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Commentary

WITH a prosperous and expanding electronic industry in this country it is not surprising that the British Component Manufacturers are themselves experiencing something resembling a boom, and evidence of this is to be found in the Twenty-Seventh Annual Report for 1959 of the Radio and Electronic Component Manufacturers' Federation which has just been published.

In actual numbers, the members of the Federation, which represents practically the entire component industry, produced during the year the staggering total of 2 400 million components valued at £120m and for the first time in the life of the industry an output of one million components was reached for every hour of the working week.

The Federation has found it no problem counting up how many 'bits and pieces' it has made, but it has run into difficulties in determining just where these components go and what is their ultimate destiny and it cannot determine with certainty whether a component ends up in a £100 television receiver or in a £100 000 computer or more humbly by being sold over a dealer's counter as a replacement part.

The trouble derives in the main in defining what is at present vaguely known as the electronic industry. What we call the electronic industry comprises much more than the domestic radio receiver industry from which it grew, and now shows every sign of outstripping.

The trouble is accentuated by the absence of suitable statistics and it is suggested that the time is ripe to provide a separate Industrial Classification and to revise the Census of Production by bringing under the appropriate headings all those various radio, electronic, audio, telecommunication and other scientific products which are now hidden under other classifications. A further step forward could be made by collecting more information from individual firms either through Board of Trade channels or through the various Trade Associations.

Even so, the component industry has had a good year, attributable in the main to the considerable increase in production of radio and television receivers for the home market, brought about by the relaxation of credit restