peak charge current consistent with the rating of VT_1.

(2) The multivibrator is not of a symmetrical type because of the output requirements, and therefore it was found necessary to have the maximum allowable current flowing through transistor VT_3 under saturation conditions and the minimum workable current through transistor VT_2 when saturated. This requirement has been met by a large difference in collector load resistors in transistors VT_2 and VT_3 and also a considerable amount of unbalance in the RC timing networks in the base circuits.

Diode MR_1 has been included to limit the reverse current in the base to emitter circuit of transistor VT_2 to a safe level. This is necessary since the cut-off base to emitter voltage exceeds the maximum allowed by the manufacturer.

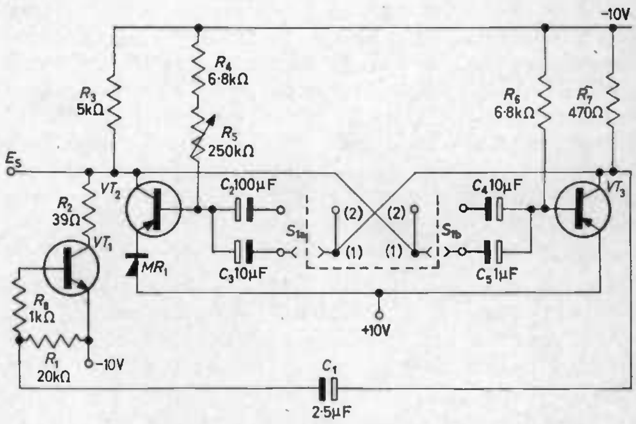


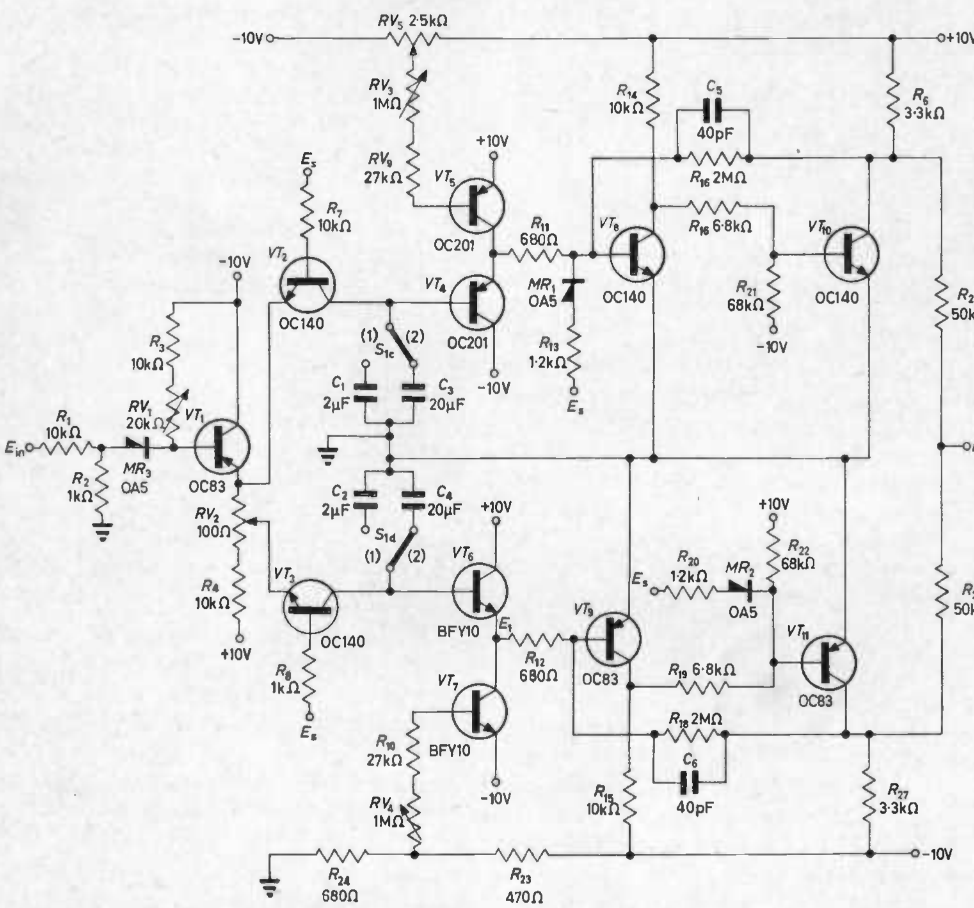
Fig. 3. Sampling interval multivibrator

PULSE-WIDTH MODULATOR CIRCUIT

Referring to Fig. 4 it will be seen that transistor VT_1 forms an emitter-follower circuit so that the charging of capacitors C_1 or C_3 and C_2 or C_4 can be achieved in the small sampling time, and also without any large loading of the input signal source.

R_1 and R_2 form a voltage divider network to reduce the input voltage to an acceptable level for the transistor circuit. RV_1, R_3 and R_2 form a divider to alter the zero input voltage condition so that no output pulses are obtained under these conditions. It is necessary to have this divider since in all emitter-follower circuits the output voltage is not exactly equal to the input voltage because of the small drop across the base to emitter junction. Also in this particular case the output voltage is not taken directly from the emitter of the transistor but at a point closer to +10V, so that a slightly larger voltage has to be injected into the base of the transistor to obtain zero output for the zero input condition.

Fig. 4. Pulse-width modulator circuit



The divider network formed by RV_2 is to obtain a balance condition which is as close as possible to the input condition of zero volts, i.e. for a small positive input voltage a small positive pulse is required and for a small negative input voltage a small negative pulse is required, and the balance adjusts the amount by which the input has to change in any direction to obtain the required output. This control is necessary because of the small voltage drops across the junctions of transistors VT_2, VT_3, VT_4 and VT_6 which tend to make the output pulse arrive only after the input changes by this voltage drop in either direction.

In the following description, operation of the positive input channel only will be described since the other channel is identical in operation but operates on negative input voltages. Reference should be made to Figs. 4 and 5.

The channel operating on positive input signals is the lower channel. Transistor VT_3 is the sampling transistor and charges capacitor C_2 or C_4 depending on the range selected during the small sampling duration determined by signals applied to its base. The capacitor C_2 or C_4 is charged to a voltage directly proportional to the input voltage E_{in}. At the end of the sampling duration transistor VT_3 is cut off since its base is now at a lower potential than the potential which any allowable input signal can produce at the emitter of transistor VT_3.

During the next period of time, capacitor C_2 or C_4 discharges through transistor VT_6, which is basically an emitter-follower stage with a transistor VT_7 substituted for the normal resistive load. Transistor VT_7 therefore effectively feeds transistor VT_6 with an almost constant current determined by the base current of VT_7. This in turn makes the base current of transistor VT_6 almost constant over quite large variations of voltages appearing across capacitor C_2 or C_4, assuming that the leakage currents of transistors VT_6 and VT_7 are neglected. Therefore effectively the voltage time relationship of the capacitor C_2 or C_4 will be a straight line and very good linearity will be obtained.

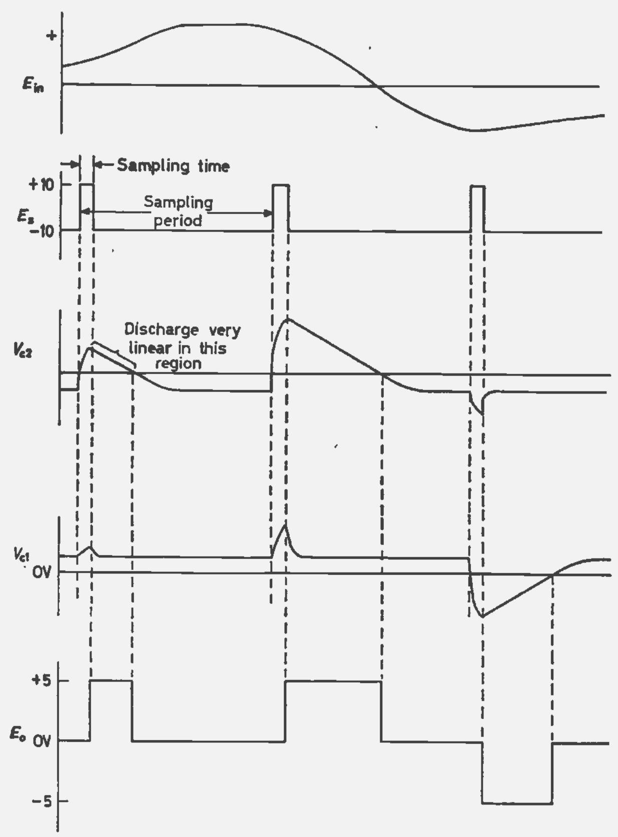


Fig. 5. Typical waveforms

The circuit formed by R_{23}, R_{24} and RV_5 in the negative channel is a balanced control circuit to adjust for variations in transistor characteristics and allow for capacitor tolerances. Potentiometer RV_5 is a pre-set control and need only be adjusted when transistors or capacitors are changed.

The effect of changes in temperature on the current amplification β of transistors VT_6 and VT_7 is minimized by the cancelling property of the circuit chosen. If VT_6 and VT_7 are reasonably well matched and their variations of β with temperature can be assumed to be equal, it is apparent that the discharge current of the capacitor C_2 or C_4 is solely determined by the base current of VT_7.

It has been assumed that VT_9 is cut off since resistor R_{12} is less than resistor R_{18} and only positive input voltages are being dealt with. This transistor will remain cut off until the voltage at the emitter of transistor VT_6 drops slightly below zero volts at which stage transistor VT_9 will start to conduct and due to feedback produced by resistor R_{18} and capacitor C_6 the transistor pair VT_9 and VT_{11} will change state.

The purpose of the diode MR_2 and the resistor R_{20} is to cancel the pulse produced at the output due to the small sampling time. It does this by cutting off transistor VT_{11} during this period irrespective of what is happening at the input terminals.

To summarize it can be said that if the input is at a positive level then capacitor C_2 or C_4 is charged to a level proportional to the input. The voltage across the capacitor causes the voltage comparator (transistors VT_9 and VT_{11}) to change state, and this state will then remain until the capacitor discharges to a small negative voltage at which stage the voltage comparator changes state again and completes the output pulse for that period. The circuit then remains in this state until the next sampling pulse arrives.

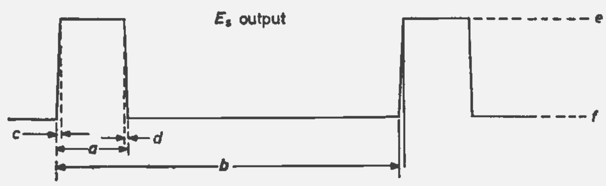


Fig. 6. Multivibrator output specification waveform

The adder consists of resistors R_{28} and R_{29} and since the stable state of the voltage comparators is with transistors VT_{10} and VT_{11} cut off, the output voltage is equal to zero volts provided that the positive and negative supplies are equal and corresponding output resistors are equal. If either of the comparators is in the 'on' state, i.e. with transistor VT_{10} or VT_{11} saturated, the output will be approximately -5V or $+5$V respectively. Consequently, for positive input signals transistor VT_{11} is saturated and the output is at $+5$V.

Design Considerations

Free Running Multivibrator

Consider the ideal output waveform shown on Fig. 6. The following listing specifies the design requirements for the multivibrator output E_s.

(1) The repetition period b must be variable from 0·1sec to 10sec.

(2) The pulse width a must not exceed 10 per cent of the period.

(3) The rise time c should be as small as practicable but not more than 1 per cent of the pulse width.

(4) The fall time of the trailing edge d should be as small as practicable but need not be as small as the rise time.

From circuit considerations it was decided that voltage levels of $+10$V and -10V should be adopted for e and f respectively.

Two pulse repetition ranges have been chosen to conveniently obtain the desired coverage.

Range (1) 0·1sec to 1sec

Range (2) 1·0sec to 10sec.

Considering range (1), i.e. with capacitors C_3 and C_5 selected in Fig. 3, it will be noted that the pulse width 'a' is governed by the time-constant R_6C_5, and that resistor R_6 is selected so that transistor VT_3 is saturated during the sampling pulse interval (b-a). A 470Ω resistor in the collector of transistor VT_3 was selected to limit the collector current to approximately 50mA which is well within the makers limit of 500mA. Resistor R_6 can be calculated in the following way:

For saturation:

$$\beta I_b > I_c$$

or:

$$I_b > (I_c/\beta) \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

By substituting values it is found that:

$$R_6 < 10\text{k}\Omega$$

A value of 6·8kΩ was chosen.

Now the time-constant R_6C_5 must be less than 10 per cent of the period b.

i.e. $C_5 < 1{\cdot}475\mu$F.

A value of 1μF was selected.

The sampling pulse interval (b-a) is governed by the time-constant $(R_4 + R_5)\,C_3$. Also the capacitor C_3 must be of such a value that the time-constant R_7C_3 is smaller than at least half the time-constant R_6C_5 so that the capacitor C_3 is charged to almost the full 20V when VT_2 is saturated and VT_3 is cut off.

$$R_7C_3 < \tfrac{1}{2}R_6C_5$$

$$\therefore\ C_3 < 15{\cdot}4\mu F$$

A value of 10μF was selected for C_3.

Therefore:

$$(R_4 + R_5) > 1/C_3 \text{ maximum time } (b\text{-}a) \quad \ldots\ldots \quad (2)$$

For range (1) the maximum time (b-a) = 1sec.

Therefore, $R_4 + R_5 > 100\text{k}\Omega$

To obtain a good range of operation a 250kΩ potentiometer was selected.

For transistor VT_2 to be saturated:

$$\beta I_b > I_c$$
$$R_3 > 4\text{k}\Omega$$

A value of 5kΩ was selected for R_3.

Resistor R_4 is determined by the smallest sampling period rquired.

$$R_4C_3 < 0{\cdot}09\text{sec.}$$
$$R_4 < 9\text{k}\Omega.$$

A value of 6·8kΩ was selected for R_4.

The recovery time of capacitor C_5 is governed by the transistor VT_1 and resistor R_2. Transistor VT_1 must be saturated when VT_3 is saturated and stay saturated until capacitor C_5 has recovered. The time-constant R_{c1} governs the time that transistor VT_3 is saturated.

Therefore time-constant

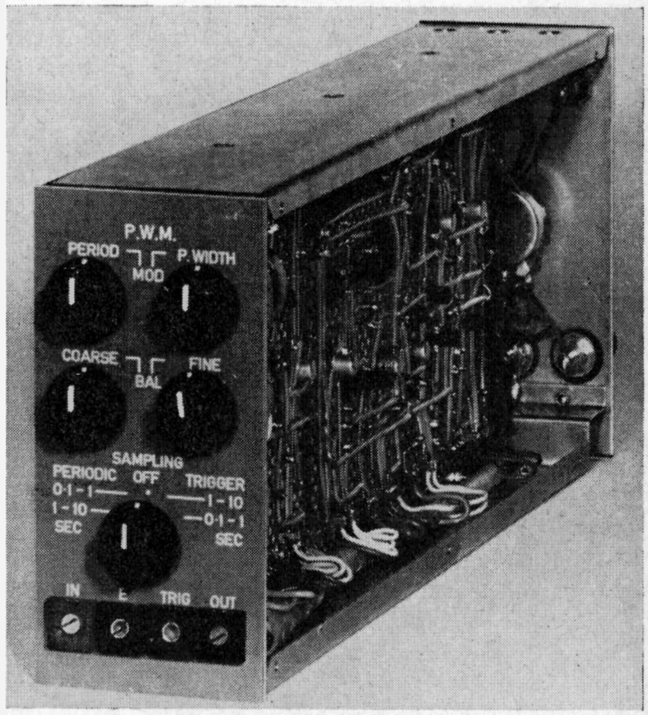
$$R_8C_1 > 4C_4R_2,$$

and:

$$R_8C_1 < \tfrac{1}{2} \times 0{\cdot}09\text{sec}$$

so that transistor VT_1 is effectively cut off when transistor VT_2 is about to conduct.

Resistor R_2 was selected to make best use of the maximum collector current of transistor VT_1. An OC140 transistor was selected and to limit the current to 500mA a 39Ω resistor was used in the collector circuit.



An engineered version of the modulator

It follows that:

$$R_8C_1 > 4 \times 5 \times 10^{-6} \times 39$$

and:

$$R_8C_1 < \tfrac{1}{2} \times 0{\cdot}09$$

Values of $C_1 = 2{\cdot}5\mu$F and $R_8 = 1\text{k}\Omega$ were chosen since they satisfy these requirements and also ensure that transistor VT_1 is saturated. Resistor R_1 was made equal to 20kΩ, and returned to −10V to ensure that transistor VT_1 is cut off during the required period.

The transistors VT_2 and VT_3 in Fig. 3 were chosen to be of the OC25 type with a maximum collector current of 500mA. Since they are either operated in a saturated or cut-off state their power dissipation was not an influencing factor, and it was just necessary to ensure that their maximum current ratings and maximum cut-off voltage ratings are not exceeded. Both these factors have been considered and the conservative values chosen above give at least a factor of 30 per cent below the maximum values stated by the manufacturer so that for normal component variations the circuits will still operate reliably.

Emitter Follower Input Stage

The following requirements were considered important.

(1) The input impedance should be greater than 10kΩ.

(2) Output impedance must be as low as possible.

A value of 10Ω for R_4 gave reasonable stability and together with RV_2 resulted in an input resistance of 500kΩ which means an input impedance of 11kΩ.

Suitable values for R_1 and R_2 gave the required base voltage of VT_1 under maximum input conditions and resulted in an output impedance of approximately 50Ω.

Sampler Circuit

Tantalum non-polarized timing capacitors have been chosen since positive and negative voltages occur across them and also they appeared better in relation to stability and leakage currents. The value of the timing capacitors should be as high as possible so that a reasonable discharge current is obtained for the maximum width setting, otherwise leakage currents of the transistors in the circuit will seriously affect the linearity of the discharge times.

Assuming from the multivibrator design that the time to charge C_2 is 5·0msec, then:

$$C_2 < \frac{5{\cdot}0 \times 10^{-3}}{4R} \quad \ldots\ldots\ldots\ldots\ldots (3)$$

where R is the equivalent charging resistance of approximately 175Ω.

i.e. $C_2 < 7\mu F$.

Corresponding values of 2μF and 20μF for ranges (1) and (2) respectively were chosen.

Timing Circuit

The discharge circuits have been designed to allow for a capacitor tolerance of ±20 per cent and a range of β from 20 to 80 for the OC201 circuit and 20 to 84 for the BFY10 circuit.

Since for practical purposes,

$$I_{bI4} = (\beta_{I5}/\beta_{I4}) \times I_{bI5} \quad \ldots\ldots\ldots\ldots (4)$$

in the extreme cases:

$$I_{bI4} = 4I_{bI5} \text{ or } \tfrac{1}{4}I_{bI5} \quad \ldots\ldots\ldots\ldots\ldots (5)$$

similarly:

$$I_{bI6} = 4{\cdot}2I_{bI7} \text{ or } (1/4{\cdot}2)I_{bI7} \quad \ldots\ldots\ldots\ldots (6)$$

If transistor VT_7 is held at a base current of 100μA it is necessary to be able to adjust the other circuit by a suitable balance control over the range of 20μA to 480μA. Also this balance control must be capable of producing this order of percentage variation over the range of discharge currents selected by the potentiometer RV_3.

The base circuit resistance of VT_7 has been returned to a potential closer to −10V rather than earth to obtain sufficient adjustment.

$$\text{The base circuit resistance} = \frac{20}{I_{b6(max)}} \quad \ldots\ldots (7)$$

$$= \frac{20}{480 \times 10^{-6}}$$

$$\text{Again the base circuit resistance} = e/I_{b7} \quad \ldots\ldots\ldots\ldots (8)$$

$$= \frac{e}{100 \times 10^{-6}}$$

Therefore:

$$e = \frac{20 \times 100 \times 10^{-6}}{480 \times 10^{-6}}$$

$$= 4{\cdot}17\text{V}$$

i.e.: The base circuit resistance of VT_7 must be returned to $-10 + 4{\cdot}17 = -5{\cdot}83$V.

Resistor R_{10} sets the maximum discharge current of capacitors C_2 or C_4, since the base currents of transistors VT_7 and VT_6 are approximately equal. It is necessary to find the maximum discharge current for the particular time involved. Assuming that the discharge current I is constant then the voltage across the capacitor is approximately given by:

$$V_o = -(I/c)\,t + 6 \quad \ldots\ldots\ldots\ldots\ldots (9)$$

assuming that the initial voltage across the capacitor is 6V. With range I selected the fastest discharge time is 0·1sec which corresponds to maximum discharge current. Since the circuit changes state when V_o is zero and C is 2μF:

$$I = \frac{6 \times 2 \times 10^{-6}}{0{\cdot}1}$$

$$= 120\mu A$$

This is the current which has to flow in the base of transistor VT_7 and:

$$R_{10} = \frac{4{\cdot}17}{120 \times 10^{-6}}$$

$$= 34{\cdot}7\Omega \text{ (preferred value of 27k}\Omega\text{)}$$

The highest resistance in the base circuits is governed by potentiometers RV_3 and RV_4; it is not critical but must be large enough to give the longest discharge time required. By similar calculations this was found to be greater than 500kΩ and since a 1MΩ dual potentiometer was readily available this was used. The resistor R_{25} was added after the circuit was built to improve tracking of the dual potentiometer and the optimum value chosen was 1MΩ. It will be noted that it has been assumed that transistors VT_8 and VT_9 are cut off.

The only other critical part of the discharge circuit R_{12} was chosen so that the emitter of transistor VT_6 is above −8V, thus allowing a reasonable 2V margin from the

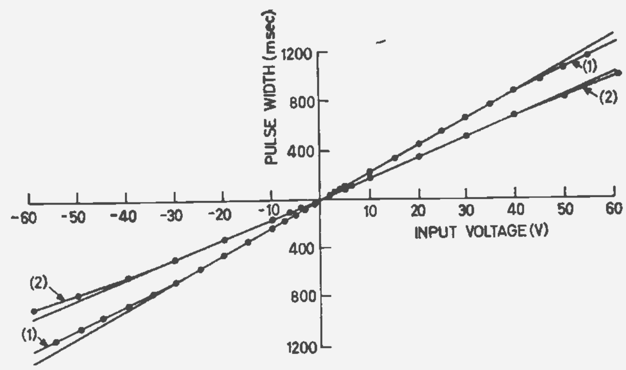


Fig. 7. Linearity of pulse-width modulator for ranges (1) and (2) at maximum and minimum settings

maximum allowable emitter to collector voltage of 20V for VT_6, a BFY10 transistor.

Voltage Comparator Circuit

Transistor VT_9 should saturate when E_1 is less than zero volts and be cut off when E_1 is above zero volts.

For saturation:

$$\beta_{I(b)} > I_o \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (10)$$

$$10/R_{15} < \beta(E_1/R_{12}) \quad \ldots\ldots\ldots\ldots (11)$$

$$\therefore\ E_1 > 10R_{12}/\beta R_{15} \quad \ldots\ldots\ldots\ldots\ldots\ldots (12)$$

From the last inequality it follows that for maximum sensitivity resistor R_{12} should be as small as possible and resistor R_{15} as large as possible. The maximum value of R_{15} is determined by the leakage current of transistor VT_9, and therefore should be made no greater than 10kΩ to prevent an undesirably high voltage drop occurring when the transistor is cut off.

When transistor VT_9 is saturated, transistor VT_{11} is cut off and consequently the voltage at the junction of resistors R_{19} and R_{23} must be greater than zero volts under the worst conditions of maximum leakage current in VT_{11}.

The feedback circuit resistor R_{18} has to be at least 2MΩ so that no appreciable loading of the discharge circuit is obtained when transistor VT_9 is cut off. Capacitor C_5 was then selected on a trial and error basis to give an optimum output pulse response. R_{20} was calculated so as to ensure that transistor VT_{11} is cut off during the sampling interval.

The voltage comparator in the negative input channel has a similar circuit except that npn transistors are used instead of pnp transistors and also the sampling time duration is cancelled at a different position in the circuit in order that the same sampling time waveform can be used.

Theoretical Accuracy and Limitations

Linearity of Pulse Width with Input Voltage

The main factors affecting linearity of the device are:

(1) Leakage currents of transistors VT_2, VT_3, VT_4, VT_5, VT_6, VT_7.

(2) The slope of the V_{ce} to I_c curves of the transistors, VT_5 and VT_7, used to discharge the timing capacitors.

Leakage currents of transistors vary exponentially with temperature as well as to a fair degree with voltage. They tend to decrease the discharge current in the timing circuit producing a more protracted capacitor discharge time.

For the OC201 transistors used, VT_4 and VT_5, the maximum collector current variation for the maximum collector voltage variation was about 0·8mA which is 10 per cent of the operating collector current. This means that the maximum current would discharge the capacitor to zero volts about 10 per cent earlier than the minimum current condition. Since in the actual circuit the capacitor voltage time curve will be exponential and intermediate between the linear curves assumed, it was considered that the actual deviation from linearity should be better than about 5 per cent. It will be noted that these two effects will tend to compensate one another.

From the specification given in the appendix it will be noted that a linearity of better than 6 per cent for a 0 to 50V input range was obtained.

Temperature Stability

The two main temperature dependent parameters in the circuit are:

(1) Leakage current I_{co}, variation in transistors.

(2) Base to emitter forward voltage drop which varies at the rate of 2·5mV/°C for germanium transistors.

Effects on the multivibrator circuit:

Leakage current causes the capacitor in the base of the transistor that is cut-off to discharge more quickly than it would if the transistor were not present. The resulting effect is that both time *b* and time *a* would decrease with temperature. However, silicon transistors have been used in the circuit and since their maximum leakage current is only 0·1μA which is less than 0·5 per cent of the minimum capacitor discharge current of 40μA, the maximum variation of time *b* and time *a* should be less than 0·5 per cent over a range of 20°C.

Since the transistors are either cut off or saturated the variation of emitter forward voltage drop has no appreciable effect on the multivibrator circuit performance.

Effects on the emitter-follower input stage:

Here the base to emitter forward voltage drop variations may make the output more positive than the actual input voltage. Since the circuit must respond to input voltages on the base of the emitter-follower of the order of 0·05V, this factor will have a large influence on the width of the output pulse. To overcome this problem a diode MR_3 has been included in the base circuit of the emitter-follower. Since the diode is made of the same material as the emitter-follower transistor, the voltage appearing across the diode will vary in almost the same way as the voltage across the emitter to base junction of the transistor and the emitter-follower input is almost independent of temperature. By using this method at least 90 per cent of the output voltage variations with temperature have been removed.

Power Supply Variations

No special precautions have been taken to make the circuit insensitive to power supply variations since it is intended that the power supplies used will be stabilized by the use of Zener diodes.

Jitter

The main cause of jitter of the output pulse assuming that the power supplies are stabilized is the slope of the discharge waveform at zero volts appearing at the base of transistors VT_8 and VT_9. If the slope is small as it must be to obtain a ten second pulse from a 5V signal then the exact point at which transistor VT_8 or VT_9 starts to conduct has to vary only over a small range to obtain variations in the trailing edge of the output pulse. To minimize this effect the sensitivity of transistor stages VT_8 and VT_9 has been made as high as practical.

Conclusion

The pulse width modulator designed achieved the specification given in the appendix. Further experimental work was performed with the modulator inserted in the error channel of a simulated feedback control system. Very close agreement was obtained between the theoretical performance obtained using Andeen's linearization procedure[8] and analogue computer results.

Acknowledgments

The authors would like to thank the staff of the analogue computing laboratory of the School of Electrical Engineering, University of New South Wales for their assistance during the testing of the design prototype.

Appendix

Specifications of Pulse-width Modulator

(a) Power supply requirements:

+ 10V ± 1 per cent at 100mA
− 10V ± 1 per cent at 100mA

(b) Input characteristics:

Input impedance	= 10kΩ
Maximum input voltage	= 60V

(c) Output characteristics:

Output impedance	= 25kΩ
Amplitude of positive pulses	= +5V
Amplitude of negative pulses	= −5V
Rise time	= 20μsec
Fall time	= 20μsec
Minimum output pulse	= 40μsec
Maximum output pulse (60V input) range (1)	= 1·5sec
Maximum output pulse (60V input) range (2)	= 10sec

Linearity better than 6 per cent for a 0 to 50V input

Jitter less than 1 per cent for 5V input

(d) Temperature stability :

No overall figure available for complete pulse-width modulator

For multivibrator—Less than 1 per cent drift for 10°C change in temperature

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Design of Small Signal Amplifiers using Field Effect Transistors

By W. Gosling*

The design of small signal amplifiers using f.e.t.'s biased near to pinch-off is considered, and selection of circuit values to achieve working-point stabilization is described. The method of design yields results which are in reasonable agreement with measured values, and the amplifiers, which typically have voltage gains up to about 50, have the usual advantage of high input impedance, characteristic of field effect transistors.

(Voir page 624 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 631)

FIELD effect, or unipolar, transistors (f.e.t.) offer many advantages in the design of electronic circuits, both in their junction gate and insulated gate (m.o.s.t.) forms. At present the former are mainly used where low noise characteristics are required, in particular in small signal amplifiers. The properties of amplifier circuits using junction f.e.t.'s have been described in the literature[1,2].

The transfer characteristic of the device is such that the drain current is an approximately parabolic function of the gate source voltage (V_{GS}); thus, writing I_D for the drain current:

$$I_D = I_{DSS}\,(1 - (V_{GS}/V_P))^2 \quad \text{............} \quad (1)$$

where I_{DSS} and V_P are constants of the transistor.
I_{DSS} is the standing drain current with the gate shorted to the source. This is non-zero for a junction f.e.t. which operates in the depletion mode.

This law is well supported by experiment, the exponent being invariably close to two for devices of the double diffused or epitaxial diffused types which are now widely used in amplifier applications.

For small values of V_{GS} equation (1) may quite well be approximated by a linear dependence[3] of the form:

$$I_D = I_{DSS} + g_{FS}\,V_{GS} \quad \text{.............} \quad (2)$$

where g_{FS} is the large-signal forward transfer conductance of the f.e.t.

To a good approximation, g_{FS} is obtained from the zero gate voltage small signal forward transfer conductance $g_{fs}(o)$, which may itself be derived by differentiation of equation (1) as:

$$g_{fs}(o) = \frac{-2I_{DSS}}{V_P} \quad \text{.............} \quad (3)$$

Equation (2) will be a valid approximation over the range $I_{DS} > I_D > 0{\cdot}4 I_{DSS}$ if g_{FS} is made equal to $0{\cdot}8 g_{fs}(o)$.

However most small signal voltage amplifiers utilizing f.e.t.'s do not operate in this region where I_D is not much less than I_{DSS}. Instead, the transistor is biased with V_{GS} almost equal to V_P, and the drain current is small. The reasons for this choice of low drain current are:

(a) Power economy.

(b) Minimal transistor temperature rise, and hence low gate leakage current and correspondingly high input impedance.

(c) Since, by inspection of equation (1), the forward path transfer conductance falls less rapidly than I_D while the maximum value of load resistor is inversely proportional to I_D (for a given load resistor d.c. voltage drop), higher gain will be obtained at lower values of drain current.

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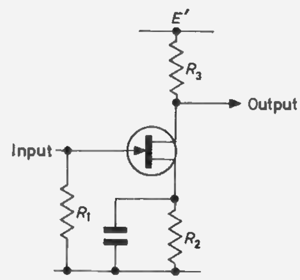
A simple, but widely used, circuit is that shown in Fig. 1. The drain current flowing through the bias resistor R_2 produces the standing d.c. bias on the transistor, and at the same time introduces negative current feedback. This stabilizes the working point of the device against the effects of temperature or the variation of device parameters between different examples of a given transistor type. Since the device is biased near to pinch-off (that is V_{GS} is only slightly smaller in magnitude than the pinch-off voltage V_P) the simple approximation of equation (2) will not hold, and analysis must proceed in terms of equation (1).

Fig. 1. A single f.e.t. resistance coupled amplifier



Relationship of Drain Current to Source Resistor

From equation (1), assuming that negligible voltage drop occurs in the gate resistor R_1:

$$I_D = I_{DSS}\,\{1 - (R_2 I_D / V_P)\}^2 \quad \text{.........} \quad (4)$$

But V_P/R_2 has the dimensions of a current. Writing this current as i, then since i must be slightly greater than I_D because $R_2 I_D$ is slightly less than V_P, one may write:

$$I_D = i\,(1 - \delta) \quad \text{...............} \quad (5)$$

where δ is a small positive dimensionless quantity.
Hence, substituting for I_D in equation (4):

$$i(1 - \delta) = I_{DSS}\,\{\delta^2\} \quad \text{.............} \quad (6)$$

or:

$$I_{DSS}\,\delta^2 + i\delta - 1 = 0$$

Solving for δ, and since δ must be positive:

$$\delta = 1/2I_{DSS}\,\sqrt{(i^2 + 4i\,I_{DSS})} - (i/2I_{DSS}) \quad \text{....} \quad (7)$$

However, i is much smaller than I_{DSS}, since it is of the same order as I_D, thus approximately:

$$\delta = \sqrt{(i/I_{DSS})} - (i/2I_{DSS}) \quad \text{...........} \quad (8)$$

The error involved in making this approximation is quite small. It leads to δ being underestimated by about 10 per cent at $I_D = 0{\cdot}5\,I_{DSS}$, by 8 per cent at $I_D = 0{\cdot}25\,I_{DSS}$, and by progressively smaller amounts at still smaller values of I_D. The errors in I_D are smaller still.

Thus an expression for I_D may be derived at once from equations (5) and (8).

$$I_D = V_P/R_2 \left\{1 - \sqrt{\frac{V_P}{R_2 I_{DSS}}} + \frac{V_P}{2R_2 I_{DSS}}\right\} \quad \ldots\ldots (9)$$

From this expression I_D may easily be calculated, or alternatively, Fig. 2 is a plot of δ against i/I_{DSS} and may be used to determine I_D in conjunction with equation (5).

WORST CASE DESIGNING

Using either of the methods described the maximum and minimum values of I_D likely to be encountered with any unit from a given batch of transistors may be determined. Since the drain current increases with both I_{DSS} and V_P (from equation (1)), maximum I_D will correspond to the maximum permitted values of these two variables. It may be worth noting that I_{DSS} and V_P tend to be positively correlated in any batch of transistors, since factors

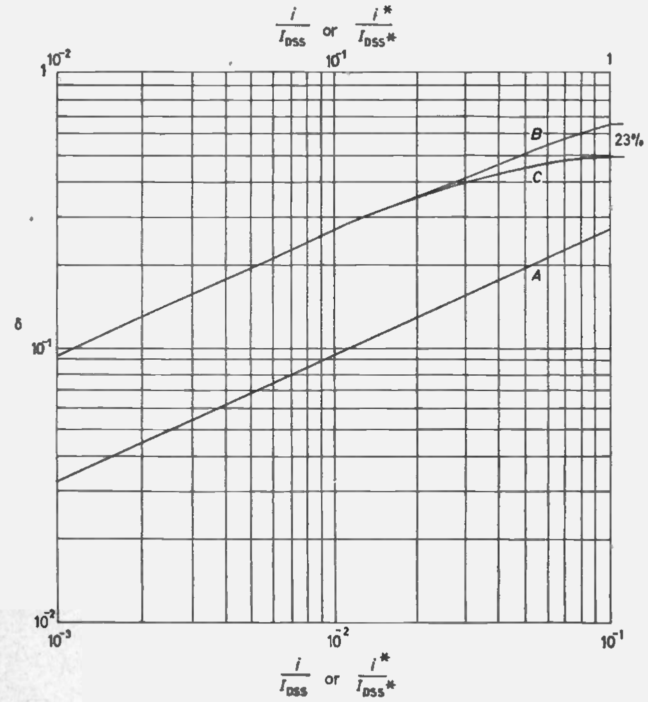


Fig. 2. The parameter S as a function of i/I_{DSS} or i^*/I_{DSS}^*
Curve A: for the abcissa range 10^{-3} to 10^{-1}
Curve B: for the abcissa range 10^{-2} to 1
Curve C: modification of curve B when the approximation of equation (8) is permitted. Note that the maximum error of Curve C is 23 per cent. but that this decreases rapidly as δ diminishes

(including high semiconductor conductivity in the channel and over-size channel dimensions) which lead to top limit values of I_{DSS} will also tend to increase V_P. Considering equation (3) this will be seen to reduce the spread of $g_{fs}(o)$ but to increase the variance of I_D, relative to that of either of the fundamental parameters.

From equation (9), if V_P was, in fact, proportional to I_{DSS}, the fractional variation of I_D would be identical with that of V_P. Generally, however, the spread of values of V_P is rather less than that of I_{DSS}, leading to a smaller overall variation in I_D.

An Example

Consider a transistor type C94 (Semitron) having the following characteristics:

$$I_{DSS(max)} = 5\text{mA}, \quad V_{P(max)} = 5\text{V}$$
$$I_{DSS(typ)} = 3\text{mA}, \quad V_{P(typ)} = 3\text{V}$$
$$I_{DSS(min)} = 1{\cdot}5\text{mA}, \quad V_{P(min)} = 2\text{V}$$

If a design is required, using standard resistor values, for $I_D = 100\mu\text{A}$ (typical) R_2 will need to be somewhat less than 30kΩ. The standard value of 27kΩ gives $i = 111\mu\text{A}$ (taking the typical value of V_P), and $i/I_{DSS} = 0{\cdot}037$ and thus δ, from Fig. 2, is 0·17 and hence $I_D = 92\mu\text{A}$. Taking top limit I_{DSS} and V_P, the value of I_D is 135μA and at bottom limit the drain current is 64μA.

Use of Gate 'Forward' Bias

The variation of drain current with the simple bias circuit of Fig. 1 is large, primarily due to the variability of the pinch-off voltage V_P. An improvement can be effected by applying a bias voltage E to the gate circuit (Fig. 3) of such a sense as to bias the gate in the direction of conduction. The voltage across R_2 will then rise by a similar amount and better d.c. stabilization will result.

Applying equation (1):

$$I_D' = I_{DSS}\left(1 - \frac{R_2 I_D - E}{V_P}\right)^2 \quad \ldots\ldots (10)$$

where I_D' is the new drain current with this bias voltage E applied, or:

$$I_D' = I_{DSS}\,(1 + (E/V_P))^2 \left\{1 - \frac{R_2 I_D}{V_P\,(1 + (E/V_P))}\right\}^2$$

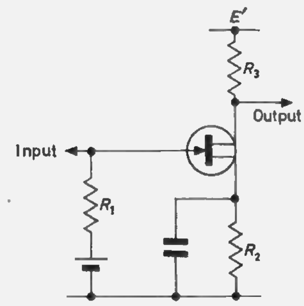


Fig. 3. Application of bias voltage to gate circuit

Writing:

$$I_{DSS}^* = I_{DSS}\,(1 + (E/V_P))^2 \quad \ldots\ldots\ldots\ldots (11)$$

and:

$$V_P^* = V_P + E \quad \ldots\ldots\ldots\ldots (12)$$

Equation (10) may be rewritten in a form identical with that of equation (4), namely:

$$I_D' = I_{DSS}^*\,(1 - (R_2 I_D / V_P^*))^2 \quad \ldots\ldots\ldots (13)$$

Thus, by analogy with equation (5), the value of the drain may immediately be written as:

$$I_D' = i^*\,(1 - \delta) \quad \ldots\ldots\ldots\ldots (14)$$

where $i^* = V_P^*/R_2$

and δ may be obtained from the ratio i^*/I_{DSS}^* using Fig. 2.

Note that:

$$i^*/I_{DSS}^* = \frac{V_P + E}{R_2} \cdot \frac{V_P^2}{I_{DSS}^2\,(V_P + E)^2} = \frac{V_P^2}{R_2(V_P + E)}$$

Usually in calculating I_D' it is most convenient to use equation (14) together with a value of δ obtained from Fig. 2. However, for purposes of analysis it will be of interest to develop an equation for I_D analogous to equation (9). By inspection, the comparable relationship is:

$$I_D' = V_P^*/R_2 \left\{1 - \sqrt{\left(\frac{V_P^*}{R_2 I_{DSS}^*}\right)} + \frac{V_P^*}{2R_2 I_{DSS}^*}\right\}$$

or:

$$= \frac{V_P + E}{R_2}\left\{1 - \sqrt{\left(\frac{V_P^2}{R_2(V_P + E)I_{DSS}}\right)} + \frac{V_P^2}{2R_2(V_P + E)I_{DSS}}\right\} \quad \ldots\ldots\ldots\ldots (15)$$

Two points about this expression are worthy of attention:

(a) Since $\dfrac{V_P^2}{R_2(V_P + E)I_{DSS}} < V_P/R^2I_{DSS}$

δ will be smaller for the case when E is non-zero, and thus variation in I_D due to variance of I_{DSS} will be smaller in this case than when E is zero, and

(b) From equations (9) and (13), if the variation of δ (due to I_{DSS} on V_P variations) is sufficienty small so that its effect on I_D or I_D' may be disregarded, then variance of drain current arises purely from changes in V_P. Against this variation the simple circuit of Fig. 1 has no stabilizing action, so that:

$$V_P/I_D \, . \, \partial I_D/\partial V_P = 1$$

but for the circuit of Fig. 3:

$$V_P/I_D' \, . \, \partial I_D'/\partial V_P = \frac{V_P}{V_P + E} < 1 \; \ldots\ldots (16)$$

In actual cases variation of δ cannot be neglected, but at least equation (13) shows that the fractional variation of drain current for a given fractional change in pinch-off voltage of the transistor is much reduced by the introduction of the bias voltage E.

Example *(continued)*

For the transistor previously described (type C94) if the circuit of Fig. 3 is used with $E = 10V$, the following characteristics result:

$$I_{DSS(max)}{}^* = 93{\cdot}5\text{mA}, \quad V_{P(max)}{}^* = 15\text{V}$$
$$I_{DSS(typ)}{}^* = 56{\cdot}0\text{mA}, \quad V_{P(typ)}{}^* = 13\text{V}$$
$$I_{DSS(min)}{}^* = 28{\cdot}1\text{mA}, \quad V_{P(min)}{}^* = 12\text{V}$$

In calculating $I_{DSS}{}^*$ a constant value of $V_P = 3V$ has been assumed. If, as is plausible, the maximum value of V_P had been used in the calculation together with the maximum value of I_{DSS} and so on, the spread of values of $I_{DSS}{}^*$ would have been much reduced. What follows, therefore, is a pessimistic estimate of the variation of drain current.

If the typical drain current is still to be about $100\mu A$, the nearest standard value for R_2 will be $120k\Omega$, making $i^* = 108\mu A$. From Fig. 2, taking $i^*/I_{DSS}{}^* = 0{\cdot}0019$, $\delta = 0{\cdot}04$, hence $I_D' = 103\mu A$. The maximum and minimum drain current values would be, respectively, $122\mu A$ and $94\mu A$.

Comparing the results obtained with those in the first example it will be seen that whereas in the latter case the maximum to minimum drain current ratio is 2·1 : 1, in the present case it is only 1·3 : 1. The improvement in stabilization is substantial.

Choice of Gate Resistor R_1

In the preceding sections it has been assumed that the potential difference between the ends of the gate continuity resistor, R_1, can be neglected. This will be the case if:

$$R_2I_D' - E \gg R_1I_G$$

where I_G = the gate current.

From this expression for the permissible maximum value of R_1 may be obtained:

$$R_1 \ll \frac{R_2I_D' - E}{I_G}$$

Since $V_P{}^*$ is only a little larger than R_2I_D', little error will result if it is substituted in the right-hand side of the expression. Similarly the inequality is adequately large if the right-hand side is about ten times the left. Hence a convenient 'rule of thumb' for R_1 is:

$$R_1 < \frac{V_P{}^* - E}{10\,I_G} = V_P/10\,I_G \; \ldots\ldots\ldots\ldots\ldots (17)$$

The value of I_G and V_P must be determined at the highest operating temperatures likely to be encountered (since I_G increases exponentially with temperature and V_P decreases slightly), and for a worst case design the highest permissible I_G and lowest V_P for the transistor type being used should be adopted. Thus the type C94 considered in the previous examples has $I_{G(max)}$ at 25°C equal to 10nA and $V_{P(min)} = 2V$. Thus at 25°C a value of R_1 of up to $20M\Omega$ would be acceptable, but for 100°C operation the equivalent value would be $300k\Omega$. Field effect transistors having much lower gate leakage currents (for example C94A having $I_{GS(max)} = 50pA$) are commercially available and would be preferred when the amplifier is required to have a high input resistance, and yet to operate at elevated temperatures.

Choice of R_3, the Drain Resistor and Supply Voltage

Because the potential difference across the gate channel junction of an f.e.t. is greatest at the drain end the normal mode of over-voltage breakdown is from drain to gate and thus manufacturers' ratings normally specify a maximum for this potential difference. When the transistor is not conducting its drain potential will rise to the supply potential, E', therefore the maximum permissible supply voltage:

$$E' < E + BV_{DGO} \; \ldots\ldots\ldots\ldots\ldots (18)$$

where BV_{DGO} = Drain-gate breakdown voltage.

To avoid bottoming the f.e.t. the drain-source voltage must not fall below V_P. Thus if the peak a.c. signal voltage at the drain is V_{sig}, to avoid bottoming:

$$E' - V_P - (R_3 + R_2)I_D' - V_{sig} > 0$$

or, since R_2I_D' is not much different from $(V_P + E)$:

$$R_3 < \frac{BV_{DGO} - 2V_P - V_{sig(pk)}}{I_D'{}_{(max)}} \; \ldots\ldots\ldots\ldots\ldots (19)$$

(using equation (18))

in the worst case. Note that some f.e.t.'s are commercially available for which $BV_{DGO} < 2V_P$. Evidently they are unsuitable for use in amplifiers of the type being considered here. It is also of interest that the value of R_3 is not dependent on E.

As well as preventing bottoming with maximum I_D', it is also necessary to ensure that the f.e.t. just fails to cut off at the opposite extreme of gate voltage excursion. For this to be so:

$$I_D'R_3 > V_{sig(pk)}$$

or:

$$R_3 > V_{sig(pk)}/I_D'{}_{(min)} \; \ldots\ldots\ldots\ldots (20)$$

in the worst case, assuming a signal waveform symmetrical about the zero line. Usually equations (19) and (20) are treated as equalities in order to permit a value for R_3 to be computed.

Voltage Gain

The voltage gain of a simple amplifier as in Fig. 1, is easily shown to be:

$$A_v = g_{fs} \, . \, \frac{R_3r_{DS}}{R_3 + r_{DS}} \; \ldots\ldots\ldots\ldots\ldots (21)$$

where r_{DS} is the incremental drain-source resistance.

The value of r_{DS} is a function of the working point, increasing with increasing V_{GS} or falling drain current. Under the working conditions assumed, in which the transistor is almost cut off, it is likely to be of the order of megohms, and thus has a negligible shunting effect on

R_3 in almost all areas. The value of g_{fs} can be derived from differentiation of equation (1) and hence:

$$A_v = \frac{-2R_3 \sqrt{I_D I_{DSS}}}{V_P} \quad \ldots\ldots\ldots\ldots (22)$$

But, substituting in a value for I_D from equation (6)

$$A_v = -\frac{2R_3 I_{DSS}}{V_P}\delta \ldots\ldots\ldots\ldots (23)$$

Similarly for the circuit of Fig. 3

$$A_v' = -\frac{2R_3 \sqrt{I_{DSS} I_D'}}{V_P} \ldots\ldots\ldots (24)$$

$$= -\frac{2R_3 I_{DSS}}{V_P}(1 + E/V_P)\delta \ldots (25)$$

These results are obtained subject to two assumptions.

(a) The capacitor C which by-passes R_2 is sufficiently large to eliminate feedback effects. This will be true if the operating frequency ω satisfies the condition.

$$\omega \gg \frac{1}{R_3 C} \ldots\ldots\ldots\ldots\ldots\ldots (26)$$

(b) That no load is connected in parallel with R_3 of such a magnitude as to shunt it significantly. This will normally be the case when f.e.t. stages are cascaded.

High Frequence Response

Transmit time effects in the f.e.t. are negligible in the audio frequency range and thus the main frequency response determining factors are the interelectrode capacitances of the device. The effective input capacitance is augmented by Miller effect and is

$$C_{in} = C_{GS} + (1+A_v)C_{GD} \ldots\ldots\ldots\ldots (27)$$

where the subscripts have the usual significance.

Typical values are of the order of a few tens of picofarads. When identical stages are cascaded the upper break frequency of the inter-stage coupling will be

$$\omega_1 = \frac{1}{R_3\{C_{GS}+[1-A_v](C_{GD}+C_{DS})\}}$$

assuming r_{DS}, $R_1 \gg R_3$

For large A_v, to fair approximation:

$$\omega_1 = \frac{-1}{R_3 A_v C_{GD}} = \frac{-1}{R_3^2 g_{fs} C_{GD}} \ldots\ldots\ldots (28)$$

and this falls as the square of R_3. Thus high gain designs have a more than proportionately restricted bandwidth.

Temperature Effects

Provided that R_1 is chosen as discussed, the main thermal factors affecting f.e.t. drain current are:

(a) I_{DSS} falls at the rate of 0·6 per cent/°C.

(b) V_P falls at the rate of 2·2mV/°C.

Both these coefficients are approximate and depend somewhat on the construction of the device. For a true worst case design, therefore, minimum values of V_P and I_{DSS} should refer to the maximum operating temperature, and maximum values to the minimum temperature. However, unless the operating temperature range is unusually large the variation of these parameters between different f.e.t.'s of the same nominal type will be much greater than those due to temperature changes.

Completion of Design Example, and Experimental Results

Using the methods detailed, for a transistor type C94 for which BV_{DGO} is 15V, the calculated value of R_3 is 22kΩ and the calculated voltage gain A_v is 9.

The circuits of Figs. 1 and 3 were constructed and in the first case the drain current had a mean of 122μA with a standard deviation of ±15 per cent for a batch of six randomly selected C94 transistors, while in the second case the mean was 103μA and the standard deviation ±4·5 per cent.

The gain in the first case had a mean of 12 and standard deviation ±8 per cent and in the second 11·7 with s.d. of ±7 per cent.

Although the batch used was too small to yield any significant statistical evidence it will be seen that the design procedure proposed yields results sufficiently close to be acceptable.

Discussion

Despite the non-linear dependence of drain current on gate-source voltage and a wide spread of device parameters, the design procedure for resistance coupled f.e.t. amplifiers presents little difficulty.

Considering equations (22) and (24) it will be noted that voltage gain is maximized (for a given value of drain current and drain resistor) if f.e.t.'s having the highest possible value of I_{DSS} and the lowest possible V_P are used. Recently transistors having I_{DSS} up to 70mA have become commercially available. With transistors of this type voltage gain per stage of up to 50 may easily be obtained.

Acknowledgments

The author would like to acknowledge the value of discussions held with his colleagues Dr. J. Watson and Mr. R. Murray-Shelley, and also the assistance of Messrs. Semitron Ltd in providing the batch of transistors used.

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A Cathode-ray Tube Display System

This new display system has been designed by the Automation Systems Division of Ferranti Ltd to meet the growing need for high definition c.r.t. displays driven directly from digital computer outputs.

The system is simple to operate and provides the following separate functions which may be used separately or combined to give complex displays.

Tabular display: A tabular format of 8, 16, 32 or 64 lines can be selected and each line can have 16 or 32 full size characters, or 32 or 64 small size characters. Large and small characters can be used together in any line and the programme is arranged to jump over blank areas so that storage of text is not wasted on large blank spaces.

Symbol display: The symbols and characters can be positioned singly or in groups anywhere on the display. The symbols are normally 4 times the character size although half size symbols can be displayed at the same time if required.

Line drawing: Vectors can be drawn from any point on a 256 × 256 matrix covering the c.r.t. to any other point on the matrix.

Circle drawing: Circles of a radius up to the full width of the c.r.t. can be described about any point on the 256 × 256 matrix.

Graph plotting 1: Graph points can be drawn anywhere on a 1024 × 1024 matrix.

Graph plotting 2: A facility is available which allows plotting on the 1024 × 1024 matrix but with automatic incrementing in the X axis. This halves the storage required for graphical information where this mode can be used.

Each mode is preceded in the text programme by a control word and the various modes can be combined in any order. Any display or part of a display can also be routed to any of the c.r.t.'s driven by the system.

The Analysis and Performance of Transistor Choppers

By D. J. Finlay*, M.Sc. (Eng.), A.M.I.E.E.

When a very small direct current or voltage is to be amplified, it is sometimes necessary to use choppers. Many circuits have been published for transistor choppers, and their mode of operation is taken for granted. However, the effects of the various defects of the chopper on overall performance are not obvious, and time can be wasted on unsatisfactory designs. This article gives a full, original analysis, with an equivalent circuit for an input transistor chopper, which enables the effects of variation of source and load resistances, temperature, chopping frequency, and transistor parameters, to be more readily understood. Typical figures are quoted and compared for the zero errors and their drift in germanium and silicon junction transistors. The error caused by transients is shown to be serious, and it is suggested that more precise information on the transients and their drift is needed when currents of the order of 1nA or voltages of the order of 10μV are to be measured.

A simple circuit for removing the distorted transients from the output, thereby linearizing the transfer characteristics, is also described.

(Voir page 624 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 631)

A CONVENTIONAL technique for amplifying a very small d.c. signal is to convert it, in a modulator, to an alternating signal of proportional size, and pass this through an a.c. amplifier. This avoids the errors caused by amplification of small drifts of the operating conditions in the input stages.

Mechanical choppers have been extensively employed for this purpose. They provide on-off switching, and convert the direct signal into square or rectangular-wave a.c. However, they are subject to trouble in the contacts and moving parts, and take appreciable power for their operation. Photo-electric choppers are also employed, in which a light source, illuminating a photocell, is periodically interrupted. Again, mechanical complexity is a drawback.

Semiconductors offer an attractive alternative to mechanical devices, since they can be made to operate as switches without the need for moving parts. An early design by Holford[1] employed four silicon diodes in a bridge modulator circuit; its performance depended on matching the diodes, and was suitable for amplifying upwards of about 2mV, 0·2μA.

Transistors make more easily controlled switches, and have been widely used in chopper amplifiers. The defects of a germanium transistor chopper, and the circuit arrangements of a complete amplifier employing it, have been described by Chaplin and Owens[2,3]. Current practice is to employ silicon transistors as choppers, and field-effect transistors are also being introduced for the purpose. A complete analysis, with measurements on these devices, is therefore desirable. The following sections show how the chopper parameters are derived in junction transistors, and how they may be used in an equivalent circuit and analysed for any transistor.

Errors in a Transistor Chopper

In Fig. 1, a pnp transistor is shown in common-emitter connexion. A direct current I is flowing, representing a d.c. signal; and the relative amounts of this current which are allowed to pass into the transistor and the amplifier input are determined chiefly by the polarity but partly by the magnitude, also, of the base current I_b. The performance may be deduced from the collector characteristics, although the makers' data sheets do not give these to a large enough scale near the origin, and they would have to be specially measured.

Chaplin and Owens[2] gave enlarged characteristics for the OC 71 near the origin, and showed how the common-emitter transistor resembled a closed switch of resistance 20Ω and contact drop 11mV, or an open switch of

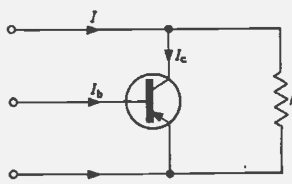

Fig. 1. Basic transistor switch

resistance 1MΩ and leakage 2μA, depending on the polarity of the base bias. They also found that interchanging the collector and emitter reduced the leakage current error to $i_{e(sim)}$, 0·4μA, and the voltage error to v_{eo}, −1mV. The physical explanation for this reduction of errors in the inverted connexion is that in a normal unsymmetrical junction transistor the collector junction has a larger area than the emitter junction; charge carriers tend therefore to diffuse more easily to the collector than to the emitter. A further development, which considerably reduced the leakage current error, was to switch off the transistor by zero base voltage instead of a reverse bias, as in Fig. 2(a). This unfortunately reduced the 'off' resistance from 1MΩ to 5kΩ at 20°C, and the value became very temperature-dependent so that it fell to 400Ω at 50°C. This could present difficulty in the choice of amplifier input resistance to follow the chopper, a major disadvantage of the germanium transistor. However, the chopper leakage current error was now reduced to about 1/150 of the diode current; with a silicon diode, the resulting error was now less than 1nA. (This current is referred to as i_L subsequently.)

A quantitative analysis of the chopper parameters, based on the classic paper by Ebers and Moll[4], showing how the current error is reduced by inversion, becoming $i_{e(sim)}$ instead of $i_{c(sim)}$, and deriving a formula for the 'off'

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resistance with base open-circuited, as measured, was also given by Chaplin and Owens. The formula for $i_{e(sim)}$ is $i_{eo}\ (1 - \alpha)/(1 - \alpha\,.\,\alpha_i)$; this can be several microamperes in a germanium transistor but about one-thousandth of this in a silicon transistor, because of the lower value of i_{eo}. Even in a silicon transistor chopper, this error can be too large and the zero-base-bias circuit is still preferred. Unfortunately, under these conditions neither the formula for $i_{e(sim)}$ nor that for the 'off' resistance has much bearing on the results. The first is invalid because of the zero bias voltage, and the second because the base current is not necessarily zero.

However, the theory has been extended to give simple approximate formulae for the voltage offset and the 'on' resistance for different values of base current (Appendix (1)). These are directly applicable to the practical circuit used. The offset voltage is given by $v_{eo} = -kT/q\beta$ and the 'on' resistance by $r_s = (kT/qi_b)((1/\alpha_i) - \alpha)$. The terms in these expressions are similar in both germanium and silicon transistors, and the magnitudes of voltage offset and 'on' resistance are therefore of the same order. Both formulae suggest that a high-gain transistor is desirable.

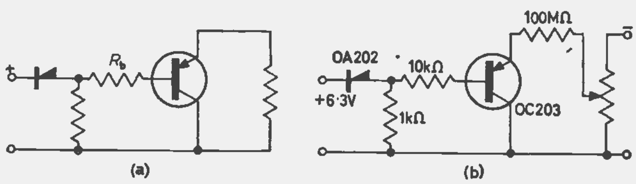

Fig. 2 (a). Switch inverted and diode added to prevent reverse bias at base (b). Measurement circuit to determine the 'off' properties of a silicon transistor

Measurements of leakage current and 'off' resistance for a silicon transistor in a practical circuit are not easy because of the very low currents and the high resistances involved. However, Dale[5] has measured the properties of an OC203 silicon junction transistor, and these are summarized with the results for the OC71 in Table 1. Inverted connexion applies except for i_{co}.

Dale's circuit for measurement of leakage current and 'off' resistance is shown in Fig. 2(b). In this, the transistor was mounted in a thermostatically-controlled oven, and the voltage at the emitter measured with a high-impedance millivoltmeter (Vibron type, $10^{13}\,\Omega$). The emitter current was then determined from the voltage across the 100MΩ resistance. Fig. 3 shows the results graphically. A point of interest is the extremely small leakage current i_L in Fig. 3(a), i.e. only 3pA at 50°C with zero emitter-collector voltage. The reverse leakage of the diode controls this to a large extent, however; if of the order of 10nA, this indicates that the ratio of current passed by the diode to the current reaching the emitter was about 3 000 to 1, compared with the 150 to 1 observed by Chaplin and Owens[3]. An improved ratio is expected, since the higher resistance of the transistor presented at the base terminal is shunted by the same value of resistance R_2, 1kΩ.

TABLE 1

QUANTITY	OC71		OC203	
TYPICAL i_{co} (makers' data)	4·5μA at 20°C	33μA at 50°C	10nA at 25°C	100nA at 100°C
	20°C	50°C	20°C	50°C
leakage current i_L	—	less than 1nA	1·5pA	3pA
'off' resistance (base open)	5kΩ	500Ω	—	—
'off' resistance (base shorted)	—	—	1250MΩ	200MΩ
v_{eo} at optimum i_b	−0·8mV	−0·8mV	−2·1mV	−2·1mV
'on' resistance r_s	20Ω	—	5Ω	—

The variation of 'off' resistance at zero base voltage is shown in Fig. 3(b). Very high values are obtained, some 1 000 times higher than indicated by the formula for 'off' resistance with zero base current. This is because base current is diverted through the external 10kΩ and 1kΩ resistances in Fig. 2(b), rather than through the high-resistance junction between base and emitter. In fact, the

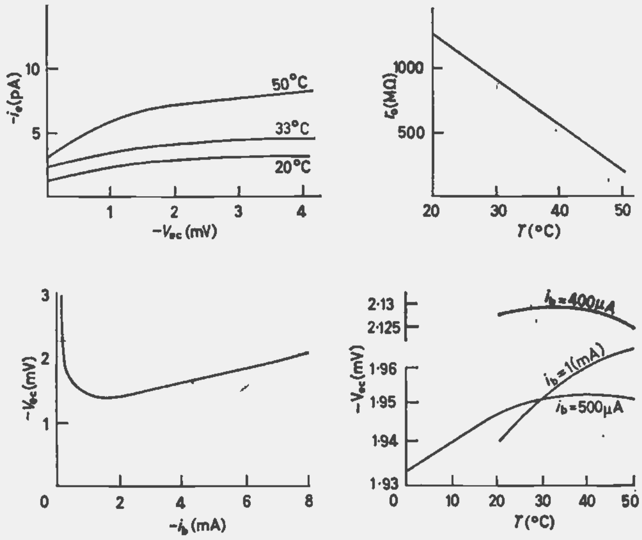

Fig. 3. Variations of switching errors in an OC203

(a) Leakage current with temperature and emitter-collector voltage
(b) 'Off' resistance with temperature
(c) Offset voltage with base current
(d) Offset voltage with temperature and base current

'off' resistance then approximates to that of the resistance of the emitter-base diode at the origin; its very high value is a major advantage of the silicon chopper transistor. In the OC71, the diversion of base current is not so marked because the resistance of the emitter-base diode is much lower, and the 5kΩ measured agrees much more closely with values obtained from the formula.

The variation of the voltage offset v_{eo} with i_b is shown in Fig. 3(c) and may also be explained theoretically. The approximate formula for this term is $v_{eo} = -k\,.\,T/q\,.\,\beta$. The only variable here is β; this rises and then falls in most transistors as the base current is increased from zero, explaining the rise and fall of the curve. A physical explanation of the rise and fall of β with current has been given by E. W. Herold[6]. The resistance of the collector material also plays a small part.

Fig. 3(d) shows the variation of offset voltage v_{eo} with temperature. An interesting point here is the levelling of the curve at base currents of 400 and 500μA. According to the expression for v_{eo}, i.e. $-kT/q\beta$, levelling can occur only if β is varying in proportion to the temperature. The level portion of this curve is at about 30°C, which indicates that the temperature and β are both increasing at about 0·33 per cent/°C; at lower temperatures, β is evidently increasing less rapidly, and at higher temperatures, more rapidly.

Transient Effects

When a switching waveform is applied to the base of a chopper transistor, large transients, often referred to as 'spikes' because of their approximately triangular shape, appear at the output. The cause of these is the finite time taken to change the charge in the base region. Analysis is not easy, since the capacitances and resistances associated with the transistor in its switching conditions are non-linear. It is found that the transients vary with the load formed by the input resistance of the a.c. amplifier, and judicious choice of this load can improve the performance. The data must be found empirically, and ranges of transistors are marketed especially for choppers, with details of transient performance for various loads. For the OC71, Chaplin and Owens[2] found typical transients at the emitter to be of 20mV magnitude, 10μsec length, when the chopper load was 500Ω. For the OC203, typical transients for a load of 1kΩ are 8·5mV, 11μsec positive, and 15mV, 5μsec negative transient, but increasing the load to 100kΩ changes the positive transient to 100mV, 30μsec and the negative one to 140mV, 1μsec. These figures indicate that a low chopper load resistance is desirable if a voltage amplifier follows the chopper. However, if the load represents the input resistance of a transistor amplifier, it is the current into this amplifier which is important, and division of voltage by resistance shows that the input current peaks are about one order less for the large load resistance. Results of similar order are published[7] for the SAC40 and 42.

In a practical chopper amplifier, the transients cause overloading of the output stages. There is little that can be done to prevent this, except to keep the transients as small as possible. To obtain more rapid switching, a base drive circuit, modified by two additional series *RC* circuits in parallel with the drive resistor R_b of Fig. 2(a), has been suggested for use with an OC44 chopper by Verster and Boothroyd[8], but choice of components for the modified circuit is dependent on measurements of parameters which are not yet quoted by the manufacturers, so that design is still difficult. The technique considerably reduces the positive transient, but does not affect the negative one.

There is a further undesirable effect of the transients which can be even more important. In all practical choppers, the input to the a.c. amplifier must be taken through a blocking capacitor which prevents the chopper in its closed condition from short-circuiting the bias circuits of the amplifier. The capacitor is charged, in opposite directions, by each successive transient. If the positive and negative transients are unequal in charge, which is proportional to the area of each, then the capacitor must discharge, when the waveform is in equilibrium, by the difference between them. In a current amplifier, it can do this effectively only during the 'on' period. Since the positive transient is usually larger in area than the negative, there is a negative current in the 'on' period, giving a step in the waveform. The difference in levels between the two parts of the waveform not carrying transients is given by $(Q_P - Q_N)/\tau$ when the 'off' resistance r_o and the source resistance of the signal are both infinite and the durations of the transients are small compared with τ, the duration of the 'on' period. Q_P and Q_N are, respectively, the positive and negative charge impulses.

Chopper Performance and Circuit Design

There are now three defects of the transistor chopper, contributing to the offset which appears at the emitter. All are affected by the source resistance of the signal and the input resistance of the amplifier. It is not always possible to make approximations, and a new and original analysis is therefore made to deduce the net effect on the input current to the amplifier. This is given in Appendix (2). The practical circuit is shown in Fig. 4(a), while the equivalent circuit, which includes all the imperfections of the chopper so far dealt with, is shown in Fig. 4(b). The analysis is only approximate as far as the transients are concerned, since it assumes rectangular pulses of current occurring wholly in the 'off' period for the positive transient, and wholly in the 'on' period for the negative transient; the magnitudes of the pulses also depend on the values of source and input resistances. The resultant waveform of current into the amplifier terminals is shown in Fig. 4(c).

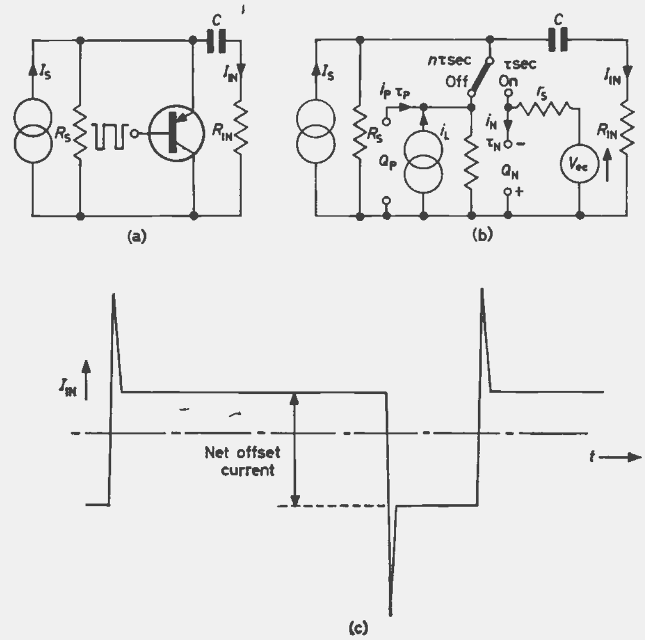

Fig. 4 (a). Current amplifier chopper with blocking capacitor at output
(b). Equivalent circuit of (a)
(c). Net waveform of current into R_{in} in (a) and (b) { **the amplitude is derived in Appendix 2**

The net difference in level between the steady portions of the waveform is

$$\frac{(n+1)\left[(I_s+i_L)R_s - v_{eo}\right]}{(n+1)R_{in}+R_s} + \frac{(Q_P-Q_N)R_s - Q_N \,.\, R_{in}}{(n+1)\tau + (R_s/R_{in})\,.\,\tau + \tau_P}\left(\frac{R_s}{R_{in}+R_s}\right)\frac{1}{R_{in}}$$

The portion of this expression due to the signal is

$$\frac{I_s \,.\, R_s(n+1)}{(n+1)R_{in}+R_s} \,.$$

This shows that, if R_s is large, the value of n chosen can influence the amplitude of the waveform, and an effective current gain can be obtained at the input. If R_s is infinite, it should be replaced by the 'off' resistance r_o in the expressions for waveform amplitude, since r_o always appears across the source in the 'off' condition, and the source is not used in the 'on' condition analysis. Making R_s infinite is the limiting condition for a current amplifier, and gives a difference in levels of the waveform, with no signal current, of $(n+1)(i_L - V_{eo}/r_o) + (Q_P - Q_N)/\tau$, assuming that r_o is much greater than R_{in}. In this expression, V_{eo} is negative, and its contribution therefore increases the waveform amplitude. The predominant term in the expression, if an OC203 is used, is $(Q_P - Q_N)/\tau$, which is, typically, $0{\cdot}05\mu A$, compared with values 3pA for i_L and 2pA *for* v_{eo}/r_o (from the values given in Table 1).

The corresponding terms for the OC71 are as follows: using the same silicon diode, the OA202, in the base drive circuit, the leakage current is some 20 times larger; becoming 60pA; the voltage offset contributes $1mV/5k\Omega = 0{\cdot}2\mu A$; and the contribution from transients is probably of the same order as for the OC203, i.e. $0{\cdot}05\mu A$. Hence the predominant term here is the voltage offset, but the transients have only a slightly smaller effect.

For a voltage amplifier, the value of R_s is small. The analysis is valid, however, if $I_s . R_s$ is replaced by V_s. The following assumptions are then made: R_s is much less than R_{in} and is also much less than $-v_{eo}/i_L$ (this last term is approximately $1M\Omega$ for OC71, or $1k\Omega$ for OC203, from Table 1); and r_o is replaced by R_s. Then the amplitude of the waveform between level portions reduces to approximately

$$\frac{V_s - v_{eo}}{R_{in}} - \frac{Q_N . f . R_s}{R_{in}},$$

where f is the chopping frequency $\left(\frac{1}{(n+1)\tau}\right)$

Taking Q_N as 40nC (based on the negative spike of 15mV, 5μsec quoted earlier) and f as 1kc/s, with R_s as

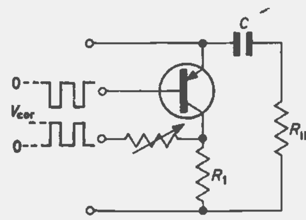

Fig. 5. Method of introducing a correcting waveform to eliminate the initial offset

50Ω, gives $Q_N . f . R_s$ as 2mV, which is of the same order as v_{eo}. Hence it may not be possible to ignore the transient offset in a voltage amplifier. However, since it is in the opposite direction to the v_{eo} error, some cancellation seems possible.

It is undesirable to pass this waveform through an amplifier, as the initial offset represents a substantial zero error which may be many times larger than the currents to be amplified. To correct the waveform, a circuit is usually provided as shown in Fig. 5. Here, the collector voltage of the chopper is raised during the 'on' period, effectively changing the contribution of the voltage offset term from

$$\frac{-(n+1) . v_{eo}}{(n+1) . R_{in} + R_s} \text{ to } \frac{-(n+1)(v_{eo} + V_{cor})}{(n+1) . R_{in} + R_s}$$

The correction waveform is derived from the current waveform through R_1. Since R_1 appears in series with r_s when the chopper is closed, it must not have too large a value. A suitable value for use with the OC203, which has a closed resistance of 5Ω, is $0{\cdot}5\Omega$. To obtain a 2mV correction amplitude would then require a current of 4mA through the $0{\cdot}5\Omega$.

In the OC71, with infinite source resistance, and assuming that $(n + 1) . R_{in} \ll r_o$, the 2mV correction voltage would allow a correction current of $-V_{cor}/r_o = -2mV/5k\Omega = 0{\cdot}4\mu A$, multiplied by $(n + 1)$, to be provided. This compares with the initial offset of $(n + 1)(60pA + 0{\cdot}2\mu A) + 0{\cdot}05\mu A$, from the figures quoted earlier. Hence this correction current is adequate for the OC71.

For the OC203, however, r_o is of the order of $1k\Omega$ at normal ambient temperatures. The corresponding correction current available is $-(n + 1) . V_{cor}$ nano-ampère. To make this equal to $-0{\cdot}1 (n + 1) \mu A$, which would in this case be sufficient, would require V_{cor} to be 100V. This is impracticable in a transistor circuit, so a method of increasing the effect of V_{cor} is required. This consists simply of connecting a resistor across the chopper, to represent a lower value of R_s or r_o in the expressions for waveform amplitude. Provided that its value is appreciably greater than $(n + 1) . R_{in}$, the signal current will be little affected. This is fairly easy to achieve in practice, since R_{in} for a transistor amplifier is often $1k\Omega$ or less. A shunt resistance of $10k\Omega$ gives $-0{\cdot}05\mu A$ correction for $V_{cor} = 5mV$. For a voltage amplifier with source resistance lower than this value of $10k\Omega$, the additional shunt is not required.

Drift

The input chopper may be adjusted for zero error at zero input, but if the temperature subsequently changes, the three offsets will drift, causing errors.

The amounts of drift, between 20 and 50°C, of leakage current and offset voltage for the OC203 have been shown in Figs. 3(a) and (d). In a current amplifier, these will contribute amplifier input drifts of $(n + 1)$ 1·5pA and

$$\frac{(n+1)\,4\mu V}{(n+1)\,R_{in} + R_s)}$$ respectively; in the latter term the $4\mu V$

is taken from the curve for optimum base current of $500\mu A$ in Fig. 3(d). The latter term is predominant if R_s is made effectively $10k\Omega$, as suggested, and then has the value $(n + 1)$ 400pA, corresponding to a signal drift of 400pA.

The corresponding signal drift of the OC71 is 30pA due to leakage current, and approximately $100\mu V/r_o$ due to offset voltage, where r_o is measured at the higher temperature and is 500Ω. This indicates a drift of $0{\cdot}2\mu A$, but this can be reduced[3] by a factor of 10 if the zero setting is carried out at 50 instead of 20°C.

It is unfortunate that no statistical information on transient drift is published, either for germanium or silicon transistors. Direct measurement of its offset is difficult, especially if a low chopper load is employed, because of the small voltage signals to be observed. Measurements by the author, on one specimen of the OC203, chopping at 1800c/s into a $40k\Omega$ load with blocking capacitor, have shown no observable drift of the 2mV offset between 20 and 50°C. The accuracy of this test was estimated at 0·1mV, in which case the drift is less than 2·5nA, i.e. a higher figure than the voltage offset drift. However, Hutcheon and Summers[9] mention that the transients decrease about 1 per cent per degree rise in temperature, which would give a drift of 0·6mV over the range tested. Further investigation would be needed on this point to minimize drift in a particular project.

In a voltage amplifier, the voltage drift is given by $-\Delta v_{eo} - \Delta Q_N f R_s$. For the OC71, the v_{eo} term would give, between 20 and 50°C, a drift of a few microvolts. (Over a batch of transistors, an optimum base current can be chosen to give a drift of not more than $100\mu V$, although this figure can be reduced for a single specimen[3]). From the transient term, the drift would, typically, be derived from $Q_N R_s = \frac{1}{2}(20\text{mV} \times 10\mu\text{sec})$ and $f = 1\text{kc/s}$, giving $Q_N f R_s = 100\mu\text{V}$ and $\Delta Q_N f R_s$ of the order of $30\mu\text{V}$ for 30°C change, based on the 1 per cent/°C quoted above. The transient term is clearly important.

Voltage and transient drifts of similar order can be expected for the OC203.

Reduction of Transient Drift in a Current Amplifier

Transient drift may be reduced at the expense of gain. The modified circuit is shown in Fig. 6. Here, the offset caused by the difference between the transients is reduced by providing an almost constant discharge path for any charge accumulated on the blocking capacitor. An analy-

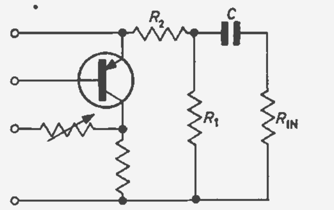

Fig. 6. Circuit giving reduced transient drift

sis, similar to that made for the simpler circuit, shows that if $R_2 \gg R_1 \gg R_{in}$, and R_s is large, the amplitude of the waveform between the level portions is approximately

$$I_s + i_L - (v_{eo}/R_2) + (Q_P - Q_N) \, . \, f \, . \, (R_1/R_2)$$

The gain is reduced by $(n + 1)$ compared with the previous circuit, but the transient error is reduced by a factor of R_1/R_2. The reduction in gain could be, therefore, typically 2 : 1 for a square-wave drive to the chopper, rising to 10 : 1 for a large mark-to-space ratio, while the reduction in transient error could be 20 : 1 or more. The transient offset drift is correspondingly reduced.

Another technique which has possibilities is to control the amplitudes of the transients independently by controlling the speed of rise of the driving waveform, which may be different from the speed of fall, e.g. by using a multivibrator drive. The transients are reduced if the rise-time or fall-time is slowed. It is remarkable that published data on transients have made no mention of the speeds employed.

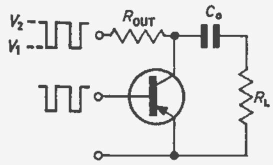

Fig. 7. Typical output chopper

Output Choppers

Larger signals must be handled by the output chopper, where the self-generated voltage and current offsets and transients are small compared with the signals being handled. These defects may be ignored. The circuit is that of Fig. 7, in which R_{out} represents the output resistance of the amplifier and R_L the load. The transistor is connected so that during the clamped period of the waveform, any signal applied to the transistor makes the collector negative with respect to the emitter, i.e. the transistor is connected as for normal operation, unlike the input chopper transistor. This is to avoid limiting the available output by current saturation in the transistor.

An analysis of the operation, assuming a perfect signal waveform, and using a technique similar to that for the input chopper, shows that R_L carries a pulsed waveform of amplitude

$$\frac{V_2 - V_1}{(n + 1)R_{out} + R_L}$$

and pulse duration $n\tau$sec, so that the mean current in the load is

$$\left[\frac{n}{n + 1}\right]\left[\frac{V_2 - V_1}{(n + 1)R_{out} + R_L}\right].$$

A fact brought out by this result is that the available output decreases as n is increased from unity. It has been

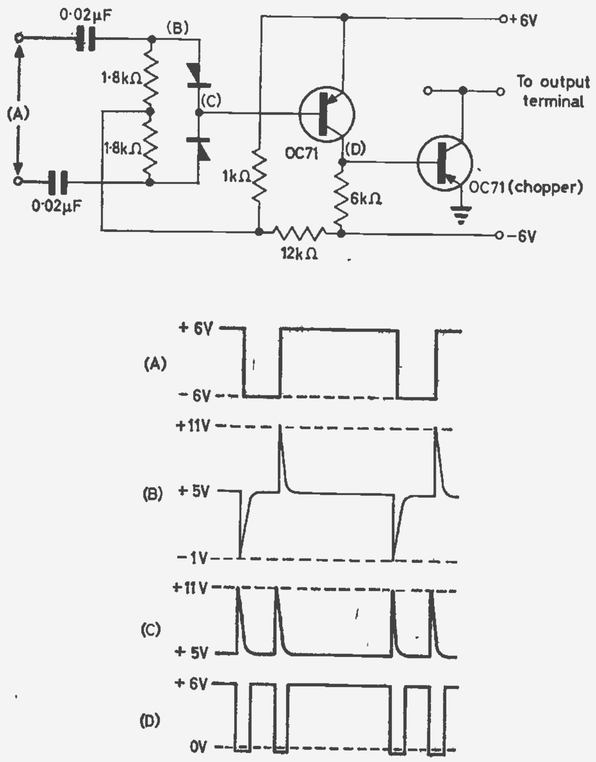

Fig. 8. Circuit and waveforms of a double-frequency chopper designed to eliminate the input positive transient

shown earlier, however, that the output of the input chopper increases as n is increased. The increased signal at the input is advantageous if noise is present. Further, if R_L is large, which may be no disadvantage if it is the resistance of a micro-ammeter, a large value of n still gives good output.

In some cases, a smoothed output is required, e.g. if overall d.c. feedback is to be applied to the system. A capacitor may then be connected across the load resistance, but to prevent its being short-circuited when the output chopper is closed, a series resistance must be added. Unlike the smoothing circuit of a conventional power pack operating from rectified sine-waves, this decreases the mean output.

A major defect of the output waveform to be handled, however, is the transients which have passed through the amplifier. In most cases, their amplitude is so high as to overload the output stages, causing clipping of the output waveform. Further, the width of the transients is larger than at the input, because of hole-storage effects in the amplifier.

Being driven synchronously with the input chopper, the output chopper clamps the same portion of the waveform to zero as does the input, and will therefore clamp the transient corresponding to the one which is negative at the input. The other is not removed at the output, however, and constitutes a large error in the output waveform and mean level. The effect is made worse by the addition of signal, because the transient peaks then change in amplitude, and the resulting output pulses become wider or narrower as a result of the hole storage phenomenon. Thus a non-linear relationship exists between input and output signals.

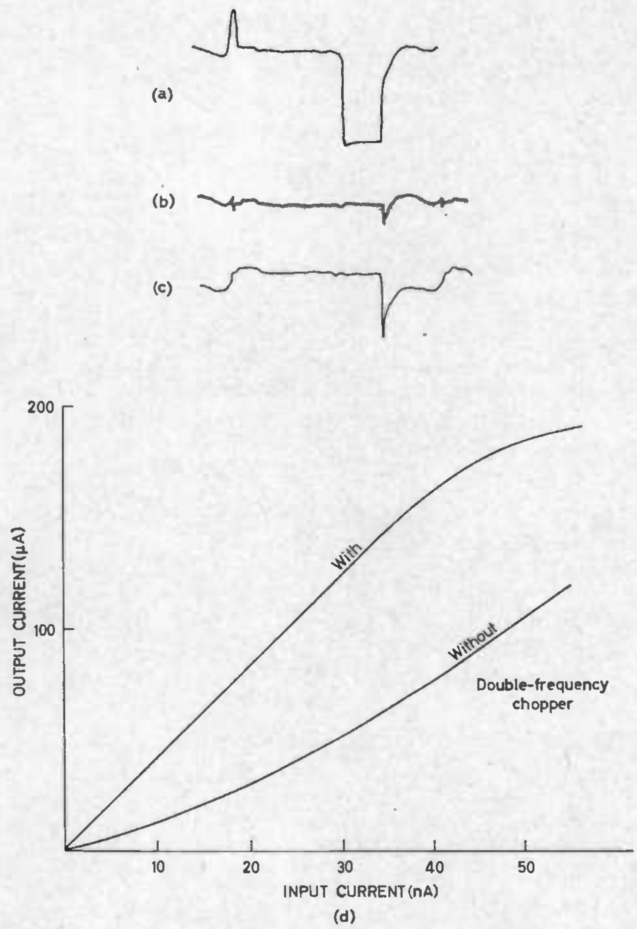

Fig. 9. Performance of a practical chopper amplifier

(a) Output waveform with no signal, before adding double-frequency chopper

(b) Output waveform with no signal, double-frequency chopper added

(c) Output waveform with signal, double-frequency chopper added

(d) Transfer characteristic with and without double-frequency chopper

A solution to the difficulty has been proposed[9] in which the opening of the output chopper is delayed by a gating and integrating circuit, so that the positive output transient is caught within the clamped period. This makes a complicated circuit, however, especially if low supply voltages are employed.

A simpler approach is to connect another chopper across the output, driven from a double-frequency waveform, which may be relatively easily derived by means of a differentiating circuit with full-wave rectifier, driven from the same source as the input chopper. The circuit and its waveforms are shown in Fig. 8. This has the slight disadvantage that the 'off' resistances of the choppers are in parallel for part of the cycle, shunting the output to some extent if these are germanium devices.

Alternatively, the double-frequency chopper may be connected into the amplifier at an intermediate stage, to clamp the signal before the transients have been amplified enough to cause limiting. This is attractive, but requires a high-impedance point for connexion, i.e. one where the impedance presented to the chopper is of the order of 1kΩ or more, which is not always easy to achieve. Also, the transients introduced by the double-frequency chopper will be amplified.

A practical point is that the double-frequency chopper is connected the opposite way round from the synchronous chopper, since it must clamp the positive transient.

Fig. 9 shows how the output waveform is improved, and the transfer characteristic linearized with increased gain, when the double-frequency chopper is added. The amplifier used in obtaining these results was a six-stage transistor version, based on designs by Chaplin and Owens[3] and employing OC71's.

Conclusions

The analysis given in this article has made possible a better understanding of transistor choppers, and has shown how the offset and drift can be predicted and minimized for junction transistors. It has been shown that suitable circuit design can give a current drift of the order of 1nA, and voltage drift of a few microvolts depending on selection of optimum base current, for both germanium and silicon choppers. More investigation is needed into transient drift, which can contribute up to about 2·5nA or 30μV. All these figures refer to a change of temperature from 20 to 50°C.

Newer devices which can be employed as choppers are the planar transistor and the various forms of field-effect transistor. Although the field-effect transistor, especially, has very different properties from those of the junction transistor, i.e. unmeasurably small offset voltage and current and poor off : on resistance ratio, the principles given are applicable, and the equivalent circuit derived for the chopper may again be used for analysis and design. It is hoped to publish performance figures for some of these devices in a future article.

Appendix

(1) Derivation of Chopper Parameters

Chaplin and Owens[2] used the basic large-signal theory of Ebers and Moll[4] to deduce relationships between the voltages and currents in a transistor in its 'cut-off' or 'saturated' conditions. Their results are as follows:

Simultaneous Leakage Currents

These are the currents flowing out of the emitter and collector when both junctions are reverse-biased:

$$i_{e(sim)} = i_{eo}\,\frac{1-\alpha}{1-\alpha\,.\,\alpha_1} \quad \ldots\ldots\ldots\ldots (1)$$

$$i_{c(sim)} = i_{co}\,\frac{1-\alpha_1}{1-\alpha\,.\,\alpha_1} \quad \ldots\ldots\ldots\ldots (2)$$

$$\text{where } \alpha\,.\,i_{eo} = \alpha_1\,.\,i_{co} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (3)$$

Collector-Emitter Potential

$$v_{ce} = \frac{k\,.\,T}{q}\ln\left[\frac{i_e(1-\alpha_1)+\alpha_1\,.\,i_b+i_{co}}{i_e(\alpha-1)+i_b+i_{eo}}\cdot\frac{i_{eo}}{i_{co}}\right] \quad \ldots (4)$$

Impedance Between Emitter and Collector for a Fixed Base Current

$$Z_{eo} = \frac{k \, . \, T}{q}\left[\frac{1-\alpha_1}{i_e(1-\alpha_1)+\alpha_1 \, . \, i_b + i_{eo}} + \frac{1-\alpha}{i_e(\alpha-1)+i_b+i_{co}}\right] \qquad \dots\dots\dots\dots (5)$$

Open-Circuit Impedance of the Chopper (Base Open-Circuited)

Putting $i_b = i_e = 0$ in equation (5) gives

$$r_o = \frac{k \, . \, T}{q \, . \, i_{co}} \, (1 + (\alpha/\alpha_1) - 2\alpha) \qquad \dots\dots\dots (6)$$

Developments from Chaplin and Owens' formulae

Putting $i_e = 0$ and assuming i_b to be much greater than i_{co} in equation (4) gives

voltage offset $v_{eo} = (k \, . \, T/q) \ln \alpha = -(k \, . \, T/q \, . \, \beta)$ approx. $\dots\dots\dots\dots (7)$

and 'on' resistance $r_s = (k \, . \, T/q \, . \, i_b) \, ((1/\alpha_1) - \alpha)$.

'Off' Resistance with Reverse Bias on Base

If $i_e = 0$, then $i_b = -i_{co}$ and the second term in equation (5) is infinite. Hence the intrinsic 'off' resistance is infinite in this condition. In practice, extrinsic leakage limits it.

(2) Analysis of Transistor Choppers

Fig. 4(b) shows the equivalent circuit of a transistor chopper. The effects of signal, voltage offset, leakage current and transients are all treated separately, and the resultant currents into the amplifier are added to give the composite waveform and its amplitudes.

Transients are represented by a positive impulse of duration τ_P, charge Q_P, and a negative impulse of duration τ_N, charge Q_N. An approximation is made by assuming the impulses to be rectangular, of current amplitudes i_P and i_N respectively.

The switch is closed for τsec, and open for $n\tau$sec.

Capacitance C is assumed very large, so that its voltage is constant.

Input Current Waveform Produced by Signal Current

Assuming $C \, . \, R_{in} \gg n \, . \, \tau$; $r_o \gg R_s$; $R_s \gg r_s$; and $R_{in} \gg r_s$; then, during the 'off' period, input current I_{in} is given by

$$I_{in} = \frac{I_s \, . \, R_s - V_{c1}}{R_s + R_{in}} \, ,$$

where V_{c1} is the voltage on the capacitor produced by signal current.

Hence the charge into C during the 'off' period is

$$I_{in} \, . \, n \, . \, \tau = \frac{n \, . \, \tau(I_s \, . \, R_s - V_{c1})}{R_s + R_{in}}$$

During the 'on' period, the input current is approximately $-V_{c1}/R_{in}$. Hence the charge leaving C during this period is $V_{c1} \, . \, \tau/R_{in}$. For a stable waveform, these charges are equal, giving

$$V_{c1} = \frac{I_s \, . \, n \, . \, R_s \, . \, R_{in}}{(n+1)R_{in} + R_s}$$

Hence the current input during the 'on' and 'off' periods may be calculated, giving a total amplitude of waveform

$$\frac{(n+1)I_s \, . \, R_s}{(n+1)R_{in} + R_s}$$

Effect of Leakage current on Input Signal

Since i_L is in parallel with I_s during the 'off' period, and both are assumed to be effectively short-circuited in the 'on' period, the formulae derived in the previous section apply except that I_s becomes i_L.

Effect of Voltage Offset v_{eo} on Input Signal

Let C carry voltage V_{c2}, caused by the voltage offset.

During the 'off' period, charge lost $= \dfrac{V_{c2} \, . \, n \, . \, \tau}{R_{in} + R_s}$

During the 'on' period, charged gained $= \dfrac{V_{eo} - V_{c2}}{R_{in}} \, . \, \tau$

Equating, and again eliminating the capacitor voltage gives a waveform amplitude of

$$\frac{-(n+1) \, . \, v_{eo}}{(n+1) \, . \, R_{in} + R_s}$$

Effect of Transients on the Input Signal

Let C carry a voltage V_{c3}, caused by the transients.

During the 'off' period transient, the amplifier input current is

$$I_{in} = \frac{i_P \, . \, R_s - V_{c3}}{R_{in} + R_s}$$

and the input charge is this quantity multiplied by τ_P.

During the remainder of the 'off' period, the discharge current is approximately

$$\frac{V_{c3}}{R_{in} + R_s}$$

and the quantity discharged

$$\frac{V_{c3} \, . \, n \, . \, \tau}{R_{in} + R_s}$$

Hence the net charge into C during the 'off' period is

$$\frac{(i_P \, . \, R_s - V_{c3})\tau_P - V_{c3} \, . \, n \, . \, \tau}{R_{in} + R_s}$$

During the 'on' period, the charge lost is

$$i_N \, . \, \tau_N + \frac{V_{c3} \, . \, \tau}{R_{in}}$$

Equating charges enables V_{c3} to be evaluated as

$$R_{in}\left[\frac{\tau_P \, . \, i_P \, . \, R_s - \tau_N \, . \, i_N \, . \, (R_{in} + R_s)}{(R_{in} + R_s) + (\tau_P + n \, . \, \tau)R_{in}}\right]$$

The amplitude of the waveform, with sign relative to the other offsets already analysed, is then

$$-\frac{V_{c3}}{R_s + R_{in}} - \left[-\frac{V_{c3}}{R_{in}}\right]$$

which reduces to

$$\left[\frac{(Q_P - Q_N)R_s - Q_N \, . \, R_{in}}{(n+1)\tau + (R_s/R_{in})\tau + \tau_P}\right]\left[\frac{R_s}{(R_{in} + R_s)R_{in}}\right]$$

with $\tau_P \, . \, i_P$ replaced by Q_P, $\tau_N \, . \, i_N$ replaced by Q_N.

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Q Measurements on Low-Loss Waveguide Cavities

By J. K. Chamberlain*, B.Sc.

The Q-factor of a chosen mode in a microwave cavity depends upon the resistive and other losses that occur in it, and, if the cavity is of right cylindrical form, is related to the attenuation constant of the waveguide of which the cylindrical walls may be regarded as composed. The article describes in outline a measuring technique using simple equipment that is suitable for the high values of Q-factor (10^5 to 10^6 and above) associated with low-loss waveguide cavities. An example is given for 2in copper waveguide at 35Gc/s, for which an attenuation constant of 3·4dB/mile is deduced.

(Voir page 625 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 632)

ATTENUATION measurements on samples of low-loss waveguide play an important part in the investigation of the different types that might be suitable for long-distance waveguide transmission, even though the values of attenuation constant so obtained may be only indirectly related to those for long lengths of guide as laid. There are two common ways of performing these measurements: by observing the attenuation of a short r.f. pulse as it travels back and forth within a closed section of the guide[1], and by measuring the *Q*-factor of a cavity whose side walls are formed by a length of the guide. Although the former technique is to be preferred because it reproduces more nearly the conditions under which the guide will ultimately be used and because it gives information about other effects such as mode conversion, it requires samples of at least several tens, and preferably some hundreds, of feet; where measurements have to be made on samples only a few feet in length the *Q*-factor method is the obvious choice.

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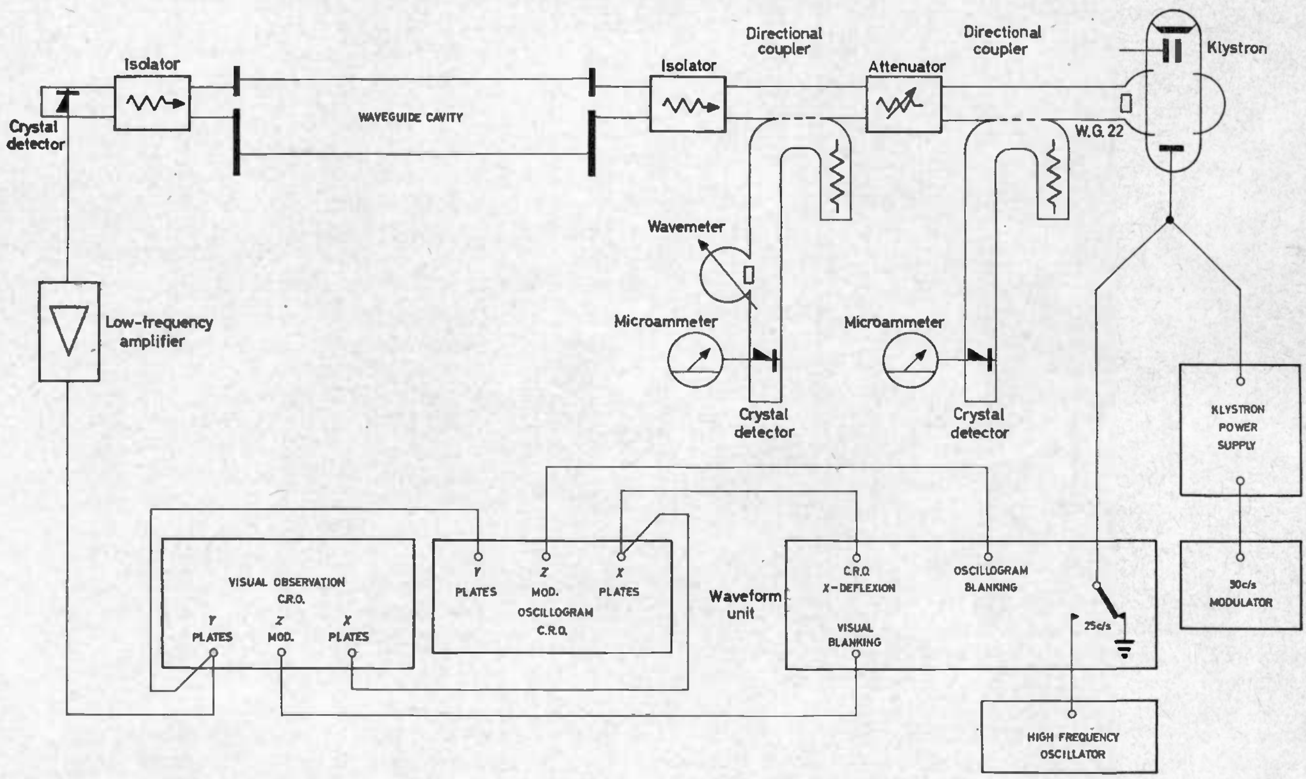
The most usual way of determining *Q* is, of course, by measuring the half-power bandwidth of an appropriate cavity resonance, taking care that coupling to other unwanted resonances is negligible. The measurement is a simple one in principle but there are practical difficulties, which increase with the value of *Q* being measured, connected with frequency instabilities of the r.f. source and the consequent uncertain correlation between instantaneous frequency and response level. A number of solutions to the problem have appeared in published accounts[2,3,4] of measurements of this sort, but the better of them tend to be rather complex. The solution described in the present note has the merit of requiring, with the exception of a small specially-built c.r.o.-blanking attachment, only standard units of r.f. and low-frequency equipment, and yields good accuracy for *Q's* in the range 10^5 to 10^6 and above.

Measuring Technique and Equipment

If the half-power bandwidth of a cavity resonance having a *Q*-factor of the order of 10^6 is to be measured

Fig. 1. Arrangement of apparatus



with an error due to frequency instability of no more than about 1 per cent, the frequency drift during the period of measurement must be limited to about one part in 10^8. It is clear that this period must be kept as short as possible if elaborate a.f.c. devices are to be avoided, and this is best effected by using a swept r.f. source (with

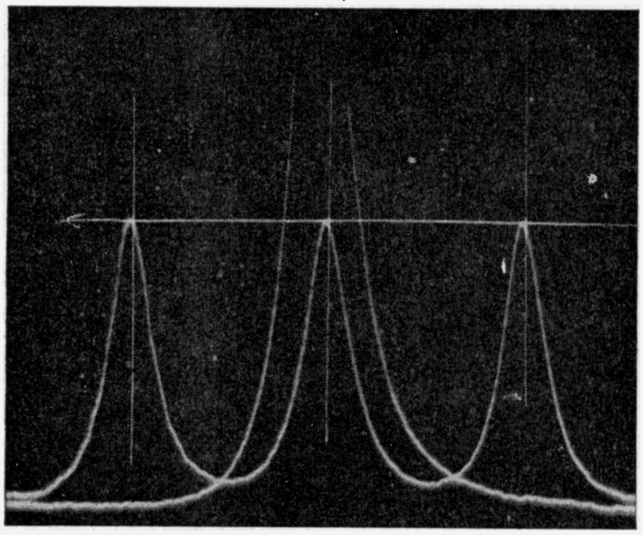


Fig. 2. Typical oscillogram: response shown is for $H_{01,228}$ resonance in 1-metre cavity of 2in diameter plain copper waveguide
Frequency spacing of calibrating peaks: 250kc/s
Resonant frequency: 34·935Gc/s
Q-factor: 5·4 × 10^5

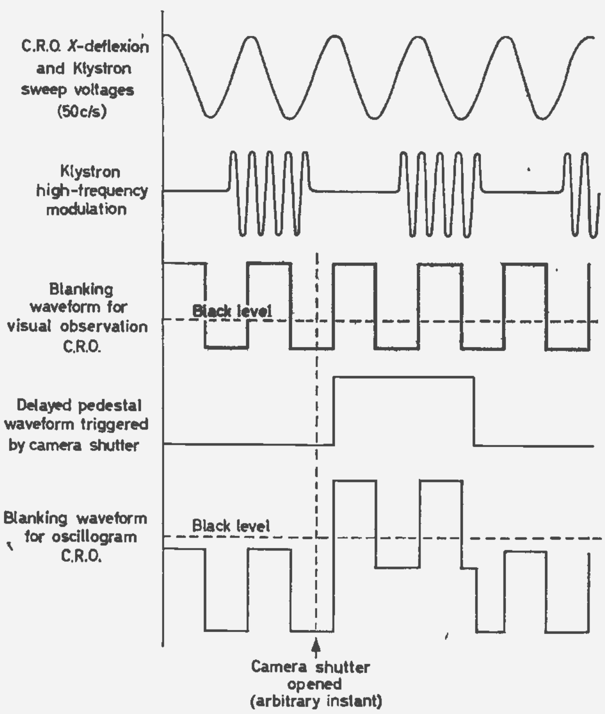


Fig. 3. Switching, sweeping, and blanking waveforms

sweep rate chosen[5] for tolerable response distortion). The response may be displayed on a c.r.t. after detection, a repetitive sweep being most convenient for selecting and observing the required cavity resonance. If normal measures are taken to stabilize power supplies and adequate warm-up time is allowed, it is found that the rate of drift of a selected response across the display is quite slow enough for observation purposes, although not usually enough for accurate response-width measurements to be made on the c.r.t. face. For the latter purpose oscillograms are recorded, preferably from a separate c.r.t. reserved for the purpose. The function of the blanking unit mentioned earlier and shown in Fig. 1 (which is otherwise assumed to be self-explanatory) is to suppress the trace on this second display at all times except during the first complete low-to-high frequency sweep after the opening of the camera shutter, and during the corresponding calibration sweep described below; this suppression is necessary to prevent the appearance on the oscillogram of successive responses with progressive lateral displacement due to frequency drift.

For the oscillograms so obtained to be of any quantitative value a frequency scale must be superimposed upon them and the half-power or some other suitable level must be indicated. Since the detector will normally operate under square-law conditions it should be possible to draw in a scale of levels without resorting to calibrating operations, although these would provide a useful check; as to the frequency scale, this need only indicate relative, not absolute, frequency. As a result of these considerations, a method which is believed to possess an element of novelty has been devised for automatically attaching to each response its own calibrating information. The method relies on the fact that if a carrier is frequency-modulated with modulation index 1·435 by a single tone the levels of the residual carrier and first f.m. sidebands become equal, each being 5·23dB below the level of the unmodulated carrier. The desired automatic calibration is achieved by switching on the modulation (which is conveniently applied to the reflector of the source klystron—Fig. 1—together with the low frequency sweeping voltage) only during alternate frequency sweeps, and choosing a modulating frequency so that three contiguous but non-overlapping responses are displayed. A typical oscillogram is shown in Fig. 2, from which is apparent the ease with which the 5·23dB bandwidth of the response, Δf, may be accurately measured.

It is readily shown that the Q-factor of the resonance is given by:

$$Q = \frac{1{\cdot}526\, f_0}{\Delta f}$$

where f_0 is the resonant frequency.

Fig. 3 shows the switching, sweeping and blanking waveforms that are generated.

The Technique in Use—an Example

The application of the technique just described is not limited to measurements on very high-Q resonators, although it is generally only in such measurements that its advantages over simpler methods become apparent. It is particularly useful for investigating resonances in low-loss microwave cavities, whether these are of interest as resonators *per se* or, as in the work for which the technique was developed, merely as means for the measurement of wall loss. In either case, the unloaded Q (Q_0) of the cavity must be deduced from the loaded (measured) Q (Q_L) through measurements of the insertion loss at resonance (L dB) and use of the well-known relation[6]

$$L = 20 \log (1 + (Q_L/Q_0))$$

To distinguish wall loss from end-plate loss it is usually necessary to perform measurements on two, otherwise-similar cavities of different lengths. In some simple cases, however, an alternative procedure is possible, which is

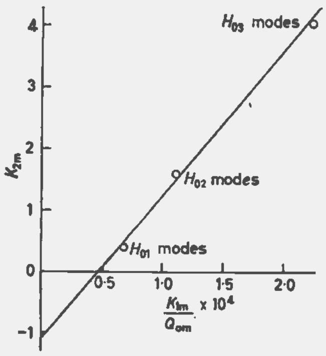


Fig. 4. K_{2m} versus K_{1m}/Q_{0m} for the H_{0mn} series of modes

illustrated by the following concluding example to the present note.

Although, in a cavity made from plain copper circular guide, the effective conductivity of the walls to r.f. currents may differ from that of the end-plates and may itself have different values for different directions of current flow, it seems reasonable to assume that a single pair of values of effective conductivity, for walls and end-plates, respectively, will apply to all the modes of a series such as H_{01}, H_{02}, etc. One can then write:

$$K_{1m}/Q_{0m} = K_{2m}\delta_w + \delta_e$$

where Q_{0m} is the unloaded Q for mode type m, K_{1m} and K_{2m} are functions (see Appendix) of frequency, mode type, and cavity dimensions, and δ_w and δ_e are the skin depths (related to the corresponding r.f. conductivities) of the walls and end-plates respectively. Plotting K_{1m}/Q_{0m} against K_{2m}, a straight line is obtained having slope δ_w and intercept δ_e.

Fig. 4 shows some results obtained for the H_{01}, H_{02}, H_{03} series of resonances at 35Gc/s with a 47·85cm cavity made from 2in diameter plain copper waveguide: δ_w and δ_e are found to be $4{\cdot}3 \times 10^{-5}$cm (corresponding to a guide attenuation coefficient of 3·7dB/mile) and $4{\cdot}9 \times 10^{-5}$cm respectively. The attenuation coefficient improved to 3·4dB/mile on polishing the inner surface of the circular guide; this compares with a theoretical value, based on the d.c. conductivity of commercial copper, of 3·1dB/mile.

Appendix

Derivation of the Factors K_{1m}, K_{2m}

For an air-filled (strictly, evacuated) cavity, the definition of unloaded Q-factor allows one to separate the two sources of dissipation by writing:

$$1/Q_0 = (1/Q_{0w}) + (1/Q_{0e})$$

Here, Q_{0w} is the component due to wall dissipation and Q_{0e} is due to dissipation in the end-plates; they should be independent of, and proportional to, the length of the cavity, respectively.

From, and in the notation of Montgomery[7], for H_{0mn} modes:

$$1/Q_{0m} = \frac{2\pi x_{0m}^2}{\lambda(x_{0m}^2 + p^2R^2)^{3/2}} \cdot \delta_w + \frac{2\pi p^2R^3}{\lambda(x_{0m}^2 + p^2R^2)^{3/2}} \cdot \delta_e$$

or $K_{1m}/Q_{0m} = K_{2m}\delta_w + \delta_e$

where $K_{1m} = \dfrac{\lambda(x_{0m}^2 + p^2R^2)^{3/2}}{2\pi p^2R^3}$

and

$$K_{2m} = x_{0m}^2/p^2R^3$$

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Computers for Air Traffic Control

A massive, computer controlled flight plan processing system, ordered from The Marconi Co. Ltd by the Ministry of Aviation at a cost of over £1·2M, will put London ahead of the world in the automation of air traffic control services.

The system is scheduled to be installed in 1969, at the new London Air Traffic Control Centre at West Drayton, near London (Heathrow) Airport. It will replace the present, interim equipment in the London centre, and will form the basis of a plan for the full automation of air traffic control services.

It will be based on a triplicated Marconi Myriad computer system which will handle flight plans and control data for all aircraft under en-route air traffic control in the southern half of the country. The Myriad computers will automatically process the vast quantity of traffic information in a fraction of the time required by present methods.

The triplicated Myriad system will give extreme reliability and greater speed, while the use of 'touch displays' will provide a unique man/machine interface, replacing the push-button controls and bringing the Air Traffic Controller into direct, finger-tip contact with the air picture.

The 'touch' displays, which will be used widely throughout this installation, provide a completely flexible input system which is fully integrated with an output display in an electronically 'written' form.

They consist of an electronic tabular display, with a matrix of very fine wires embedded in the lower half of the display face plate. The computer writes details of the traffic on the tube in such a way that those items which the controller may wish to modify, or about which he may require more information, coincide with the touch-wires. The controller can then request information from the computer, or modify the flight plans or make decisions, by simply touching the relevant items on the display. The computer will then change the display, either to provide the information called for, or to show the results of a modification, or the effect of a decision. This display must then be confirmed before the new instructions are passed to the system and on to the pilot.

The end of each touch-wire is exposed in the surface of the face plate, while the other end is connected to a sensitive trigger circuit in the display back-up unit. When the operator touches a wire, he upsets the electrical balance of this trigger circuit. The trigger is designed to make a positive input to the computer when the wire is touched firmly. Particular care has been taken to ensure that a hesitant touch cannot cause jitter, and that adjacent touch-wire circuits are not affected.

The function of each touch-wire is determined by the computer programme as necessary, and appropriate labels can be written, from the programme, next to each touch-wire.

In this way, the touch-wires can be made to fit the requirements of the displayed information, and their functions and labels can be made to simplify the task of the operator by dictating his operational procedure in a logical sequence. In a simple case, for example, the display will list all aircraft flying in a given sector of the airspace. The touch-wires will enable the controller to select more detailed information on any one of these aircraft. Seconds later, these same touch-wires might be used to call up a display of reporting points, and the times at which this and other aircraft will pass them.

The principle of this unique display/control system was originally established by the Royal Radar Establishment, and practical equipment has subsequently been developed under licence.

A Waveform Regenerator for Amplitude Sampled Systems

By T. I. Mitchell*, B.Sc., and V. J. Phillips*, Ph.D., B.Sc. (Eng.).

One intuitively obvious way to reconstruct a signal which has been amplitude-sampled is to "join the tops" of the samples, a process which, in practice, is not nearly as simple as it sounds. The present article describes the apparatus built by the authors for this purpose.

(Voir page 625 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 632)

ALTHOUGH the growing popularity of pulse-modulation techniques in communications has led to intensive development work on such systems, little attention seems to have been paid to methods of regenerating the input waveform from the modulated pulse train. Pulse modulation systems normally utilize sampling at regular time intervals according to the well-known sampling theorem[1]. This states that if the frequency spectrum of a signal is limited to the range 0 to W cycles/second, then it is necessary to take the amplitude samples at a rate not less than $2W$ samples/second. Thus, for example, a speech signal $s(t)$ containing frequencies in the range 0 to 4kc/s must be sampled at a rate not less than 8kc/s if all the information contained in the signal is also to be conveyed by the train of sample pulses.

The sampling theorem also states that the original speech signal can be regenerated from the sample pulse train by passing it through an ideal low-pass filter whose cut-off frequency is equal to W cycles/second (i.e. 4kc/s in the above example). Since it is not possible to realize such an ideal filter physically some distortion of the signal, however small, must occur in practice. An alternative regenerator which is sometimes used makes use of a holding circuit (also known as a 'box car' circuit) as illustrated by Fig. 1(a). The purpose of this circuit is to hold the amplitude of each sample pulse constant until the next sample occurs thus converting the pulse train into a stepped approximation to the original signal $s(t)$. This process can be thought of as a crude form of low-pass filter, the frequency response falling (according to a $\frac{\sin x}{x}$ law) to zero at the frequency[2] W cycles/second. Since the filtering action of this circuit is so crude, it is often supplemented by a further stage of low-pass filtering in order to smooth out the waveform.

Intuitively, the most obvious way of regenerating the signal is to join up the tops of the samples as illustrated in Fig. 1(b). This is clearly a better approximation to the signal than the step waveform, but the practical realization of a circuit to perform this function is more complicated than would appear at first sight. This article describes a method of achieving the desired 'joining of the tops'.

Principle of Operation

If the waveform produced by the holding circuit (Fig. 1. (a)) were to be passed through an integrating circuit such as an RC network of suitable time-constant the output would consist of a series of triangular sections or ramp functions of various slopes similar to those in the required waveform of Fig. 1(b). However, a little thought will show that such a waveform would be an approximation to $\int s(t).dt$ instead of an approximation to $s(t)$ itself. It follows from this that in order to obtain the required output the integration must be carried out on a step waveform which is itself an approximation to the differential

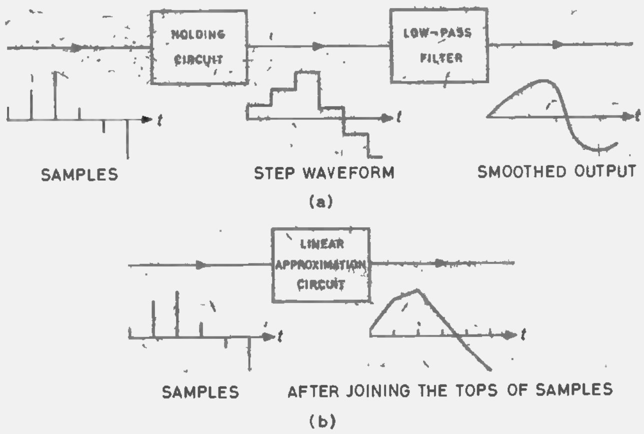

Fig. 1. Methods of regenerating amplitude sampled signals
(a) Using holding circuit
(b) Joining tops of samples

of $s(t)$. Fig. 2 illustrates just such a process. Fig. 2(a) shows a signal $s(t)$ and the pulses which result after sampling this signal, and Fig. 2 (b) shows the ideal stepped approximation to $s(t)$. This is an idealized waveform because it will be assumed for the present that it can change from level to level in an infinitesimal interval of time. It is, of course, impossible to produce such a signal in practice, and the implications of this are considered later.

The second stage is to produce from this waveform the signal of Fig. 2(c) consisting of a series of pulses whose heights are proportional to the changes in step height at each transition. These are then applied to a second holding circuit, resulting in the signal of Fig. 2(d). Finally, passing this through a suitable integrating network produces the signal of Fig. 2(e). The slope of the output waveform at any time is proportional to the height of the pulse at the integrator input. Since these are, in turn, proportional to the differences between adjacent samples, the output waveform is equivalent to that produced by joining the tops of the samples and is exactly the 'linear' approximation required. Note that waveform 2(e) is delayed one sampling interval relative to 2(a).

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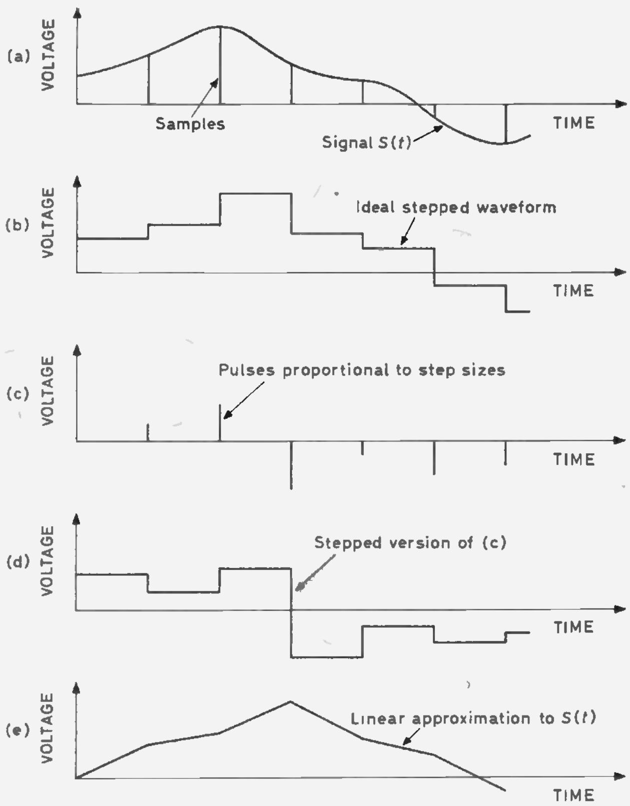


Fig. 2. Method of producing linear approximation

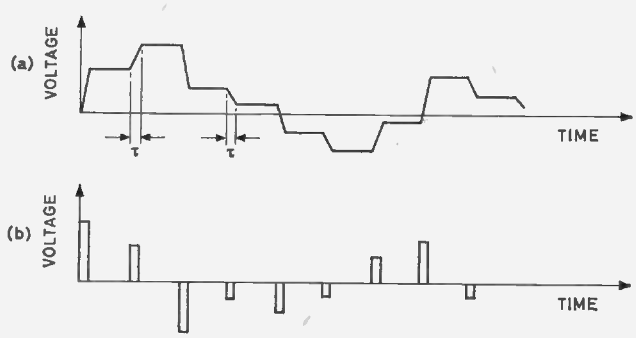
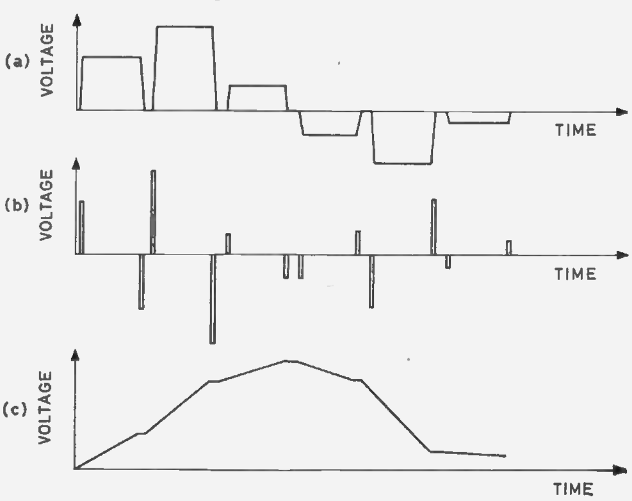


Fig. 3. (a) Stepped waveform with finite rise/fall time
(b) Waveform obtained by differentiation of (a)

Fig. 4. (a) Practical form of stepped waveform
(b) Differentiated version of (a)
(c) Integrated version of (a)



Although all these operations would seem to be very straightforward, they are complicated by the non-ideal nature of the first step waveform (Fig. 2(b)). As previously mentioned, the change from level to level cannot take place in infinitesimal time, but if this were the only imperfection in the waveform it might be possible to turn it to good advantage. If the signal were to change from one level to another in a linear manner, and if each change occupied

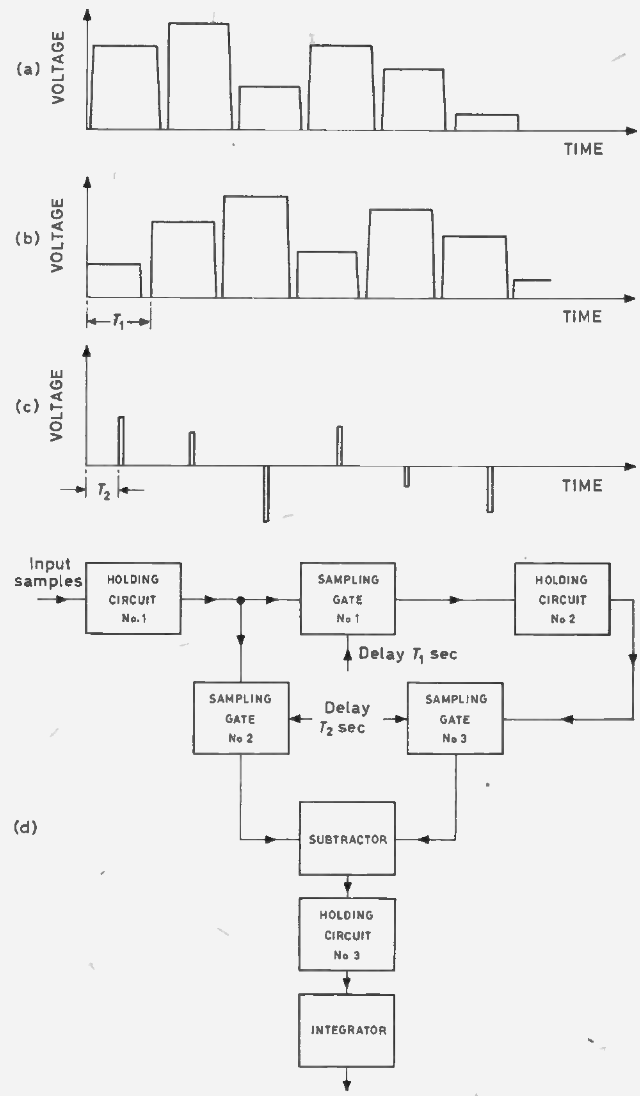


Fig. 5. (a) Stepped version of input samples
(b) Delayed version of (a)
(c) Signal produced by sampling (a) and (b) simultaneously
(d) Circuit arrangement

some definite and constant interval of time, then it would be possible to obtain the required pulses of Fig. 2(c) by differentiation. This is illustrated in Fig. 3. However, the situation is further complicated by the practical aspects of the operation of the holding circuit.

This works by charging a capacitor to a voltage proportional to the height of the input sample voltage, the time-constant associated with this capacitor being such that very little of the charge leaks away in the interval between samples. If charges due to successive samples are not to accumulate on the capacitor it is clearly necessary to discharge it completely before a new sample value is stored. Thus the actual stepped waveform produced is of the form shown in Fig. 4(a), consisting of a train of isolated pulses of varying heights separated by short gaps. In normal circumstances these gaps are of no importance, and their effects are removed by any low-pass filtering which may follow the box-car circuit. In the present case, differentiation of this waveform would produce the set of double pulses shown in Fig. 4(b), and these are quite useless as the input to the second holding circuit. It is perhaps worth mentioning here that when the second stepped waveform is integrated to form the final output signal the imperfections give rise to small 'plateaus' between the linear portions (Fig. 4(c)), but these are of quite negligible importance. To give some idea of the magnitudes of the time involved; if the sampling rate is 8kc/s, then the interval between the samples is 125μsec. With the transistors used by the authors the discharge of the capacitor can be completed in 5μsec. so that the duration of the plateaus is less than 5 per cent of the duration of the linear segments.

It is necessary therefore to find some means of generating the waveform of Fig. 2(c) which is not affected by the discharge sections of the stepped signal. It will be recalled that the amplitudes of the pulses required are equal to the differences in amplitudes between successive sample heights, and the proposed method, which is illustrated in Fig. 5 obtains these differences by direct subtraction. The original input samples are applied to holding circuit No. 1 giving the usual stepped waveform. This stepped waveform is itself then sampled, using pulses which are delayed by some time interval T_1 relative to the original samples. These new samples are then applied to holding circuit No. 2 producing what is, in effect, a delayed replica of the first stepped waveform. The magnitude of the delay T_1 must be less than one sampling interval, and if this is so, then for some portion of the time between samples the first waveform will have a value corresponding to the n^{th} sample, while the second will have a value corresponding to the previous or $(n-1)^{th}$ sample. Both of these waveforms are again sampled simultaneously at some instant of time T_2, where $0 < T_2 < T_1$. These simultaneous samples are then subtracted from one another thereby generating the desired pulse signal free from disturbances due to discharge of the holding capacitors. An alternative

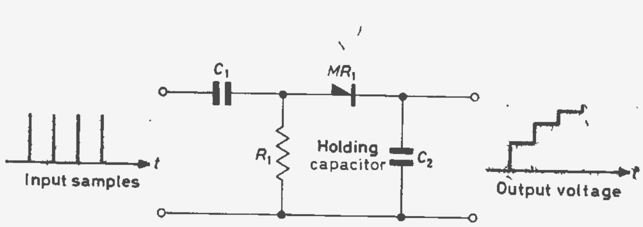


Fig. 6. Basic holding circuit

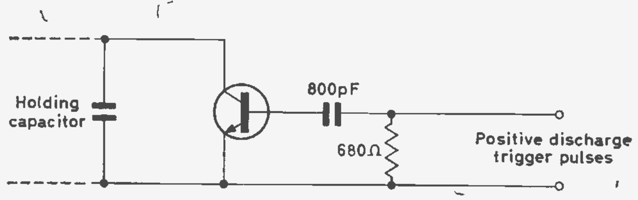


Fig. 7. Discharge circuit

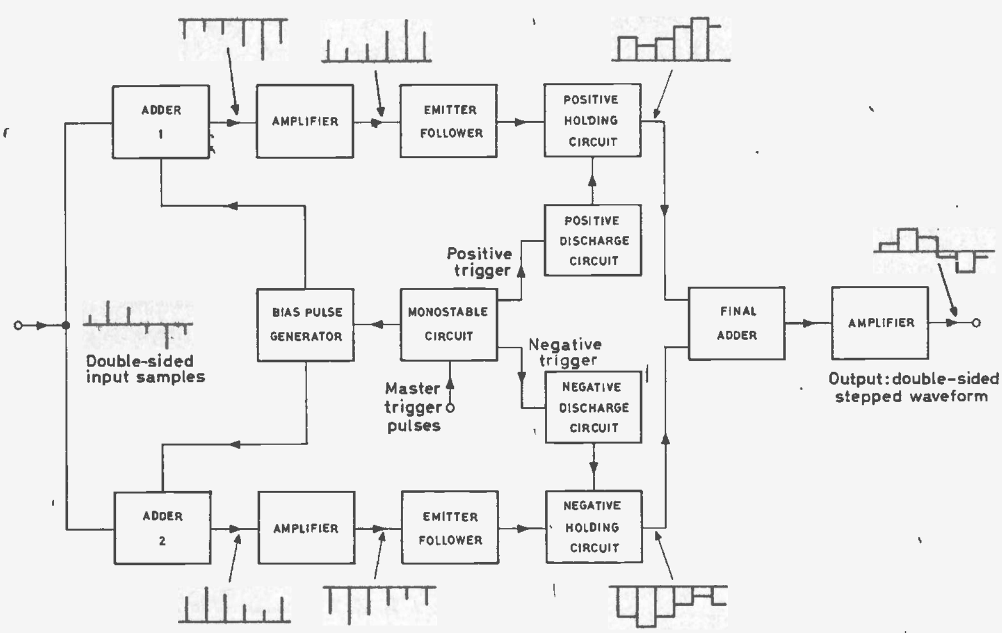
method is to subtract the two stepped waveforms, and to sample the resulting difference signal—both methods work satisfactorily. Holding circuit No.3 generates the final stepped signal which is then applied to the integrator to produce the linear approximation to the input signal $s(t)$.

Practical Circuit Details

Holding Circuits

The basic configuration of a holding circuit is shown in Fig. 6. When a sample pulse occurs at the input terminals diode MR_1 conducts, and C_2, the holding capacitor, charges up via C_1. The voltage to which C_2 charges is proportional to the amplitude of the input pulse and is determined by the ratio of C_1 to C_2. After the input pulse has ceased the diode is left in a reversed biased condition so that the charge on C_2 can only leak away at a very slow rate, and thus a voltage proportional to the sample voltage is held on C_2. During the intersample period the charge remaining on C_1 is removed by the flow of current through R_1. If further samples were now to occur, more charge would be stored on C_2—this is the familiar diode pump action. However, this is not what is required for the present purpose, and it is necessary to arrange that the existing stored charge is completely removed. A suitable discharg-

Fig. 8. Arrangement of holding circuit with double-sided output



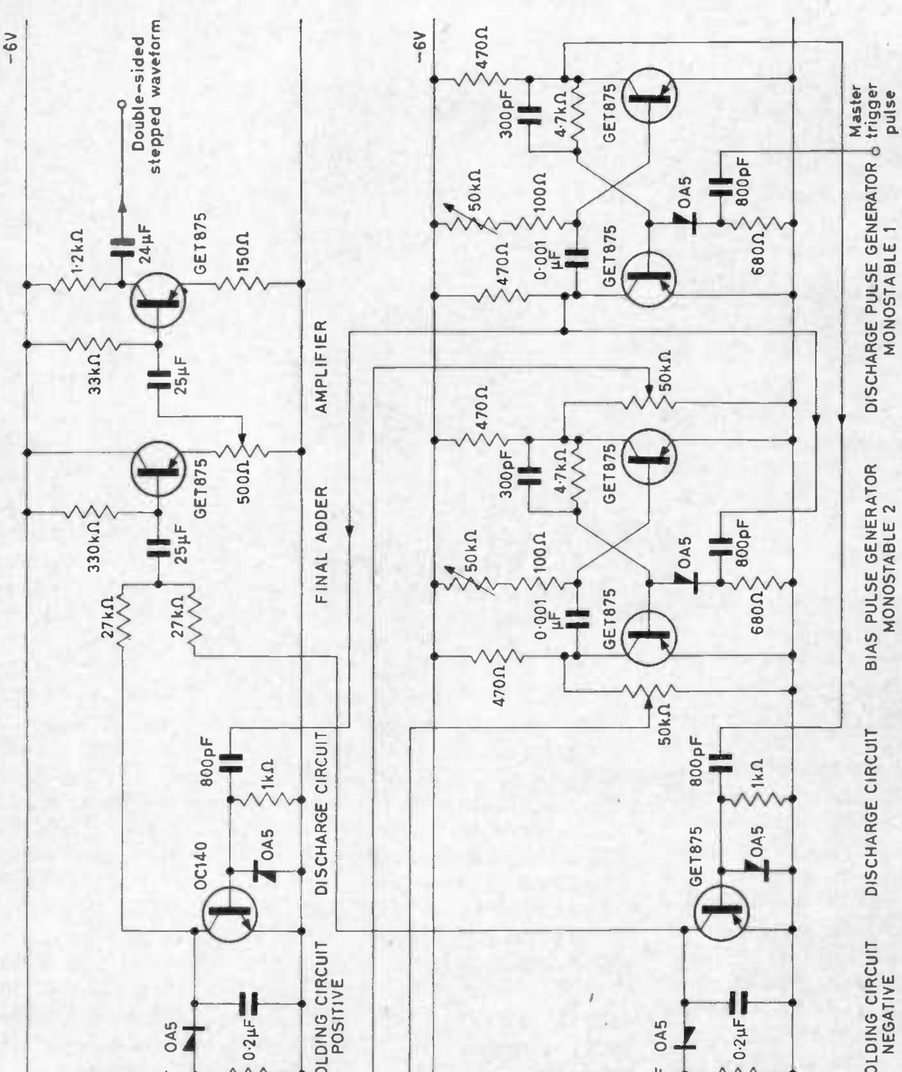


Fig. 9. Double-sided input and output holding circuit

ing circuit is shown in Fig. 7 consisting simply of a transistor connected across the holding capacitor. When a short pulse is applied to the base, the transistor conducts and completely discharges C_2.

The simple holding circuit described so far has two major limitations. First, it is only capable of handling pulses of one polarity (the so-called single - sided samples). Secondly, the diodes used in practice are not ideal, and non-linearity in their characteristics at low voltages causes distortions to occur in the smaller amplitude samples. It would be possible to overcome the first of these difficulties by connecting two diode charging circuits to the same holding capacitor, one circuit dealing with the positive samples, and the other with negative samples. Experience has shown that this is not really a very satisfactory arrangement, and the quite serious second difficulty is still present. Both limitations can be overcome at the expense of some increase of complexity in the circuit in the manner shown in block form in Fig. 8. The input samples are double-sided (i.e. positive and negative going), and to these are added bias pulses which consist of pulses identical in duration to the sample pulses themselves. Negative bias pulses are added to the input in adder No. 1, and positive bias pulses in adder No. 2. These bias pulses are of constant amplitude so that the effect is to convert the input into two sets of single-sided pulses of opposite polarities as shown. Since the adders are simple resistive networks attenuation is introduced, and the single-sided samples must be amplified before application to the holding circuits. Notice that the amplifiers introduce a phase inversion so that holding circuits of appropriate polarity must be used. The holding circuits need a low impedance input if they are to function correctly, and emitter-followers are inserted to provide this. It is clearly necessary to ensure that the bias pulses and the discharge pulses for the holding circuits occur at exactly the right instant of time. This is achieved by driving the circuits which generate them from the master trigger pulses which are synchronized to the samples themselves. In the particular apparatus constructed by the authors the master trigger pulses were in fact 'pre-pulses'—i.e. were pulses which occurred just before the samples. These were applied to a monostable multivibrator, and the leading edges of the two output pulses produced were used to operate the discharge circuits. The duration of the monostable pulse was adjustable, and the trailing edge was used to trigger the second monostable circuit which produced the bias pulse.

The outputs of the two holding circuits are now added in the final adder circuit, producing an output which is the required double-sided stepped waveform. The amplifier is used to compensate for attenuations in the various stages and to make the final signal level correspond exactly with that of the input samples (if this is required), and to produce an output of the same polarity as the input. The important point to note about this scheme of operation is that, as well as producing a double-sided output, the diodes in the individual holding circuits are always operating at a high level due to the bias pulses so that distortions at low voltages caused by non-linearity cannot occur.

It will be recalled (Fig. 5) that three holding circuits are required for the complete system. It is not necessary for the first two circuits to be double sided since their outputs are to be subtracted anyway and any d.c. bias will be removed in this way. For these circuits it is therefore sufficient to use one half of the double-sided circuit of Fig. 8. Bias pulses are added to counteract non-linearity as before of course. The final holding circuit does need to be operated in a truly double-sided manner since any overall d.c. level in the signal would cause the voltage on the capacitor to rise indefinitely. If the step waveform were perfect, and had no discharge pulses associated with it it might be possible to use a single-sided circuit in conjunction with a d.c. blocking capacitor. If a positive-going single-sided circuit was used, the discharge pulses would produce fairly large negative peaks after the blocking capacitor and this would result in the plateaus turning into regions of appreciable negative slope. It is thus much better to use a double-sided circuit at this stage.

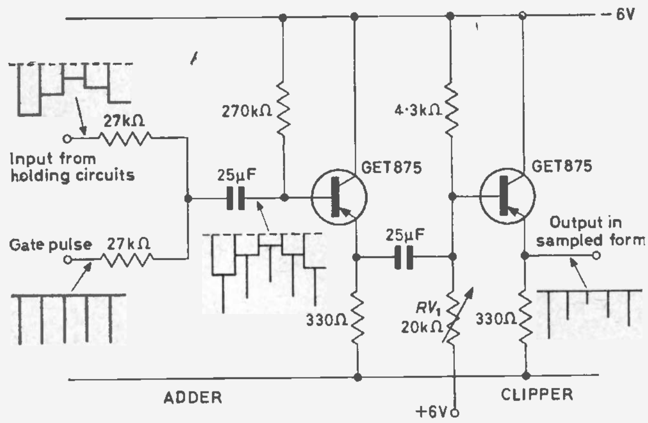


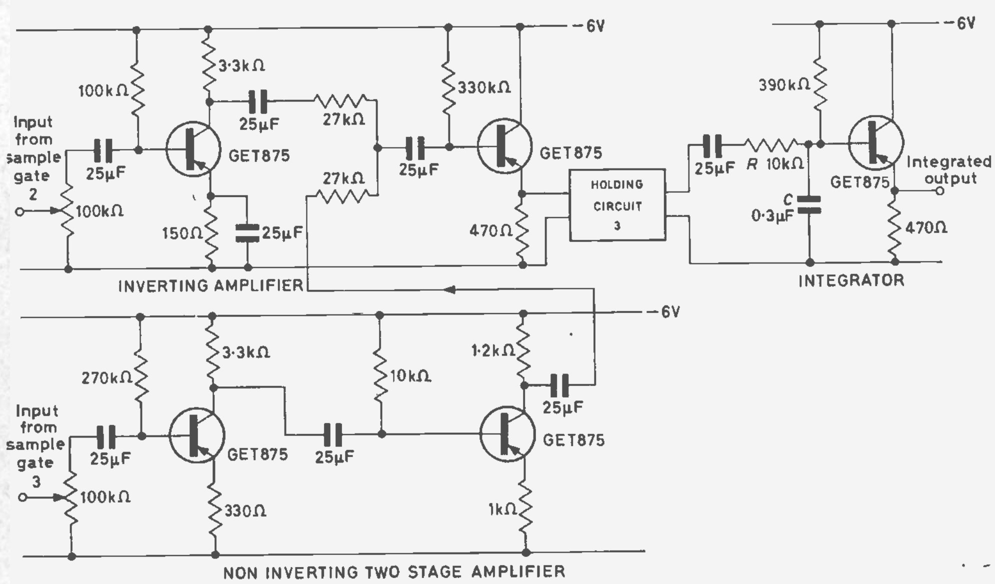
Fig. 10. Practical gate circuit

The circuit diagram of a double-sided holding circuit is given in Fig. 9. The master trigger pulses are used to trigger off a monostable circuit, the trailing edge of whose output pulse is used to trigger a further monostable stage. The duration of the first pulse is adjustable so that a variable delay is produced, and the duration of the second pulse is also adjustable so that it can be matched exactly to the width of the incoming samples. Two polarities of bias pulse are available from the two sides of the monostable, pulses of one polarity being added to the input signal in the upper adder, and the opposite polarity in the lower adder. The single-sided samples thus formed are applied to holding capacitors via amplifiers and emitter-followers in the manner previously described, and the step waveforms are finally added and amplified as shown. The discharge pulses for the holding capacitors are also obtained by differentiation of the leading edges of the outputs of the first monostable. Balancing controls are provided which can be used to equalize the heights of the two sets of bias pulses and equalize the gains of the two amplifiers so that a symmetrical step waveform is produced.

Sampling Gates

These gates are used to sample the single-sided step wave-

Fig. 11. The complete subtractor and integrator



forms, and the circuit is shown in Fig. 10. Pulses are added to the step waveform in the manner shown, and the clipping level is set so as to remove the step portion of the resulting signal, leaving only the tops of the pulses which are then effectively samples of the step waveform. If the system were being designed to work with input signals of fixed maximum excursion, then it would not be necessary to provide variable bias by means of RV_1. In this particular case, however, it was to be used with various different input signals, and it was thought advisable to make this a pre-set control. This method of sampling has the merit that the mean level of the output samples can be controlled independently of the amplitude of the input step waveform by adding different heights of bias pulse. The pulses required for this circuit are generated by a set of two monostable multivibrators similar to those in Fig. 9, so that the correct phasing of pulses relative to steps can be achieved.

Subtractor and Associated Amplifiers

The sampling gates Nos. 2 and 3 of Fig. 5 (d) produce two sets of samples which have then to be subtracted from one another, this being accomplished in the circuit of Fig. 11. One set of pulses is amplified and inverted by the upper single-stage amplifier, while the other set is amplified by the same amount, but not inverted, in the two-stage lower amplifier. The difference signal is obtained at the output of the adding network and applied to the final emitter-follower.

Integrator

This integration is carried out using an RC circuit of the usual kind, this being also shown in Fig. 11. The essential components are those marked R and C having values of 10kΩ and 0·3μF. The choice of these values is governed by the usual considerations; viz. for low values of time-constant RC the exponential nature of the charging/discharging cycle is apparent and the circuit does not act as a perfect integrator. For larger values the integration is good but the output level is low, and the final values chosen represent a compromise between these two factors. Any resistance connected in parallel with the capacitor will obviously upset the integrating action, and the output is therefore taken from the capacitor through a high impedance emitter-follower isolating stage. The particular emitter-follower used in this case was designed to work into an impedance of 600Ω, and the output level is of the order 100mV which is adequate for feeding into audio amplifiers etc. A large capacitor (25μF) is inserted at the input to the integrator to ensure that no d.c. voltage can reach the RC network.

Initial Adjustment of the Circuits

Various pre-set controls are provided in the circuits described. It has been found in practice that once these have been set to the correct values initially they require very little further attention. The most critical setting is on the inputs to the subtractor circuit, and it is worthwhile monitoring these signals occasionally to ensure that correct balance is being maintained. Since the whole system is a.c. coupled throughout the question of d.c. drift does not arise, but any changes in the gains of the earlier stages is cumulative, and results in unbalance at the subtractor output. Experience with the system has shown that very little trouble of this sort occurs, and it has not proved at all difficult to use.

The individual circuits of Fig. 5(d) are adjusted as follows:

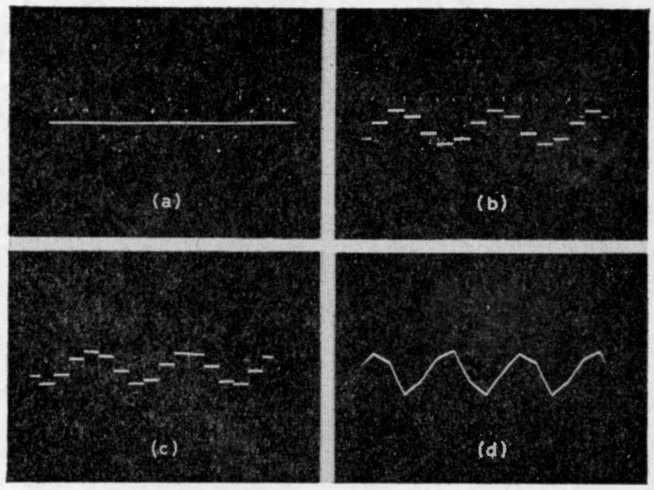


Fig. 12. (a) Sampled sinusoid. (b) Single-sided step waveform (c) Output of double-sided step circuit (d) Final output of integrator

Single-sided holding circuit No. 1; unmodulated sample pulses are fed in at the input, and the controls in the monostable circuits are operated to ensure that the bias pulses occur at the correct instants of time and have the correct duration. The amplitude of the bias pulses is set so that when the input samples are modulated, a correct single-sided set of samples will be produced. The setting of the gain control in the first amplifier is not critical, and it is merely used to ensure that the amplitude of the output step waveform corresponds roughly with that of the input samples. Sampling gates No. 1 and 2; the bias controls are set so that the clipper removes all the step waveform itself, and so that the outputs from the two gates are approximately equal.

Single-sided holding circuit No. 2; Adjusted in the same way as No. 1. Sampling gate No. 3; bias control is adjusted so that the output is equal in amplitude to that of gate No. 2.

Subtractor; a final balance is carried out on the two input gain controls of this circuit.

Double-sided holding circuit; Bias pulses of correct duration and timing are obtained by means of the controls on the multivibrators and the amplitudes of the bias pulses to the two halves of the circuit are adjusted to be equal. Input samples of constant amplitude are then applied to holding circuit No. 1, and the final output is made equal to zero by means of the two amplifier gain controls in the subtractor. The gain control on the final amplifier is merely used to obtain a convenient output from the integrator stage, and acts as a 'volume' control for the whole system. Waveforms at various stages in the system are shown in Fig. 12. Photograph (a) shows a regularly sampled sinewave which, in this case, is the input to the system. Holding circuits 1 and 2 produce single-sided step waveforms, one of which is shown in photograph (b). The output of double-sided holding circuit No. 3 is of the form shown in photograph (c). Photograph (d) shows a typical output from the integrator. For this photograph a low sampling frequency relative to the frequency of the input sinusoid has been chosen in order to illustrate the nature of the 'linear approximation'.

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Transfer Function Measurement Using Fast Pulses

(Part 2)

By R. C. French*, A.M.I.E.R.E.

(Voir page 553 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 560)

Experimental Model

An experimental model working in the slow mode was built to test the principles discussed above. A frequency range up to 100Mc/s was chosen so that existing equipment and conventional circuit techniques could be used. A block diagram of the system is shown in Fig. 4. The sampling is done in a sampling oscilloscope and the a.d.c. in a digital voltmeter. The circuits are described below.

FAST PULSE GENERATOR AND SAMPLING FILTER

The circuit diagram is shown in Fig. 5. A negative going 4·5V step from the data processing unit is amplified by the emitter-follower (ASZ21) and used to trigger the avalanche transistor (ASZ23). The avalanche pulse is positive going and has an amplitude of 1·2V, a half height width of 2·5nsec and a rise-time of 0·8nsec. The fast pulse waveform is shown in Fig. 6.

The filter circuit used to band-limit the spectrum of the fast pulse is a seven element Tchebycheff low-pass filter with a cut-off frequency of 97·5Mc/s and a reflection coefficient of 15 per cent (i.e. *T*, 07, 15)[4]. The filter has a monotonic stop-band attenuation characteristic giving 40dB loss at 155Mc/s. It is estimated that the spectrum of the filtered pulse is 60dB down at 200Mc/s so that the minimum sampling frequency is 400Mc/s. The filtered pulse waveform is shown in Fig. 7. The pulse has a width, at 50 per cent amplitude, of 5nsec and a rise-time of 2nsec. The pulse amplitude is reduced to about 0·8V by the diode (BAY38) clipping circuit.

The avalanche pulse is followed by a negative overshoot below the base line of 0·24V amplitude which continues at constant amplitude for 220nsec when a sharp return to the base line occurs. This very long pulse contains more energy than the fast pulse. This spurious pulse must either be recorded in its entirety or removed, if the record of the input and output pulse is to be valid. In the experimental model a sampling oscilloscope is used as the sampler. If the time-base speed is made fast enough to accurately sample the fast pulse, then the time-base duration or length is less than 220nsec. The spurious pulse is therefore recorded as having a length equal to the time-base length. In Fig. 8 the apparent spectral content of the fast pulse is shown calculated from measurements at two different time-base speeds. At the fast speed of 1nsec/div the spurious pulse is almost excluded and a typical short pulse spectrum is obtained. At the slower time-base speed of 10nsec/div the spurious pulse has an apparent length of 65nsec, i.e. the length of the time-base neglecting the time before the fast pulse began. A pulse having a length of 65nsec would produce a spectrum with a minima at the frequency 1/65nsec = 15·4Mc/s, and its harmonics 30·8Mc/s, 46·2Mc/s as is shown in Fig. 8. Since the time-base length cannot be long enough to completely enclose the spurious pulse, the diode clipping circuit (BAY 38) was used to suppress voltages less than +0·5V. The discussion of the effects of this spurious pulse serves to illustrate the effect of failing to record the complete waveform. The clipping circuit must precede the filter circuit because it generates harmonics of its input due to its non-linear characteristic.

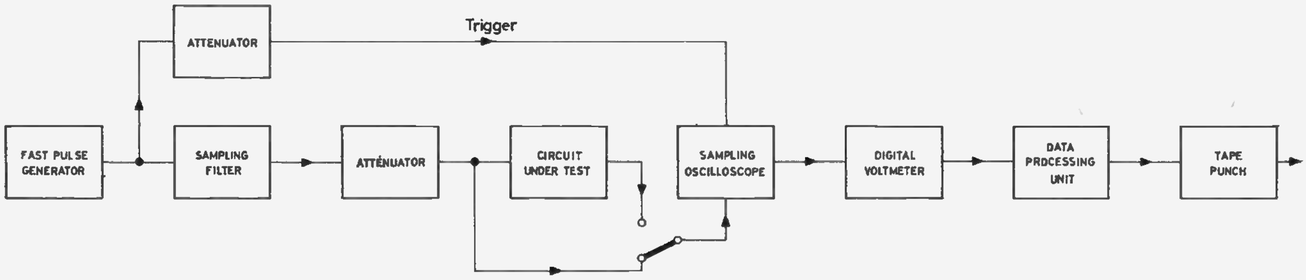

Fig. 4. Experimental model

Fig. 5. Fast pulse generator and sampling filter

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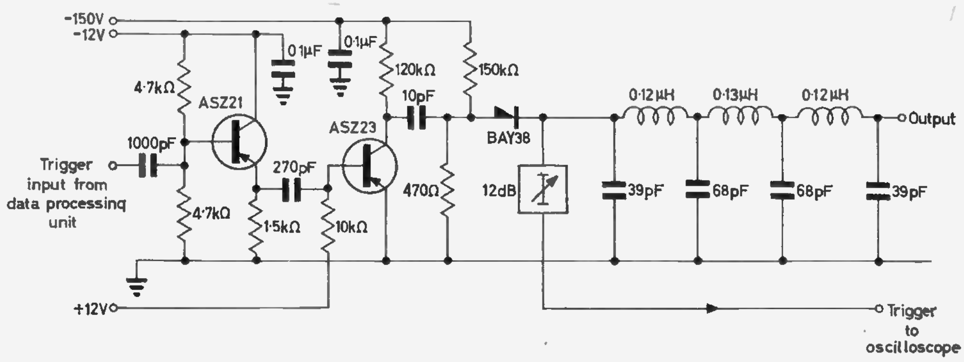

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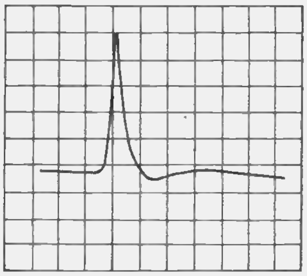


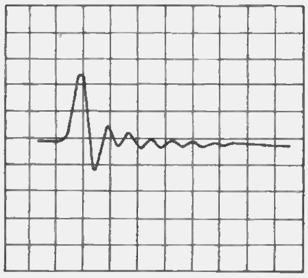


Fig. 6. Fast pulse generator output
Y = 200mV/div X = 5nsec/div

Fig. 7. Fast pulse generator output after filtering
Y = 200mV/div X = 10nsec/div

Data Processing Unit

The purpose of the data processing unit is to accept and store the output of the digital voltmeter and feed it, in serial form, to the punch at the correct time and rate. The output from the digital voltmeter is a twelve bit word presented in parallel, the words consisting of three decades, each described by four bits. The block diagram of the unit is given in Fig. 9 and the timing diagram in Fig. 10.

The measurement is started by pressing the start button on the oscilloscope time-base. Nothing happens until the 'conversion complete pulse' goes positive at time $t = 0$. At time $t = 0$ a trigger pulse is produced by the divide by two bistable BS2 which fires the fast pulse generator which in turn triggers the oscilloscope time-base. The oscilloscope takes the first sample of the fast pulse waveform and the digital voltmeter reads the sample voltage and begins to convert it into digital form. The conversion is completed at time 8msec. The negative going edge of the 'conversion complete pulse' fires the monostable MS1 via BS1. MS1 resets the store and its trailing edge fires MS2, which opens the input AND gates and reads the output from the digital voltmeter into the store. The trailing edge of MS2 passes through AND 13, which is held open for the duration of the time-base sweep, and sets BS3. With BS3 set, AND 14 passes the first punch sync pulse that arrives, say at 12msec (it may arrive at any time between 8msec and 17·1msec). The punch sync pulse passes through AND 14 and fires MS3 (4msec duration) and MS6 (9·1msec). MS3 then resets BS3 which holds AND 14 shut excluding further punch sync pulses. MS4 (4msec) is fired 9·1msec

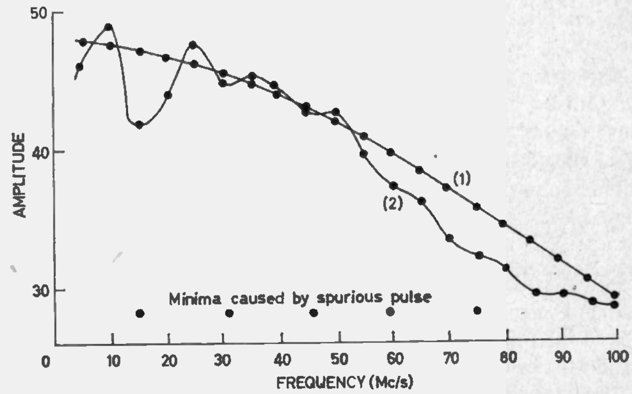
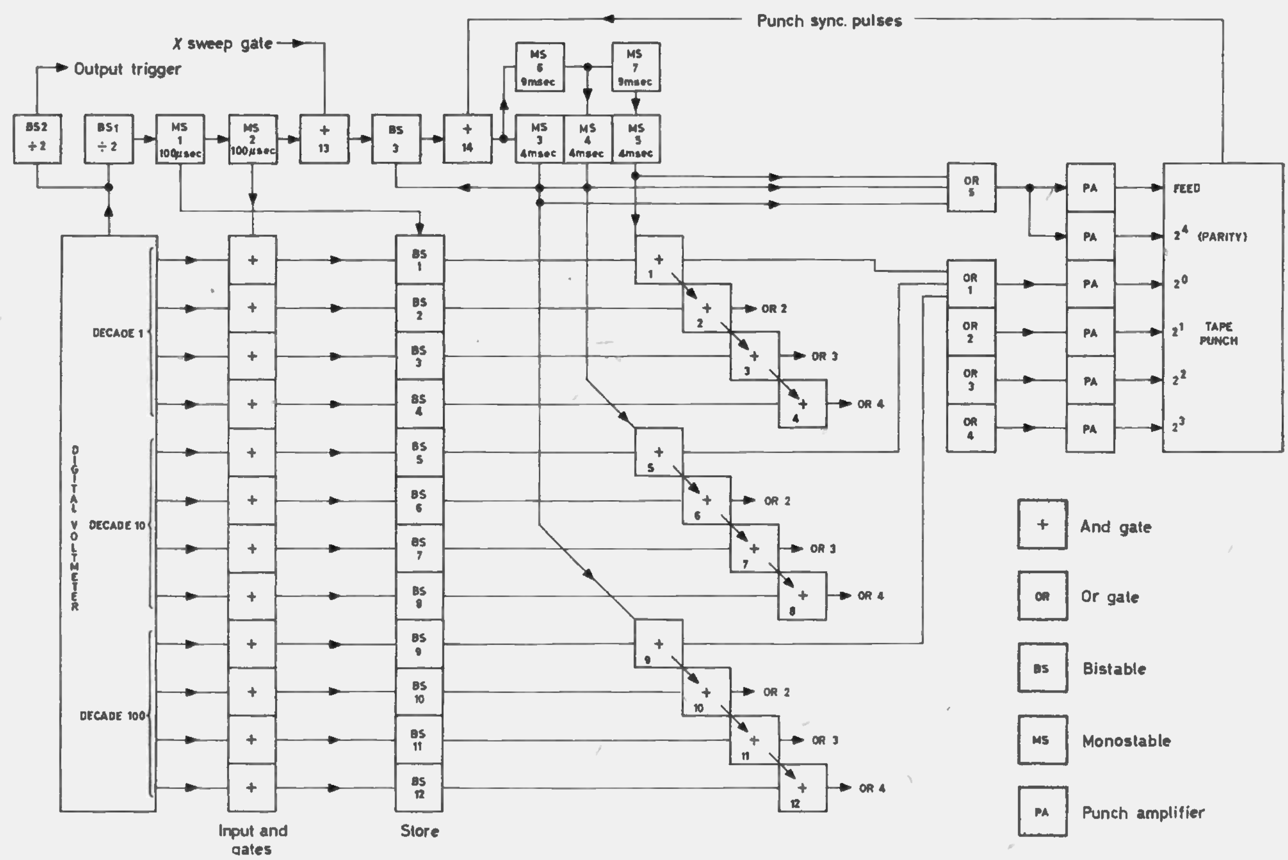


Fig. 8. Fourier spectrum of fast pulse and effect of spurious pulse
Curve 1 Time-base speed = 1nsec
2 Time-base speed = 10μsec
Length of spurious pulse = 65μsec
Zeros in spectrol contour = n/65nsec = n.15Mc/s
n = 1, 2, 3

Fig. 9. Arrangement of data processing unit



after MS3 and MS5 (4msec) is fired 9·1msec after MS4. The timing of MS3, 4 and 5 are therefore correct for punching the three decades 100, 10 and 1 respectively. When MS3 (4msec) fires, AND 9, 10, 11 and 12 pass the contents of the store describing decade 100 through the OR gates OR 1, 2, 3, 4 to the punch amplifiers PA, and cause the character to be punched. 9·1msec later MS4 fires and decade 10 is punched via AND 5, 6, 7, 8, and then 9·1msec later decade 1 is punched via AND 1, 2, 3, 4.

Each time a character is punched a feed pulse is required and these are derived by passing MS3, 4 and 5 through OR 5 to the feed punch amplifier. The feed pulses from OR 5 are also fed to the unused 2^4 level of the 5 hole tape, to punch a bit whenever a character is punched. This enables the character zero to be distinguished from a blank.

The three decades have been punched by time 34msec and 6msec elapse before the second sample is taken at time 40msec. BS1 and BS2 suppress the unwanted digital voltmeter reading at time 20msec.

The data processing unit was built with M.E.L. Equipment Ltd circuit blocks supplemented with some additional circuits. Fifty-five circuit blocks were used and the extra circuits required thirty-one transistors.

Sampler and A.D.C. Circuits

A Tektronix type 564 storage oscilloscope with plug-in sampling sweep unit 3T77 was used as the sampler. The storage facility is needed as the basic p.r.f. is only 25c/s, so that a 100 dot trace takes 4sec.

The time interval t between successive samples must be accurately known so that in the Fourier transform the spectral frequencies are correct. The method of calibration used was to connect a sinusoidal signal of frequency $1/t$ to the input of the oscilloscope. The sine wave and dot structure interfere and in the ideal case produce a horizontal row of dots. Non-linearity of the time-base shows up as departure from a straight line. A typical interference pattern is shown in Fig. 11. As can be seen from the figure the linearity is good apart from the first division where the time-base speed is high by 10 per cent. The internal delay was used to position waveforms in the linear region before recording them. The input frequency was adjusted to measure t in the second and third division where the main part of the waveform occurs.

The oscilloscope was slightly modified to provide a gate pulse lasting the length of the time-base and the time-base set to single sweep. A digital voltmeter was used for the a.d.c. The binary coded decimal output from three decades was used giving a maximum of one thousand quantization levels. The voltmeter was a Digital Measurements DM2001 set in the continuous mode and deriving its input from the vertical amplifier output of the sampling oscilloscope.

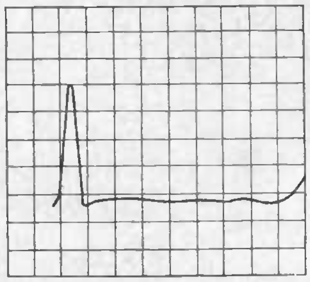


Fig. 11. Calibration of oscilloscope time-base
$Y = 200mV/div$ $X = 20nsec/div$

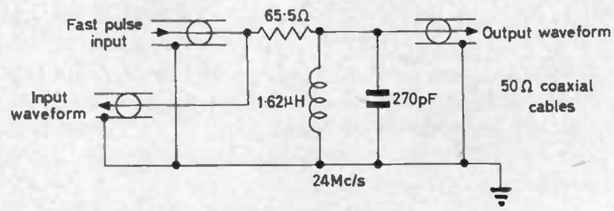
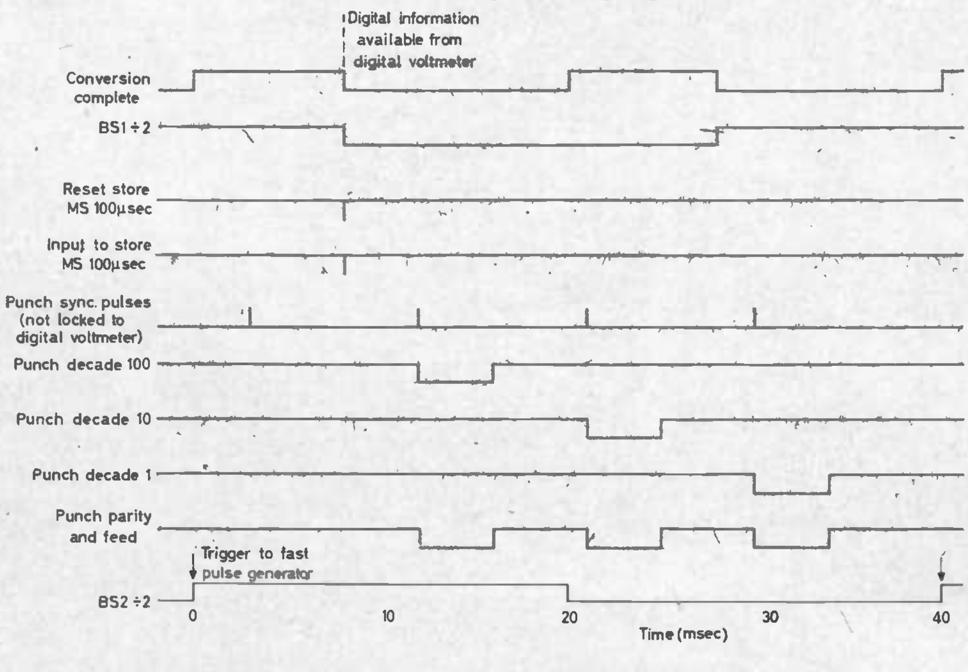


Fig. 12. Tuned circuit

Experimental Results

Initial experiments were made to verify the validity of the measurement system and find the accuracy obtainable with the equipment used. In the first test two records of the fast filtered pulse were made. One record was called an input pulse and the other an output pulse. The computed transfer function should ideally be unity as the pulses should be identical. Over the band up to 100Mc/s the worst error observed in a number of such tests was $\pm 0{\cdot}2$dB gain and $\pm 1{\cdot}5°$ phase. In this test the dynamic range of the 'test circuit' was zero so that the above figures show the errors when working at the optimum signal-to-noise ratio over the whole band. In a measurement on a real circuit some variation in gain over the band would occur, so that some frequencies would be attenuated yielding a lower signal-to-noise ratio and therefore a lower accuracy at those frequencies.

Fig. 10. Data processing unit, timing diagram



The transfer function of a number of circuits was measured as described below.

Transfer Function of Coaxial Cable

The transfer function of a 12ft length of matched 50Ω cable was measured. The gain differed from unity by ±0·3dB and the phase from a linear characteristic by ±2°.

Transfer Function of Tuned Circuit

The circuit shown in Fig. 12 was constructed and its transfer function measured. The theoretical characteristics were also determined and the two are compared in Figs. 13 and 14. The phase response was also measured with an Ad Yu precision phase detector type 205B4, and the result is shown in Fig. 14. The correspondence between the theoretical and pulse measured characteristics is worst at the higher frequencies and is consistent in the phase and amplitude characteristics. The accurcy of the Ad Yu measurements is better at the higher frequencies where a

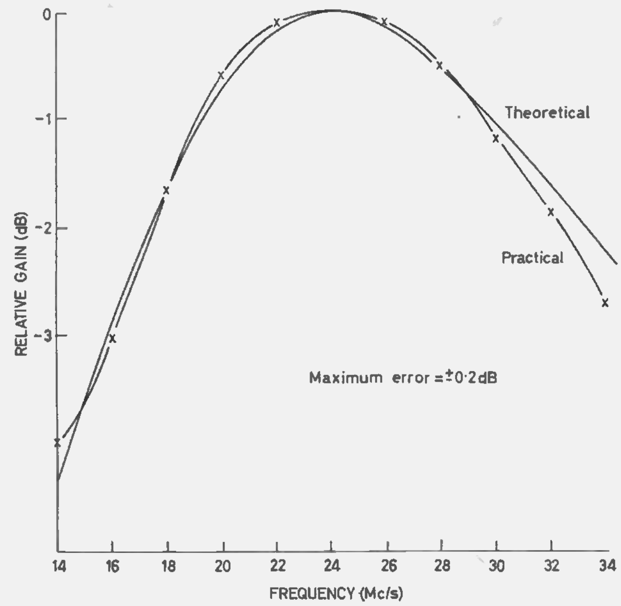
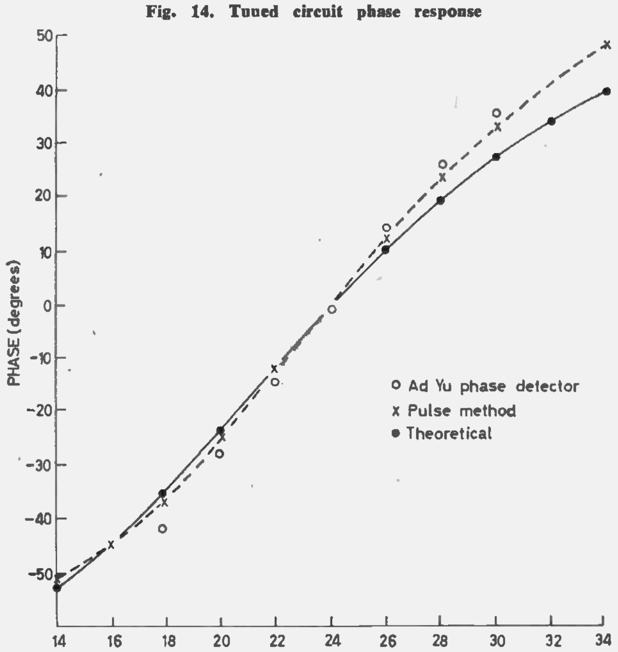


Fig. 13. Tuned circuit amplitude response

Fig. 14. Tuned circuit phase response



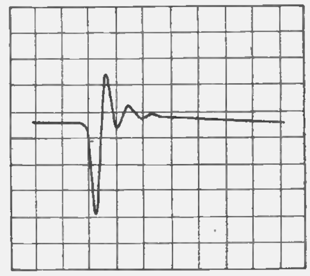


Fig. 15. Power amplifier input waveform
Y = 100mV/div X = 20nsec/div

Fig. 16. Power amplifier output waveform
Y = 100mV/div X = 20nsec/div

sharper null was obtained on the bridge. The accuracy of the pulse measurement is difficult to determine because the main difference between the phase characteristics is in the slope at the resonant frequency. These small differences in slope could be due to slight differences in the Q of the tuned circuit. Since the tuned circuit load was formed by the 50Ω cables terminated in the various test equipment, such differences are possible.

Transfer Function of Power Amplifier

A wideband transistorized power amplifier[5] was tested, and the input and output waveforms are shown in Figs. 15 and 16. The 60Mc/s cut-off frequency of the amplifier has slowed down the fast pulse and suppressed the high frequency ringing. The low frequency response is rather poor as shown by the low frequency ringing after the pulse which has a slow decay. This is caused by the low frequency peak in the gain characteristic and the corresponding peak in group delay. The gain and group delay characteristics are shown in Figs. 17 and 18. The gain characteristic was also measured on a Rohde and Schwarz Polyskop and found to agree with the pulse measurement within ±0·5dB, this is as accurate as the Polyskop can measure.

Effect of Noise in Sampler

The effect of noise in the sampling oscilloscope sampling circuits was tested by repeating the measurement of the 12ft coaxial cable using progressively smaller input pulses. The fast pulse was attenuated in a wideband variable attenuator (Hatfield Instruments LE10) before being applied to the cable. The results are shown in Fig. 19 where the peak error in the band 0 to 100Mc/s is plotted against input signal level. The curve shows that quantization noise is negligible for signal levels above about 0·3V but for smaller signals the error rises rapidly. The situation cannot be improved by increasing the input to the circuit as the oscilloscope will not accept inputs larger than about 1·5V peak-to-peak. In the pass-band of circuits where the gain is set to unity by attenuators the accuracy will be good, but in the regions of low gain where the output signal is small the accuracy will be less. An accuracy of about ±1dB of gain and ±3° of phase can be achieved with a dynamic range of 17dB which is adequate in many applications.

Averaging Technique to Improve Signal-to-Noise Ratio

The noise produced by the sampler can be partially suppressed by taking several records of the input and output waveforms and taking an average of the transfer functions computed from them. This technique was used with a certain amount of success. The scatter on the gain response was reduced by a greater amount than the scatter on the phase data and the reason for this is not understood. The technique is partly unsatisfactory as it increases the computation time. A further disadvantage is that if one record, for some reason, is very inaccurate its effect on the transfer function is partially obscured by the other records. Therefore, the results are inaccurate but not so inaccurate as to be obviously so. Such averaging techniques will be effective in suppressing random

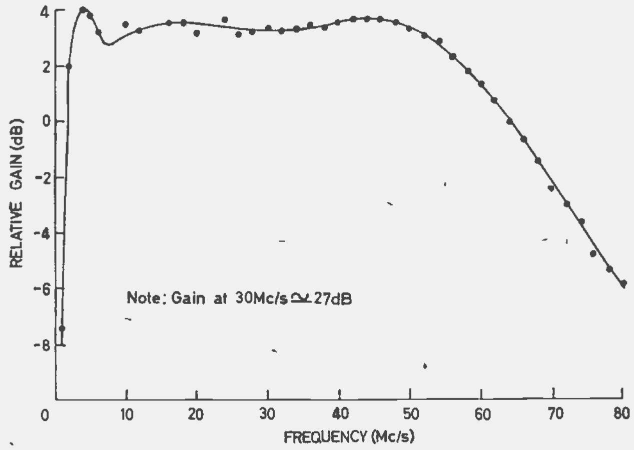


Fig. 17. Power amplifier gain response

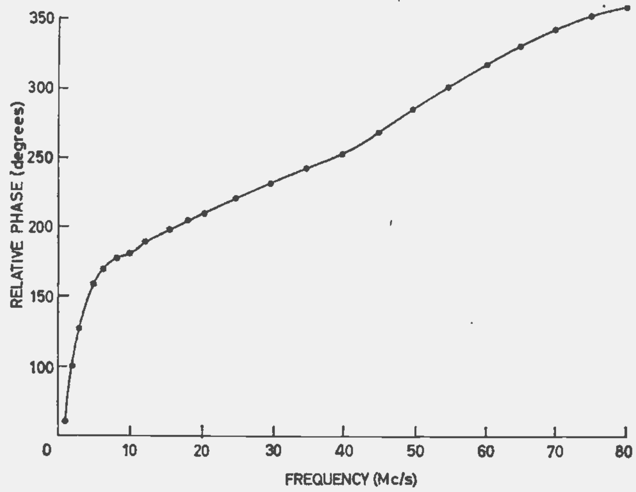


Fig. 18. Power amplifier phase response

errors but not in suppressing systematic errors such as time-base non-linearity in the sampling oscilloscope.

Conclusions

The experimental work described demonstrates that the method is an effective way of measuring the transfer function of electric networks. Examples of circuits which can be measured are narrow- and wide-band amplifiers, ultrasonic delay lines, and filters. The method may also be used on complete systems containing modulator-demodulator pairs. The word pair is included because the input and output waveforms must not be frequency shifted relative to one another. The circuits measured must be able to pass the input pulse without non-linear distortion.

The experimental model has a frequency range up to 100Mc/s and when measuring circuits with attenuation ranging over 10dB, an accuracy of 0·3dB in gain and $\pm 2°$ in phase. With a dynamic range of 20dB the accuracy is reduced to ± 1dB in gain and $\pm 4°$ in phase. This accuracy could be improved by using a custom built sampler and a.d.c. instead of the oscilloscope and digital voltmeter. The accuracy of the group delay characteristic is a function of the accuracy of the phase data and also of the spacing of the spectral lines chosen. In practice as wide a spacing as possible should be chosen consistent with following the group delay characteristic. In a test on a length of coaxial line the spectral line spacing was made 5Mc/s and the variations in group delay were found to be $\pm 0{\cdot}75$nsec. The group delay accuracy could be improved by smoothing the phase characteristic in the computer before calculating the group delay.

The measurement of phase by existing techniques is only possible on a point by point basis. The measurement time is therefore large and the number of readings made is limited. The method described is of comparable or

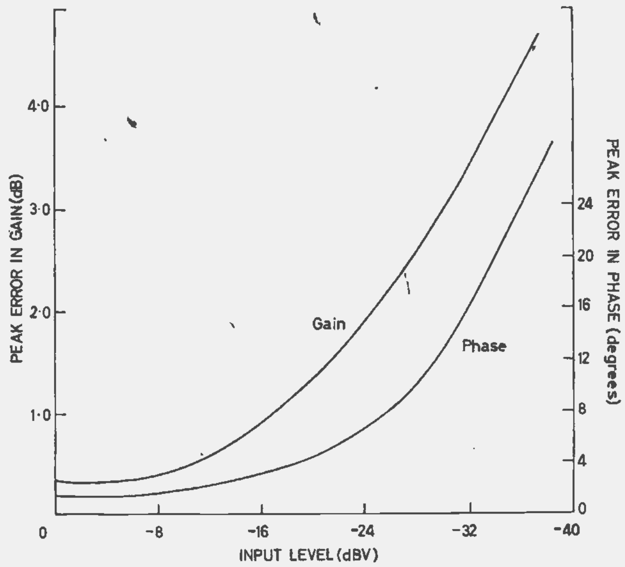


Fig. 19. Effect of quantization noise on accuracy

greater accuracy and very much quicker. The reduced measurement time avoids the great difficulty, inherent in point by point measurements, of drift in the whole characteristic due to variation of supply voltages or temperature with time. These drifts of the phase characteristic can obscure the variation with frequency which is being measured.

Since the transfer function is stored in the computer it can be manipulated easily by suitable programming. Thus, results can be plotted as graphs or printed out, they can be normalized, or compared with a theoretical transfer function. Also, the data can be used in existing computer programmes to determine amplitude or phase equalizers, and for synthesis of networks having the measured characteristics. The latter use would be valuable in determining equivalent circuits.

Acknowledgments

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Spectrum Analysis of a Train of Modulated Trailing Edge Variable Slope Pulses

By O. E. Kruse* and R. W. Montgomery*

A mathematical analysis of the spectrum of a train of triangularly shaped pulses having their trailing edges modulated is developed. The mathematically predicted spectrum is then compared with the experimentally determined spectrum.

(Voir page 625 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 632)

MODULATION schemes in which the intelligence to be transmitted is carried in the leading or trailing edges of a train of pulses have been developed and described[1,2,3,4]. Fig. 1 illustrates two trains of pulses so modulated.

The spectrum of a leading-edge variable slope pulse modulated wave has been developed and checked experimentally[5]. It is the intention here to present a mathematical analysis of a trailing-edge variable slope pulse modulation wave and to present an experimental check of the results so obtained. Refer to the train of triangularly-shaped trailing edge pulses in Fig. 2.

The voltage waveform of the train of unmodulated waves in Fig. 2 may be represented analytically by a Fourier series as follows:

$$V(t) = A_o + \sum_{n=1}^{\infty} (A_n \cos 2\pi nct + B_n \sin 2\pi nct) \quad \ldots\ldots \quad (1)$$

where:

$$A_o = (1/T_o) \int_0^{T_o} V(t)\, dt \quad \ldots\ldots \quad (2)$$

$$A_n = 2(1/T_o) \int_0^{T_o} V(t) \cos 2\pi nct\, dt \quad \ldots\ldots \quad (3)$$

$$B_n = 2(1/T_o) \int_0^{T_o} V(t) \sin 2\pi nct\, dt \quad \ldots\ldots \quad (4)$$

In the above equations:

$$V(t) = \begin{cases} -pt + E & \text{for } 0 < t < \text{(pulse duration)} \\ 0 & \text{for (pulse duration)} < t < T_o, \end{cases} \quad \ldots\ldots \quad (5)$$

where p is considered to be positive.

When $V(t)$ first becomes zero,

$$t = E/p; \text{ therefore, pulse duration} = E/p \quad \ldots\ldots \quad (6)$$

Employing equations (5) and (6), and since $T_o = 1/c$, equation (2) gives:

$$A_o = cE^2/2p \quad \ldots\ldots \quad (7)$$

Similarly, equation (3) becomes:

$$A_n = (p/2\pi^2 n^2 c)(1 - \cos 2\pi ncE/p) \quad \ldots\ldots \quad (8)$$

and equation (4) gives:

$$B_n = (p/2\pi^2 n^2 c)\,[2\pi ncE/p - \sin(2\pi ncE/p)] \quad \ldots\ldots \quad (9)$$

Equations (7), (8), and (9) could now be substituted into equation (1) to give the Fourier expansion of $V(t)$ for the unmodulated pulse train with p a constant. Now if this pulse train is modulated by some intelligence signal, the slope p will no longer be a constant but will vary linearly with the intelligence.

The slope may be written

$$p(t) = P_o(1 + m \sin 2\pi at) \quad \ldots\ldots \quad (10)$$

where:

$-P_o$ = slope of pulses with no modulating signal

m = degree of modulation

a = modulating frequency

Even though it can be seen that the slope is a function of

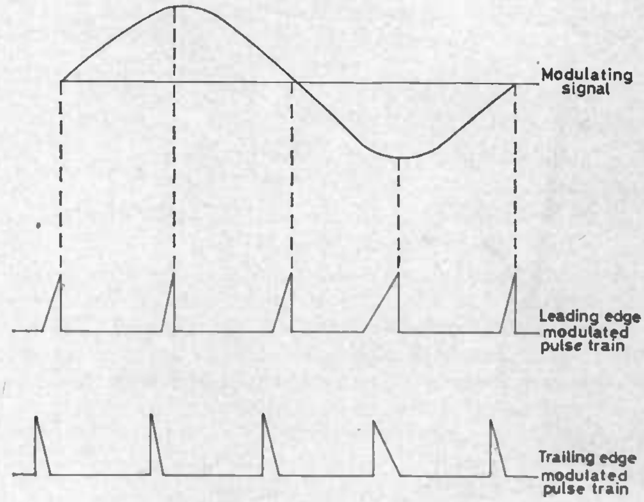

Fig. 1. Variable slope pulse modulation schemes

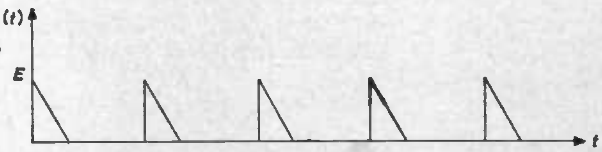

Fig. 2. Unmodulated triangularly shaped trailing edge pulses

E = peak value of the pulses
c = pulse repetition frequency
$-p$ = slope of the trailing side of the pulses
T_o = time between start of successive pulses

time, it will change very little during the duration of a single triangular shaped pulse, provided the pulse duration is much less in time than the period of the modulating signal.

Fig. 3 shows a triangular pulse which has a time duration that does not meet this requirement, and the effects of these extended sampling times on the modulated pulse are illustrated in the figure.

Obviously, a large sampling time introduces a non-linear slope and is therefore undesirable. Keeping the sampling times short in duration, the Fourier coefficients for the modulated case may be obtained by direct substitution of equation (10) into equations (7), (8), and (9).

Substituting equation (10) into equation (7) yields:

$$A_o = cE^2/(2P_o(1 + m \sin 2\pi at)) \quad \ldots\ldots \quad (11)$$

The analysis will be carried out for degrees of modulation,

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m, small enough so that:

$$1/(1 + m \sin 2\pi at) \simeq 1 - m \sin 2\pi at \quad \ldots\ldots \ (12)$$

Equation (11) then becomes:

$$A_o = (cE^2/2P_o)(1 - m \sin 2\pi at) \quad \ldots\ldots\ldots \ (13)$$

Substituting equation (10) into equation (8) and using equation (12) gives:

$$A_n = \frac{P_o(1 + m \sin 2\pi at)}{2\pi^2 n^2 c} \times \{1 - \cos[(2\pi ncE/P_o)(1 - m \sin 2\pi at)]\} \quad \ldots \ (14)$$

For convenience, let:

$$H = 2\pi ncE/P_o \quad \ldots\ldots\ldots\ldots\ldots\ldots \ (15)$$

Then equation (14) may be written as:

$$A_n = (P_o/2\pi^2 n^2 c)\,[1 + m \sin 2\pi at - \cos(H - Hm \sin 2\pi at) - m(\sin 2\pi at)\cos(H - Hm \sin 2\pi at)] \quad \ldots\ldots \ (16)$$

Using the familiar trigonometric expansion for the cosine of the difference of two angles, equation (16) becomes:

$$\begin{aligned}A_n = (P_o/2\pi^2 n^2 c)\,\{&1 + m \sin 2\pi at - \cos H \cos(Hm \sin 2\pi at) \\ &- \sin H \sin(Hm \sin 2\pi at) - m \sin 2\pi at\,[\cos H \\ &\times \cos(Hm \sin 2\pi at) + \sin H \sin(Hm \sin 2\pi at)]\}\end{aligned} \quad \ldots\ldots\ldots\ldots \ (17)$$

It can be shown[6] that:

$$\cos(Hm \sin 2\pi at) = J_0(Hm) + 2J_2(Hm)\cos 4\pi at + 2J_4(Hm)\cos 8at + 2J_6(Hm)\cos 12\pi at + \text{---} \quad \ldots\ldots \ (18)$$

and that:

$$\sin(Hm \sin 2\pi at) = 2J_1(Hm)\sin 2\pi at + 2J_3(Hm)\sin 6\pi at + 2J_5(Hm)\sin 10\pi at + \text{---} \quad \ldots\ldots \ (19)$$

where $J_0(Hm)$, $J_1(Hm)$, $J_2(Hm), \ldots$, are the Bessel functions of the first kind and of orders $0, 1, 2, \ldots$, and of argument Hm. In further notation, the argument, Hm, of the Bessel functions will be omitted, for convenience. Employing the relations expressed in equations (18) and (19), equation (17) may be written as:

$$\begin{aligned}A_n = (P_o/2\pi^2 n^2 c)\,\{&1 + m \sin 2\pi at \\ &- \cos H\,[J_0 + 2J_2 \cos 4\pi at + 2J_4 \cos 8\pi at \text{---}] \\ &- \sin H\,[2J_1 \sin 2\pi at + 2J_3 \sin 6\pi at + \text{---}] \\ &- m \cos H\,[J_0 \sin 2\pi at + 2J_2 \sin 2\pi at \cos 4\pi at \\ &\quad + 2J_4 \sin 2\pi at \cos 8\pi at + \text{---}] \\ &- m \sin H\,[2J_1 \sin 2\pi at \sin 2\pi at + 2J_3 \sin 2\pi at \sin 6\pi at \\ &\quad + 2J_5 \sin 2\pi at \sin 10\pi at + \text{---}]\}\end{aligned} \quad \ldots\ldots \ (20)$$

Using the trigonometric identities:

$$\begin{aligned}\sin x \cos y &= \tfrac{1}{2}[\sin(x + y) + \sin(x - y)] \\ \sin x \sin y &= \tfrac{1}{2}[\cos(x - y) - \cos(x + y)]\end{aligned} \quad \ldots\ldots\ldots\ldots \ (21)$$

Equation (20) may be written as:

$$\begin{aligned}A_n = (P_o/2\pi^2 n^2 c)\,\{&1 + m \sin 2\pi at \\ &- \cos H\,[J_0 + 2J_2 \cos 4\pi at + 2J_4 \cos 8\pi at + \text{---}] \\ &- \sin H\,[2J_1 \sin 2\pi at + 2J_3 \sin 6\pi at + \text{---}] \\ &- m \cos H\,[J_0 \sin 2\pi at + J_2 \sin 6\pi at - J_2 \sin 2\pi at \\ &\quad + J_4 \sin 10\pi at - J_4 \sin 6\pi at + \text{---}] \\ &- m \sin H\,[J_1 - J_1 \cos 4\pi at + J_3 \cos 4\pi at - J_3 \cos 8\pi at \\ &\quad + J_5 \cos 8\pi at - J_5 \cos 12\pi at + \text{---}]\}\end{aligned} \quad \ldots\ldots\ldots\ldots \ (22)$$

Equation (22) gives the value of A_n for a triangular shaped pulse train modulated a small amount by a sinusoidally varying signal. It is now desirable to obtain $A_n \cos 2\pi nct$.

$$\begin{aligned}A_n \cos 2\pi nct = (P_o/2\pi^2 n^2 c)\,\{&\cos 2\pi nct + m \sin 2\pi at \cos 2\pi nct \\ &- \cos H\,[J_0 \cos 2\pi nct + 2J_2 \cos 2\pi nct \cos 4\pi at \\ &+ 2J_4 \cos 2\pi nct \cos 8\pi at + 2J_6 \cos 2\pi nct \cos 12\pi at + \text{---}] \\ &- \sin H\,[2J_1 \sin 2\pi at \cos 2\pi nct + 2J_3 \sin 6\pi at \cos 2\pi nct \\ &\quad + 2J_5 \sin 10\pi at \cos 2\pi nct + \text{---}] \\ &- m \cos H\,[J_0 \sin 2\pi at \cos 2\pi nct + J_2 \sin 6\pi at \cos 2\pi nct \\ &\quad - J_2 \sin 2\pi at \cos 2\pi nct + J_4 \sin 10\pi at \cos 2\pi nct \\ &\quad - J_4 \sin 6\pi at \cos 2\pi nct + \text{---}] \\ &- m \sin H\,[J_1 \cos 2\pi nct - J_1 \cos 4\pi at \cos 2\pi nct \\ &+ J_3 \cos 4\pi at \cos 2\pi nct - J_3 \cos 8\pi at \cos 2\pi nct \\ &+ J_5 \cos 8\pi at \cos 2\pi nct - J_5 \cos 12\pi at \cos 2\pi nct + \text{---}]\}\end{aligned} \quad \ldots\ldots\ldots\ldots \ (23)$$

Using the following trigonometric identity:

$$\cos x \cos y = \tfrac{1}{2}[\cos(x + y) + \cos(x - y)]$$

and equation (21), equation (23) becomes:

$$\begin{aligned}A_n \cos 2\pi nct = (P_o/2\pi^2 n^2 c)\,\{&\cos 2\pi nct + \tfrac{1}{2}m \sin 2\pi(nc + a)t \\ &- \tfrac{1}{2}m \sin 2\pi(nc - a)t \\ &- \cos H[J_0 \cos 2\pi nct + J_2 \cos 2\pi(nc + 2a)t + J_2 \cos 2\pi(nc - 2a)t \\ &+ J_4 \cos 2\pi(nc + 4a)t + J_4 \cos 2\pi(nc - 4a)t + \text{---}] \\ &- \sin H\,[J_1 \sin 2\pi(nc + a)t - J_1 \sin 2\pi(nc - a)t \\ &+ J_3 \sin 2\pi(nc + 3a)t - J_3 \sin 2\pi(nc - 3a)t + \text{---}] \\ &- \tfrac{1}{2}m \cos H\,[J_0 \sin 2\pi(nc + a)t - J_0 \sin 2\pi(nc - a)t \\ &+ J_2 \sin 2\pi(nc + 3a)t - J_2 \sin 2\pi(nc - 3a)t - J_2 \sin 2\pi(nc + a)t \\ &+ J_2 \sin 2\pi(nc - a)t + J_4 \sin 2\pi(nc + 5a)t - J_4 \sin 2\pi(nc - 5a)t \\ &- J_4 \sin 2\pi(nc + 3a)t + J_4 \sin 2\pi(nc - 3a)t + \text{---}] \\ &- \tfrac{1}{2}m \sin H\,[2J_1 \cos 2\pi nct - J_1 \cos 2\pi(nc + 2a)t - J_1 \cos 2\pi(nc - 2a)t \\ &+ J_3 \cos 2\pi(nc + 2a)t + J_3 \cos 2\pi(nc - 2a)t - J_3 \cos 2\pi(nc + 4a)t \\ &- J_3 \cos 2\pi(nc - 4a)t + J_5 \cos 2\pi(nc + 4a)t + J_5 \cos 2\pi(nc - 4a)t + \text{---}]\}\end{aligned} \quad \ldots\ldots\ldots\ldots \ (24)$$

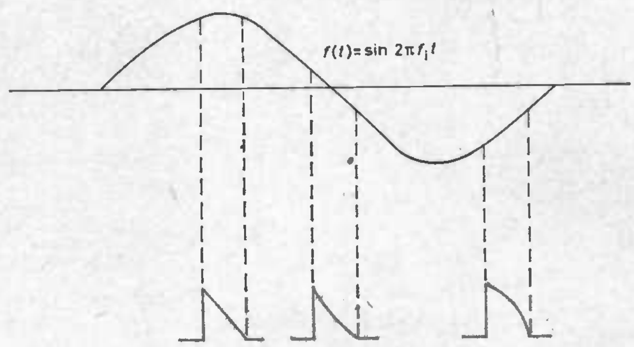


Fig. 3. Illustrating effects of large sampling times

Equation (24) can be put in a more convenient form by collecting and rearranging terms, showing very clearly the frequency components present in $A_n \cos 2\pi nct$.

$$\begin{aligned}A_n \cos 2\pi nct = (P_o/2\pi^2 n^2 c)\,\{&[1 - J_0 \cos H - J_1 m \sin H] \cos 2\pi nct \\ &+ [+\tfrac{1}{2}m - J_1 \sin H + (J_2 - J_0)\tfrac{1}{2}m \cos H] \sin 2\pi(nc + a)t \\ &+ [-\tfrac{1}{2}m + J_1 \sin H - (J_2 - J_0)\tfrac{1}{2}m \cos H] \sin 2\pi(nc - a)t \\ &+ [-J_2 \cos H - (J_3 - J_1)\tfrac{1}{2}m \sin H] \cos 2\pi(nc + 2a)t \\ &+ [-J_2 \cos H - (J_3 - J_1)\tfrac{1}{2}m \sin H] \cos 2\pi(nc - 2a)t \\ &+ [-J_3 \sin H + (J_4 - J_2)\tfrac{1}{2}m \cos H] \sin 2\pi(nc + 3a)t \\ &+ [+J_3 \sin H - (J_4 - J_2)\tfrac{1}{2}m \cos H] \sin 2\pi(nc - 3a)t \\ &+ [-J_4 \cos H - (J_5 - J_3)\tfrac{1}{2}m \sin H] \cos 2\pi(nc + 4a)t \\ &+ [-J_4 \cos H - (J_5 - J_3)\tfrac{1}{2}m \sin H] \cos 2\pi(nc - 4a)t + \text{---}\}\end{aligned} \quad \ldots\ldots\ldots\ldots \ (25)$$

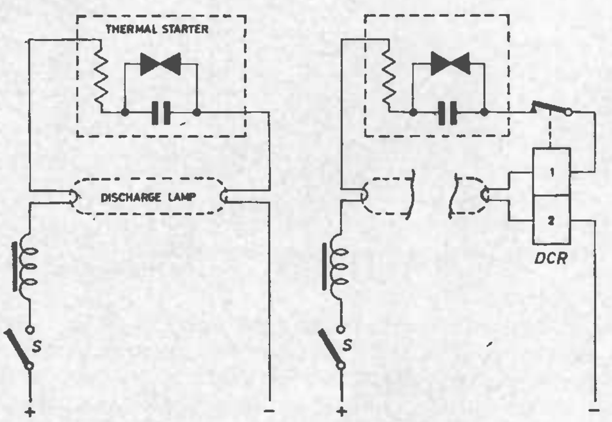
Substituting equation (10) into equation (9) and using equation (12) yields:

$$\begin{aligned}B_n &= \frac{P_o(1 + m \sin 2\pi at)}{2\pi^2 n^2 c}\left\{\frac{2\pi ncE}{P_o(1 + m \sin 2\pi at)} - \sin[(2\pi ncE/P_o)(1 - m \sin 2\pi at)]\right\} \\ &= \frac{P_o}{2\pi^2 n^2 c}\,\{2(\pi ncE/P_o) - \sin[(2\pi ncE/P_o) - (2\pi nce/P_o)m \sin 2\pi at] \\ &\quad - m \sin 2\pi at \sin[(2\pi ncE/P_o) - (2\pi ncE/P_o)m \sin 2\pi at]\}\end{aligned} \quad (26)$$

Using equation (15), this may be written as:

cal lamp circuit is shown at Fig. 3(a). On closing the mains switch (*S*), the current drawn by the emitters at each end of the tubular lamp also energizes the heating element of the thermal starter. When the emitters reach their working temperature, the starter contacts open, inducing a sharp rise in voltage across the inductor which establishes the discharge path through the gas and the lamp displays a characteristic glow. As the voltage required to maintain the discharge is very much lower than the supply voltage, there is sometimes a tendency, due to poor regulation, for the starter contacts to reclose and extinguish the arc as the temperature of the heating element falls. This is overcome, in the circuit of Fig. 3(b), by the introduction of a d.c.r. (with coils in opposition) having a normally closed contact which isolates the thermal starter when the lamp is alight.

On closing *S*, since both coils are energized, the d.c.r. will not operate and its contact remains closed so as to energize the heating element of the thermal starter. When



(a) (b)

Fig. 3. (a) Discharge lamp circuit (b) Discharge lamp circuit with d.c.r.

the latter breaks the circuit, current ceases to flow in coil 1, but the lamp strikes and draws current via coil 2; this unbalances the d.c.r. which opens and disconnects the starter.

Simple Photocell Circuits

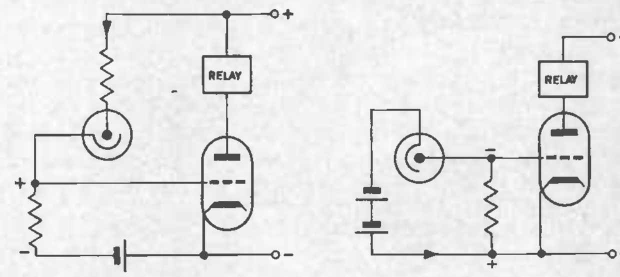
A photoelectric cell can be used to control a relay through the agency of a thermionic valve, the relay being arranged to switch on an alarm circuit, operate a counter or start and stop an electric motor. The basic circuits are shown in Figs. 4(a) and 4(b).

Fig. 4(a) shows a circuit for closing the relay when light falls on the photocell. In the absence of light the grid of the triode is sufficiently negative to inhibit the flow of anode current and the relay cannot operate. Illumination renders the photocell conductive and results in a change of grid potential which allows the triode to conduct and operate the relay.

Fig. 4(b) shows a circuit in which illumination of the photocell causes the relay to fall out, which is the reverse of what happens in Fig. 4(a). When the photocell is inactive (in the absence of light) the valve grid is at cathode potential and the anode current operates the relay. When light is transmitted to the photocell, the mesh consisting of the battery, grid leak *R* and the photocell itself conducts in an anti-clockwise direction, making the control grid negative to the cathode and so inhibiting the flow of anode current. As long as the photocell is illuminated the relay remains inoperative but it closes as soon as the light is extinguished.

A need may arise in industry for the automatic detection of any change in the transparency of a liquid as, for instance, in turbidity control. The circuit used is often a combination of Figs. 4(a) and 4(b) in that one photocell is connected between grid and anode and the other between grid and cathode of a single triode, the cells forming two elements of a Wheatstone bridge. By dividing the light into two beams, one illuminating the first photocell and the other passing through a test cell, containing a sample of the liquid, on its way to the second photocell, a d.c.r. connected as in Fig. 5 will detect any change in light intensity at the second photocell in relation to that at the first.

With both cells P_1 and P_2 illuminated the steady current in coil 1 and the current in coil 2 are balanced by means of the resistor *R*. When the light distribution changes and equilibrium is disturbed the current in coil 1 will increase



(a) (b)

Fig. 4 (above). (a) Relay to operate on incident light (b) Relay to operate on extinction of light

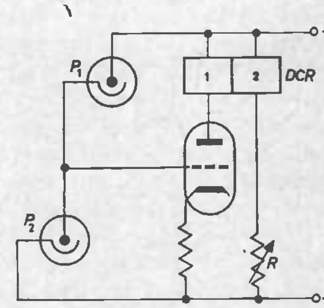


Fig. 5 (right). Double coil relay for comparison of two illuminations

or decrease to upset the balance, causing the relay to operate and make or break an auxiliary circuit. The diagram shows that the circuit is to some extent self-stabilizing, since fluctuations in the supply voltage will be reflected in both coil currents without disturbing the balance.

Delayed Switching Using A Thermionic Valve

The circuits operate so that either (a) the grid charges negatively and the charge is allowed to leak away through a high resistance until the change of grid potential is sufficient for the rising anode current to operate the d.c.r., or (b) the process is reversed, the grid assuming a more negative potential on the charging cycle which reduces the anode current in one coil of the d.c.r. The instant at which the relay opens or closes depends upon the voltage across the grid circuit capacitor, which charges or discharges through the high resistance grid leak, thereby introducing the time factor.

Fig. 6(a) shows a circuit which generates an impulse of fixed duration regardless of the length of the starting impulse. Manual operation of the push-button *PB* energizes the valve filament and coil 2 of the d.c.r. which thereupon closes (note that the relay is self-holding via its own changeover contact). Now, since a voltage is developed across *R* with the polarity shown, the capacitor

C commences to charge. When the grid is sufficiently positive for the anode current in coil 1 to balance that in coil 2, the d.c.r. opens. If *PB* is still closed nothing will happen until it is released, when *C* discharges via the back contact of *PB* and that of the d.c.r., thus restoring the circuit for the next operation. Another valve timing circuit with improved repetition accuracy, which can accommodate supply voltage variations of ± 15 per cent, is illustrated in Fig. 6(b). It is sometimes preferable to use a tetrode rather than a triode since the screen voltage may be adjusted for fine control of the time delay.

Simple Thyratron Circuit

Fig. 7 shows an arrangement in which a gas-filled triode maintains a subsidiary circuit closed for a fixed interval after the cessation of the starting impulse. Closing *PB* energizes the valve filament in series with coil 2 of the d.c.r. which closes and locks in via its contact DCR_2. The second contact of *PB* applies a negative voltage to the valve grid. When *PB* is released (and not before) the capacitor starts to discharge through the grid leak. At the end of the discharge period the 'valve 'fires' and the anode current in coil 1 sets up a magnetic field in the core which neutralizes the field due to coil 2 and releases the d.c.r. Contact DCR_1 then breaks the anode circuit of the thyratron and the circuit is ready for the next impulse.

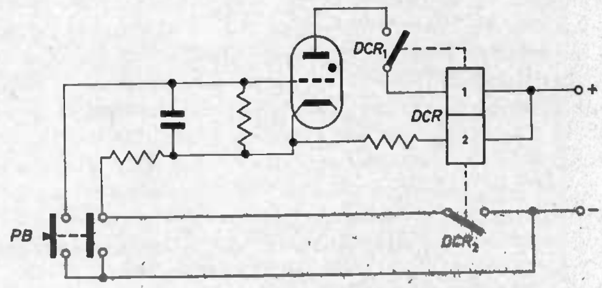


Fig. 7. Delayed switching with a thyratron

The Use of Thyratrons for Synchronizing A.C. Supplies

Fig. 8 is a simplified diagram of a circuit for switching an emergency three-phase alternator in parallel with the normal three-phase power supply prior to taking the load off the latter. Before the interconnector can be closed, both sources of voltages must correspond in magnitude, frequency and phase.

A pair of gas-filled triodes derive their anode voltages from one phase of the supply mains. The anode conducts when it is positive to the cathode, provided the grid is not negative, and the discharge is extinguished at the end of each positive half-cycle. The grid circuits are not shown but the exciting voltages are derived from a simple bridge circuit, supplied from two transformers, the secondary windings of which are connected in series opposition. The primary windings are connected to the respective three-phase systems marked *A* and *B* on the diagram. The output of the bridge feeds the thyratron grids in such a way that the voltage of system *A* tends to trigger valve V_1 and that of system *B* tends to 'fire' valve V_2. When the phase difference is in one direction V_2 will 'fire' first causing relay *E* to open and isolate the circuit of V_1. When the phase difference is in the other direction so that V_1 'fires' first, relay *C* operates and completes the anode circuit of V_2 via relay *D* which is the interconnector

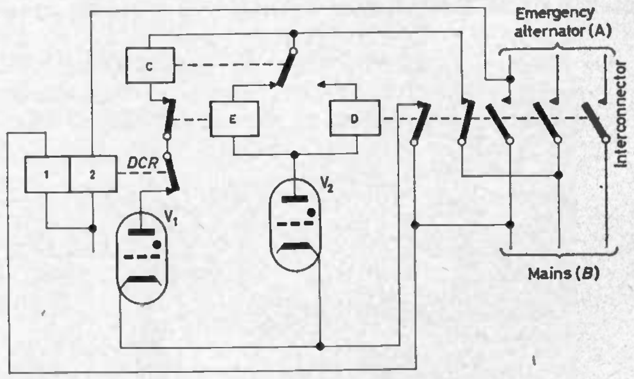
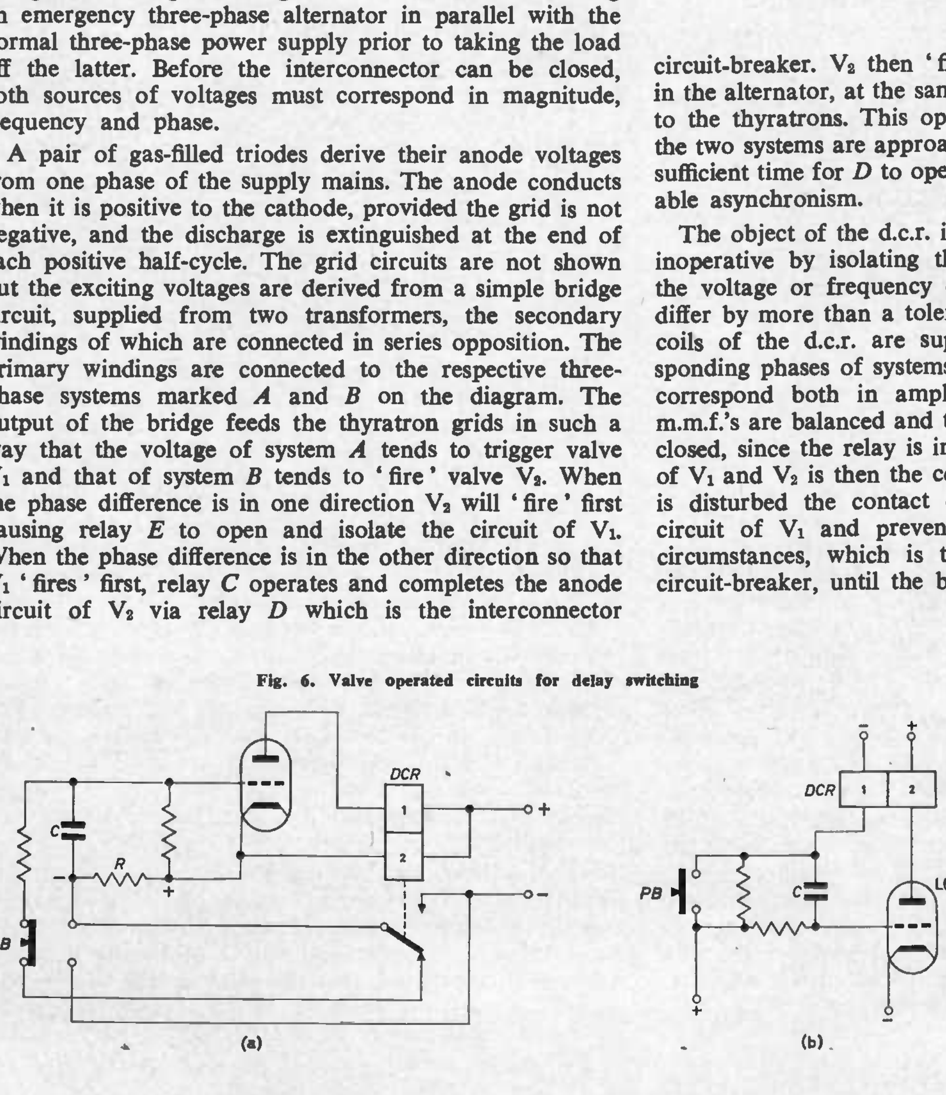


Fig. 8. Synchronizing a.c. supplies

circuit-breaker. V_2 then 'fires' and operates *D* to switch in the alternator, at the same time interrupting the supplies to the thyratrons. This operation takes places only when the two systems are approaching phase equality and allows sufficient time for *D* to operate before there is any appreciable asynchronism.

The object of the d.c.r. is to render the thyratron circuit inoperative by isolating the anode circuit of V_1 should the voltage or frequency of the two three-phase systems differ by more than a tolerable margin. The two identical coils of the d.c.r. are supplied respectively from corresponding phases of systems *A* and *B*. When their voltages correspond both in amplitude and frequency, the coil m.m.f.'s are balanced and the contact on the d.c.r. remains closed, since the relay is inoperative; the order of 'firing' of V_1 and V_2 is then the controlling factor. If this balance is disturbed the contact on the d.c.r. opens the anode circuit of V_1 and prevents it from 'firing' under any circumstances, which is tantamount to locking out the circuit-breaker, until the balance is restored.

Fig. 6. Valve operated circuits for delay switching



(a) (b)

Coils in Unison

When the fluxes are additive and the coils assist each other, there are two distinct modes of operation:

(a) Energizing the low resistance coil will close the relay, which is retained by its high resistance coil, the latter being excited via a contact on the relay itself. This contact must close just

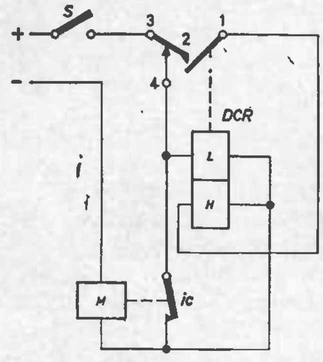


Fig. 9 (above). Basic circuit of d.c.r. with 'line' and 'holding' coils

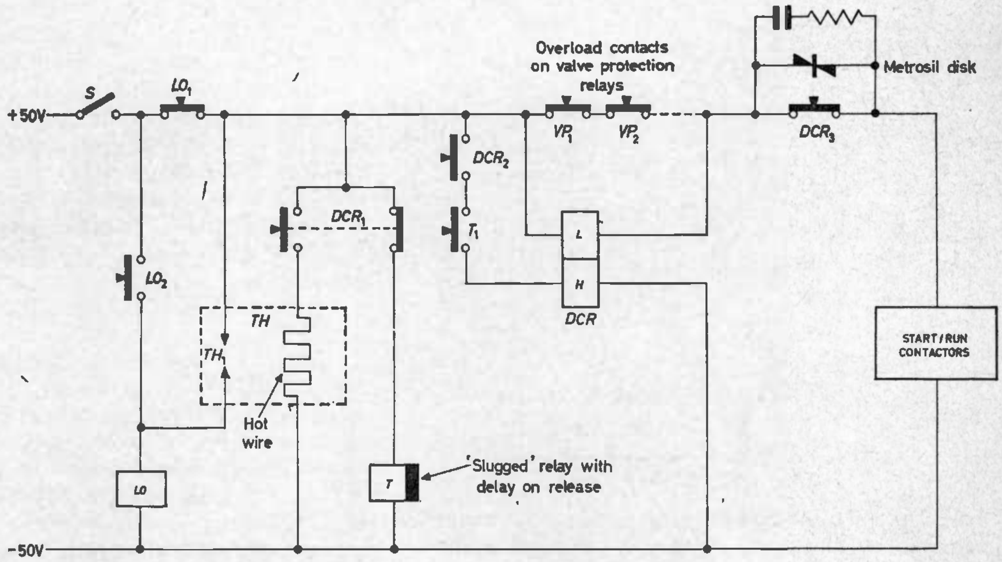


Fig. 10 (right). Automatic reclosing device for overload protection

before the armature reaches the end of its travel (early make) or the relay will fall out.

(b) The relay will not close unless both coils are energized simultaneously. The applications of such a relay are fairly well known, whereas (a) is the principal ingredient of a novel circuit for automatic overload protection, which has been operating for many years and merits further description.

Automatic Overload Protection

The d.c.r. is designed to close following momentary excitation of the series or 'line' coil; as the relay closes, a contact in series with the shunt or 'holding' coil causes the relay to lock 'on', and it is then said to be 'self-holding'. Evidently each coil should develop the same number of ampere-turns (m.m.f.); the series coil may have only a few hundred turns, carrying several amperes, whereas the shunt coil will have several thousand turns, carrying a few hundred milliamperes, a small fraction of the current in the series coil.

Fig. 9 shows a simple circuit for this type of operation. *M* is a master relay, having one closed interrupter contact (*ic*): the d.c.r. has a line coil (*L*) and a holding coil (*H*) and carries a make (12) before break (34) contact connected as shown. Its function is to ensure that the master relay is energized just long enough (and no longer) to initiate an operation of some kind in a subsidiary circuit.

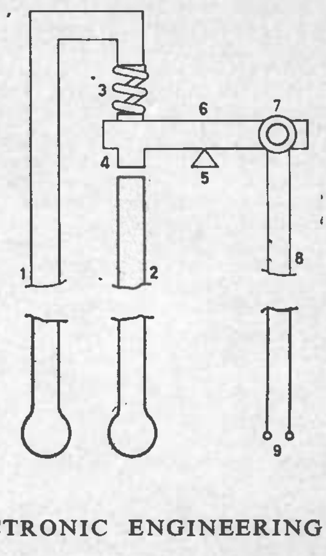
Closing switch *S* energizes the master *M* via its contact *ic* which initially short-circuits coil *L* of the d.c.r. *M* then operates, opens *ic* and diverts the current through *L*, which pulls in the d.c.r. This closes contact 12 before opening contact 34 by mechanical action as the diagram indicates. The d.c.r. is thus energized via coil 8 and the opening of contact 34 isolates the master relay which cannot be reclosed until *S* is opened to release the d.c.r. The latter is then said to be locked out.

Fig. 10 shows a modification of this circuit for an overload device which recloses automatically three times in quick succession before locking out. The circuit may be used to control the operation of a modern high-tension rectifier supplying the anodes of several high-power transmitting valves. Each valve is protected by an overload relay having one closed contact and these contacts VP_1, VP_2, etc. (i.e. valve protection) are all connected in series. The d.c.r. has a changeover contact (DCR_1) an open contact (DCR_2) and two closed master contacts in series (DCR_3) which are protected by a Metrosil suppressor to inhibit sparking.

TH is a hot-wire vacuum switch which serves as a thermal delay relay, having one open contact TH_1 and a heating cycle of 4·5sec duration. The action of the switch is shown diagrammatically in Fig. 11. Its operation relies upon the absence of an arc when an electric current is interrupted by the separation of two surfaces in a vacuum. The gap required is only about one-thousandth of an inch so that the contacts are very light and the movement itself minute. Such a small movement is readily provided by the thermal expansion of a wire through which the control current passes. 1 and 2 are fixed conductors the lower extremities of which are the terminals of the switch contacts (4). The hot wire 8 consists of a number of strands of special steel wire wound between two Steatite insulating bobbins (7). A spring 3 attached to a lever 6 resting on the fulcrum 5 is held in compression by the tension (adjustable) of the resistance wire looped between the insulating bobbins and rigidly fixed to the terminals 9. By tensioning this wire the spring is compressed to open the tungsten contacts (open or 'off' position of switch). When current is passed through the steel wire, it is heated and expands, so releasing the spring and forcing the lever to close the tungsten contacts (closed or 'on' position of switch).

The rate of expansion of the steel wire and hence the time required to close the contacts depends upon the total time during which current is flowing in the wire. In this particular application of the hot-wire switch the current is intermittent in three short spells of 1·5sec each which, in the aggregate, amounts to the 4·5sec of heating time necessary for the wire to expand and close the contacts.

Fig. 11. Hotwire vacuum switch



T is a slugged relay having one open contact T_1 which releases in

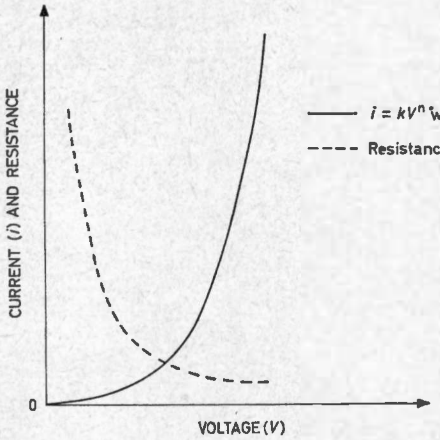


Fig. 12. Typical 'Metrosil' characteristics

1·5sec after it is de-energized. The delay on release is achieved by means of a copper ring called a 'heel slug', which surrounds the core at the end remote from the armature. When the coil circuit is broken, current is induced in the copper ring which retards the decay of the flux in the core.

T is normally energized via the changeover contact of the d.c.r. Should an overload occur at any one valve anode, its particular *VP* contact opens and the current in the coil of the 'run' contactor, in the starting circuit, is diverted to the series coil *L* of the d.c.r., which operates almost immediately*. The d.c.r. is retained by its 'hold-on' coil *H*, the circuit of which is completed by its own 'early make' contact DCR_2 via contact T_1 of the 'slugged' relay. Contacts DCR_3 interrupt the supply to the 'run' contactor and remove the high tension from the valve anodes before any damage can occur. In the absence of anode current the *VP* contact recloses and short-circuits coil *L* of the d.c.r. but this is of no consequence since the latter cannot release so long as coil *H* is energized.

Contacts DCR_1 serve to disconnect the coil of the 'slugged' relay *T* and energize the thermal relay *TH*. After 1·5sec relay *T* drops out and contact T_1 opens the circuit of coil *H*, causing the d.c.r. itself to release. This reconnects the starting circuit to the 50V supply and initiates the normal starting sequence. Contacts DCR_1 re-energize the 'slugged' relay *T* and reset the hot-wire circuit. Should the overload persist, the whole cycle of operations is repeated. On the third successive cycle the thermal relay opens contact TH_1, having been energized for three periods of 1·5sec each, in quick succession, equivalent to 4·5sec altogether. Contact TH_1 then closes the lock-out relay *LO* which holds itself on via LO_2 while LO_1 isolates the starting circuit from the 50V supply. The circuit cannot now be re-energized until the operator manually opens switch *S* to release the lock-out relay and reclose contact LO_1. Before remaking the switch the operator must of course open up the equipment in order to ascertain the nature of the fault and proceed to clear it. The advantage of a reclosing device of this description is that when the fault is intermittent, the equipment is protected and normal operation restored without the intervention of an operator.

The object of the Metrosil suppressor is to minimize 'sparking' at the double-break contact DCR_3. 'Metrosil' is the registered trade mark of a resistive material containing silicon carbide which has a non-linear voltage-current characteristic of the type illustrated in Fig. 12. The shape of this curve is so far removed from the linearity of Ohm's Law that, by doubling the voltage across the disk, the current it passes can increase twenty-fold or by trebling it, a hundredfold. This means that the resistance of a 'Metrosil' disk falls sharply as the current rises so that it restricts the voltage surge at the opening contacts to about one tenth of its normal value.

Under running conditions the current from the 50V source is only about 3A but the starting circuit is so inductive that the momentary rise in voltage across the opening contacts would result in a rapid deterioration of their surfaces without some form of 'spark' suppression. The inclusion of a series capacitor-resistor combination in parallel with the 'Metrosil' disk reduces the duration of the 'spark' to such an extent that the contact surfaces are not impaired and should last for years with ordinary routine maintenance.

* *It would require a finite time initially for the operative current to rise from zero in an inductive circuit.*

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Microminiaturized S.S.R. Equipment

Cossor Electronics Ltd are now producing what they describe as third generation secondary surveillance radar equipments. These consist of the microminiaturized airborne transponder type 2100 which operates in conjunction with the type CR 1600 microminiaturized interrogator/responsor.

In the new transponder, conventional delay-line encoding and decoding circuits are replaced by digital shift registers which are housed in the SSR 2101 control unit. This configuration simplifies aircraft installation wiring since the transponder/control unit interconnexions are largely reduced to detected video and modulation signals. For an installation requiring 30ft of inter-unit wiring, the cable weight is 1 lb 8oz compared with 5 lb for a conventional system. The fewer pin connexions necessary also improve reliability.

Approximately fifty flat-pack silicon integrated circuits are located on four printed circuit boards plus a number of thin-film and T0-5 packaged devices. Some ten transistors are incorporated in the power supply and modulator sections. An m.t.b.f. of at least 10 000 hours is predicted due in part to electroweld assembly techniques in preference to soldered joints. In the event of failure of an integrated circuit, total breakdown of the system is avoided by compensating redundancy circuits.

The SSR 2100 is readily adaptable to meet M.I.L. or I.C.A.O. requirements or any other combination. The basic unit provides 4096 codes on any two modes with three-pulse s.l.s. on all modes and these operational features are achieved at a substantially lower cost compared with other conventional transponders. A self-test monitor circuit is incorporated and visually indicates satisfactory performance of receiver sensitivity. and centre frequency. transmitter power, decoding functions and reply code fidelity. The versatility of the SSR 2100 makes it suitable for application in light military aircraft as well as general aviation.

For interrogating the SSR 2100 there is the new CRI 600 micro-miniaturized i.f.f. Mk. 10 ground system. This equipment, which can fit into a small suitcase, weighs only 12½ lb for the transmitter receiver and 6 lb for the decoder. Solid state and integrated circuit techniques are used throughout ensuring high reliability with low volume packaging.

The design allows easy integration into a large variety of radar environments. The transmitter receiver can rotate with the primary aerial using slip rings or be remoted to work through an i.f.f. channel in the rotating joint and the decoder may be readily accommodated close to display equipment or in the radar office.

NEW BOOKS

Semiconductors

By F. J. Hyde. 324 pp. Demy 8vo. Macdonald & Co. 1965. Price 65s.

IN the preface Dr. Hyde, who is very well known for his many contributions on the various aspects of semiconductor device physics, states that it is hoped that this book will bring the engineer and the physicist closer together. In this objective the author has certainly succeeded.

The coverage is much wider than the title of the book may suggest. About half the book is devoted to a thorough study of the electrical properties of semiconductors as influenced by such external factors as heat, electric and magnetic fields, and radiation. The remaining half is devoted to a study of numerous semiconductor devices. A thorough discussion of the junction diode and the transistor is given; the treatment going as far as the development of equivalent circuit representations. In this way a bridge is established between the physical electronic aspects of semiconductor devices and their circuit theory as expounded in other textbooks.

A very notable feature of the book is that particular attention is given to the practical behaviour of devices, and the departure from idealized behaviour is carefully explained. The technology and limitations of the devices are stressed.

The only criticism of the book is that a number of mathematical relations are stated without proof, and it would have been helpful to the motivated reader if appropriate references were given.

The book is strongly recommended for use in final year electronic engineering and applied physics courses. It should also prove useful to the practising electronic engineer wishing to acquaint himself with semiconductor device physics.

S. S. HAKIM

Feedback Circuit Analysis

By S. S. Hakim. 392 pp. Demy 8vo. Iliffe Books Ltd. 1966. Price 95s.

THE subject of feedback has been covered in a comprehensive way at the level of the final year undergraduate or postgraduate student. The author has taken some trouble to show the identity of the subject with the theory of control systems by adopting terminology common to both, but the treatment is biased towards amplifier design.

A short introduction is followed by a survey of linear circuit analysis based on the differential form of the circuit equations, active devices like valves and transistors being included by the addition of current and voltage controlled sources. The Laplace transform is introduced at a very early stage as a method of manipulating the circuit differential equations, and is applied to the evaluation of the steady state and transient response of systems. The 'two port black box' together with the standard tools for manipulation, and matrix theory follow. Feedback is dealt with in the classical case of separate forward and feedback paths, and then in the more complex general system which is represented by flow graphs. The stability of systems is considered in three ways which illuminate the problem from different directions. Firstly the transfer function in the *p*-plane, then the steady state response to a sinusoidal signal and finally in terms of the input and output impedances. A further chapter deals with the design of compensating networks. The book is completed by three chapters on the design of wideband amplifiers, operational amplifiers, and oscillators respectively.

The presentation of each topic is in the form of a survey of the mathematical theory and techniques with derivations in many cases. These sections are of necessity concise, but are not always easy to follow for a reader unfamiliar with the subject, and it was noticed that on page 15, oversimplification had led to the fallacious proof of the derivation of the Laplace transform of the Unit Impulse. Each topic is followed by a very comprehensive and detailed application of the methods to commonly used circuits, and it is these examples which are the main strength of the book. Each chapter has a short list of references and problems with answers.

While the book will be of interest to students, the strength of the presentation in the sections where the ideas are applied to actual circuits may make it of greater value to the practising engineer.

R. V. LEEDHAM

'Modern Radar'—Analysis, Evaluation, and System Design

By Raymond S. Berkowitz. 660 pp. Med. 8vo. John Wiley. 1965. Price 147s.

THIS multi-author book will undoubtedly become a reference of international standing.

'Modern Radar' is a massive work comprising some 26 chapters by 14 authors. A great deal of material has been compressed into this compass to achieve the "intensive exposition" of the preface. Much of this material is highly mathematical, and it is felt that the greatest appeal of the book will be to the mathematician rather than to the technologist or even to the mathematical physicist who has become involved in this wide-ranging form of engineering.

This is largely because the authors, presumably restricted by space limitations, rarely link their conclusions directly with relevant engineering application. The result is a perspective which could be misleading, particularly in the systems design context where it is vital that practical boundary conditions should be taken into account.

Nevertheless this criticism must not be allowed to diminish the credit that must be given to the contributors and their Editor for bringing together—almost certainly for the first time—so much of the theory of complex signals and the associated statistical and specific probability concepts. In this connexion, full commendation must be given to the inclusion of a number of tables in the text, such as the tabular list of some 'Useful Fourier Integral "Mates" or "Pairs",' which hitherto have not been collated in this way.

Tribute must also be paid to the uniformly high standard of writing and presentation generally which is evident throughout the book. This is always difficult to accomplish in such circumstances; but in this case, with each author writing as a specialist on a highly complex subject, such uniformity would have appeared almost impossible to attain. At the same time, however, a system of chapter division has been adopted which it is not uncharitable to describe as confusing. There are two groups of three parts each, the first group being almost entirely theoretical in nature, the second somewhat more concerned with technique; and each part takes its own chapter numbering, beginning with 1. Consequently there are, for instance, six Chapter 2's, a situation which, despite careful associated part numbering, creates a sense of irritation which cannot be understood until it has been experienced.

There are also certain other minor points of workmanship which invite comment (the final 'e' in 'envelope' sometimes appears and sometimes is omitted), and which tend to introduce unnecessary complexity. One of the most outstanding of these is the printing of a series of binary '1's' and '0's', 63 in all, across a page which is removed from the autocorrelation diagram to which they refer. It should be added, however, that a first reading of the book failed to reveal any obvious technical misprints, thus implying a feat which would be remarkable for one of quarter the size.

One major criticism which must be put forward is with regard to the references, of which the average is slightly under 16 per chapter, ranging from a maximum of 115 for a technique paper on 'Modern Low-Noise Devices' (L. S. Nergaard) to (understandably) zero for certain other chapters. The point at issue here is the low number of references other than U.S. in origin, which appear in the total of just over 400. According to a fairly close check, there are nine from British sources, two Canadian, and one each from Switzerland, France, and India.

It must be stressed that there is full understanding in the United Kingdom of the enormous effort put into radar by the U.S. Nevertheless, the British contribution of the cavity magnetron, and the U.K. development of the operational use of radiolocation are surely such as to merit some reference in a work of this nature. From the point of view of historical perspective alone, their inclusion would appear to be justified. Also in relation to more recent work, British papers on atmospheric attenuation and other propagation effects, notably in the ionosphere, have been of sufficient calibre to be included.

R. E. YOUNG

Electrical Circuits

By L. A. Manning. 567 pp. Med. 8vo. McGraw-Hill. 1966. Price 96s.

THE author of this book, who is Professor of Electrical Engineering at Stanford University, has sought those topics from modern network analysis which will most quickly bring students into contact with contemporary thinking. Starting at the very beginning of the subject new ideas are presented individually and with prolific use of examples (totalling nearly 200) so that it becomes surprisingly easy to approach concepts which until recently were in postgraduate courses. The book is concise and lucid and text explanations are supplemented with summaries of the main results at the end of each chapter. Over 300 carefully devised problems will help students gain confidence and understanding of the material presented.

The chapters are grouped into three relatively discrete parts. The first ten constitute a self-contained introduction to electrical networks and are suitable for students of all engineering branches. Units, definitions and circuit laws are first given; careful organization then leads quickly to an understanding of impedance, complex numbers, phasors, resonance, power, pole-zero diagrams, coupled circuits, mesh and nodal analysis, transients and natural response. One chapter treats analogies of electrical with mechanical, acoustical and other systems. In the second part (seven chapters) step and impulse notation is introduced and the transfer function is discussed. By using a time-domain approach, tables of impulse responses are derived and their use in transients problems explained. The serial-product method is developed for use with numerical, graphical and difficult analytic inputs; through a limiting process it leads to the superposition integral. The Laplace transformation integral is derived from the superposition integral and is shown to relate a transfer function to its unit impulse response.

The last five chapters constitute the third part. Properties of the Laplace transformation are developed and used. The Fourier series and integral are employed to introduce frequency-domain concepts and to derive the direct and inverse Laplace integrals. An introduction is given to sampling, correlation and interpolation.

This clearly written book, with over 600 figures, can be highly recommended for University undergraduates taking courses in electrical and electronic engineering.

F. A. BENSON

Semiconductors and their Circuits Vol. 1. Selected Semiconductor Theory

By N. F. Moody. 345 pp. Demy 8vo. Electrical Engineering Series, English Universities Press. 1966. Price 42s.

THIS book is the first of a two-volume series on semiconductor devices and circuits.

The headings of the 11 chapters follow the conventional pattern: crystal structure, electronic conduction, pn junctions, diodes, transistors, etc. but the contents unfortunately do not. The volume starts with a short chapter "Historical background to semiconductor devices" which turns out to be a rather poor outline of transistor action. The next chapter on "Crystalline structure and conduction in crystals" is written on a very primitive level. The treatment throughout the book is sketchy and contains numerous inaccuracies, loose statements and misleading illustrations. For instance, on page 24: "From a mathematical viewpoint it is convenient to ascribe a different mass to carriers in different regions of their bands", no hint being given of the true concept of an effective mass. Or in Table 53 headed "Some properties of degenerate and non-degenerate semiconductors at 25°C", one value each is given for electron concentration and resistivity of degenerate and non-degenerate germanium and silicon, as if a unique value could be assigned to these parameters. This is followed abruptly by a statement that "the Einstein relationship ($D\alpha\mu$) . . . . is invalid for degenerate materials", without further discussion. In describing Zener tunnelling (page 118), "the electrical fields must be sufficiently great to cause the bands to cross, for only then is it geometrically possible for an electron to travel horizontally (*sic*) from one band to the other". This list could be continued *ad lib.* Such statements must be utterly confusing to the unwary reader, and even the expert will hardly recognize from such scientific slang what the author had in mind. The general impression is that the book has been written in a hurry without much consideration for the student or the practising engineer. Thus an entire chapter, No. 8, "Mathematical analysis of intrinsic diffusion- and drift-transistors under incremental conditions" according to the author may be omitted if the reader is not interested in formal analysis; so why include it at all? Even for a student new to the electronic theory of solids and semiconductors the treatment is far too sketchy to be of any value. Merely taking a few standard equations, graphs and illustrations from recognized texts without proper explanations will hardly help. The general editor's claim that "the text would be of special interest to research workers, postgraduates and students, etc." is very much doubted by this reviewer.

E. BILLIG

Theorie et Calcul des Reseaux de Transport D'Energie Electrique

By H. Edelmano. 309 pp. Med. 8vo. Dunod 1966

The original version of this book was written in German by the author in 1963 during his professorship at the Technical High School at Darmstradt and published under the title "Berechnnung Elektrischer Verbandnetze".

It has now been translated into French by Professors Boland, Borte and Gregoire of the Polytechnique at Mons (Belgium).

Information Processing Machines

Edited by V. Rubenik. 249 pp. Crown 4to. Iliffe Books Ltd. 1966. Price 55s.

This book contains the 21 papers presented at the six-session symposium held at Prague in September 1964 and organized by the Czechoslovak Scientific and Technical Society and the Research Institute of Mathematical Machines.

British Instruments Directory and Data Handbook

6th Edition

302 pp. Demy 4to. United Science Press Ltd. 1966. Price £8 8s.

The sixth edition of this directory now lists over 1 800 British manufacturers of instruments and components together with the overseas agents. Included also is a short dictionary in French, German, Spanish, Russian and Italian.

Conversion de l'energie

By S. S. L. Chang. 222 pp. Med. 8vo. Dunod Editeur. 1966. Price 46 F.

Originally published in America in 1963 under the title "Energy Conversion" this book has now been translated into French by J. B. Morea. It deals with the five main methods of energy conversion, namely, thermoelectric and thermonic conversion, magneto-hydrodynamic generators, solar and fuel cells.

Progress in Nuclear Energy Analytical Chemistry

By H. A. Elion and D. C. Stewart. 157 pp. Med. 8vo. Pergamon Press. 1966. Price 10s.

Library Planning for Automation

Edited by A. Kent. 195 pp. Med. 8vo. Macmillan. 1976. Price 84s.

This book is based on a conference held at the University of Pittsburgh in June 1964 and contains not only the papers presented at the conference but the lengthy discussions which followed.

Brains and Computers

By A. M. Andrew. 79 pp. Pott 4to. George G. Harrap & Co. Ltd. 1966. Price 10s. 6d.

RCA Transistor Manual

480 pp. Demy 8vo. 1966. Electronic Components and Devices, Radio Corporation of America. Price $1.50

Basic Industrial Electricity Part 1 and 2

By Van Valkenborgh, Nooger and Neville Inc. Part 1: 136 pp. Part 2: 132 pp. Med. 8vo. 1966. The Technical Press Ltd. Price 18s. each

This two part training manual was originally published in America and has now been revised for British and Commonwealth readership.

Analytical Chemistry Volume 4

Editors C. E. Crouthamel, D. C. Stewart and H. A. Elion. 116 pp. Med. 8vo. Pergamon Press. 1966. Price 70s.

High Speed Photography

By R. F. Saxe. 137 pp. Crown 4to. Focal Press. 1966. £3 3s.

The non-military applications of high-speed photography is the subject of this book. It deals with high-speed cine cameras, with streak and framing cameras together with single-shot, short-exposure picture taking devices. A chapter is given on electronic devices which make possible high-speed photographic methods.

Synchro Engineering Handbook

By A. R. Upson and J. H. Batchelor. 235 pp. Med. 8vo. Hutchinson Publishing Group. 1966. Price 63s.

This book has been written for the user of synchronous motors and similar rotating components and deals with the principles of operation and construction of the basic components.

A number of applications is described together with information on the methods of mounting, zeroing, adjustment and handling.

Electrical Who's Who 1966-1967

Compiled by 'Electrical Review'. 549 pp. Med. 8vo. Iliffe Books Ltd. 1966. Price 65s.

In the 1966/67 edition 1 250 new names have been added to this directory which now includes a total of 8 250 entries.

Junction Transistors

By J. S. Sparkes. 246 pp. Crown 8vo. Pergamon Press. 1966. Price 25s.

The object of this book is to explain the operation and characterization of junction transistors to the point from which detailed circuit analysis and design can be undertaken. It analyses the behaviour of semiconductors, pn junctions and all types of bipolar transistors from the standpoint of classical physics, together with the high-speed operation of transistors.

Electronics Reliability-Calculation and Design

By G. W. A. Dommer and N. B. Griffin. 238 pp. Crown 8vo. Pergamon Press. 1966. Price 25s.

This practical textbook deals with the reliability of modern electronic equipment and provides a reference to the various aspects contributing towards increased reliability of equipment and complete systems.

Fundamentals of Reliable Circuit Design Volume 1

By M. Xlander. Iliffe Books Ltd. Demy 8vo. 1966. Volume 1, 197 pp. Price 30s. Volume 2, 138 pp. Price 27s. 6d.

Volume 1 of this title introduces the five basic elements in electronics, voltage and current sources, resistance, capacitance and inductance followed by the design of elementary circuits.

An Introduction to Electrotechnology

By S. J. Kowalski. 354 pp. Demy 8vo. Chapman & Hall Ltd. 1966. Price 35s.

The second edition of this book has been confined to minor alterations and corrections of the original issue published in 1960. It also includes an additional chapter on transformers.

Mathematics for Electronics

ITT Federal Electric Corporation. 598 pp. Med. 8vo. Prentice-Hall International. 1966. Price 96s.

This book is designed as a self-instructional manual on the theory and practice of classical algebra, analytical geometry and complex algebra related to electronics.

Instrument and Chemical Analysis Aspects of Electronic Microanalysis and Macroanalysis

By D. C. Stewart and H. A. Elion. 256 pp. Med. 8vo Pergamon Press. 1966. Price 90s.

The general scope of this book covers electron optics, X-ray optics, detection and detectors. It also includes specimen preparation and observation X-ray emission, electron backscatter and absorption.

Electron Tubes

By R. G. Kloeffler. 262 pp. Med. 8vo. John Wiley & Sons. 1966. Price 45s.

This book is intended to serve as an introduction to the study of electronics in technical institutes and colleges. It covers the construction, theory of operation, and the characteristics of thermionic tubes, both vacuum and gaseous.

Induction Machines for Special Purposes

By E. R. Laithwaite. 337 pp. Med. 8vo. George Newnes Ltd. 1966. Price 84s.

The principles dealt with in this book deal with the construction and design of induction machines of unusual shape, purpose and mechanical arrangement, more frequently termed linear machines producing thrust and motion in a straight line.

Artificial Intelligence Techniques

By E. B. Carne. 149 pp. Med. 8vo. Macmillan. 1966. Price 57s. 6d.

Intended for use by the electronic engineer or scientist, this book provides a summary of electronic techniques for simulating human intelligence.

NEW EQUIPMENT

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

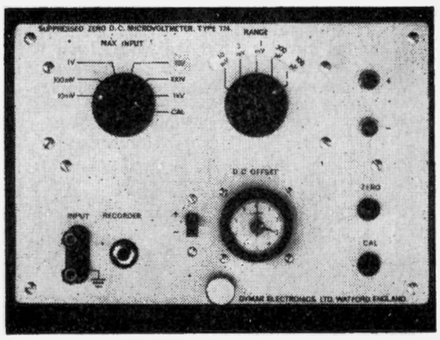
(Voir page 619 pour la traduction en français; Deutsche Übersetzung Seite 626)

SUPPRESSED ZERO D.C. MICROVOLTMETER

Dymar Electronics Ltd, Rembrandt House, Whippendell Road, Watford, Hertfordshire

(Illustrated below)

An addition to the Dymar range of plug-in units is the suppressed zero d.c. microvoltmeter type 724. This unit permits the measurement of incremental changes of large voltage. Typically a change of 1μV in 10mV or 100mV in 1 000V can be measured and recorded. The zero suppression is obtained by means of a 10-turn helical potentiometer having 1 000 scale divisions and the meter range scale is mechanically coupled to the input attenuator switch to give unambiguous indication of meter full scale sensitivity. A recorder output is provided and a typical application is the measurement of regulation and stability of power supplies.



An input of 10mV f.s.d. to 1 000V f.s.d. is covered in six ranges, while the voltmeter covers 100μV to 1 000V in 15 ranges of 1, 3, 10 sequence. The input impedance is 1MΩ on the 10mV, 100mV and 1V ranges and 100MΩ on the 10V, 100V and 1 000V ranges.

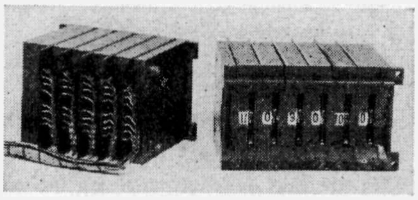
EE 97 751 *for further details*

PRE-WIRED SWITCH ASSEMBLIES

Digitizer Techniques Ltd, 26 Sheen Road, Richmond, Surrey

(Illustrated above right)

The 'Digitswitch' made by Digitizer Techniques Ltd is now available as a pre-wired assembly tailored to the user's connexion requirements. The switch itself is of the stacking thumbwheel type allowing an almost unlimited choice of arrangements within very compact dimensions. This new pre-wiring service provides a complete harness of connexions terminating in a multiple plug or plugs, edge connector or other device



to suit the user's equipment.

In this way the Digitswitch can be installed in a matter of minutes no matter how many connexions are involved, and subsequent circuit modifications and servicing become simple.

EE 97 752 *for further details*

H.F. BETA TESTER

Cathodeon Electronic Ltd, Bircham Road, Southend-on-Sea, Essex

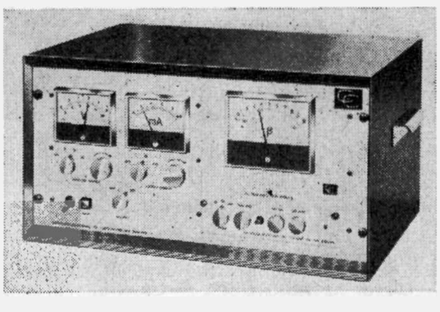
(Illustrated below)

The Precision Measurements Division of Cathodeon Electronic Ltd is now producing a new β tester, the PMD 401.

The unit is a direct reading instrument for the measurement of high frequency β gain of transistors in the common emitter mode at a frequency of 20Mc/s or 100Mc/s.

Indication of collector voltage, emitter current and high frequency β is by separate meters. The testing, under small-signal conditions, of transistors of either polarity can be carried out. Once bias conditions for a particular type of transistor are set, subsequent transistors of a like type may be measured merely by plugging in the transistor and reading off β.

Designed for the laboratory testing of batch samples of transistors by semi-skilled personnel, the PMD 401 provides, in a single unit, the functions of a high frequency oscillator, calibrated gain amplifier, detector and transistor power supplies.



EE 97 753 *for further details*

D.C. MICROVOLTMETER/ NANOAMMETER

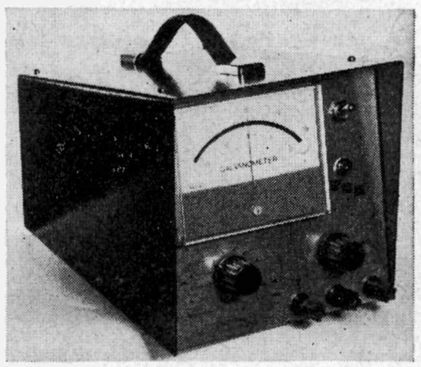
Test Equipment Repair, Leigh Road, Leigh, Lancashire

(Illustrated below)

The 'chopper galvo' is a transistorized centre reading microvoltmeter, nanoammeter, and null detector, with fully floating input, having six full scale voltage ranges from 100μV to 1V at 10MΩ/V and six full scale current ranges from 100nA to 500μA, with a maximum input resistance of 1kΩ.

The instrument consists of a field effect solid state chopper, a highly stable a.c. amplifier, and a phase sensitive detector. This method of d.c. amplification produces great stability, and zero drifts of less than 0·5μV/°C.

To improve its usefulness the instrument has been designed for measure-



ments into both high and low impedances, and due to the type of chopper employed the offset voltage is negligible; therefore only one zero control is necessary.

A recorder output and attenuator is included making the instrument useful as a very stable low level pre-amplifier for potentiometer recorders etc.

Overload conditions, up to 6V on the 100μV range are claimed.

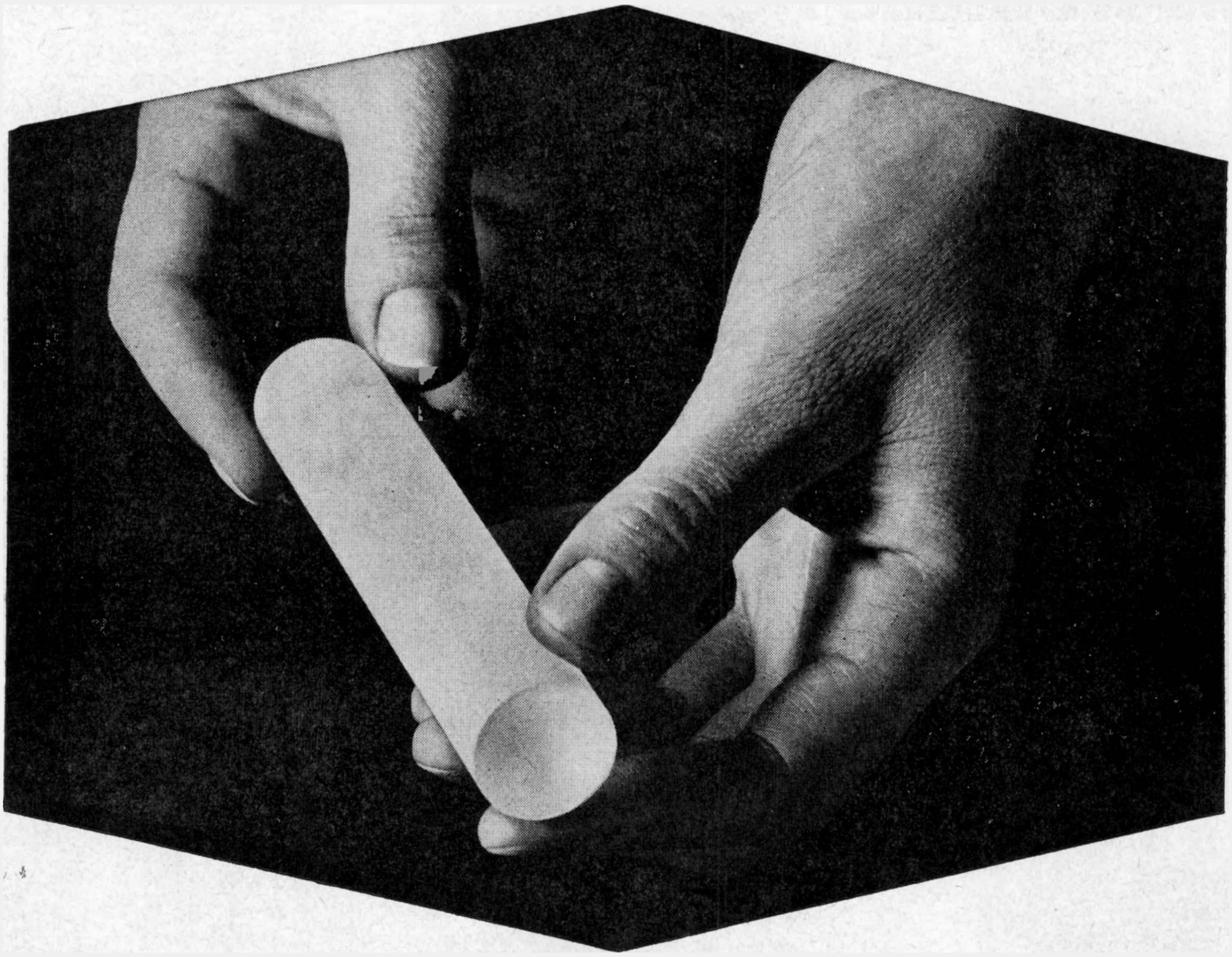
EE 97 754 *for further details*

COMPONENT BRIDGE

The Wayne Kerr Co. Ltd, Sycamore Grove, New Malden, Surrey

(Illustrated on page 611)

Wayne Kerr has produced a low-cost bridge giving 1 per cent measurements of *L*, *C* and *R* over an extremely wide measurement range. Two dials, both direct-reading, give simultaneous values for the *R*· and *C*/*L* terms. A novel feature is the provision of an internal



The ultimate in laser rubies

50 per cent of the output energy contained within 0.1 x 10-3 radians per linear inch

In terms of physical properties and performance characteristics the Developmental Quality Laser Ruby now available from Union Carbide Limited is virtually perfect. Produced by the unique Linde process it has a beam divergence which contains 50 per cent of the output energy within 0.1×10^{-3} radians per linear inch of rod and no less than 90 per cent within 0.2×10^{-3} radians per linear inch of rod. Internal scattering is less than one per cent. Dislocation density is negligible. And the fringe count as determined by interferometric analysis is less than 1.5 fringes per linear inch—regardless of rod diameter. Moreover, the homogeneity and hexagonal crystalline structure of the host material approximates to the theoretical ideal. As a result, the Developmental Quality Ruby combines high energy density and spot brightness with high durability and a long useful life.
For applications where the use of a Developmental Quality Ruby would not be justified, Union Carbide supply a number of other high quality rubies, foremost among which are the S.I.Q. and the Standard grade rubies. There are also the revolutionary new low threshold YAG crystals suitable for continuous operation at room temperature. YAG is doped singly with Neodymium or double doped Neodymium and chromium. In addition, the company markets electro-optical crystals for use in modulation, Q-switching, high efficiency frequency doubling, tunable lasers and other critical applications, and electronic sapphire substrates, silicon monoxide and alumina powders. Having been involved in laser research since its inception, and now being among the world's foremost suppliers of lasers and laser equipment, Union Carbide is in a unique position to advise on the application of synthetic crystals and associated materials. If you are currently engaged upon work in this field a discussion with Union Carbide could save you a great deal of time and quite possibly reduce your development costs by a very considerable margin.

Synthetic Crystals

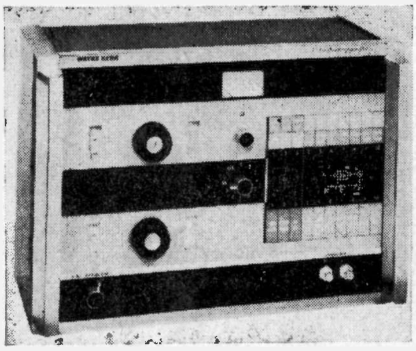
another practical product of

UNION CARBIDE

UNION CARBIDE LIMITED · KEMET DEPARTMENT · 8 GRAFTON STREET · LONDON W1 · MAYFAIR 8100

CHEMICALS · CRYOGENICS · DYNEL FIBRE · ELECTRIC WELDING · ELECTRONICS · FERRO-ALLOYS · FLAME PLATING · PRESTONE CAR POLISH · SILICONES · SUPERALLOYS · UCON LUBRICANTS

ECR88



frequency-doubling circuit for the power-derived source. Thus measurements can be made at 50 and 100c/s (or 60 and 120c/s) at the turn of a switch.

Simplicity of operation has been achieved by careful layout of the few operating controls and the provision of connexion diagrams, operating sequences and a table of ranges on the front-panel plaque. The transistorized detector amplifier has a logarithmic response and is so designed that under no circumstances can the associated null meter give an ambiguous deflexion due to overload conditions. This feature permits rapid location of the most suitable measurement range and a precise determination of the final balance point.

Special provision is made for evaluation of the *L*, *C* and *R* constants of very low impedance circuits. Third and fourth measurement connexions are available when components are to be checked while in-circuit or when attenuators and filters are under test. The overall coverage is 1MΩ to 1 000MΩ, 1pF to 5F and 1μH to 500kH.

EE 97 755 *for further details*

MODULAR COUNTING SYSTEM

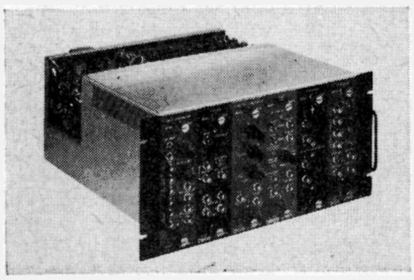
Distributed by: High Volt Linear Ltd, 67 Dudley Street, Luton, Bedfordshire

(Illustrated below)

Now available from High Volt Linear Ltd is the E.G. & G. (U.S.A.) M100 modular counting system.

This is a general purpose real time data handling system designed specifically for applications where high data rates, short resolving times and insensitivity to noise are of utmost importance. Since the system is d.c. coupled and therefore rate insensitive, it is capable of truly asynchronous operation, moreover it is capable of operating at continuous and aperiodic rates in excess of 100Mc/s.

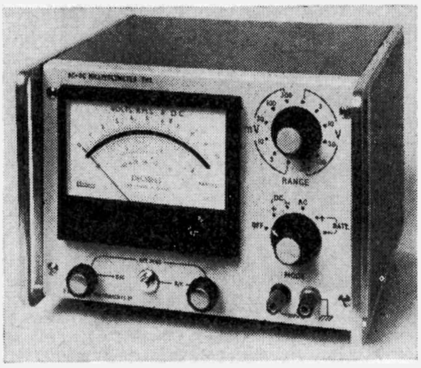
The M100 modular counting system is composed of a broad range of modular instruments and accessories including discriminator/triggers, coincidence AND/OR units, linear gates, amplifiers, stretchers, pre-scalers, time-to-height convertors, etc.



The discriminator/triggers and logic modules are entirely direct coupled. This, and the absence of regenerative elements in the logic modules means that the system is capable of random-input real time operation at 100 per cent efficiency (duty factor) from d.c. to 100Mc/s. Included in the M100 system are a number of versatile stable direct-coupled linear instruments for signal processing which feature wide bandwidth and excellent behaviour under overload.

Typical applications of the system include high energy nuclear physics research, nuclear structure investigation, laser research, optical astronomy, radar system studies and computer interface work.

EE 97 756 *for further details*



TRANSISTORIZED MILLIVOLTMETER

Farnell Instruments Ltd, Sandbeck Way, Wetherby, Yorkshire

(Illustrated above)

The transistorized millivoltmeter type TM1 is a general purpose, battery operated instrument for a.c. or d.c. measurements. It operates from two PP11 (or equivalent) batteries and covers measurements from 1mV f.s.d. to 300V f.s.d. in twelve ranges.

The frequency range on a.c. is 10c/s to 100kc/s and the input impedance below 30mV is 100kΩ, from 30mV to 1V is 1MΩ and above 1V is 10MΩ shunted by 40pF.

On d.c. the input resistance is 1MΩ/V to 10MΩ maximum.

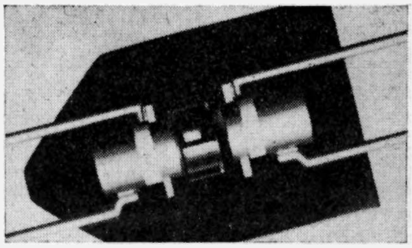
The accuracy of a.c. is 4 per cent of f.s.d. and on d.c. is 3 per cent of f.s.d.

EE 97 757 *for further details*

OPTOELECTRONIC ISOLATOR

Texas Instruments Ltd, Manton Lane, Bedford

(Illustrated above right)



A new optoelectronic coupling device which permits economical high-voltage electrical isolation up to 5kV has been introduced by Texas Instruments Ltd. The new optical isolator, designated TIML101, combines a TI planar silicon light sensor (LS600) with a gallium arsenide light source (TIXL101) in a single opaque epoxy package.

The new device is designed to provide electrical isolation where circuit feedback problems exist. As a replacement for electromechanical relays, it offers significant advantages in switching speed, reliability, mechanical ruggedness, and small size. Though capable of handling very high voltages, the TIXL101 is sensitive to small signal changes Therefore it is particularly suitable for application in high-voltage low-current telecommunications relay lines.

Switching speed greatly exceeds that of the fastest relays permitting the transmission of more information with fewer devices. Capable of flat response beyond 10kc/s, the TIXL101 has a typical reverse switching time (t_r) of 1·5μsec and a forward switching time (t_f) of 15μsec. Input current rating is 50mA. Output is 250μA minimum.

The incorporation of two proven hermetically sealed components in a solid, one-piece epoxy package results in a physically rugged component suitable for heavy-duty industrial applications where they are subjected to high vibration, shock, and other environmental extremes. Contact 'chatter' or 'bounce', often encountered when relays are subjected to high vibration, is completely eliminated. The device provides stable performance over a broad temperature range from −55° to +125°C.

The TIXL101 is encased in an electrically isolated cylindrical package measuring only 0·22 by 0·35in.

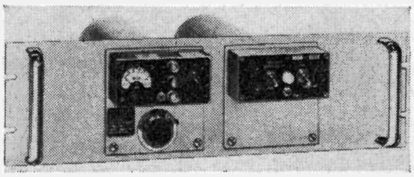
EE 97 758 *for further details*

PRECISION FREQUENCY STANDARD

Distributed by: Racal Instruments Ltd, Dukes Ride, Crowthorne, Berkshire

(Illustrated below)

This new crystal frequency standard produced by Sulzer of U.S.A. the model 2·5C, has a typical frequency stability of at least 1 part in 10^{11} per day, with output frequencies of 5Mc/s, 1Mc/s and 100kc/s. Spectral purity filters are fitted to the 5Mc/s and 1Mc/s outputs, which provides output frequencies with spectral purity better than 80dB to allow a high degree of frequency multiplication of these frequencies.



The proven features of Sulzer frequency standards are retained. All-silicon solid-state circuits are used, and both inner and outer ovens have proportional control. Inner oven temperature is adjusted to the zero-temperature coefficient of the crystal to ensure maximum stability. Designed for military application and capable of withstanding storage temperatures between −28°C and +70°C, and of withstanding shocks of 30*g* with a frequency change of less than 1 in 10^8, hermetically sealed for operation at 90 per cent relative humidity.

A precise, temperature-controlled a.v.c. system maintains crystal power within a few hundredths of a decibel at approximately 1μW for optimum stability. The high stability, extreme reliability, with m.t.b.f. exceeding 20 000 hours, and wide temperature range make this frequency standard a suitable frequency source for standard-frequency broadcasting, laboratory measurement, precise time systems, microwave spectroscopy and observatory time-keeping.

Sulzer products are available from Racal Instruments Ltd who are agents on a near-world-wide basis for the Tracor Group of which Sulzer is a member.

EE 97 759 *for further details*

MODULAR POWER SUPPLIES

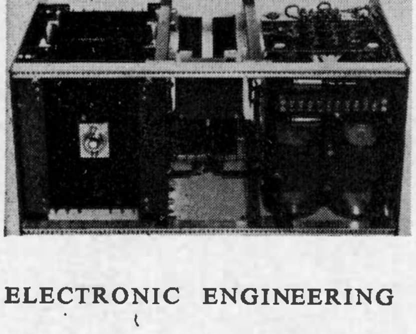
Standard Telephones & Cables Ltd, Components Group, Footscray, Sidcup, Kent

(Illustrated below)

Standard Telephones & Cables Ltd has recently introduced a series of high-performance stabilized power supply sub-units specially designed for incorporation into transistorized electronic equipment. Modular in construction, they can readily be arranged to fulfil any constant voltage requirement up to 10A, 50V d.c., both in performance and mounting flexibility, thus saving the considerable amount of time, effort and money involved in the design of special units.

Each of the new units consists of four basic modules (including connector) housed in a standard frame fitted with captive nuts for ease of installation. Modules can be removed from the frame and regrouped to fit virtually any space available.

Suitable for operation from either 110 to 125 or 200 to 250V a.c. mains 45 to 65c/s, units are available in three pre-set voltage ranges; 0 to 16, 0 to 30, and 0 to 50V d.c., outputs being pre-set at the factory. Output can be readily re-adjusted on site to suit changing requirements.



Output stability is such that fluctuations of ±10 per cent in mains input only alter output level by 0·001 per cent, even at 65°C (139°F) ambient.

All units utilize silicon semiconductor devices and high quality components of ample rating for high reliability. Full overload protection is incorporated, the standard circuit being manually reset.

The photograph shows the largest unit in the range.

EE 97 760 *for further details*

STROBOSCOPE

Lunartron Electronics Ltd, Chester Works, Chester Avenue, Luton, Bedfordshire

(Illustrated below)

The model 1209 is an addition to the range of stroboscopes manufactured by Lunartron and is designed as an industrial instrument for use in conditions where occasional mechanical shock is unavoidable. All controls and facilities are readily accessible including a socket for operating an external lamp. A wide range of frequencies is covered and the speed of operation is indicated by a drum read-out meter inside a window on the upper face of the instrument.

The equipment is fully transistorized and stabilized against fluctuations in mains voltages.

The flashing rate of 60 to 15 000 flashes/min is covered in three ranges, the accuracy being ±3 per cent of f.s.d. The flash duration is 5 to 10μsec.

EE 97 761 *for further details*

CHARGE AMPLIFIERS

Distributed by: Technitron Ltd, Walmgate Road, Perivale, Greenford, Middlesex

(Illustrated above right)

Available from Technitron Ltd are the Data Control Systems models GCA-1 and GCA-2 all solid-state charge amplifiers designed to measure the output of piezoelectric transducers and high-level devices without being affected by the length of the connecting cable between the transducers and the amplifier input. These amplifiers have exceptionally broad band response, low noise output, excellent linearity, and are designed so that no restrictions are placed on the leakage of the transducer. The amplifiers employ the latest techniques to measure a wide dynamic range of input charge from 10pC to 30 000pC with shunt cable capacitances which are in excess of 0·1μF.

Both the GCA-1 and GCA-2 provide dual output amplifiers; a voltage output for magnetic tape recording and a high current output for galvanometer drive purposes. The model GCA-2 additionally incorporates an average-of-peaks meter (120 per cent full scale) and a normalized output for servo-control.

Cable lengths of up to 10 miles (1μF) may be used. The calibrated dial permits the setting of the gauge factor of the transducer used. Built-in facilities provide gain calibration within ±1 per cent.

EE 97 762 *for further details*

PHOTO-ELECTRIC CONTROL UNIT

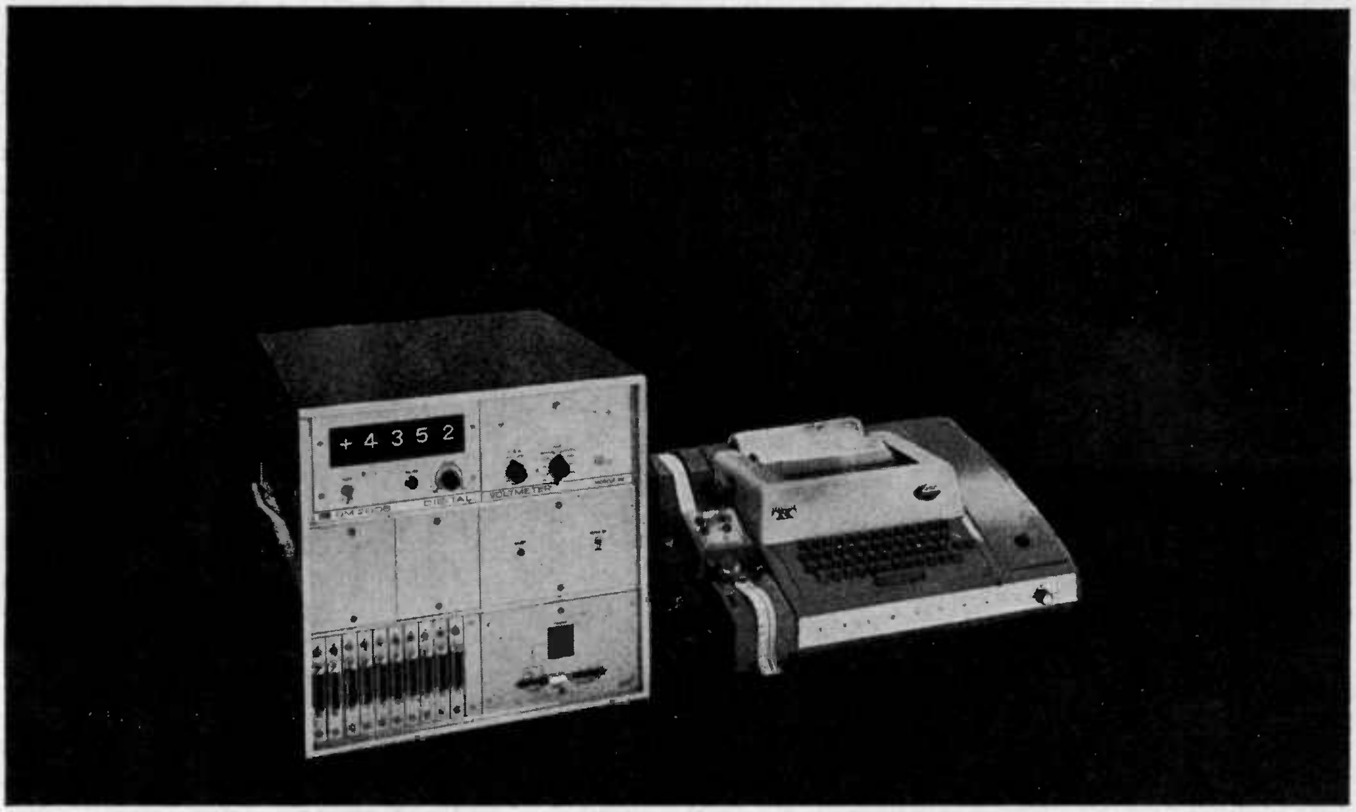
Kappa Electronics Ltd, 159 Hammersmith Road, London, W.6

(Illustrated on page 613)

The model FA15 transistor control unit has been developed as a general purpose device with numerous applications, these include, detection of registration markers, limit switches, electronic adjustment of machinery, safety fences, detecting conveyor belt stoppages, operating gate mechanisms, sizing, sorting, level detectors, alarms, and other general supervisory duties, etc.

The conventional on/off output control contacts are provided upon the breaking and the re-establishment of the light beam. This switching action can be supplemented by input paralysis timing (i.e. delay of control action for a preset time) or by timed interval functions where the output is maintained for a pre-set time. Special plug-in boards can be provided to differentiate input pulse widths, and response times down to 1msec can be catered for. In addition various output devices such as solid-state switching are available in place of the standard relay.

Fail-safe (relay energized 'light' or 'dark') conditions can be obtained, and the provision of a variable control enables adjustment of the switching point to meet ambient lighting conditions. A second variable potentiometer gives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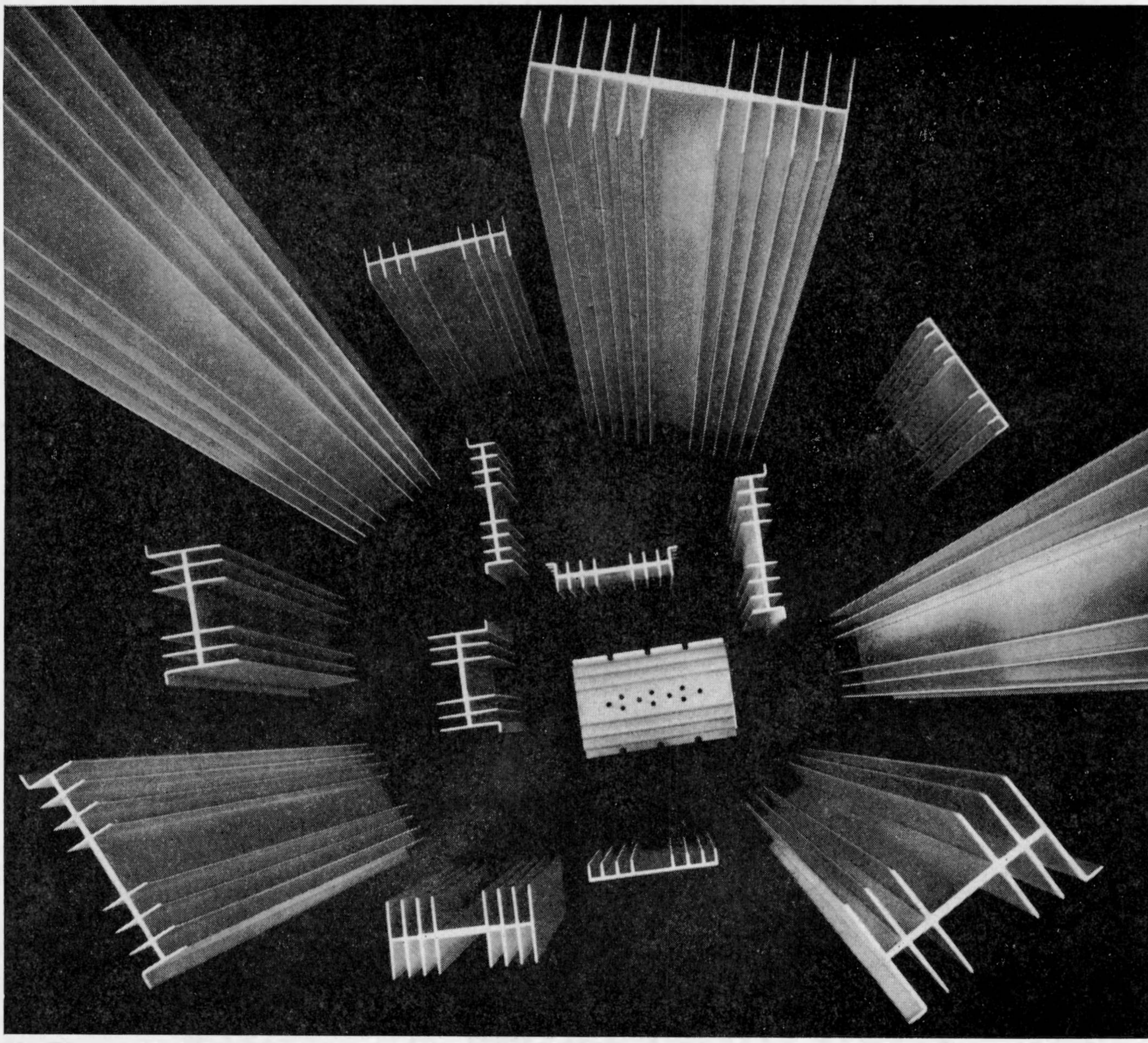
DIGITAL VOLTMETER

EE 97 118 *for further details*



From man-size to midget-size

The need for progressively smaller and more compact equipment creates its own cooling problems, and Marston Heat Sinks offer the electronic engineer outstanding advantages in his approach to design.
Marston Heat Sinks, cooled by natural air convection, are specially designed for use with power transistors and semi-conductor devices.
Marston Heat Sinks have low thermal resistances, are compact and light in weight. They have been extensively tested and are performing satisfactorily in many different types of equipment.
Marston Heat Sinks are supplied in a wide selection of lengths, hole patterns and surface finishes.
Marston Heat transfer equipment has the reliability which is the result of 50 years experience in the design and manufacture of heat exchangers.
Fill in the coupon for further information on Marston Heat Sinks.

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Please send me full details of **Marston Heat Sinks**

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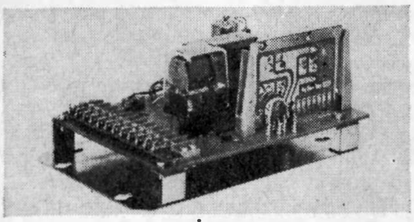
MAR 360



control of the time function. Any, or both, can be mounted remote from the control unit.

The unit consists of a printed circuit base board, containing a mains stabilized power supply, provision to plug-in the selected circuit board, and also an octal base to accommodate the output device. All connexions are brought out to a 12-way quick make-and-release terminal strip to facilitate easy removal for maintenance purposes.

The control unit can be supplied for use with the standard cadmium sulphide cell, or the fast-response silicon photo duo-diode, and is available with a keyhole mounting plate for incorporation into customers' own comprehensive control systems, or as a self-contained unit in a weatherproof sheet metal housing 9in × 6in × 4in.

Should a multi-channel unit be required to incorporate a number of circuits performing separate functions, then a special base board holding up to six plug-in boards, and a 6-way relay board, in a compact housing 12in × 9in × 6in, are available. This, of course, is also suitable for a complete control system when used in conjunction with the many other types of circuit boards such as timing, channel selection, sequencing, etc., which can be supplied. It will also accommodate a range of transducer amplifying circuits to meet the requirements of the system.

EE 97 763 *for further details*

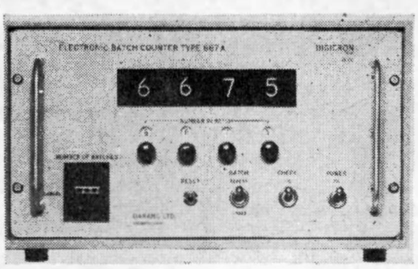
BATCH COUNTERS

Darang Electronics Ltd, Restinor Way, Hackbridge Road, Hackbridge, Surrey

(Illustrated below)

Nine new models are included in the 667 series of digital electronic batch counters recently introduced by Darang Electronics Ltd. The standard range has 2, 3 and 4 decade versions, accommodating maximum batches of 99; 999 and 9 999 respectively. These are available with or without a total batch indicator and with or without the in-line digital display.

All are fitted with pre-batch facilities, which may be preselected by the



user to meet varying requirements. The pre-batch signals may be controlled from any one of the decades as required.

The designs are based on a system of low cost plug-in boards and use extremely reliable cold cathode techniques of advanced design.

Single batch or automatically repeating modes are provided and a self-checking facility is incorporated. Batch selection is made by ten-way rotary switches, with numbered dials which are protected by the front panel.

Input signal may be between 2V and 300V. The input circuit is d.c. coupled and responds to positive going signals from earth. The input resistance is 50kΩ. Counts may also be actuated by external contacts. The output contacts are rated at 5A, 230V a.c.

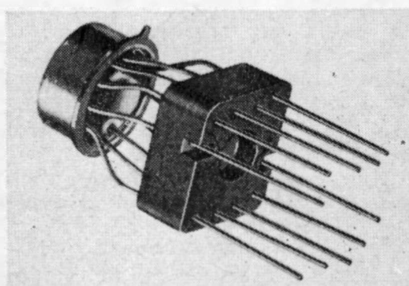
EE 97 764 *for further details*

MICROCIRCUIT MOUNTING PAD

Jermyn Industries, Vestry Estate, Vestry Road, Sevenoaks, Kent

(Illustrated below)

The mounting pad type AE-10L is for use with ten lead TO5 microcircuits.



Made of I.C.I. A.100 Nylon (melting point 200°C) the pad spreads the TO5 leads to two parallel lines each containing 5 leads to 0·1in matrix. Thus the lead pattern is converted to conform to standard matrix printed circuit boards and TO5 microcircuits can be mounted more easily than flat packages.

EE 97 765 *for further details*

INFRA-RED DEVICES

M.C.P. Electronics Ltd, Station Wharf Works, Alperton, Wembley, Middlesex

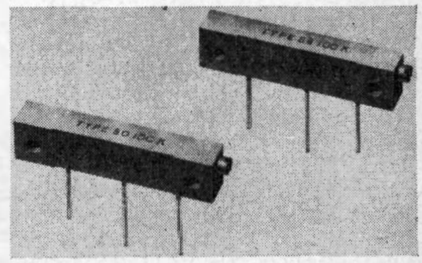
To supplement their MGA 600 range of gallium arsenide infra-red light emitting diodes, M.C.P. Electronics Ltd is now manufacturing a series of photo-receiving devices known as types MSP/3 and MSP/6.

These devices provide extremely high sensitivities; output currents up to 100mA enable direct operation of a conventional electromagnetic relay.

They are of silicon structure having a peak spectral response at 9 000Å but may also be used under visible light conditions.

For low current operation the 30F2 series of silicon planar microphotodiodes offer the advantage of a compact physical size and are therefore suitable for stacking in punched-tape reader applications.

EE 97 766 *for further details*



POTENTIOMETERS

Morganite Resistors Ltd, Bede Trading Estate, Jarrow, Co. Durham

(Illustrated above)

Two cermet linear-motion trimming potentiometers for printed circuits are now included in the Morganite range. Known as types 80 and 88, the area of board space which they occupy is less than a quarter of a square inch (about 1·5cm²). They are available in selected values between 10Ω and 2MΩ.

Type 80, rated at 0·75W at 70°C, provides a reliable and inexpensive means of circuit adjustment for commercial use.

Type 88 is designed for operation under the most testing service conditions. It is robust, water-resistant, and meets the requirements of MIL-R-22097B Characteristic C with a rating of 1·0W at 85°C.

Both types have terminal pins plated with noble metal and spaced for the standard 0·1in grid. The resistance track is unaffected by humidity and is resistant to chemicals. Circuit adjustment can be made to extremely fine limits, and the method of manufacture makes catastrophic failure impossible.

EE 97 767 *for further details*

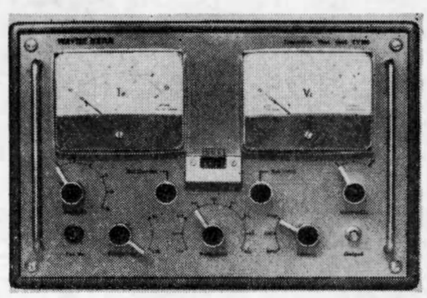
TRANSISTOR TEST UNIT

The Wayne Kerr Co. Ltd, Sycamore Grove, New Malden, Surrey

(Illustrated below)

Wayne Kerr has introduced a new test unit for the measurement of all *h* parameters of pnp and npn transistors, at 1 000c/s, in the common-emitter or common-base modes of operation. The test unit, TT100, includes facilities for setting-up and monitoring emitter current (0 to 30mA in six ranges) and collector voltage (0 to 30V in three ranges).

Operation is based on the use of an oscillator having very high amplitude-stability, accurate resistive networks to apply the 1kc/s signal, at high impedance, to all three connexions of the transistor, and a precision a.f. milli-



voltmeter. A measurement accuracy of ±3 per cent f.s.d. has been achieved for all parameters. Voltage and current reading accuracy is ±1 per cent f.s.d.

A single connecting block (incorporating a fourth socket for a screen lead) is used for all measurements, the appropriate connexions being established automatically by operating one parameter selector switch. A jack is provided for use when collector voltages exceeding 30V are required, permitting external supplies and metering to be employed.

EE 97 768 *for further details*

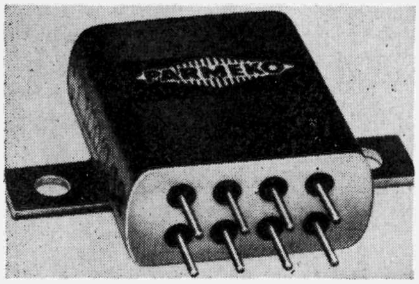
LATCHING RELAYS

Parmeko Ltd, Percy Road, Aylestone Park, Leicester

(Illustrated below)

Parmeko Ltd announces an extension to its G100 range of miniature sealed relays to include single coil two-pole changeover latching relays. Known as type G100 L, they are intended for applications where high performance and reliability are essential factors.

With a sensitivity of only 30mW these relays are available in coil voltages from 1·5 to 48V. Contact life is greater than 10^6 operations at less than 0·3A



and 10^5 operations at 0·3A to 1A. Type G100 L relays can withstand a linear acceleration of 100*g*, vibration of 25*g* peak acceleration 10c/s to 3 500c/s with a relay hard mounted, and will operate within the temperature range of −55°C to +125°C.

Basic dimensions are 0·920in × 0·820in (crystal can), and the unit weighs approximately 20g. Maximum reliability has been achieved by sealing the coils and contacts in separate compartments, and the use of a gold alloy for the contact material.

Parmeko Ltd can also supply a series of double coil latching relays.

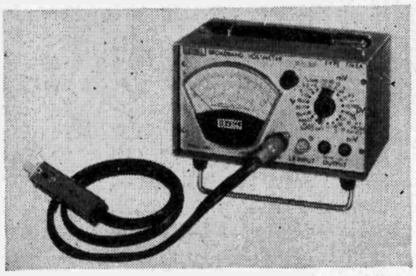
EE 97 769 *for further details*

WIDEBAND VOLTMETER

Levell Electronics Ltd, Park Road, High Barnet, Hertfordshire

(Illustrated above right)

This portable instrument (type TM6A) measures signals down to microvolt levels over the frequency range from 1c/s to over 100Mc/s. Size and appearance are similar to Levell transistor a.c. microvoltmeter type TM3A except for the addition of an h.f. probe. Eight h.f. ranges are provided of 1mV f.s.d. to 3V f.s.d. with a sensitivity of 300µV over the band 300kc/s to 50Mc/s and



3mV at 400Mc/s. Fourteen l.f. ranges are provided from 50µV f.s.d. to 500V f.s.d. for measurements from 10µV over the band 1c/s to 3Mc/s. These ranges are the same as those provided on type TM3A except for the omission of the 15µV and 150µV ranges.

The h.f. ranges use all semiconductor circuits to convert the h.f. signal into a square wave of frequency about 20c/s and amplitude proportional to the square of the h.f. signal. No mechanical chopper is used and the circuits are adequately temperature compensated. The efficiency is high resulting in a power consumption of only 10mA from a 9V battery.

The instrument responds to true r.m.s. on all the h.f. ranges but on the l.f. ranges the instrument responds to the mean and is calibrated in terms of r.m.s. for a sinusoidal input.

EE 97 770 *for further details*

DIGITAL TYPEWRITERS

Hilger & Watts Ltd, 98 St. Pancras Way, Camden Road, London, N.W.1

(Illustrated below)

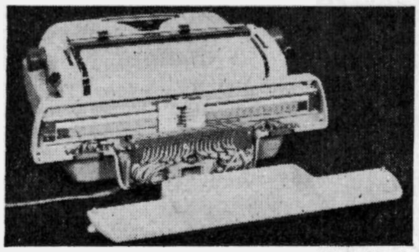
Electrically-operated digital typewriters, with or without programming facilities, are announced by Hilger & Watts Ltd for use with automatic data-processing systems. The basic typewriter FD 574 has a 17in carriage and an elite type face, and is for remote operation under the control of electronic circuits.

Character and function keys adapted for solenoid operation are the numerals 0 to 9, tabulator, full stop, space, carriage return/line feed, asterisk, letters A and R, and red/black ribbon change.

Additional or alternative keys can be modified for solenoid operation to special requirements, up to a total of 26 keys and the 'space' function.

A commutator bar and patchboard unit FD 576 is available for fitting to the basic machine, so that variations in the layout of printed information can be selected on the typewriter, by interchangeable plug-in programming units made up to suit customers' requirements.

The 'combar' acts as a serializer and



ensures that the printed readout remains synchronized with the source of the information. It also allows the typewriter to run at its optimum speed of 8 to 10 characters/sec.

The maximum capacity of the combar is 190 columns, and the typewriters accept a maximum of 50 input lines.

The photograph shows a rear view of the typewriter with the 'combar' cover plate removed.

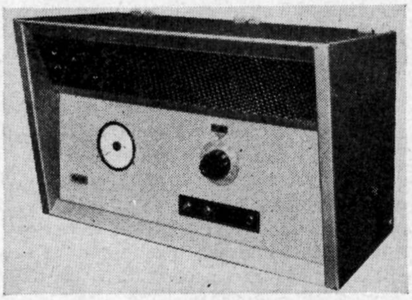
EE 97 771 *for further details*

THYRISTOR SPEED CONTROLLERS

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

(Illustrated below)

The speed of d.c. electric motors can be accurately controlled from rest to base speed with a new range of controller units announced by the Automation Division of The M.E.L. Equipment Co. Ltd. Known as the 'Ergotrol' range, the units enable the speed of d.c. motors to be varied manually or by a control signal from a parent equipment or system. Silicon controlled rectifiers (thyristors) are used to control the motor armature supply, and operation is claimed to be smooth, efficient and reliable.



The units operate from a.c. mains, and the range comprises seven models designed for use with d.c. shunt wound motors from 1 to 40 h.p. to British Standard 2613. Regulation at the set speed is determined by feedback from the armature supply and is within 2½ per cent of base speed. If greater accuracy is required, a separate tachogenerator can be employed to provide the feedback signal, the regulation then being within 1 per cent of base speed. An important advantage of the tachogenerator method is that its accuracy is unimpaired by field winding heating.

Starting is controlled by solid state circuits which safeguard against excessive currents even if the drive is started when the speed control is set to maximum. Also, there is armature current limiting while allowing an adequate margin for acceleration.

A slow motion drive to the speed control, and start and stop controls, are fitted as standard. A number of optional facilities are available, including a speed indicator calibrated in percentage speed or to individual requirements. A load indicator can also be fitted calibrated in current. Reverse running is another optional facility; this allows reverse running to be selected manually and includes automatic braking to a halt before reverse running commences. Dynamic braking for rapid stopping can also be provided. In addition, all controls and indicators can be supplied on a separate panel for remote working.

Units up to 10 h.p. operate from single-phase mains and are wall-mounted, while units above 10 h.p. operate from three-phase mains and are enclosed in floor standing consoles. If required, units can be supplied uncased for incorporation in other equipment.

EE 97 772 *for further details*

OXYGEN ANALYSER

Servomex Controls Ltd, Crowborough, Sussex

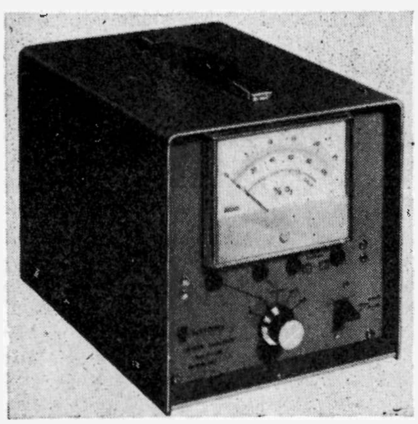
(Illustrated below)

Servomex Controls Ltd has introduced a new battery powered portable—type OA.150—to supplement its range of oxygen analysers. The new instrument has two switched ranges, 0 to 25 per cent and 0 to 100 per cent, with an accuracy of ± 1 per cent f.s.d. on each range.

Simplicity of operation is the key feature of the instrument, the main controls consisting of a range selector switch and an operating key. The reading is presented on a built-in indicating meter. The selector switch has further positions in which the battery voltages and correct circuit adjustment may be checked, using the same meter. A graphic display on the front panel makes the operating and checking procedure very obvious, and symbols are used wherever possible in place of words which makes the instrument equally suitable for overseas markets.

The analyser may be used with flowing or static samples, the reading being unaffected by sample flow rates within the range 0 to 150 ml/min. A sintered glass disk filter and valves for adjusting the analyser and by-pass flows are built into the instrument, and a hand aspirator and drying tube are included as standard equipment.

The analyser uses the same measuring cell as previous Servomex oxygen analysers. A quartz dumb-bell is suspended on a platinum filament in a non-uniform magnetic field. It experiences a torque proportional to the magnetic susceptibility of the sample gas, and this is measured by maintaining an equal and opposite restoring torque produced by current flow in a single turn coil mounted on a dumb-bell. A light source, twin photocell, and difference amplifier maintain the null-balance condition automatically. The output meter measures the restoring current, which is directly proportional to the oxygen content. Silicon photocells and transistors are used and the instrument is extremely rugged and reliable. A temperature compensating system maintains the specified accuracy over variations of ± 5°C.



The measurement is virtually unaffected by all common gases other than oxygen, and the instrument may thus be checked on nitrogen and oxygen, or air, and subsequently used to measure mixtures including, for example, nitrous oxide, carbon dioxide or hydrogen.

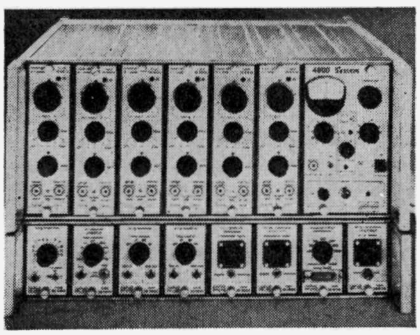
EE 97 773 *for further details*

MULTI-CHANNEL AMPLIFIER

S.E. Laboratories (Engineering) Ltd, Astronaut House, Feltham, Middlesex

(Illustrated below)

S.E. Laboratories (Engineering) Ltd has introduced a rack or table mounted multi-channel amplifier system SE.4000 which can accept up to eight different or identical high performance amplifiers.



The system is specifically designed for use with thermocouples, resistance temperature bulbs, strain gauges, differential transformer devices, half or full bridge resistive or variable reluctance transducers, velocity pick-ups, self-generating transducers, flowmeters, millivolt signals, etc.

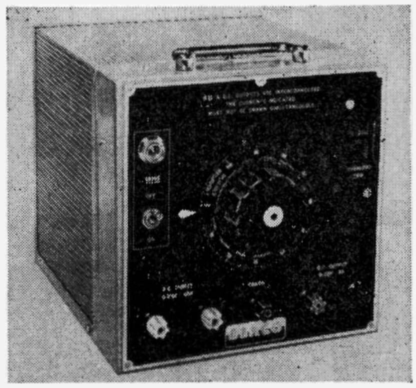
A feature is the optionally available built-in calibration system which can be manually or automatically operated. A continuous mode of calibration signals therefore can be programmed from a computer or data logger which can act as a reference signal from any input data.

Each amplifier module is fitted with a panel-mounted overload indicator. The drive circuit can also be connected to an external alarm device energized when input data deviates from pre-set limits.

All amplifier channels have a standardized output of $\pm 1{\cdot}4$V 10mA. A pick-a-back plug-in drive amplifier board can be plugged into each amplifier module raising the output to ± 10V ± 100mA.

Four basic types of amplifier module, all interchangeable within the Series 4000 rack housing, are available.

EE 97 774 *for further details*



LOW VOLTAGE POWER SUPPLY

The British Electric Resistance Co. Ltd, Queensway, Enfield, Middlesex

(Illustrated above)

The BERCO low voltage power supply unit is a convenient source of low voltage a.c. and d.c. power for use with educational apparatus and developed by BERCO in association with The Nuffield Foundation.

A double wound rotary 'REGAVOLT' variable transformer and a selenium full wave bridge rectifier are mounted in a ventilated steel case with a front panel carrying a mains on/off switch, mains indicator neon, the 'Regavolt' an overload trip, 2 a.c. output terminals, 2 d.c. output terminals and earth terminal.

The supply is brought in through a double pole mains switch to a pilot indicator lamp and thence to the primary winding of the double wound 'Regavolt' variable transformer. This is provided with supply voltage tappings to enable any supply voltage between 200 and 250V to be accommodated in steps of 10V.

The secondary winding of the 'Regavolt' fully isolates the output from the supply and the brush provides a smooth infinitely variable output of 0 to 25V.

A circuit-breaker in series with the output brush of the 'Regavolt' transformer protects the equipment from the effects of short-circuits, or continuous overloads. The output from the circuit-breaker is taken to a full wave bridge rectifier of the selenium type having a very high overload capacity.

No attempt has been made to smooth the d.c. output as the intention is to provide a d.c. supply at the lowest possible cost for schools for experimental use. The amount of smoothing may vary with different applications.

Each output is taken to a pair of insulated screw terminals. The a.c. output terminals are provided with yellow heads while the d.c. output terminals have one red for positive and the other black for negative for simple visual identification of their polarity. The earth terminal is coloured green. Additional identification is provided by labels indicating the type of output voltage and current rating.

The back of the case is fitted with cable stowage for the 5ft of three-core cable. Rubber mounting feet are provided for bench use.

EE 97 775 *for further details*

SHORT NEWS ITEMS

The Scientific Instrument Manufacturers' Association (S.I.M.A.) announces in its 1965 annual report that deliveries of scientific measuring instruments rose to a total of £180 281 000 in 1965, a 10 per cent increase on the 1964 figures. The proportion of exports increased substantially to £63·2M, an advance of 8·6 per cent over the previous year.

Australia remains the largest single market for British measuring instruments, followed closely by West Germany, France and the U.S.A. The Commonwealth is still the main overseas market and last year exports to constituent countries rose to £17·4M, as opposed to the previous year's figure of £16·3M. EFTA's share remained constant at 11 per cent. The quantity of instruments exported to China has increased substantially and the number exported to Spain almost doubled. Imports of instruments into the U.K. increased by £5·7M as compared with 1964, standing at a total of £43·7M.

The Society of Cardiological Technicians is to hold its eighteenth Exhibition of Cardio-Pulmonary Apparatus at the Piccadilly Hotel, London, W.1, on 2-3 December this year.

Complimentary tickets are available on application to the Exhibition Secretary, c/o Cardiac Research Department, Guy's Hospital, London, S.E.1.

The General Post Office has placed an order valued at nearly £1M with Standard Telephones & Cables Ltd and the General Electric Company Ltd for terminal and line equipment incorporating the pulse code modulation (p.c.m.) system.

This equipment is to be installed on junction telephone cables in the London area, to be followed by similar installations in the North, Midlands and the West Country.

Included in the order are some hundreds of 24-circuit system terminals and well over 1 000 repeaters, the latter installed underground in watertight housings spaced at intervals of about 2 000 yards to coincide with existing G.P.O. manhole positions. A housing accommodates 24 repeaters measuring only 3½ by 1¼ by 9½in, each repeater handling 24 two-way telephone conversations over two pairs of wires. Since previously one pair of wires dealt with one two-way conversation, the capacity of a cable can be increased by 12 times assuming that all pairs can be used for p.c.m. working.

The terminal equipment for the p.c.m. systems makes extensive use of modern microcircuits to give a significant increase in reliability and a reduction in size.

Power for the repeaters is supplied from the terminal stations over the wires carrying the coded speech.

The first European Technical Conference on Production and Inventory Control is to be held at the Grand Hotel, Brighton, on 1 to 13 November this year.

This conference is being sponsored by the London Chapter of the American Production and Inventory Control Society.

An Exhibition of Production Planning, Control Tools, Aids and Systems will be held during the conference.

Further details of registration and accommodation are available from the Organizing Secretary, Conference Services Ltd, 11 Whitehall Court, London, S.W.1.

The Road Research Laboratory of the Ministry of Transport is to set up an experimental road traffic control scheme in the centre of Glasgow.

An area of about 1 square mile in the central business and shopping district has been selected for this experiment. It includes some 80 traffic signals and four bridges over the Clyde. The traffic signals will all be connected directly to a computer, although normal methods of control will still be available.

The experiment is intended to assess the advantages of the various known types of traffic control and to investigate new methods.

For this purpose an order has been placed with the Marconi Co. Ltd for its Myriad computer to provide the process control facilities in the scheme where various traffic sensing devices will be used. The computing system has to be able to accept 'real-time' information from a variety of sources. Very little provision will be made for manual operation of any part of the system which will have sufficient flexibility and spare capacity to enable entirely new methods of traffic control to be tried, including some which may be developed during the course of the experiments.

Results will be assessed largely by the conventional methods of measuring time reductions in set journeys, using 'floating car' techniques, but it is hoped that new automatic methods will be developed, using the potential of the computer in conjunction with the sensing devices.

The Marconi Company will also supply an analogue scanning unit and an analogue-to-digital convertor to feed data from measuring instruments directly into the computer, together with an incremental plotter which will provide a rapid means of extracting data from the processing system in graphical form.

The Tellurometer Company—part of the Plessey Organization—has produced a commercial version of the military electronic distance measuring equipment.

The equipment—known as the MRA 101 tellurometer—meets the need for a commercial (non-military) instrument which would have sufficient accuracy to enable it to be used for long and short ranges for geodetic work, civil engineering and other survey work.

The MRA 101, which is fully compatible with its military counterpart MRA 3, is particularly suitable for urban use as its large reflector (13in in diameter), together with advanced but simple design, enables a high degree of accuracy to be achieved. The beamwidth of only 6° ensures that reflections from passing traffic are held to a minimum, and accurate measurements are possible even in busy city streets.

New features include the all-transistor circuit design using silicon transistors; the single printed wiring board, which carries almost all the electronic components and the virtual elimination of interconnecting plugs and sockets.

G.E.C. Road Signals Ltd (formerly SGE Signals Ltd), a subsidiary of the General Electric Co. Ltd, has received a contract valued at £40 000 from the Ministry of Transport for a closed-circuit television surveillance system for use with an area traffic control experiment.

The system will be used in the West London Area Traffic Control experiment, due to begin early next year, and will provide the control centre in Victoria with a continuous visual survey of traffic flow and behaviour at several key intersections.

Compact remotely-controlled transistorized cameras will be mounted 40ft above the junctions on buildings or on slim poles. Pavement-mounted termination units will house remote control, transmission and line termination equipment.

Vision signals will be transmitted to the control centre over G.P.O. cables, while remote control signals to alter the 'pan', 'tilt', aperture and focus of the cameras will be carried on ordinary telephone circuits by a G.E.C. 'Teleshift' frequency shift signalling system.

In the control centre, pictures from the cameras will be displayed on a bank of 11in transistorized monitors, and each operator working in the centre will be able to select any picture from this bank and display it on a 17in monitor on his own desk. In doing so he will automatically link the remote control panel on his desk to the appropriate camera.

The Ministry of Technology has placed a contract valued at £14 500 with Ferranti Ltd for a six months' study of Ferranti's automatic drawing machine in order to develop its use in conjunction with computers. The work will be carried out in collaboration with the Ministry's National Engineering Laboratory, East Kilbride, Glasgow.

The use of the machine with computers and associated peripheral equipment will be investigated to meet design office requirements such as the production of detail and sectional drawings and perspective views, so covering the needs of as many industries as possible for drawings for all types of machining, including sheet metal work.

Recommendations on the storage and retrieval of information, including the provision of displays for individual draughtsmen, will also be made. Detailed specifications will be prepared for mechanical devices able to produce to specfications to be established. Recommendations will also be made on the minimum size of computer and peripherals to carry out all these jobs.

The work will include an investigation of the feasibility of automatically up-dating drawings and of maintaining a system for this purpose which would need minimum input information and effort. A detailed specification will be prepared for a visual display unit capable of producing the required drawings.

A conference on m.f., l.f. and v.l.f. radio propagation is to be held at the Institution of Electrical Engineers on 8 to 10 November 1967.

The conference, sponsored jointly by the IEE Electronics Division and the Institution of Electronic and Radio Engineers, and will include:

(1) Navigational aids; communication.

(2) The effect of ground, coastline and climate.

(3) Ionosphere—irregularities, absorption, movement and non-linear effects.

(4) Waveguide mode and ray theories.

(5) Excitation factors.

(6) Phase and field strength stability.

(7) Atmospherics.

(8) V.L.F. in the magnetosphere.

(9) Subsurface propagation.

Those wishing to offer material are invited to submit synopses of about 200 words by 31 January 1967 to The Joint Conference Secretariat, IEE, Savoy Place, London, W.C.2.

Further details and registration forms will be available from the Joint Conference Secretariat a few months before the conference.

Marconi radio communication and navigation equipment worth over £¼M has been re-ordered for the 15 new Hawker Siddeley Trident 2E aircraft which are currently on order for British European Airways. This equipment provides the basic guidance information for the automatic landing system pioneered in the existing BEA Trident fleet.

The Marconi v.h.f. navigational equipment, type AD260, provides the autopilot with a definition of the runway centre line and glide slope position throughout the approach and flareout as the aircraft comes in to touch-down, with an accuracy and reliability which exceeds all the international requirements which are being proposed for automatic landing.

The v.h.f. communications equipment, type AD160, and the automatic direction finder, type AD360, will also be fitted in each aircraft, as in the existing Trident fleet. All of this equipment will be duplicated in each aircraft, with an additional i.l.s. localizer to provide the fully triplicated automatic landing system, which ensures the maximum possible reliability.

The Post Office has decided to place a contract with English Electric Leo Marconi Ltd for five Systems 4/70 computers worth £3M, and is taking an option on another four valued at over £2M. The five machines are due for delivery from 1968 to 1970.

This order is the largest placed by the Post Office for computer equipment and is the second occasion in just over 18 months on which a British computer manufacturer has won a large Post Office contract in straight competition with overseas firms.

The Post Office is in process of building up a network of large computer centres, spanning a wide range of operations in its regions and departments. Centres equipped with Leo 326 machines are already in operation in London and in Lytham St. Annes. Others are to open in Portsmouth, Derby and Edinburgh early in 1967. Two of the Systems 4/70 machines will go to Bootle to serve the Giro office, due to open there in the second half of 1968; two more will be installed in Glasgow in the new Savings Bank building, also due to come into use in 1968; and the fifth is for another of the Post Office's Regional Centres which will, in the next few years, handle a growing number of postal and telecommunications projects throughout the country.

The new Giro office will use computers from the outset. Instructions, in the form of Giro documents, will be received in the Giro centre daily. These instructions will be converted into language the computers can accept so that the electronic records of customers' accounts, kept on reels of magnetic tape, can be up-dated, and statements of accounts, printed on computer-controlled high-speed printers, at a rate of 5 000 an hour per machine. Simultaneously, the documents will be sorted ready to be associated with statements and finally both statements and documents will be enveloped and dispatched.

A special feature of the Giro system is that, for large users, statements will be produced in printed form or in machine language (e.g. paper tape or magnetic tape), which can then be used without further treatment as input to the customer's own computer system. The compatibility of magnetic tapes in the Systems 4 range with tapes used in other makes of computer makes this attractive facility possible. Discussions are proceeding with manufacturers on the choice of input devices for Giro; possibilities being studies include character-reading equipment and direct communication links with the computer by means of keyboard devices. The Systems 4/70 will be able to cope with any of these forms of input equipment which may finally be chosen.

Plans for the Giro were announced in a White Paper at the end of last year when it was first forecast that it would have more than one million account holders and would be run from a new building in Bootle.

The Giro will provide a simple, cheap and fast money transfer service. It will make available to the whole adult population, current account banking facilities for the transfer of credit within the system and for the exchange of payments with people outside the system. Firms will be able to integrate the Giro accounting system with their own computer systems by receiving information on magnetic tape direct from the Giro.

The Valve Division of Standard Telephones and Cables Ltd at Paignton, Devon, has recently produced the company's 10 000th travelling wave tube.

This particular tube is of the STC type W3/26 providing an output of 12W in the microwave communication band of 10·7 to 13·2Gc/s.

The company's activities in the microwave communications equipment field began in the early 1930's with experiments in line-of-sight communications systems.

In 1931 STC, in conjunction with the French associate company LCT (Laboratoire Central de Telecommunications), established the first 'microray' communications across the English Channel between Dover and Calais, with experimental equipment working on a wavelength of 17·6cm (1 136·3Mc/s).

This was followed in 1934 by the establishment of the world's first commercial microray link between Lympne and St. Inglevert.

The STC company has pioneered the development of travelling wave tubes for microwave communications, having designed the first operationally successful tube, the VX7030 in 1949, and produced the first tube to be installed in a main line microwave link. This was the W7/2D, a 4 000Mc/s 1W tube which was installed in the 250-mile Manchester-Kirk O'Shotts television link in 1952.

The Digital Systems Department of Ferranti Ltd, at Bracknell, Berkshire, has installed a high speed data link system between Imperial College, South Kensington, and London University's Atlas computer at Gordon Square, London.

The system enables a computer to be used for on-line, real-time working by a user located several miles away, and is at present being used by the college to transmit data and receive the computed results of adaptive control experiments, and also to conduct a normal programming service.

The link effectively provides full duplex working at 5 000 blocks/sec in both directions which gives actual data rates of 12 data bits/block, i.e. 60 000 data bits/sec in the adaptive control mode, and a peak capacity of one paper tape character/block, i.e. 5 000 characters/sec in the paper tape mode.

The transmission in each direction is over a coaxial cable, and in the adaptive control mode the link data rate is matched to the interrupt rate of Atlas by a special buffer store which can alternatively be used to attach a local peripheral when the data link is in program mode.

Kent Precision Electronics Ltd, manufacturers of advanced solid state industrial control instrumentation, announce a change of name. The company will henceforth be known as KPE Ltd, of Vale Road, Tonbridge, Kent.

The new name represents the company's initials, by which it has always been known, and states clearly the field of operations. It further eliminates any confusion that might arise in relation to the George Kent Group, with which KPE Control Ltd has no connexion.

G.E.C. (Electronics) Ltd is to supply a transistorized vidicon closed circuit television system for installation in the 400 yard long tunnel forming the first part of the new Leeds inner ring road system.

The north and south-bound dual carriageways running through the tunnel will each be covered by four cameras and another camera will be mounted on a mast overlooking the tunnel's southern approach. Monitors will be installed in the Leeds police headquarters half a mile away.

A symposium on Numerical Control of Production Processes is to be held at the University of Aston, Birmingham on 30 November this year.

This symposium, sponsored by the United Kingdom Automation Council, is being organized by the Society of Instrument Technology in co-operation with the Institution of Mechanical Engineers, the Institution of Electrical Engineers, the Institution of Production Engineers and the British Computer Society.

The object of this symposium is to examine the present state of techniques of numerical control and their application in manufacturing processes as a whole, the emphasis being on uses other than for the control of machine tools, and to discuss such applications as automatic assembly and testing, textile production and metal forming.

Further details are obtainable from the Secretary, the Society of Instrument Technology, 20 Peel Street, London, W.8.

A Redifon 80W m.f. radio beacon is to be installed as a marker in conjunction with the i.l.s. system at Woolsington airport, near Newcastle-upon-Tyne. The equipment will comprise dual Redifon G.142 transmitters in a single housing together with a changeover unit, which, in the event of failure of transmission, will automatically switch the service from one transmitter to the other.

Other airports in the U.K. now using G.142 radio beacons include Sunderland, Swansea, Bristol (Municipal), Leeds and Bradford, Largs (Dundee), Tees-side (Middleton) and Enniskellin (Co. Down). This type of radio beacon is frequently employed as a locator and can be supplied in a weatherproof kiosk. The dual version of the beacon, in the kiosk, is the standard installation at R.A.F. airfields.

A number of G.142 equipments have also been supplied to the Posts and Telegraphs Department in Eire and to private operators such as Martin Baker Aircraft Ltd and B.A.C. (Filton), and to oil rigs in the North Sea as an aid for helicopters.

Elliott-Automation's Power Generation Division has received orders worth over £600 000 for instrumentation and automatic control systems for the Central Electricity Generating Board's new 1 000MW Rugeley 'B' power station.

The existing power station at Rugeley, opened in October 1963, which is also controlled by Elliott-Automation equipments, has five 120MW generating sets. The fact that the new station will generate 66 per cent more power using only two 500MW generating units illustrates the rate at which technology is advancing in this field and emphasizes the necessity for the use of the most advanced automatic control techniques and the vital importance of the highest standards of accuracy and reliability in the control of these large and expensive units.

Orders for the control room equipment for the two 500MW generating sets have been received from the Northern Projects Group of the C.E.G.B. and for the instrumentation and automatic controls for the re-heat boilers from Foster Wheeler John Brown.

The East African Posts & Telecommunications Administration has placed an order with The Marconi Co. Ltd for a new type of low capacity, radio communications system, which will provide six telephone channels between Mwanza and Bukoba, some 112 miles apart across Lake Victoria.

This new system, which is known as Thin Line Tropospheric Scatter, can provide a limited number of very reliable telephone or telegraph circuits over distances of up to 200 miles without the need for repeater stations. The system is free from the interference and distortion which are normally associated with high frequency radio circuits, while the cost of the equipment is considerably below that of the high capacity tropospheric scatter systems which have formerly been used for this type of work.

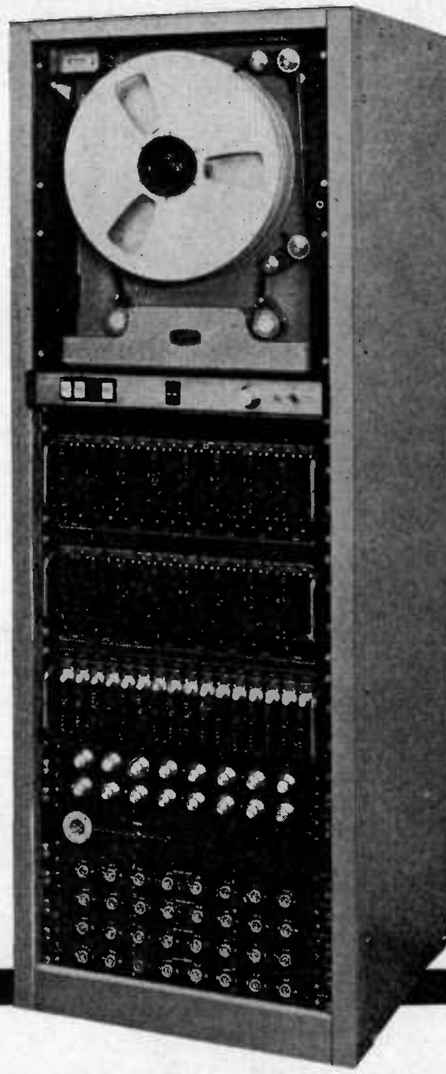
This new telephone route across Lake Victoria will replace an indirect land-line route. It will increase the number of channels available between these two important centres, and will provide a considerable improvement in the speech quality. These new circuits will be integrated into the normal E.A.P. & T. telephone network.

The new equipment has a narrow band solid state transmitter with an output of 7 to 10W in the frequency range 790 to 960Mc/s. The frequency is crystal controlled and sensitive solid state low-noise receivers are employed with 30ft diameter dish aerials at both the transmitting and receiving stations.

Frequency diversity is used to provide additional reliability in the system, without increasing the transmitter power.

'PLUMBICON' is the registered trade mark of N.V. Philips Gloeilampenfabrieken, Eindhoven, Holland, and should be used in connexion with their television camera tubes only. It should not be used in a generic sense and we apologize for having done so on page 81 of the February 1966 issue.

EE 97 119 *for further details*



designed for industry

EPSYLON MR 1400 SIX-SPEED TAPE DECK

The new MR.1400 tape deck has a tape capacity of 1,200 feet (2,200 metres) but due to the use of concentric spools a panel space of no more than 22.75 inches (57.5 cm.) is required.

Two alternative models cover six speed ranges between $\frac{15}{16}$ and 30 or $1\frac{7}{8}$ and 60 inches/second.

The MR.1400 is available with up to sixteen record-reproduce channels or separately.

FEATURES

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NOUVELLES Réalisations

Traduction des pages 610 à 615

Une description basée sur des renseignements fournis par les fabricants de nouveaux composants, accessoires et instruments d'essai

MICROVOLTMÈTRE À COURANT CONTINU DE SUPPRESSION DE ZÉRO

Dymar Electronics Ltd, Rembrandt House, Whippendell Road, Watford, Hertfordshire

(Illustration à la page 610)

La gamme des éléments interchangeable Dymar vient d'être complétée par le microvoltmètre type 724 à courant continu de suppression de zéro. Cet appareil permet de mesurer les changements différentiels de grande tension. Il peut, par exemple, mesurer et enregistrer un changement de 1 μV en 10 MV ou de 100 mV en 1000 V. La suppression de zéro s'obtient au moyen d'un potentiomètre hélicoïdal à dix tours ayant 1000 divisions d'échelle et l'échelle de l'instrument de mesure est accouplée mécaniquement au commutateur d'atténuation d'entrée afin de donner une indication claire de la sensibilité de l'instrument de mesure sur la totalité de l'échelle. Une sortie d'enregistrement est prévue et une des applications typiques consiste à mesurer la regulation et la stabilité d'alimentation.

Une entrée de 10 mV (sur la totalité de l'échelle) à 1000 V (sur la totalité de l'échelle) est couverte en six gammes, cependant que le voltmètre couvre 100 μV à 1000 V en 15 gammes de 1, 3, 10 séquences. L'impédance d'entrée est de 1 MΩ sur les gammes de 10 mV, 100 mV et 1 V et de 100 MΩ sur les gammes de 10 V, 100 V et 1000 V.

EE 97 751 *pour plus amples renseignements*

ASSEMBLAGES DE COMMUTATION PRÉBOBINÉS

Digitizer Techniques Ltd, 26 Sheen Road, Richmond, Surrey

(Illustration à la page 610)

Le "Digitswitch" fabriqué par la société Digitizer Techniques Ltd est maintenant fourni comme assemblage pré-bobiné et dessiné suivant la demande de l'utilisateur. Le commutateur lui-même est du type à roue molletée permettant un choix presque illimité de combinaisons à l'intérieur de dimensions très compactes. Ce nouveau dispositif de pré-bobinage constitue un harnais complet de connexions terminant en une fiche ou en fiches multiples, en connecteur de bord ou en tout autre dispositif convenant au matériel de l'utilisateur.

On peut ainsi installer le "Digitswitch" en l'espace de quelques minutes, quel que soit le nombre de connexions voulues et les modifications de circuit ultérieures ainsi que l'entretien deviennent fort simples.

EE 97 752 *pour plus amples renseignements*

BÊTAMÈTRE HF

Cathodeon Electronic Ltd, Bircham Road, Southend-on-Sea, Essex

(Illustration à la page 610)

La Division de Mesure de Précision de la société Cathodeon Electronic Ltd produit maintenant un nouveau bêtamètre, le PMD 401.

Il s'agit d'un instrument à lecture directe pour la mesure de gain bêta haute fréquence de transistors dans le mode émetteur commun à une fréquence de 20 MHz ou 100 MHz.

L'indication de la tension au collecteur, du courant émetteur et de rayons bêta HF s'effectue par des instruments de mesure séparés. Le contrôle dans des conditions de signaux réduits de transistors de l'une ou l'autre polarité peut être effectué. Une fois établies les conditions de polarisation pour un type particulier de transistors, les transistors subséquents d'une même type peuvent être mesurés simplement en branchant le transistor et en lisant l'indication bêta.

Conçu pour le contrôle en laboratoire d'échantillons de lots de transistors par un personnel semi-entraîné, le PMD 401 assure, à lui seul, les fonctions d'un oscillateur HF, d'un amplificateur de gain étalonné, d'un détecteur et d'une alimentation à transistors.

EE 97 753 *pour plus amples renseignements*

MICROVOLTMÈTRE/ NANOAMPÈREMÈTRE À COURANT CONTINU

Test Equipment Repair, Leigh Road, Leigh, Lancashire

(Illustration à la page 610)

Cet appareil est un microvoltmètre transistorisé à lecture centrale, un nanoampèremètre et un détecteur de zéro avec une entrée entièrement flottante. Il comporte six gammes de tension sur la totalité de l'échelle allant de 100 μV à 1 V à 10 MΩ/V et six gammes de courant sur la totalité de l'échelle allant de 100 nA à 500 μA, avec une résistance d'entrée maxima de 1 kΩ.

L'instrument se compose d'un relais modulateur constitué de corps solides à effet de champ, d'un amplificateur de courant alternatif à haute stabilité et d'un détecteur sensible aux phases. Cette méthode d'amplification de tension continue produit une grande stabilité et des dérives de zéro inférieures à 0,5 V/°C.

Pour améliorer son utilité l'instrument a été conçu pour effectuer des mesures tant de basse impédance que de haute impédance, et en raison du type de relais modulateur utilisé la tension décalée est négligeable; une seule commande de zéro est dont nécessaire.

Il comprend une sortie d'enregistrement et un atténuateur ce qui en fait un préamplificateur à niveau réduit très stable pour les enregistreurs potentiométriques. Les conditions de surcharge vont jusqu'à 6 V dans la gamme de 100 μV.

EE 97 754 *pour plus amples renseignements*

PONT DE COMPOSANTS

The Wayne Kerr Co. Ltd, Sycamore Grove, New Malden, Surrey

(Illustration à la page 611)

La société Wayne Kerr a produit un pont à bas prix effectuant des mesures de 1 % de L, C, et R dans une gamme de mesures extrêmement étendue. Deux cadrans, tous deux à lecture directe, donnent des valeurs simultanées pour les indices de R et C/L. Le pont comporte en outre un circuit interne de doublage de fréquence pour la source entraînée par courant force. On peut ainsi effectuer des mesures à 50 et 100 Hz (ou 60 et 120 Hz) par simple tour de commutateur. La simplicité du fonctionnement a été réalisée grâce à une disposition judicieuse des quelques commandes de fonctionnement et par la fourniture de diagrammes de connexions de séquences de fonctionnement et d'une table de gammes sur la plaque du panneau frontal. L'amplificateur-détecteur transistorisé a une réponse logarithmique et a été conçu de telle manière que l'indicateur de zéro ne puisse en aucune circonstance donner une déviation ambigüe due à des conditions de surcharge. Cette particularité permet le repérage rapide de la gamme de mesure la plus appropriée et une détermination précise du point d'équilibre final.

Le pont permet en outre d'effectuer une évaluation des constantes L, C et R de circuits à très faible impédance. Des connexions de troisième et de quatrième mesure sont prévues lorsque des composants doivent être vérifiés pendant qu'ils sont en circuits ou lorsque des atténuateurs ou des filtres sont soumis à un essais. La gamme totale va de 1 MΩ à 1000 MΩ, de 1 pF à 5 F et de 1 μH à 500 kH.

EE 97 755 *pour plus amples renseignements*

SYSTÈME DE COMPTAGE MODULAIRE

Distributeurs: High Volt Linear Ltd, 67 Dudley Street, Luton, Bedfordshire

(*Illustration à la page 611*)

La société High Volt Linear Ltd fournit maintenant le système de comptage modulaire M 100, de E. G. & G. (U.S.A.). Il s'agit d'un système universel de traitement de données en temps réel, conçu spécifiquement pour les applications où les taux élevés de données, les temps de résolution courts et l'insensibilité au bruit sont d'une importance extrême. Ce système étant à couplage de courant continu est donc insensible au taux; il est capable d'un fonctionnement véritablement asynchrone. De plus, il peut fonctionner à des vitesses continues et apériodiques dépassant 100 MHz.

Le système de comptage modulaire M100 se compose d'une série étendue d'instruments et d'accessoires modulaires comprenant des déclencheurs/discriminateurs, des éléments ET/OU de coïncidence, des portes linéaires, des amplificateurs, des prééchelles, des convertisseurs de temps/hauteur, etc.

Les déclencheurs/discriminateurs et les modules logiques sont entièrement à couplage direct. Cet avantage ainsi que l'absence d'éléments régénératifs dans les modules logiques rendent le système capable de fonctionnement en temps réel à entrée aléatoire d'une efficacité totale, du courant continu à 100 MHz. Le système M100 comprend également un certain nombre d'instruments linéaires à couplage direct stable et d'une grande souplesse d'utilisation pour le traitement des signaux. Ces instruments se caractérisent par une largeur de bande étendue et une excellente performance sur les surtensions. Les applications typiques du système comprennent les recherches de physique nucléaire à haute énergie, les études de structures nucléaires, les recherches de laser, l'astronomie optique, les études de systèmes de radar et les travaux par calculatrice.

EE 97 756 *pour plus amples renseignements*

MILLIVOLTMÈTRE TRANSISTORISÉ

Farnell Instruments Ltd, Sandbeck Way, Wetherby, Yorkshire

(*Illustration à la page 611*)

Le millivoltmètre transistorisé type TM1 est un instrument universel fonctionnant sur batterie pour la mesure du courant alternatif ou du courant continu. Il est alimenté par deux batteries PP11 (ou équivalentes) et effectue des mesures de 1 mV de déviation sur la totalité de l'échelle à 300 V de déviation sur la totalité de l'échelle en douze gammes.

La gamme de fréquence sur courant alternatif va de 10 Hz à 100 kHz et l'impédance d'entrée au-dessous de 30 mV est de 100 kΩ; de 30 mV à 1 V elle est de 1 MΩ et au-dessus de 1 V elle est de 10 MΩ shuntés par 40 pF.

Sur courant continu la résistance d'entrée est de 1 MΩ/V à 10 MΩ au maximum.

La précision sur courant alternatif est de 4 % de déviation sur la totalité de l'échelle et sur courant continu elle est de 3 % de déviation sur la totalité de l'échelle.

EE 97 757 *pour plus amples renseignements*

ISOLATEUR OPTOÉLECTRONIQUE

Texas Instruments Ltd, Manton Lane, Bedford

(*Illustration à la page 611*)

Un nouveau dispositif de couplage optoélectronique qui assure l'isolement électrique économique à haute tension jusqu'à 5 kV a été introduit par la Texas Instruments Ltd. Le nouvel isolateur optique, type TIML101, allie un instrument planaire sensible à la lumière au silicium, le T1 (LS600) à une source lumineuse à l'arsenure de gallium (TIXL101) en un seul coffret d'époxyde opaque.

Le nouveau dispositif a été conçu pour assurer l'isolement électrique lorsqu'on se trouve en face de problèmes de réaction de circuit. En tant qu'instrument de remplacement des relais électromécaniques, il offre des avantages importants en ce qui concerne la vitesse de commutation, la fiabilité, la robustesse mécanique et l'encombrement. Bien que capable d'isoler des tensions très élevées, le TIXL101 est sensible aux petits changements de signaux. Il est donc particulièrement indiqué pour application aux lignes de relais de télécommunications à faible courant et à tension élevée. La vitesse de commutation dépasse largement celle des relais les plus rapides et permet la transmission de beaucoup plus d'informations avec moins de dispositifs. Capable d'une réponse linéaire allant au delà de 10 kHz, le TIXL101 a un temps de commutation inversée typique (t_r) de 1,5 μsec et un temps de commutation avant (t_f) de 15 μsec. La puissance nominale du courant d'entrée est de 50 mA. La sortie minima est de 250 μA.

En intégrant deux composants hermétiquement scellés en un seul élément en époxyde solide on a pu obtenir un composant d'une grande robustesse physique pour les applications industrielles de fatigue où ils sont soumis à de fortes vibrations et à d'autres conditions sévères d'environnement. Les défauts de contact qui se vérifient fréquemment dans le cas des relais ordinaires lorsqu'ils sont soumis à de fortes vibrations sont complètement éliminés dans le cas du nouvel isolateur. Ce dispositif garantit une performance stable dans une gamme de températures étendues allant de −55° C à +125° C.

Le TIXL101 est logé dans un coffret cylindrique à isolement électrique ne mesurant que 0,22 pouce × 0,35 pouce.

EE 97 758 *pour plus amples renseignements*

ETALON DE FRÉQUENCE DE PRÉCISION

Distributeurs: Racal Instruments Ltd, Dukes Ride, Crowthorne, Berkshire

(*Illustration à la page 611*)

Le nouvel étalon de fréquence à cristal, modèle 2.5C, réalisé par la société Sulzer des Etats-Unis, a une stabilité de fréquence typique d'au moins 1 partie dans 10^{11} par jour, avec des fréquences de sortie de 5 MHz, 1 MHz et 100 kHz. Les sorties de 5 MHz et 1 MHz sont munies de filtres de pureté spectrale qui assurent des fréquences de sortie d'une pureté spectrale supérieure à 80 dB, permettant un degré élevé de multiplication de ces fréquences.

La qualité des caractéristiques des étalons de fréquence Sulzer est maintenue. Des circuits constitués de corps solides et entièrement au silicium sont utilisés. Tant les fours intérieurs que les fours extérieurs sont à commande proportionelle. La température de four intérieure est réglée en fonction du coefficient de température nulle du cristal afin de garantir une stabilité maxima. L'étalon a été conçu pour des applications militaires et il peut résister à des températures d'emmagasinage de −28°C à +70°C; il peut, en outre, résister à des chocs de 30*g* avec un changement de fréquence inférieur à 1 dans 10^8. Enfin, il est hermétiquement scellé pour l'utilisation à une température relative de 90 pour cent.

Un système à commande automatique de volume et de température, d'une précision à toute épreuve, maintient la puissance de cristal à quelques centièmes de décibel à environ 1 μW pour une stabilité maxima. La haute stabilité, l'extrême fiabilité, le degré d'utilisation normale dépassant 20 000 heures et la gamme étendue de températures font de cet étalon de fréquence une source de fréquence appropriée pour les émissions de radiodiffusion à fréquence normale, pour la mesure en laboratoire, pour les système à minutage précis, pour la spectroscopie micro-ondes et le minutage d'observatoire.

Les produits Sulzer peuvent être obtenus de la société Racal Instruments Ltd que représente sur une base presque mondiale le Groupe Tracor dont Sulzer fait partie.

EE 97 759 *pour plus amples renseignements*

BLOCS D'ALIMENTATION MODULAIRES

Standard Telephones & Cables Ltd, Components Group, Footscray, Sidcup, Kent

(Illustration à la page 612)

La société Standard Telephones & Cables Ltd a récemment lancé une série de sous-éléments d'alimentation stabilisée à haute performance spécialement conçus pour pouvoir être incorporés au matériel électronique transistorisé. De construction modulaire, ils peuvent être aisément montés pour fournir n'importe quelle tension constante jusqu'à 10 A, 50 V c.c., grâce à leur flexibilité de montage. Ils permettent ainsi d'effectuer une économie considérable de temps, d'effort et d'argent que nécessiterait la mise au point d'appareils spéciaux.

Chacun des nouveaux blocs se compose de quatre modules de base (connecteur y compris) logés dans un châssis standard muni d'écruns captifs pour faciliter l'installation. Les modules peuvent être retirés du châssis et regroupés pour s'adapter pratiquement à n'importe quel espace disponible.

Ces éléments sont prévus pour trois gammes de tension préréglées et peuvent être utlisés soit sur 110 à 125 V soit sur 200 à 250 V c.a. secteur de 45 à 65 Hz. Les sorties de 0 à 16, 0 à 30 et 0 à 50 V c.c. sont préréglées à la fabrique. Elles peuvent, cependant, être facilement ajustées sur place pour les adapter à des conditions différentes.

La stabilité de la sortie est telle que des fluctuations de +10% dans l'entrée de courant secteur ne modifient le niveau de la sortie que de 0,001%, même à une température ambiante de 65°C (139°F).

Tous les blocs sont munis de dispositifs semiconducteurs au silicium et de composants de haute qualité et à caractéristiques nominales étendues pour garantir une haute fiabilité. Ils sont également équipés de disjoncteurs à maxima, le circuit standard étant à réglage manuel.

Notre gravure montre le plus grand de ces blocs.

EE 97 760 *pour plus amples renseignements*

STROBOSCOPE

Lunartron Electronics Ltd, Chester Works, Chester Avenue. Luton, Bedfordshire

(Illustration à la page 612)

Le modèle 1209 vient d'être ajouté à la gamme de stroboscopes fabriqués par la société Lunartron. Il s'agit d'un instrument industriel conçu pour être utilisé dans des conditions où les chocs mécaniques occasionnels sont inévitables. Toutes les commandes sont d'un accès facile, y compris la prise pour une lampe extérieure. Une gamme étendue de fréquences est couverte et la vitesse de fonctionnement est indiquée par un instrument de lecture à tambour se trouvant à l'intérieur d'une fenêtre sur la face supérieure de l'instrument.

Le stroboscope 1209 est entièrement transistorisé et stabilisé contre les fluctuations de tension secteur.

Le taux d'éclats de 60 à 15 000 éclats/minute est couvert en trois gammes, la précision étant supérieure à ±3% de la déviation sur la totalité de l'échelle. La durée de l'éclat est de 5 à 10 μsec.

EE 97 761 *pour plus amples renseignements*

AMPLIFICATEURS DE CHARGE

Distributeurs: Technitron Ltd, Walmgate Road, Perivale, Greenford, Middlesex

(Illustration à la page 612)

La société Technitron Ltd fournit maintenant les Systèmes de Contrôle de Données, modèles GCA-1 et GCA-2. Il s'agit d'amplificateurs de charge entièrement constitués de corps solides, conçus pour mesurer la sortie de transducteurs piézoélectriques et de dispositifs à niveau élevé sans être affectés par la longueur du câble de connexion entre les transducteurs et l'entrée de l'amplificateur. Ces amplificateurs ont une réponse de bande exceptionnellement large, une faible sortie de bruit, une excellente linéarité et ils sont étudiés de manière à ce qu'aucune restriction ne soit placée sur la fuite du transducteur. Ils utilisent les méthodes les plus récentes pour mesurer une gamme dynamique étendue de charges d'entrée allant de 10 pC à 30 000 pC avec des capacités de câble shunt dépassant 0,1 μF.

Le modèle GCA-1 et le modèle GCA-2 constituent tous deux des amplificateurs à sortie double: une sortie de tension pour l'entregistrement sur bande magnétique et une sortie de courant élevé pour l'entraînement de galvanomètres. Le modèle GCA-2 comprend, en outre, un instrument de mesure de la moyenne des pointes (120% de la totalité de l'échelle) et une sortie normalisée pour la commande asservie.

Des longueurs de câble allant jusqu'à 16 km (1 μF) peuvent être utilisées. Le cadran étalonné permet le règlage du facteur de jauge du transducteur utilisé. Des dispositifs incorporés assurent l'étalonnage du gain à ±1%.

EE 97 762 *pour plus amples renseignements*

ÉLÉMENT DE CONTRÔLE PHOTOÉLECTRIQUE

Kappa Electronics Ltd, 159 Hammersmith Road, London, W.6

(Illustration à la page 613)

L'élément de contrôle de transistors, modèle FA15, constitue un appareil universel aux nombreuses applications, à savoir: la détection de marqueurs d'enregistrement, les commutateurs de limite, les dispositifs de sécurité, la détection d'arrêts de chaînes de montage, l'entraînement de mécanismes de porte, la mise aux dimensions, le triage, la détection de niveaux, l'alarme ainsi que d'autres fonctions générales de vérification.

Les contacts classiques de contrôle de sortie arrêt/marche sont déclenchés au moment de l'interruption et du rétablissement du faisceau lumineux. La commutation peut être complétée par le minutage de la paralysie d'entrée (c'est à dire le retard de l'action de contrôle pendant une durée de temps préréglée) ou par des fonctions d'intervalles minutés lorsque la sortie est maintenue pendant une durée de temps préréglée. Des plaquettes spéciales à fiches peuvent être fournies pour différencier les largeurs d'impulsions d'entrée et des temps de réponse allant jusqu'à 1 msec peuvent être prévus. De plus, divers dispositifs de sortie tels que les commutateurs constitués de corps solides peuvent être fournis à la place du relais standard.

Des conditions à l'épreuve de toute panne peuvent être obtenues (relais amorcés) et une commande variable permet de régler le point de commutation pour répondre aux conditions d'éclairage ambiantes. Un deuxième potentiomètre variable assure le contrôle de la fonction de temps. L'un ou l'autre, ou les deux, peuvent être montés à distance de l'élément de contrôle.

L'appareil se compose d'une plaquette de base à circuit imprimé contenant une alimentation stabilisée secteur et offrant la possibilité de brancher la plaquette de circuit choisie, ainsi qu'une base octale pour recevoir le dispositif de sortie. Toutes les connexions sont reliées à une plaquette de bornes à 12 directions qui facilite considérablement l'entretien.

L'appareil peut être fourni pour l'emploi avec la cellule standard au sulphure de cadmium ou avec la duodiode au silicium à réponse rapide. Il est livrable avec une plaque de montage à trou de serrure pour pouvoir être incorporé au système de contrôle du client ou sous forme d'élément autonome dans un coffret métallique étanche mesurant 22,86 cm × 15,24 cm × 10,16 cm.

Au cas où le client aurait besoin d'un élément multivoies incorporant un certain nombre de circuits effectuant des fonctions distinctes, il peut lui être livré une plaquette de base spéciale comportant jusqu'à six plaquettes à fiches et une plaquette de relais à 6 directions dans un coffret compact de 30 cm × 22,86 cm × 15,24 cm. Cette plaquette de base convient, bien entendu, également pour un système de contrôle complet lorsqu'il est utilisé en liaison avec les nombreux autres types de plaquettes de circuits pour le minutage, le choix des voies, les séquences, etc. Cet élément multivoies peut aussi recevoir toute une série de circuits amplificateurs de transducteurs répondant aux besoins du système.

EE 97 763 *pour plus amples renseignements*

COMPTEURS DE LOTS

Darang Electronics Ltd. Restinor Way, Hackbridge Road, Hackbridge, Surrey

(Illustration à la page 613)

La série 667 des compteurs de lots électroniques numériques que vient de lancer la société Darang Electronics Ltd comprend neuf nouveaux modèles. La gamme standard comporte des versions à 2, 3 et 4 décades pouvant compter des lots maxima de 99, 999 et 9 999 respectivement. Ces modèles sont livrables avec ou sans indicateur de lot total et avec ou sans affichage numérique en ligne.

Tous ces compteurs sont munis de dispositifs de précomptage pouvant être choisis par l'utilisateur pour répondre à diverses nécessités. Les signaux de précomptage peuvent être commandés à partir de n'importe laquelle des décades, suivant les besoins. Les dessins sont basés sur un système de plaquettes à fiches d'un faible coût ainsi que sur l'emploi de méthodes extrêmement sûres et perfectionnées d'utilisation de cathodes froides.

Le comptage s'effectue soit par lots uniques soit suivant le mode de répétition automatique. Ce dernier comporte d'ailleurs un dispositif d'autovérification. Le choix des lots se fait par commutateurs rotatifs à 10 directions, avec cadrans numérotés protégés par le panneau frontal.

Le signal d'entrée peut varier de 2 V à 300 V. Le circuit d'entrée est à couplage au courant continu et il répond à des signaux positifs provenant de la masse. La résistance d'entrée est de 50 kΩ. Le comptage peut être déclenché également par contacts extérieurs. Les contacts de sortie sont d'une puissance nominale de 5 A, 230 V c.a.

EE 97 764 *pour plus amples renseignements*

SUPPORT À MICROCIRCUITS

Jermyn Industries, Vestry Estate, Vestry Road, Sevenoaks, Kent

(Illustration à la page 613)

Le support type AE-10L a été étudié pour l'emploi avec des microcircuits TO5 à dix conducteurs. Fabriqué en Nylon A.100 de la I.C.I. (point de fusion: 200°C), le support divise les conducteurs TO5 en deux lignes parallèles contenant chacune 5 conducteurs suivant une matrice d'un dixième de pouce. Ainsi, le dessin des conducteurs est converti pour le rendre conforme aux plaquettes de circuit imprimé à matrice standard et les microcircuits TO5 peuvent être montés plus facilement que les ensembles plats.

EE 97 765 *pour plus amples renseignements*

DISPOSITIFS À RAYONS INFRAROUGES

M.C.P. Electronics Ltd, Station Wharf Works, Alperton, Wembley, Middlesex

Afin de compléter sa gamme MGA 600 de diodes émettrices de rayons lumineux infrarouges à l'arsenure de gallium, la société M.C.P. Electronics Ltd fabrique maintenant une série de dispositifs récepteurs photoélectriques sous les références MSP/3 et MSP/6.

Il s'agit de dispositifs d'une sensibilité extrêmement élevée; des courants de sortie allant jusqu'à 100 mA permettent l'utilisation directe d'un relais électromagnétique classique.

Ces dispositifs sont à structure de silicium et ont une réponse spéctrale de pointe à 9000 Å. Ils peuvent cependant être utilisés également dans des conditions de lumière visible.

Pour le fonctionnement à courant faible, la série 30F2 de diodes microphotoélectriques planaires au silicium présente l'avantage de dimensions compactes et elle est donc indiquée pour l'empilage dans les lecteurs à bande perforée.

EE 97 766 *pour plus amples renseignements*

POTENTIOMÈTRES

Morganite Resistors Ltd, Bede Trading Estate, Jarrow, Co. Durham

(Illustration à la page 613)

La gamme Morganite comprend maintenant deux potentiomètres de réglage à mouvement linéaire pour circuits imprimés, à savoir les modèles 80 et 88. La surface qu'ils occupent sur la plaquette est d'environ 1,5 cm^2. Ils sont prévus en valeurs choisies entre 10 Ω et 2 MΩ.

Le type 80, dont la valeur nominale est de 0,75 W à 70°C, représente un élément sûr et peu coûteux de réglage de circuit destiné à des fins commerciales.

Le type 88 a été conçu pour l'emploi dans les conditions de service les plus astreignantes. Il est robuste, résistant à l'eau et conforme à la norme britannique MIL-R-22097B, caractéristique C, que lui assure sa puissance nominale de 1 W à 85°C. Les deux modèles comportent des broches de bornes revêtues de métal noble et espacées pour la grille standard de 0,1 pouce. La piste de résistance n'est pas affectée par l'humidité et elle résiste aux matières chimiques. Le réglage de circuit peut s'effectuer dans des limites extrêmement précises et la méthode de fabrication exclut tout risque de panne grave.

EE 97 767 *pour plus amples renseignements*

CONTRÔLEUR DE TRANSISTORS

The Wayne Kerr Co. Ltd, Sycamore Grove, New Malden, Surrey

(Illustration à la page 613)

La société Wayne Kerr a mis au point un nouveau contrôleur pour mesurer tous les paramètres *h* des transistors pnp et npn, à 1 000 Hz, dans les modes de fonctionnement à émetteur commun ou à base commune. Le contrôleur TT100 permet de régler et de contrôler le courant émetteur (0 à 30 mA dans six gammes) ainsi que la tension au collecteur (0 à 30 V en trois gammes).

Le fonctionnement est basé sur l'emploi d'un oscillateur ayant une très grande stabilité d'amplitude, des réseaux résistifs précis pour appliquer le signal de 1 kHz à une impédance élevée aux trois connexions du transistor et un millivoltmètre BF de précision. Une précision de mesure de ±3% sur la totalité de l'échelle a été réalisée pour tous les paramètres. La précision de lecture du courant et de la tension est de ±1% sur la totalité de l'échelle.

Une seule barrette de raccordement est utilisée pour toutes les mesures. Cette barrette comporte une quatrième prise pour un conducteur d'écran. Les raccordements voulus s'effectuent automatiquement en actionnant un seul sélecteur de paramètre. Un jack peut être utilisé lorsque des tensions au collecteur dépassant 30 sont exigées. Il permet d'employer des alimentations extérieures.

EE 97 768 *pour plus amples renseignements*

RELAIS DE VERROUILLAGE

Parmeko Ltd, Percy Road, Aylestone Park, Leicester

(Illustration à la page 614)

La société Parmeko Ltd vient d'ajouter à sa gamme G100 de relais scellés miniature des relais de verrouillage à permutation bipolaires et à bobine unique. Ces relais, type G100L, ont été conçus pour les applications exigeant un comportement et une fiabilité de tout premier ordre.

La sensibilité de ces relais n'est que de 30 mW et les tensions de bobine prévues vont de 1,5 à 48 V. La durée de contact est supérieure à 10^6 opérations à moins de 0,3 A et à 10^5 opérations de 0,3 A à 1 A. Les relais type G100L peuvent résister à une accélération linéaire de 100 *g*, à des vibrations d'une accélération de pointe de 25 *g* entre 10 Hz et 3 500 Hz (le relais étant monté sur une surface solide) et ils peuvent être utilisés dans des gammes de température de −55°C à +125°C.

Les dimensions de base sont de 0,920 pouce × 0,820 pouce × 0,380 pouce et le poids est d'environ 20 *g*. Une fiabilité maxima a été obtenue en scellant les bobines et les contacts dans des compartiments séparés et par l'emploi d'alliage d'or pour le matériau de contact.

La société Parmeko Ltd peut également fournir une série de relais de verrouillage à deux bobines.

EE 97 769 *pour plus amples renseignements*

VOLTMÈTRE À LARGE BANDE

Levell Electronics Ltd, Park Road, High Barnet, Hertfordshire

(Illustration à la page 614)

Cet instrument portatif (type TM6A) mesure les signaux jusqu'aux niveaux de microvolts dans la gamme de fréquence de 1 Hz à plus de 1 000 MHz. Ses dimensions et son aspect sont semblables au microvoltmètre c.a. à transistors Levell, type TM3A, sauf qu'il comprend une sonde HF. Huit gammes HF sont prévues, de 1 mV sur la totalité de l'échelle à 3 V sur la totalité de l'échelle, avec une sensibilité de 300 μV dans la bande de 300 kHz à 50 MHz et 3 mV à 400 MHz. Quatorze gammes BF sont prévues de 50 μV sur la totalité de l'échelle, à 500 V sur la totalité de l'échelle, à partir de 10 μV dans la bande de 1 Hz à 3 MHz. Ces gammes sont les mêmes que celles du type TM3A sauf qu'elles ne comportent pas les gammes de 15 μV et de 150 μV.

Les gammes HF sont à circuits "tout transistors" pour convertir le signal HF en une onde carrée de fréquence d'environ 20 Hz et d'une amplitude proportionnelle au carré du signal HF. Aucun interrupteur mécanique n'est utilisé et les circuits sont à compensation de température adéquate. Le degré d'efficacité atteint est élevé en raison d'une consommation électrique sur batterie de 9 V se réduisant à 10 mA seulement.

L'instrument répond à la valeur efficace réelle sur toutes les gammes HF mais, sur les gammes BF, il répond à la moyenne et il est étalonné en valeurs efficaces pour une entrée sinusoïdale.

EE 97 770 *pour plus amples renseignements*

MACHINES À ÉCRIRE NUMÉRIQUES

Hilger & Watts Ltd, 98 St. Pancras Way, Camden Road, London, N.W.1

(*Illustration à la page 614*)

Des machines à écrire numériques à fonctionnement électrique, avec ou sans programmation, viennent d'être réalisées par la société Hilger & Watts Ltd pour l'emploi avec des systèmes automatiques de traitement des données. La machine à écrire de base, type FD574, a un chariot de 43,18 cm et des caractères du type "élite". Elle est étudié pour le fonctionnement à distance par commande des circuits électroniques.

Les caractères et les touches sont adaptés pour le fonctionnement par solénoïdes qui comprend les chiffres de 0 à 9, le tabulateur, le point, l'espacement, le retour du chariot, l'astérisque, les lettres A et R, et le changement de ruban rouge/noir.

Des touches supplémentaires ou différentes peuvent être modifiées pour le fonctionnement par solenoïdes pour répondre à des besoins particuliers. On peut ainsi modifier jusqu'à un total de 26 touches ainsi que la fonction d'espacement.

Une barre de commutation et un répartiteur FD 576 peuvent être fournis pour pouvoir être montés sur la machine de base afin de permettre le choix sur la machine à écrire de variations dans la disposition d'informations imprimées au moyen d'éléments de programmation à fiches interchangeables constitués suivant les besoins du client.

La barre de commutation sert de sérialisateur et assure la synchronisation de l'indication imprimée avec la source de l'information. Elle permet également à la machine à écrire de fonctionner à sa vitesse optimum de 8 à 10 caractères par seconde.

La capacité maxima de la barre de commutation est de 90 colonnes et les machines à écrire peuvent recevoir un maximum de 50 lignes d'entrée.

Notre gravure montre l'aspect arrière de la machine à écrire sans le couvercle de la barre de commutation.

EE 97 771 *pour plus amples renseignements*

CONTRÔLEURS DE VITESSE À THYRISTORS

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

(*Illustration à la page 614*)

La vitesse des moteurs électriques à courant continu peut être contrôlée avec précision, du repos à la vitesse de base, avec la nouvelle gamme de contrôleurs créés par la Division d'Automatisation de la société The M.E.L. Equipment Co. Ltd. La gamme "Ergotrol", ainsi qu'elle s'appelle, permet de faire varier à la main la vitesse des moteurs à courant continu ou par un signal de commande d'un système principal. Des redresseurs à commande par silicium (thyristors) sont utilisés pour contrôler l'alimentation de l'induit du moteur et leur fonctionnement est régulier, efficace et sûr.

Les nouveaux contrôleurs fonctionnent sur courant alternatif de secteur et la gamme comprend sept modèles conçus pour l'emploi avec des moteurs inducteurs shunt à courant continu de 1 à 40 chevaux suivant la norme britannique 2613. Le réglage à la vitesse voulue s'effectue par réaction à partir de l'alimentation d'induit et elle est à 2,5% près de la vitesse de base. Lorsqu'une plus grande précision est nécessaire, un tachygénérateur séparé peut être utilisé qui fournit le signal de réaction, le réglage étant alors à 1% près de la vitesse de base. Un avantage important de la méthode du tachygénérateur est que sa précision n'est pas affectée par l'échauffement du bobinage de champ.

La mise en marche est contrôlée par des circuits constitués de corps solides qui protègent contre les courants excessifs même si l'entraînement commence lorsque le contrôle de vitesse est réglé au maximum. Il y a, en outre, une limite de courant d'induit qui permet, cependant, une marge suffisante d'accélération.

Le contrôleur est muni d'un mécanisme d'entraînement à action lente relié à la commande de vitesse, ainsi que de commandes d'arrêt/marche. Ce sont là des pièces standard. Un nombre de composants facultatifs peuvent être fournis dont un indicateur de vitesse étalonné en pourcentages de vitesse ou suivant les besoins de l'utilisateur. Un indicateur de charge, étalonné en fonction du courant, peut également être ajouté à l'appareil. Le dispositif de marche arrière constitue aussi une pièce facultative. Il permet de mettre en marche arrière à la main et comporte le freinage automatique à l'arrêt complet avant que la marche arrière ne commence. Le freinage dynamique pour l'arrêt rapide peut être prévu en variante. Enfin, toutes les commandes et tous les indicateurs peuvent être fixés sur un panneau à part pour l'emploi à distance.

Les contrôleurs d'une puissance maxima de 10 chevaux fonctionnent sur courant secteur monophasé et sont à montage mural, tandis que les contrôleurs de plus de 10 chevaux fonctionnent sur courant triphasé et sont enfermés dans des consoles fixées au sol. Des contrôleurs sans coffret peuvent être fournis sur demande pour pouvoir les incorporer à d'autres installations.

EE 97 772 *pour plus amples renseignements*

ANALYSEUR D'OXYGÈNE

Servomex Controls Ltd, Crowborongh, Sussex

(*Illustration à la page 615*)

La société Servomex Controls Ltd a réalisé un nouvel analyseur portatif fonctionnant sur batterie, le type OA.150, pour compléter sa gamme d'analyseurs d'oxygène. Le nouvel appareil comporte deux gammes de commutation, soit de 0 à 25 % et de 0 à 100 %, dont la précision sur chaque gamme est de ±1 % sur la totalité de l'échelle.

La simplicité du fonctionnement est une des principales caractéristiques de l'appareil, les commandes essentielles consistant en un sélecteur de gamme et en une clef de fonctionnement. L'indication est présentée sur un instrument de lecture incorporé. Le sélecteur comporte également des positions au moyen desquelles la tension de la batterie et le réglage correct du circuit peuvent être vérifiés, toujours à l'aide du même instrument de lecture. Un affichage graphique sur le panneau frontal rend le processus de fonctionnement et de vérification très simple, des symboles étant utilisés autant que possible à la place de mots, ce qui rend l'appareil également utilisable pour les marchés étrangers.

L'analyseur peut être employé avec des échantillons statiques ou en flux, la lecture n'étant pas affectée par des vitesses de flux de 0 à 150 ml/min. Un filtre à disque de verre aggloméré et des valves pour ajuster l'analyseur et le flux sont incorporés à l'instrument. En outre, un aspirateur à main et un tube de séchage constituent des pièces standard de l'appareil.

L'analyseur utilise la même cellule de mesure que les analyseurs d'oxygène Servomex précédents. Une cloche d'aérage en quartz est suspendue à un filament en platine dans un champ magnétique non-uniforme. Elle subit un couple proportionnel à la sensibilité magnétique du gaz soumis au contrôle, ce dernier étant mesuré en maintenant un couple de rétablissement égal et opposé produit par le flux de courant dans une bobine à un seul tour montée sur la cellule d'aérage. Une source lumineuse, une cellule photoélectrique double et un amplificateur de différence maintiennent l'état d'équilibre nul de manière automatique. L'enregistreur de sortie mesure le courant de rétablissement qui est directement proportionnel au contenu en oxygène. Des cellules photoélectriques au silicium et des transistors sont utilisés dans l'analyseur qui est de construction extrêmement robuste et sûre. Un système de compensation de la température maintient la précision voulue en dépit de variations de ±5° C.

La mesure reste pratiquement insensible à tous les gaz communs autres que l'oxygène et l'instrument peut donc être contrôlé par l'azote et l'oxygène, ou même l'air, puis utilisé pour mesurer des mélanges comprenant, par exemple, du protoxyde d'azote, du gaz carbonique ou de l'hydrogène.

EE 97 773 *pour plus amples renseignements*

AMPLIFICATEUR MULTIVOIES

S.E. Laboratories (Engineering) Ltd, Astronaut House, Feltham, Middlesex

(*Illustration à la page 615*)

La société S.E. Laboratories (Engineering) Ltd vient de créer un amplificateur multivoies, système SE.4000, pouvant être monté sur table ou bâti et recevoir un maximum de huit amplificateurs de haute performance identiques ou différents.

Ce système a été spécifiquement conçu pour l'emploi en liaison avec des thermocouples, des ampoules thermiques de résistance, des extensomètres, des transformateurs différentiels, des transducteurs résistifs ou à réluctance variable à demi-pont ou à pont entier, des transducteurs auto-générateurs, des débitmètres, des signaux de millivolts, etc.

Il se caractérise par le dispositif à étalonnage incorporé, fourni sur demande et pouvant être actionné à la main ou automatiquement. On peut ainsi exécuter la programmation d'un train continu de signaux d'étalonnage à partir d'une calculatrice ou d'un appareil de concentration de données servant de signal de référence à partir de n'importe quelles données d'entrée.

Chacun des modules d'amplification est muni d'un indicateur de surcharge monté sur panneau. Le circuit d'entraînement peut également être relié à un mécanisme d'alarme extérieur déclenché par toute déviation des données d'entrée de leurs limites préréglées.

Tous les canaux d'amplification ont une sortie normalisée de ±1,4 V 10 mA. Une plaquette d'amplificateur d'entraînement à fiches peut être branchée sur chacun des modules amplificateurs, portant ainsi la sortie à ±10 V ±100 mA.

Quatre types de modules amplificateurs, tous interchangeables dans la série 4000 des appareils pour montage sur bâti, sont prévus.

EE 97 774 *pour plus amples renseignements*

BLOC D'ALIMENTATION À FAIBLE TENSION

The British Electric Resistance Co. Ltd, Queensway, Enfield, Middlesex

(*Illustration à la page 615*)

Le bloc d'alimentation à faible tension BERCO est une source pratique de faible tension continue ou alternative pouvant être utilisée avec des appareils d'enseignement. Il a été réalisé par BERCO en collaboration avec la Nuffield Foundation.

Un transformateur variable rotatif "REGAVOLT" à double bobinage et un redresseur en pont à deux alternances au sélénium sont montés dans un coffret en acier aéré avec un panneau frontal portant un interrupteur secteur arrêt/marche, un indicateur au néon, un disjoncteur à maxima, deux bornes de sortie à courant alternatif, deux bornes de sortie à courant continu et une borne de terre.

L'alimentation est transmise à travers un commutateur de secteur bipolaire à une lampe-témoin, puis au bobinage primaire du transformateur variable à double bobinage "Regavolt". Ce dernier est muni de prises de tension d'alimentation permettant de recevoir n'importe quelle tension d'alimentation entre 200 et 250 V en plots de 10 V.

L'enroulement secondaire du "Regavolt" isole entièrement la sortie de l'alimentation et le balai fournit une sortie régulière à variation infinie de 0 à 25 V.

Un disjoncteur en série avec le balai de sortie du transformateur "Regavolt" protège l'équipement contre les effets de court-circuits ou de surcharges continues. La sortie du disjoncteur est injectée à un redresseur en pont à deux alternances du type à sélénium à très grande capacité de surcharge.

La sortie de tension continue n'a pas été filtrée le but étant de fournir une alimentation en tension continue au plus bas prix possible à l'usage des expériences scolaires. Le degré de filtrage varie suivant les applications.

Chacune des sorties est reliée à une paire de bornes à vis isolées. Les bornes de sortie de courant alternatif sont munies de têtes jaunes tandis que les bornes de sortie de courant continu ont une tête rouge pour le courant positif et une tête noire pour le courant négatif, pour l'identification visuelle simple de leur polarité. La borne de terre est de couleur verte. Une identification supplémentaire est assurée par des étiquettes indiquant le type de tension de sortie et la puissance du courant.

L'arrière du coffret est muni d'une niche d'arrimage pour le câble à trois âmes de 152,4 cm. Des pieds montés sur caoutchouc sont prévus pour l'usage sur banc.

EE 97 775 *pour plus amples renseignements*

Résumés des Principaux Articles

Le dessin d'un modulateur de largeur d'impulsion à transistors pour les applications de contrôle par R. D. Bell et K. E. Tait

Résumé de l'article aux pages 562 à 567

Le dessin d'un modulateur de largeur d'impulsion à transistors pouvant être utilisé pour l'étude par calcul analogique des systèmes de contrôle de la modulation de largeur d'impulsion est analysé par les auteurs. Ils formulent également une prédiction théorique des limites et de la précision de l'appareil et ces limites sont comparées de manière critique avec les résultats pratiques.

L'étude d'amplificateurs de signaux faibles à transistors à effet de champ par W. Gosling

Résumé de l'article aux pages 568 à 571

Cet article traite de l'étude d'amplificateurs de signaux faibles utilisant des transistors à effet de champ polarisés. Il décrit ensuite le choix des valeurs de circuit permettant d'obtenir la stabilisation du point de régime. Cette méthode de réalisation donne des résultats qui s'accordent raisonnablement avec les valeurs mesurées. Les amplificateurs dont les gains de tension caractéristiques sont d'environ 50x ont l'avantage particulier d'une caractéristique d'impédance élevée propre aux transistors à effet de champ.

Analyse du comportement des relais modulateurs à transistors par D. J. Finlay.

Résumé de l'article aux pages 572 à 578

Il est parfois nécessaire d'utiliser des relais modulateurs pour pouvoir amplifier un très faible courant continu ou une très faible tension continue. De nombreux circuits pour relais modulateurs à transistors ont été réalisés et leur mode de fonctionnement est fort connu. Ce qui est moins connu, cependant, c'est l'effet des divers défauts du relais modulateur sur son comportement général et une certaine perte de temps peut donc être occasionnée par des réalisations inadéquates. Cet article constitue une analyse détaillée et originale, complétée par le dessin d'un circuit équivalent, d'un relais modulateur à transistors d'entrée qui facilite considérablement l'étude des effets de variations de résistance de source et de charge, de température, de fréquence d'interruption ainsi que des paramètres des transistors. Des chiffres caractéristiques sont indiqués par rapport aux erreurs de zéro et à la dérive des transistors à jonction au germanium et au silicium. Il est démontré que les erreurs dues aux phénomènes transitoires peuvent être sérieuses et qu'il serait souhaitable de pouvoir disposer d'informations plus précises sur les phénomènes transistoires et leur dérive lorsqu'il s'agit de mesurer des courants de l'ordre de 1nA ou des tensions de l'ordre de 10μV.

L'article décrit en outre un circuit simple pour supprimer les phénomènes transitoires déformés à la sortie et, par conséquent, de linéariser les caractéristiques de transfert.

La mesure du coefficient de surtension dans les cavités de guides d'ondes à faible perte par J. K. Chamberlain

Résumé de l'article aux pages 579 à 581

Le coefficient de surtension d'un mode déterminé dans une cavité microondes dépend des pertes résistives et autres qui s'y produisent; si la cavité est de forme cylindrique droite, le coefficient de surtension est en rapport avec la constante d'atténuation du guide d'ondes dont les parois cylindriques peuvent être considérées comme en faisant partie. Cet article décrit les grandes lignes d'une méthode de mesure utilisant un matériel simple et approprié aux valeurs élevées de coefficient de surtension (10^5 à 10^6 et au dessus) caractérisant les cavités de guide d'ondes à faible perte. La méthode est illustrée par son application à un guide d'ondes de cuivre de 5cm à 35GHz dont on déduit une constante d'atténuation de 3,4dB/1,609km.

Un régénérateur de forme d'onde pour systèmes d'échantillonnage d'amplitude par T. I. Mitchell et V. J. Phillips

Résumé de l'article aux pages 582 à 587

On peut, en théorie, reconstruire de manière fort simple un signal échantillonné en amplitude en "joignant les sommets" des échantillons; en pratique, cependant, ce processus est loin d'être aussi simple. Cet article décrit l'appareil construit par les auteurs dans ce but.

Analyse spectrale d'une train d'impulsions à pente variable et à durée d'affaiblissement modulée par O. E. Kruse et R. W. Montgomery

Résumé de l'article aux pages 593 à 595

Les auteurs traitent d'une analyse mathématique du spectre d'un train d'impulsions triangulaires dont l'affaiblissement est modulé. Le spectre mathématiquement prédit est alors comparé au spectre déterminé expérimentalement.

Fréquence de balayage pour contrôles généraux de réception par J. F. Golding

Résumé de l'article aux pages 596 à 601

L'utilisation du générateur de balayage—ou wobbulateur—en liaison avec un oscilloscope pour l'alignement de la réception est tellement connue qu'il n'est guère nécessaire de la décrire. Il existe, cependant, d'autres utilisations pour un signal d'essai à balayage de fréquence dans l'évaluation de la performance d'ensemble d'un récepteur. Par exemple, la sensibilité et le rapport signal/bruit peuvent, parfois, être mesurées beaucoup plus facilement à l'aide d'un générateur de signaux à balayage de fréquence que par des moyens classiques.

Quelques applications des relais à deux bobines dans les circuits électroniques par H. Biggar

Résumé de l'article aux pages 602 à 606

Le relais à deux bobines est un composant électromagnétique d'une grande souplesse d'emploi. Chacune des bobines est excitée séparément et contribue au champ magnétique composé. Les applications de ce relais se répartissent en deux groupes distincts suivant que les champs individuels sont additifs ou soustractifs. Les relais à deux bobines remplissent de nombreuses fonctions de grande utilité qui ne sont pas effectuées par les relais à bobine unique.

La gamme des applications peut être étendue lorsque le relais devient partie intégrante d'un circuit électronique utilisant des cellules photoélectriques ou des tubes à trois électrodes. L'article conclut par la description d'un circuit peu habituel pour la commande et la protection automatiques d'un certain nombre de tubes électroniques de grande puissance fonctionnant à partir d'une source commune de très haute tension.

NEUE AUSRÜSTUNGEN

Übersetzung der Seiten 610 bis 615

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Gleichstrom-Mikrovoltmesser mit Nullpunktunterdrückung

Dymar Electronics Ltd, Rembrandt House, Whippendell Road, Watford, Hertfordshire

(Abbildung Seite 610)

Das Programm von Dymar-Einschüben wurde durch das Gleichstrom-Mikrovoltmeter mit Nullpunktunterdrückung Typ 724 erweitert, das Zusatzänderungen hoher Spannungen misst. Typische Änderungen, die sich messen und registrieren lassen, sind 1 μV in 10 mV oder 100 mV in 1000 V. Die Nullpunktunterdrückung wird mittels eines mit einer 1000teiligen Skala versehenen 10gängigen Wendelpotentiometers eingestellt. Die Bereichskala des Messgerätes ist automatisch mit dem Eingangsabschwächerschalter gekuppelt und gibt eine eindeutige Anzeige der Empfindlichkeit bei Vollausschlag. Für Anschluss eines Registriergerätes ist ein Ausgang vorhanden. Eine typische Anwendung ist das Messen der Regelung und Konstanz von Stromversorgungen.

Sechs Eingangsbereiche haben Skalenendwerte von 10 mV bis zu 1000 V, und das Voltmeter überstreicht 100 μV... 1000 V in 15 Bereichen und 1-3-10 Folge. Die Eingangsimpedanz ist 1 MΩ für die 10-mV-, 100-mV- und 1-V-Bereiche und 100 MΩ für die 10-V-, 100-V- und 1000-V-Bereiche.

EE 97 751 *für weitere Einzelheiten*

Vorverdrahtete Schalterbaugruppe

Digitizer Techniques Ltd, 26 Sheen Road, Richmond, Surrey

(Abbildung Seite 610)

Der von Digitizer Techniques Ltd hergestellte "Digiswitch" ist nunmehr als vorverdrahtete Baugruppe nach Kundenvorschrift lieferbar. Der Schalter selbst wird durch Daumenrad betätigt und lässt sich stapeln, so dass innerhalb sehr kompakter Abmessungen fast unbegrenzte Anordnungsmöglichkeiten bestehen. Dieser neue Vorverdrahtungsdienst umfasst Anbringung eines kompletten Leitungssatzes, der den Anforderungen des Kunden entsprechend mit einem Mehrfachstecker, Steckern, Federleisten oder dergleichen abgeschlossen wird.

Auf diese Weise kann ein Digiswitch in Minuten und ohne Berücksichtigung der Anzahl der Anschlussverbindungen angebaut werden. Nachfolgende Schaltungsabwandlungen und Service werden vereinfacht.

EE 97 752 *für weitere Einzelheiten*

HF-Betatester

Cathodeon Electronic Ltd, Bircham Road, Southend-on-Sea, Essex

(Abbildung Seite 610)

Der Bereich Präzisionsmessungen der Cathodeon Electronic Ltd hat die Fertigung eines neuen β-Testers PMD 401 aufgenommen.

Das Gerät ist ein direktanzeigendes Instrument zum Messen der hochfrequenten β-Verstärkung von Transistoren in der Emitterschaltung bei einer Frequenz von 20 MHz oder 100 MHz.

Anzeige von Kollektorspannung, Emitterstrom und HF-β erfolgt auf getrennten Messgeräten. Das Testen von Transistoren jeder Polarität ist unter Kleinsignalbedingungen durchführbar. Wenn die Vorspannungsbedingungen für einen bestimmten Transistortyp eingestellt sind, können nachfolgende Transistoren des gleichen Typs einfach durch Einstecken des Transistors und Ablesen von β gemessen werden.

Der PMD 401 wurde für das Testen von Partiemustern von Transistoren im Labor durch angelernte Kräfte entwickelt und kombiniert in einem Gerät die Funktionen eines HF-Oszillators, eines Verstärkers mit geeichter Verstärkung, eines Demodulators und einer Transistorstromversorgung.

EE 97 753 *für weitere Einzelheiten*

Gleichstrom-Mikrovoltmeter-Nanoammeter

Test Equipment Repair, Leigh Road, Leigh, Lancashire

(Abbildung Seite 610)

Das "Chopper-Galvanometer" ist ein transistoriertes Mikrovoltmeter mit beiderseitigem Ausschlag, Nanoammeter und Abgleichindikator, völlig erdfreiem Eingang und mit sechs Spannungsbereichen von 10° μV...1 V Vollausschlag bei 10 MΩ/V und sechs Strombereichen von 100 nA...500 μA Vollausschlag bei 1 kΩ Höchsteingangswiderstand.

Das Gerät besteht aus einem Feldeffekt-Festkörper-Chopper, einem hochkonstanten Wechselstromverstärker und einem phasenempfindlichen Anzeigegerät. Diese Methode der Gleichstromverstärkung gibt hohe Stabilität und unter 0,5 μV/° C Nulldrift.

Zur Verbesserung der Nützlichkeit wurde das Gerät so entworfen, dass man sowohl an hoher wie niedriger Impedanz messen kann. Dank des angewendeten Choppertypes ist die erforderliche Kompensationsspannung klein und daher nur ein Nullregler erforderlich.

Anschluss für ein Registriergerät und einen Abschwächer ist vorhanden, so dass man das Gerät auch als hochkonstanten Kleinsignal-Vorverstärker für Kompensatoren usw. einsetzen kann.

Überlastung bis zu 6 V im 100-μV-Bereich ist zulässig.

EE 97 754 *für weitere Einzelheiten*

Bauelement-Messbrücke

The Wayne Kerr Co. Ltd, Sycamore Grove, New Malden, Surrey

(Abbildung Seite 611)

Wayne Kerr hat eine billige Messbrücke der Klasse 1 mit äusserst grossem Messumfang für L, C und R herausgebracht. Zwei direktanzeigende Skalen geben gleichzeitig die Werte für R und C/L. Ein neues Merkmal ist die interne

Frequenzdopplerschaltung für die vom Netz hergeleitete Speisespannung, so dass Messungen durch Umlegen eines Schalters bei 50 und 100 Hz (oder 60 und 120 Hz) vorgenommen werden können.

Sorgfältige Anordnung der wenigen Bedienelemente, Mitlieferung von Anschlussdiagrammen, Betätigungsfolge und eine Bereichtabelle auf der Frontplatte erleichtern die Bedienung. Der transistorierte Verstärker für den Abgleichindikator hat eine logarithmische Kurve und ist so ausgelegt, dass das zugehörige Anzeigegerät selbst unter Überlastungsbedingungen keinen mehrdeutigen Ausschlag geben kann. Dadurch wird die Wahl des geeignetsten Messbereiches und genaue Bestimmung des endgültigen Abgleichpunktes erleichtert.

Besondere Vorkehrungen gestatten die Bestimmung der L-, C- und R-Konstanten sehr impedanzarmer Schaltungen. Dritte und vierte Messverbindungen stehen für den Fall zur Verfügung, dass in eine Schaltung eingebaute Bauelemente oder Abschwächer und Filter zu prüfen sind. Der Gesamtmessumfang ist 1...1000 MΩ, 1 pF...5 F und 1 μH...500 kH.

EE 97 755 *für weitere Einzelheiten*

Modulares Zählsystem

Vertrieb: High Volt Linear Ltd, 67 Dudley Street, Luton, Bedfordshire

(Abbildung Seite 611)

Das modulare Zählsystem M100 der E.G. & G. (USA) kann jetzt von High Volt Linear Ltd bezogen werden.

Es ist ein Mehrzweck-Datenaufbereitungssystem, das in Real-time arbeitet und besonders für Anwendungszwecke entwickelt wurde, in denen hohe Datenraten, kurze Auflösungszeiten und Störfestigkeit von grösster Bedeutung sind. Da das System galvanisch gekoppelt und damit gegen Raten unempfindlich ist, kann es wirklich asynchron und darüber hinaus bei kontinuierlichen und aperiodischen Raten über 100 MHz arbeiten.

Das modulare Zählsystem M100 besteht aus einer breiten Auswahl modularer Messgeräte und Zubehör einschliesslich Diskriminator/Trigger, Koinzidenz UND/ODER-Einheiten, linearer Tore, Verstärker, Dehner, Voruntersetzer, Zeit-Höhe-Umsetzer usw.

Da die Diskriminator-Trigger-Moduln und Logik-Moduln ganz direkt gekoppelt sind und die Logik-Moduln keine regenerierenden Elemente haben, kann das System mit willkürlicher Eingabe und in Real-time von 0...100 MHz mit 100 % Wirkungsgrad (Impulskennziffer) arbeiten. In das System M100 einbegriffen sind eine Anzahl vielseitiger, konstanter, direktgekoppelter linearer Geräte für Signalverarbeitung mit grosser Bandbreite und ausgezeichnetem Verhalten unter Überlastung.

Typische Anwendungsmöglichkeiten für das System bestehen u.a. in der energiereichen Kernphysikforschung, Untersuchung der Kernstrukter, Laserforschung, optischen Astronomie, Arbeiten an Radarsystemen und Schnittstellen der Datenverarbeitung.

EE 97 756 *für weitere Einzelheiten*

Transistoriertes Millivoltmeter

Farnell Instruments Ltd, Sandbeck Way, Wetherby, Yorkshire

(Abbildung Seite 611)

Das transistorierte Millivoltmeter TM1 ist ein batteriebetriebenes Mehrzweckinstrument für Wechsel- oder Gleichspannungsmessungen. Es wird von zwei Batterien PP11 (oder Äquivalent) gespeist und hat einen Messumfang von 1 mV Skalenendwert bis zu 300 V Skalenendwert in 12 Bereichen.

Der Frequenzbereich bei Wechselstrom ist 10 Hz...100 kHz; unter 30 mV ist die Eingansimpedanz 100 kΩ, von 30 mV...1 V 1 MΩ und über 1 V 10 MΩ parallel mit 40 pF.

Bei Gleichstrom ist der Eingangswiderstand 1 MΩ/V bis zu höchstens 10 MΩ.

Die Messunsicherheit ist 4 Prozent des Skalenendwertes für Wechselspannungen und 3 Prozent des Skalenendwertes für Gleichspannungen.

EE 97 757 *für weitere Einzelheiten*

Optoelektronische Trennstufe

Texas Instruments Ltd, Manton Lane, Bedford

(Abbildung Seite 611)

Eine wirtschaftliche elektrische Hochspannungsisolation bis zu 5 kV wurde von Texas Instruments mit der Einführung einer neuen optronischen Kopplung ermöglicht. Die neue optische Trennstufe TIML101 besteht aus einem Silizium-Planar-Lichtsensor T1 (LS600) mit einer Gallium-Arsenid-Lichtquelle (TIXL101) in einem undurchsichtigen Epoxyd-Baustein.

Der neue Baustein wurde als elektrische Trennstufe für Schaltungen, in denen Rückkopplungsprobleme bestehen, entwickelt. Er bietet gegenüber elektromagnetischen Relais wesentliche Vorteile hinsichtlich Schaltgeschwindigkeit, Zuverlässigkeit sowie mechanischer Haltbarkeit und hat kleine Abmessungen. Obwohl die TIXL101 sehr hohe Spannungen verarbeiten kann, spricht sie auf kleine Signaländerungen an und ist daher hauptsächlich für Anwendung in Fernmelderelaisleitungen, die mit hohen Spannungen und niedrigen Strömen arbeiten, geeignet.

Die wesentlich über den schnellsten Relais liegenden Schaltgeschwindigkeiten gestatten Übertragung von mehr Information mit weniger Einrichtungen. Bei linearem Frequenzgang über 10 kHz hinaus hat die TIXL101 eine Umkehrschaltzeit (t_r) von 1,5 μs und eine Durchlassschaltzeit (t_f) von 15 μs. Sie ist für Eingangsströme von 50 mA ausgelegt und hat einen Mindestausgangsstrom von 250 μA.

Einbau von zwei erprobten, hermetisch abgeschlossenen Bauelementen in einen Epoxyd-Block ergibt einen mechanisch robusten Baustein, der sich für hochbeanspruchte industrielle Anwendung eignet, in der er schweren Vibrationen, Stoss und anderen Umgebungsextremen ausgesetzt wird. Das bei starker Vibration in Relais oft auftretende Kontaktprellen ist völlig beseitigt. Der Baustein hat über den breiten Temperaturbereich von −55° C...+125° C konstante Eigenschaften.

Die elektrisch isolierte Einkapselung der TIXL101 hat Zylinderform und Abmessungen von nur 5,6 mm × 8,9 mm.

EE 97 758 *für weitere Einzelheiten*

Präzisions-Frequenznormal

Vertrieb: Racal Instruments Ltd, Dukes Ride, Crowthorne, Berkshire

(Abbildung Seite 611)

Das neue, von Sulzer in USA hergestellte Quarzfrequenznormal Modell 2,5 C hat eine typische Frequenzkonstanz von mindestens 1×10^{-11} pro Tag mit Ausgangsfrequenzen von 5 MHz, 1 MHz und 100 kHz. Die Ausgänge für 5 MHz und 1 MHz sind mit spektralen Reinheitsfiltern ausgerüstet, die Ausgangsfrequenzen mit spektraler Reinheit von besser als 80 dB ergeben, wodurch hohe Frequenzmultiplikation dieser Frequenzen möglich wird.

Die erprobten Merkmale der Sulzer-Frequenznormale wurden beibehalten. Silizium-Festkörperschaltungen finden Anwendung, und sowohl der innere wie auch der äussere Thermostat wurde proportional geregelt. Zur Erzielung höchster Konstanz wird die Temperatur des inneren Thermostaten auf die des Null - Temperaturkoeffizienten des Quarzes eingeregelt. Das für militärische Anwendungszwecke konstruierte Gerät kann Umgebungstemperaturen von −28° C...+70° C vertragen und Stösse von 30 g mit Frequenzänderung von weniger als 1×10^{-8} aushalten; es ist hermetisch dicht für Betrieb in bis zu 90% relativer Feuchtigkeit.

Für optimale Konstanz wird die Kristalleistung mittels genauer, temperaturgeregelter Verstärkungsautomatik innerhalb von ein paar Hundertstel Dezibel bei etwa 1 μW gehalten. Durch die hohe Konstanz, extreme Zuverlässigkeit mit einer Durchschnittszeit von über 20 000 Stunden zwischen Ausfällen und den breiten Temperaturbereich eignet sich dieses Frequenznormal als Frequenzquelle für Normalfrequenzsendungen, Labormessungen, genaue Zeitsysteme, Mikrowellenspektroskopie und Sternwarten-Zeitangaben.

Sulzer-Erzeugnisse werden von Racal Instruments Ltd vertrieben, die die fast weltweite Vertretung der Tracor-Gruppe innehat, zu der Sulzer gehört.

EE 97 759 *für weitere Einzelheiten*

Modulare Stromversorgungen

Standard Telephones & Cables Ltd, Components Group, Footscray, Sidcup, Kent

(*Abbildung Seite 612*)

Standard Telephones and Cables Ltd hat vor kurzem eine Reihe von Hochleistungs - Konstantstromversorgungen in Bausteinform für Einbau in transistorierte elektronische Geräte herausgebracht. Durch die modulare Konstruktion können die Bausteine jeder beliebigen Forderung nach Konstantspannung bis zu 10 A, 50 V– in bezug auf Leistung sowie Anpassungsfähigkeit genügen, was beim Entwurf von Sonderzweckgeräten bedeutende Aufwandeinsparungen an Zeit, Personal und Kosten ermöglicht.

Jeder der neuen Bausteine besteht aus vier Grundmoduln (einschliesslich Steckverbindung), die in einem Standardrahmen untergebracht sind, der zwecks leichterer Montage mit unverlierbaren Muttern ausgerüstet ist. Moduln lassen sich aus dem Rahmen entfernen und zur Anpassung an fast jeden verfügbaren Raum umgruppieren.

Sie sind für Anschluss an Wechselstromnetze von 110 ... 125 V oder 200 ... 250 V, 45 ... 65 Hz ausgelegt und für drei vorbestimmte Spannungsbereiche von 0 ... 16, 0 ... 30 und 0 ...50 V– lieferbar; die Ausgangsspannungen werden im Werk eingestellt. Die Ausgänge lassen sich leicht an Ort und Stelle neu einstellen, um geänderten Bedingungen nachzukommen.

Bei Eingangsspannungen von 10 Prozent wird selbst bei 65° C Umgebungstemperatur nur eine Änderung des Ausgangspegels um 0,001 Prozent auftreten.

Alle Einheiten sind mit Silizium-Halbleitern bestückt und die Qualitäts-Bauelemente für hohe Zuverlässigkeit konservativ bemessen. Voller Überlastungsschutz ist eingebaut; bei den Standardschaltungen erfolgt Rückstellung manuell.

Die Abbildung zeigt den grössten Baustein des Programmes.

EE 97 760 *für weitere Einzelheiten*

Stroboskop

Lunartron Electronics Ltd, Chester Works, Chester Avenue, Luton, Bedfordshire

(*Abbildung Seite 612*)

Das Fertigungsprogramm für Lunartron-Stroboskope wurde durch das für industrielle Anwendung—bei der gelegentliche mechanische Stösse unvermeidlich sind — entwickelte Modell 1209 erweitert. Alle Bedienknöpfe und Einrichtungen, einschliesslich einer Buchse für Betrieb einer externen Lampe, sind leicht zugänglich. Das Gerät hat einen grossen Frequenzumfang, und die Betriebsfrequenz wird auf einem Messgerät mit Trommelanzeige unter einem Fenster in der Oberseite abgelesen.

Das Gerät ist volltransistoriert und gegen Schwankungen der Netzspannung stabilisiert.

Die Lichtblitzfolge von 60 ... 15 000 Blitzen je Minute wird in drei Teilbereichen mit einer Unsicherheit von ±3 Prozent des Skalenendwertes überstrichen. Die Blitzdauer ist 5 ... 10 μs.

EE 97 761 *für weitere Einzelheiten*

Ladungsmessverstärker

Vertrieb: Technitron Ltd, Walmgate Road, Perivale, Greenford, Middlesex

(*Abbildung Seite 612*)

Technitron Ltd vertreibt die von Data Control Systems hergestellten Festkörper-Ladungsmessverstärker GC-1 und GC-2, die für das Messen des Ausgangs piezoelektrischer Messwandler und hochpegliger Elemente ohne Beeinflussung durch die Verbindungskabellänge zwischen den Wandlern und dem Verstärkereingang bestimmt sind. Diese Verstärker haben einen ungewöhnlich breiten Frequenzgang, geräuscharmen Ausgang sowie ausgezeichnete Linearität und sind so konstruiert, dass sie durch die Streuung des Wandlers nicht beeinflusst werden. In den Verstärkern finden die neusten Methoden zum Messen eines breiten dynamischen Eingangsladungsbereiches von 10 ... 30 000 pC mit Nebenschluss-Kabelkapazitäten von über 0,1 μF Anwendung.

Sowohl der GC-1 als auch der GC-2 sind Doppelausgangverstärker mit einer Ausgangsspannung für Magnetbandaufzeichnung und einem Hochstromausgang für Galvanometeranzeige. Das Modell ist ausserdem mit einem Messgerät für mittlere Spitzenanzeige (120 Prozent des Skalenendwertes) und einem normierten Ausgang für Servoregelung ausgerüstet.

Bis zu 16 km lange Kabel (1 μF) können Verwendung finden. Die geeichte Skala gestattet Einstellen des Eichfaktors des verwendeten Wandlers. Eingebaute Einrichtungen ermöglichen Verstärkungseichung innerhalb ±1 Prozent.

EE 67 762 *für weitere Einzelheiten*

Fotoelektrisches Schaltgerät

Kappa Electronics Ltd, 159 Hammersmith Road, London, W.6

(*Abbildung Seite 613*)

Das Transistor-Schaltgerät FA15 für fotoelektrische Einrichtungen wurde als Mehrzweckgerät für zahlreiche Anwendungsmöglichkeiten, unter anderem Aufsuchen von Registriermarken, Grenzschalter, elektronisches Nachstellen von Maschinen, Schutzzäune, Warnung beim Stoppen von Förderbändern, Betätigung von Toreinrichtungen, Klassieren, Füllstandanzeiger, Warnungen und andere Überwachungseinrichtungen, entwickelt.

Im Ausgang werden bei Unterbrechung und Wiederherstellung des Lichtstrahls die konventionellen Ein-Aussteuerkontakte betätigt. Diese Schaltaktion kann durch zeitweise Eingangsverriegelung (d.h. Verzögerung der Schaltaktion um eine vorgewählte Zeit) oder—wo der Ausgang für eine vorgewählte Zeit aufrecht erhalten wird—durch zeitbegrenzte Intervallfunktionen ergänzt werden. Es gibt Spezialsteckkarten, die Differenzieren zwischen Eingangsimpulsbreiten gestatten; Ansprechzeiten bis zu 1 ms herunter sind erreichbar. Anstelle der Standard-Relais lassen sich auch verschiedene andere Ausgangseinrichtungen oder Festkörperschalten vorsehen.

Selbstschutz kann für "hell"- oder "dunkel"-erregte Relais erreicht werden, und die Anpassung des Schaltpunktes an die Lichtbedingungen der Umgebung erfolgt mittels eines vorgesehenen Reglers. Ein weiteres Drehpotentiometer regelt die Zeitfunktion. Jeder der Regler oder beide können vom Schaltgerät entfernt installiert werden.

Das Gerät besteht aus einer Druckschaltungsgrundplatte mit der Konstantstromversorgung, Steckeinrichtungen für die gewählte Schaltungskarte und ausserdem einer Oktalfassung zur Aufnahme der Ausgangsvorrichtung. Alle Verbindungen werden über eine schnell lösbare und anschliessbare 12polige Klemmleiste hergestellt, was den Ausbau für Wartung erleichtert.

Das Schaltgerät ist für Betrieb mit der Standard-Kadmium-Sulfidbatterie oder der schnellansprechenden Silizium-Fotodoppeldiode lieferbar und kommt entweder mit Montageplatte zum Einbau in ein umfassendes Steuersystem des Anwenders oder als in sich geschlossenes Gerät in einem wetterfesten Stahlgehäuse von 229 × 152 × 102 mm.

Wenn Bedingungen ein mehrkanaliges Gerät erfordern, in dem eine Anzahl von Schaltungen getrennte Funktionen ausüben, kann eine Sonderzweckgrundplatte mit bis zu sechs Einsteckkarten und einer sechswegigen Relaiskarte in einem Gehäuse von 309 × 229 × 152 mm geliefert werden. Diese Ausführung ist natürlich auch für komplette Steuersysteme geeignet, da sie mit vielen anderen lieferbaren Steckkarten wie z.B. Zeitgeber, Kanalwahl, Folgeschaltungen usw. zu bestücken ist. Ausserdem kann auch eine Anzahl von Messwandler-Verstärkerschaltungen untergebracht werden, falls das System es verlangt.

EE 97 763 *für weitere Einzelheiten*

Vorwahlzähler

Darang Electronics Ltd, Restinor Way, Hackbridge Road, Hackbridge, Surrey

(*Abbildung Seite 613*)

In der von Darang Electronics Ltd neu eingeführten Baureihe 667 für elektronische Digital-Vorwahlzähler gibt es neun Modelle. Die Standardmodelle sind mit zwei, drei oder vier Dekaden bestückt und können Partien bis zu 99, 999 bzw. 9999 zählen. Sie sind mit oder ohne Anzeige der Gesamtzählung und mit oder ohne einzeilige Digitalanzeige lieferbar.

Alle sind mit Partie-Vorsignalen ausgerüstet, die der Anwender entsprechend seinen Wünschen wählen kann. Die Vorsignale können auf Wunsch von jeder der Dekaden gesteuert werden.

Die Konstruktion beruht auf einem System preisgünstiger Einsteckkarten, und eine äusserst zuverlässige moderne Kaltkathodentechnik findet Anwendung.

Es kann mit Einzelpartie-Zählung oder wiederholter Zählung gearbeitet werden, und eine automatische Kontrolle ist vorgesehen. Die Partievorwahl erfolgt mittels 10poliger Drehschalter mit bezifferten Skalen, die durch die Frontplatte geschützt werden.

Eingangssignale können zwischen 2 und 300 V liegen. Die Eingangsschaltung ist galvanisch gekoppelt und spricht auf gegenüber Erde positive Signale an. Der Eingangswiderstand ist 50 kΩ. Zählungen können auch über externe Kontakte ausgelöst werden. Die Ausgangskontakte sind für 5 A, 230 V~ bemessen.

EE 97 764 *für weitere Einzelheiten*

Montagehilfe für Mikroschaltungen

Jermyn Industries, Vestry Estate, Vestry Road, Sevenoaks, Kent

(*Abbildung Seite 613*)

Die Montagehilfe AE-10L ist für Mikroschaltungen im TO5-Gehäuse mit zehn Anschlüssen bestimmt und aus ICI-Nylon A.100 (Schmelzpunkt 200° C) hergestellt. Sie spreizt die Anschlüsse des TO5 in zwei parallele Linien von je fünf Anschlüssen mit 0,25 mm Rastermassabstand, was die Anschlussanordnung so umformt, dass sie mit dem genormten Raster der Druckschaltungen übereinstimmt und TO5 - Mikroschaltungen leichter eingesetzt werden können als Flachgehäuse.

EE 97 765 *für weitere Einzelheiten*

Infrarote Elemente

M.C.P. Electronics Ltd, Station Wharf Works, Alperton, Wembley, Middlesex

Zur Ergänzung ihres Programmes infrarotemittierender Gallium-Arsenid-Dioden MCA 600 hat M.C.P. Electronics Ltd die Fertigung einer Reihe lichtempfangender Elemente MSP/3 und MSP/6 aufgenommen.

Diese Elemente sind äusserst empfindlich; Ausgangsströme von bis zu 100 mA gestatten direkte Steuerung eines konventionellen elektromagnetischen Relais.

Sie haben eine Siliziumstruktur mit einer Spitzenfarbempfindlichkeit bei 9000 Å, können aber auch bei sichtbarem Licht benutzt werden.

Für Betrieb mit niedrigem Strom bietet die Silizium-Planar-Mikrophotodiode 30F2 den Vorteil kompakter Abmessungen und kann in optischen Lochstreifenlesern gestapelt werden.

EE 97 766 *für weitere Einzelheiten*

Potentiometer

Morganite Resistors Ltd, Bede Trading Estate, Jarrow, Co. Durham

(*Abbildung Seite 613*)

Das Morganite-Programm wurde durch zwei Abgleichpotentiometer in Metallkeramik mit geradliniger Schleiferbewegung für gedruckte Schaltungen erweitert. Sie werden als Typ 80 und 88 angeboten, nehmen auf der Druckschaltungsplatte nur etwa 1,5 cm² Raum in Anspruch und sind in bestimmten Werten lieferbar.

Typ 80 hat eine Nennbelastbarkeit von 0,75 W bei 70° C und stellt ein zuverlässiges und preisgünstiges Mittel für Schaltungsabgleich bei kommerzieller Anwendung dar.

Typ 88 wurde für Betrieb unter den härtesten Bedingungen im militärischen Einsatz entwickelt. Er ist robust, wasserbeständig und genügt den Forderungen von MIL-R-22097B Charakteristik C mit einer Nennbelastbarkeit von 1,0 W bei 85° C.

Beide Typen haben mit Edelmetall plattierte Anschlussstifte in 2,54 mm Rasterabstand. Die Widerstandsbahn wird nicht durch Feuchtigkeit beeinflusst oder durch Chemikalien angegriffen. Schaltungsabgleich kann in sehr engen Grenzen erfolgen, und die Fertigungsmethoden machen einen Ausfall unmöglich.

EE 97 767 *für weitere Einzelheiten*

Transistor-Tester

The Wayne Kerr Co. Ltd, Sycamore Grove, New Malden, Surrey

(*Abbildung Seite 613*)

Wayne Kerr hat einen neuen Tester herausgebracht, der alle h-Parameter von pnp- und npn-Transistoren bei 1000 Hz in Emitter- oder Basisschaltung misst. Ausserdem hat der Tester TT100 Einrichtungen, mit denen sich der Emitterstrom (0 ... 30 mA in sechs Bereichen) und die Kollektorspannung (0 ... 30 V in drei Bereichen) einstellen und überwachen lassen.

Die Arbeitsweise beruht auf Verwendung eines Oszillators mit sehr hoher Amplitudenkonstanz, genauer Widerstandsnetzwerke zum Anlegen des 1-kHz-Signals an die drei Transistoranschlüsse bei hoher Impedanz sowie eines Präzisions-NF-Millivoltmeters. Für alle Parameter wurde eine Messunsicherheit von ±3 Prozent des Vollausschlags erreicht. Die Anzeigegenauigkeit bei Strom- und Spannungsmessungen ist ±1 Prozent des Skalenendwertes.

Für alle Messungen wird nur eine Anschlussleiste (mit einer vierten Buchse für eine Leitungsabschirmung) benutzt, und die erforderlichen Verbindungen werden automatisch durch Betätigung eines Parameterwählers hergestellt. Für Kollektorspannungen über 30 V ist eine Klinke zum Anschluss externer Speisequellen und Messgeräte vorgesehen.

EE 97 768 *für weitere Einzelheiten*

Stromstossrelais

Parmeko Ltd, Percy Road, Aylestone Park, Leicester

(*Abbildung Seite 614*)

Parmeko Ltd hat die Ergänzung ihrer Baureihe G100 für abgedichtete Miniaturrelais durch Stromstossrelais mit Einzelspule und zweipoliger Umschaltung bekanntgegeben. Der mit G100L bezeichnete Typ ist für Anwendungszwecke bestimmt, bei denen Leistungsfähigkeit und Zuverlässigkeit unbedingt erforderlich sind.

Diese Relais sprechen bei nur 30 mW an und sind für Spulenspannungen von 1,5 ... 48 V lieferbar. Die Lebensdauer der Kontakte ist höher als 10^6 Schaltvorgänge bei unter 0,3 A und 10^5 Vorgänge bei 0,3...1 A. Das Relais G100L kann lineare Beschleunigungen von 100 g, bei hart aufmontiertem Relais Schwingungen bis zu Spitzenbeschleunigungen von 25 g bei 10 ... 3 500 Hz aushalten und innerhalb des Temperaturbereiches −55°C ... +125°C arbeiten.

Grundabmessungen sind 280 × 250 × 116 mm (Kristall-Gehäuse) bei etwa 20 g Gewicht. Höchste Zuverlässigkeit wird durch Abdichten von Spule und Kontakten in getrennten Abteilen und Verwendung einer Goldlegierung als Kontaktmaterial erreicht.

Parmeko kann auch Stromstossrelais mit Doppelspule liefern.

EE 97 769 *für weitere Einzelheiten*

Breitband-Voltmeter

Levell Electronics Ltd, Park Road, High Barnet, Hertfordshire

(*Abbildung Seite 614*)

Das tragbare Instrument TM6A misst Signale bis zum Mikrovoltpegel herunter über einen Frequenzbereich von 1 Hz bis zu 100 MHz. Es ist in Grösse und Ausstattung dem Levell-Wechselstrom-Mikrovoltmeter TM3A ähnlich, hat jedoch einen zusätzlichen HF-Messkopf. Insgesamt sind acht HF-Bereiche von 1 mV Vollausschlag bis zu 3 V Vollausschlag mit einer Ansprechempfindlichkeit von 300 µV im Band 300 kHz ... 50 MHz und von 3 mV bei 400 MHz vorgesehen. Die vierzehn NF-Bereiche haben Skalenendwerte von 50 µV bis zu 500 V für Messungen von 10 µV ab im Band 1 Hz ... 3 MHz. Diese Bereiche sind dieselben wie im Modell TM3A mit Ausnahme der ausgelassenen 15-µV- und 150-µV-Bereiche.

In den HF-Schaltungen werden ausschliesslich Halbleiterschaltungen zur Umwandlung der HF-Signale in Rechteckwellenform von etwa 20 Hz und dem Quadrat des HF-Signals proportionaler Amplitude verwendet. Es wird kein mechanischer Zerhacker benutzt, und die Schaltungen sind ausreichend temperaturkompensiert. Durch den hohen Wirkungsgrad wird der 9-V-Batterie nur ein Strom von 10 mA entnommen.

In allen HF-Bereichen spricht das Instrument auf echte Effektivwerte, in NF-Bereichen jedoch auf Mittelwerte an und ist in Effektivwerten für sinusförmigen Eingang geeicht.

EE 97 770 *für weitere Einzelheiten*

Digital-Schreibmaschine

Hilger & Watts Ltd, 98 St. Pancras Way, Camden Road, London, N.W.1

(*Abbildung Seite 614*)

Digitale elektrische Schreibmaschinen mit oder ohne Programmiereinrichtung wurden von Hilger & Watts Ltd für Anwendung in automatischen Datenverarbeitungssystemen angekündigt. Die Grundausführung hat als Schreibmaschine FD574 einen 43,2 cm langen Wagen, Zeichen im Elite-Schriftbild und ist für elektronische Fernsteuerung ausgelegt.

Für Magnetspulenbetätigung ausgelegte Schriftzeichen- und Funktionstasten gibt es für die Ziffern 0 bis 9, Tabulator, Punkt, Leertaste, Wagenrücklauf-Zeilensprung, Stern, Buchstaben A und R und Farbbandumschaltung rot-schwarz.

Zusätzliche oder alternative Tasten können im Sonderauftrag bis zu insgesamt 26 Tasten plus Leertaste auf Magnetspulenbetätigung umgestellt werden.

Die Grundmaschine kann mit Kommutator und Steckfeld FD576 nachgerüstet werden, was Änderungen im Layout der gedruckten Information durch Auswechseln der nach Anwenderwünschen erstellten Steck - Programmierbausteine an der Schreibmaschine ermöglicht.

Der Kommutator dient als Parallel-Serien-Umsetzer und gewährleistet, dass die ausgedruckten Daten mit der Informationsquelle synchronisiert bleiben. Ferner gestattet er, dass die Schreibmaschine mit ihrer optimalen Geschwindigkeit von 8 bis 10 Zeichen/Sekunde arbeitet.

Die maximale Kapazität des Kommutators ist 190 Kolonnen, und die Schreibmaschine kann bis zu 50 Eingangsleitungen aufnehmen.

Die Abbildung zeigt eine Rückansicht der Schreibmaschine, von der die Deckplatte des Kommutators entfernt ist.

EE 97 771 *für weitere Einzelheiten*

Drehzahlregelung

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

(*Abbildung Seite 614*)

Ein neues Fertigungsprogramm für Steuergeräte zur genauen Drehzahlregelung von Gleichstrommotoren vom Stillstand zur Grundgeschwindigkeit hat der Bereich Automatisierung der M.E.L. Equipment Co. Ltd angekündigt. "Ergotrol" ermöglicht Änderung der Drehzahl von Gleichstrommotoren entweder manuell oder durch ein Steuersignal von einer Mutterausrüstung oder einem System. Die am Anker des Motors liegende Spannung wird mittels steuerbarer Siliziumgleichrichter geregelt, und die Anordnung soll gleichförmig wirksam und zuverlässig regeln.

Das Programm der vom Wechselstromnetz betriebenen Geräte umfasst sieben Modelle für Gleichstrom-Nebenschlussmotoren von 1 ... 40 PS nach der britischen Norm B.S. 2613. Regelung der eingestellten Geschwindigkeit wird durch Rückkopplung von der Ankerspeisung bestimmt und ist innerhalb 2,5 Prozent der Grunddrehzahl. Wenn grössere Genauigkeit erforderlich ist, kann einem getrennten Tachogenerator ein Rückkopplungssignal entnommen werden, wodurch die Regelung innerhalb 1 Prozent der Grunddrehzahl gebracht werden kann. Die Tachogenerator-Methode hat den weiteren wichtigen Vorteil, dass ihre Genauigkeit von der Erwärmung der Feldwicklung nicht beeinflusst wird.

Das Anlaufen wird durch Festkörperschaltungen gesteuert, die selbst dann gegen Überstrom schützen, wenn die Drehzahlregelung beim Anlaufen auf höchste Geschwindigkeit eingestellt ist. Obwohl der Ankerstrom begrenzt ist, besteht genug Spielraum für die Beschleunigung.

Ein Feintrieb für Drehzahleinstellung, Start- und Stopp-Tasten gehören zur Standardausrüstung. Ein in Prozent der Drehzahl oder nach Sonderwünschen geeichter Geschwindigkeitsanzeiger gehören auch zu einer Reihe wahlweiser Einrichtungen. Auch einen in Strom geeichten Lastanzeiger kann man anbauen. Ein weiterer wahlweiser Zusatz ist eine Einrichtung für Richtungsumkehr, in der die Umkehr manuell eingestellt, dann aber automatisch bis zum Stillstand gebremst wird, ehe das Laufen in umgekehrter Richtung beginnt. Dynamisches Bremsen für schnelles Stoppen kann auch vorgesehen werden. Ausserdem sind alle Regler und Anzeiger auf einer getrennten Tafel für Fernbedienung lieferbar.

Die Steuergeräte bis zu 10 PS werden vom Einphasennetz gespeist und sind für Wandmontage konstruiert; Geräte für über 30 PS sind für Drehstrom ausgelegt und in eine auf dem Fussboden stehende Konsole eingebaut. Auf Wunsch sind Geräte ohne Gehäuse für Einbau in andere Ausrüstungen lieferbar.

EE 97 772 *für weitere Einzelheiten*

Sauerstoffanalysator

Servomex Controls Ltd, Crowborough, Sussex

(*Abbildung Seite 615*)

Servomex Controls Ltd hat ihr Programm für Sauerstoffanalysatoren durch ein tragbares, batteriegespeistes Modell OA.150 ergänzt. Das neue Instrument hat zwei umschaltbare Bereiche 0 ... 25 Prozent und 0 ... 100 Prozent mit einer Messunsicherheit von ±1 Prozent des Skalenendwertes für jeden Bereich.

Das Hauptmerkmal des Instrumentes ist seine einfache Betriebsweise, denn als Bedienelemente kommen nur der Bereichschalter und die Betätigungstaste in Frage. Die Anzeige erfolgt auf einem eingebauten Messgerät. Der Bereichschalter hat eine weitere Position, in der mit demselben Messgerät Batteriespannung und Schaltungsabgleich überprüft werden können. Eine grafische Darstellung auf der Frontplatte macht die Arbeitsweise und Prüfmethode sehr offensichtlich; wo immer möglich, werden Symbole anstelle von Worten benutzt, damit das Instrument in gleicher Weise für Exportmärkte geeignet ist.

Der Analysator kann mit fliessenden oder statischen Proben arbeiten; Probendurchfliessungen bis zu 150 ml/min beeinflussen die Anzeige nicht. Ein Scheibenfilter aus Sinterglas und Ventile zur Einregelung des Analysators und Umgehungsstroms sind im Instrument eingebaut; ein Handaspirator und ein Trockenrohr gehören zur Standardausrüstung.

Dieser Analysator ist mit derselben Messzelle ausgerüstet wie die früheren Servomex-Sauerstoffanalysatoren. Ein Quarz-Hantelmodell ist in einem ungleichförmigen Magnetfeld über einem Platinheizfaden aufgehängt und erhält ein der magnetischen Suszeptibilität der Gasprobe proportionales Drehmoment. Es wird durch Aufrechterhaltung eines gleichen, entgegengesetzten Rückstelldrehmomentes gemessen, das durch den in einer eingängigen Spule auf dem Hantelmodell fliessenden Strom erzeugt wird. Eine Lichtquelle, Zwillingsfotozellen und ein Differenzverstärker stellen den Nullabgleich automatisch her. Der Ausgangsanzeiger misst den Rückstellstrom, der dem Sauerstoffgehalt direkt proportional ist. Silizium-Fotozellen und Silizium-Transistoren finden Anwendung, und das Instrument ist ausserordentlich robust und zuverlässig. Ein Temperaturausgleichssystem hält die Messunsicherheit bei Abweichungen von ±5° C auf dem spezifizierten Wert.

Die Messung wird praktisch durch die gewöhnlichen Gase ausser Sauerstoff nicht beeinflusst, und man kann das Instrument daher mit Stickstoff und Sauerstoff oder Luft prüfen und nachfolgend für die Analyse von Mischgasen einschliesslich beispielsweise Stickstoffoxyd, Kohlensäure oder Wasserstoff einsetzen.

EE 97 773 *für weitere Einzelheiten*

Mehrkanalverstärker

S.E. Laboratories (Engineering) Ltd, Astronaut House, Feltham, Middlesex

(*Abbildung Seite 615*)

Das von S.E. Laboratories (Engineering) Ltd eingeführte Mehrkanal-Verstärkersystem SE.4000 für Gestelleinbau oder in Arbeitsplatzausführung kann bis zu acht verschiedene oder identische Verstärker hoher Leistungsfähigkeit aufnehmen.

Das System eignet sich vor allem für Einsatz mit Thermoelementen, Widerstandsthermometern, Dehnungsmessstreifen, Differentialübertragern, Messwandlern mit Widerstand oder mit regelbarer

Reduktanz in Halb- oder Vollbrückenschaltung, Geschwindigkeitsaufnehmern, selbsterregten Messwandlern, Durchströmungsmessern, Millivoltsignalen und dergleichen.

Ein Merkmal ist das wahlweise lieferbare eingebaute Eichsystem, das man von Hand oder automatisch betreiben kann. In der kontinuierlichen Betriebsweise können daher Eichsignale über einen Rechner oder ein Datenerfassungssystem programmiert werden und so als Bezugssignal für beliebige Eingangsdaten dienen.

Jeder Verstärkermodul ist mit einer Frontplatte ausgerüstet, auf die eine Überlastungsanzeige montiert ist. Die Treiberschaltung kann man auch mit externen Alarmvorrichtungen verbinden, die bei Abweichen der Eingangsdaten von vorgegebenen Toleranzen erregt werden.

Alle Verstärkerkanäle haben einen Standardausgang von ±1,4 V, 10 mA. Ein Huckepack-Treiberverstärker kann in jeden Verstärker eingesteckt werden und erhöht die Ausgangsleistung auf ±10 V, ±110 mA.

Verstärkermoduln kommen in vier Grundtypen, die alle innerhalb des ebenfalls lieferbaren Gestells Baureihe SE.4000 austauschbar sind.

EE 97 774 *für weitere Einzelheiten*

Niederspannungs-Stromversorgung

The British Electric Resistance Co. Ltd, Queensway, Enfield, Middlesex

(*Abbildung Seite 615*)

Die BERCO-Niederspannungs-Stromversorgung ist eine zweckdienliche Quelle für Gleich- und Wechselstrom niedriger Spannung zur Speisung von Ausbildungsapparaten und wurde von BERCO und der Nuffield-Stiftung gemeinsam entwickelt.

Ein bifilarer, regelbarer Drehtransformator "REGAVOLT" und ein Selen-Vollwegbrückengleichrichter sind in einem belüfteten Stahlgehäuse untergebracht, auf desen Frontplatte ein Netzschalter, eine Neonanzeigelampe, der Überlastungsschütz des "Regavolt", zwei Wechselstromausgangsklemmen, zwei Gleichstromausgangsklemmen sowie eine Erdklemme angeordnet sind.

Die Netzspannung wird über einen zweipoligen Schalter zu einer Anzeigelampe und dann zur Primärwicklung des bifilaren Regavolt - Drehtransformators gespeist. Die Wicklung hat Abgriffe für Anschluss jeder beliebigen Netzspannung zwischen 200 und 250 V in 10-V-Stufen.

Die Sekundärwicklung des "Regavolt" isoliert den Ausgang völlig vom Netz, und die Bürste ermöglicht eine glatte, kontinuierliche Ausgangsregelung von 0 ... 25 V.

Ein in Serie mit der Ausgangsbürste des "Regavolt"-Transformators liegender Schütz sichert das Gerät gegen Kurzschlüsse und Dauerüberlastung. Der Ausgang des Schütz liegt an einem Vollwegbrückengleichrichter der Selentype mit sehr hoher Überlastungskapazität.

Es wird nicht versucht, den Ausgangsgleichstrom zu glätten, da die Aufgabe war, für Schulen und Versuche eine Gleichstromversorgung zu liefern, deren Kosten so gering wie möglich sein sollten. Für verschiedene Anwendungsmöglichkeiten kann die erforderliche Glättung unterschiedlich sein.

Jeder Ausgang liegt an einem Schraubklemmenpaar. Die Wechselstromklemmen sind gelb kodiert, die Gleichstromklemmen rot für positiv und schwarz für negativ zur einfachen Sichtkennzeichnung der Polarität. Die Erdklemme hat grüne Farbkennzeichnung. Zusätzliche Information wird durch Beschriftung zur Kennzeichnung der Ausgangsspannungsart und des Nennstroms gegeben.

Auf der Rückseite des Gerätes ist ein längliches Abteil zur Aufnahme des 1,50 m langen, dreiadrigen Kabels vorgesehen. Für Einsatz auf einem Arbeitstisch sind Gummifüsse vorgesehen.

EE 97 775 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Entwurf eines Transistor-Pulsbreitenmodulators für Steuerzwecke von R. D. Bell und K. E. Tait

Zusammenfassung des Beitrages auf Seite 562-567

Der Entwurf eines Pulsbreitenmodulators, der für Anwendung in Analogrechneruntersuchungen pulsbreitenmodulierter Steuersysteme geeignet ist, wird besprochen. Eine theoretische Vorhersage der Grenzen und Genauigkeit des entwickelten Gerätes wird gegeben und diese Grenzen mit praktischen Ergebnissen kritisch verglichen.

Entwurf eines Kleinsignalverstärkers mit Feldeffekttransistoren von W. Gosling

Zusammenfassung des Beitrages auf Seite 568-571

Der Entwurf eines Kleinsignalverstärkers mit bis fast zur Sperrung vorgespannten FETs wird besprochen und die zur Stabilisierung des Arbeitspunktes erforderliche Wahl der Bauelementwerte beschrieben. Die Entwurfsmethode führt zu Ergebnissen, die mit den gemessenen Werten annähernd übereinstimmen; die Verstärker haben einen typischen Spannungsverstärkungsfaktor von etwa 50 sowie den ungewöhnlichen Vorteil hoher Eingangsimpedanz, der Feldeffekttransistoren charakterisiert.

Analyse und Leistungswerte eines Transistorchoppers von D. J. Finlay

Zusammenfassung des Beitrages auf Seite 572-578

Wenn es sich um die Verstärkung eines sehr kleinen Gleichstroms oder einer sehr niedrigen Gleichspannung handelt, wird manchmal die Anwendung von Chopperschaltungen erforderlich. Viele Transistor-Chopperschaltungen sind veröffentlicht worden, und ihre Arbeitsweise wird als bekannt vorausgesetzt. Der Einfluss der verschiedenen Mängel des Choppers auf seine Gesamtleistung ist jedoch nicht offensichtlich, und die sich dadurch ergebenden unzulänglichen Entwürfe können sehr zeitverschwendend werden. Dieser Beitrag gibt eine volle Analyse mit Ersatzschaltungen für einen Eingangs-Transistorchopper, mit deren Hilfe der Einfluss der Änderungen in Quellen- und Lastwiderstand, Temperatur, Chopperfrequenz und Transistorparameter besser verständlich gemacht wird. Typische Werte werden angegeben und mit den Nullfehlern und ihrer Drift in Germanium- und Siliziumflächentransistoren verglichen. Es wird gezeigt, dass der durch Einschwingvorgänge hervorgerufene Fehler ernst zu nehmen ist und nahegelegt, dass für das Messen von Strömen in der Grössenordnung von 1 nA und Spannungen von etwa 10 µV genauere Information über Einschwingvorgänge und ihre Drift erforderlich sei.

Eine einfache Schaltung, die zur Linearisierung der Übertragungscharakteristik die verzerrenden Einschwingvorgänge vom Ausgang entfernt, wird auch beschrieben.

Q-Messungen an verlustarmen Wellenleiterhohlräumen von J. K. Chamberlain

Zusammenfassung des Beitrages auf Seite 579-581

Der Q-Faktor einer ausgewählten Schwingungsart in einem Mikrowellenhohlraum ist von den ohmschen und anderen Verlusten abhängig, mit denen er behaftet ist; wenn der Hohlraum durch einen geraden Zylinder gebildet wird, hängt der Faktor mit der Dämpfungskonstanten des Wellenleiters zusammen, der durch die zylindrischen Wände gebildet wird. Der Beitrag umreisst eine Messtechnik, die bei Einsatz einfacher Geräte für die hohen Q-Werte (10^5. . .10^6 und höher) in verlustarmen Wellenleiterhohlräumen geeignet ist. Als Beispiel wird ein Kupferwellenleiter von 2 Zoll (50,8 mm) Durchmesser bei 35 GHz aufgeführt, für den eine Dämpfungskonstante von 3,4 dB/Meile (2,1 dB/km) angegeben wird.

Ein entzerrender Wellenformübertrager für amplitudenabgetastete Systeme von T. I. Mitchell and V. J. Phillips

Zusammenfassung des Beitrages auf Seite 582-587

Der intuitiv offenkundige Weg zur Rückgewinnung eines Signals, dessen Amplituden abgetastet wurden, ist das "Verbinden der Probenspitzen"; dieses Verfahren ist jedoch in der Praxis gar nicht so einfach wie es aussieht. Die Autoren beschreiben im vorliegenden Beitrag Geräte, die sie für diesen Zweck konstruiert haben.

Spektrumanalyse einer durch variable Steilheit der Hinterflanke modulierten Impulsreihe
von O. E. Kruse und R. W. Montgomery

Zusammenfassung des Beitrages auf Seite 593-595

Eine mathematische Analyse des Spektrums einer Reihe dreieckig geformter Impulse, deren Hinterflanken moduliert sind, wird entwickelt. Das mathematisch vorhergesagte Spektrum wird dann mit dem versuchsmässig bestimmten verglichen.

Wobbelfrequenz für die Gesamtprüfung von Empfängern von J. F. Golding

Zusammenfassung des Beitrages auf Seite 596-601

Die Anwendung eines Wobbulators in Zusammenarbeit mit einem Oszillografen für den Abgleich von Empfängern ist so weit verbreitet, dass es sich kaum lohnt, darüber zu berichten. Es gibt jedoch noch weitere Anwendungsmöglichkeiten für das gewobbelte Testsignal in der Bestimmung der Gesamtleistungskennwerte eines Empfängers. So können z.B. Empfindlichkeit und Rauschabstand oftmals einfacher mit einem Wobbulator als mit konventionellen Mitteln gemessen werden.

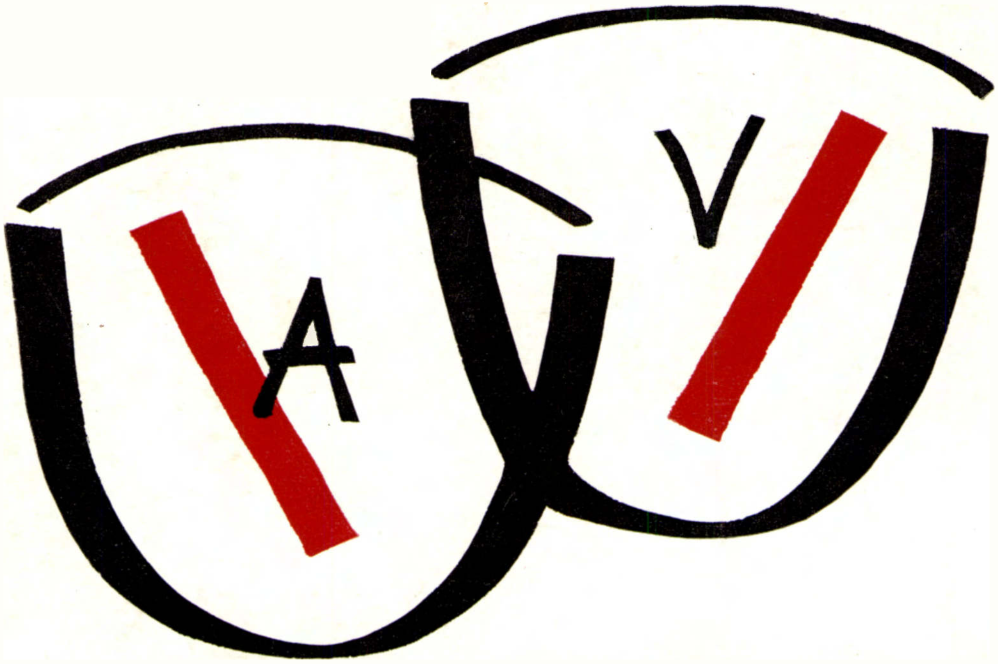
Einige Anwendungsmöglichkeiten für zweispulige Relais in elektronischen Schaltungen von H. Biggar

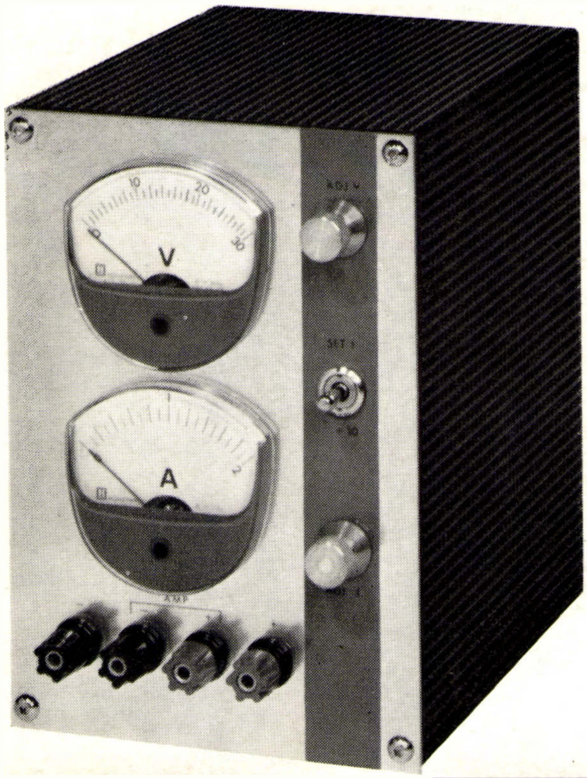
Zusammenfassung des Beitrages auf Seite 602-606

Das zweispulige Relais ist ein sehr vielseitiges elektromagnetisches Bauelement. Jede Spule wird getrennt erregt und trägt zum zusammengesetzten Magnetfeld bei. Die Anwendungsmöglichkeiten für ein solches Relais fallen—je nachdem, ob es additiv oder subtraktiv ist—in klar abgegrenzte Gruppen. Zweispulige Relais haben viele nützliche Funktionen, die mit einspuligen Relais nicht so leicht zu erreichen sind.

Der Anwendungsumfang kann dadurch erweitert werden, dass man das Relais in eine elektronische Schaltung mit Fotozellen oder Trioden integriert. Der Beitrag schliesst mit der Beschreibung einer ungewöhnlichen Schaltung für Steuerungs- und Schutzautomatik einer Anzahl von Hochleistungselektronenröhren, die mit einer gemeinsamen Höchstspannungsquelle betrieben werden.

EE 97 004 for further details





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LA200	0-30	2amp					£79
LA400	0-15	4amp					£81
LB200	0-50	2amp	8½" 216mm	5⅝" 143mm	14½" 368mm	18lbs 8.2kg	£105
LB500	0-30	5amp					£110
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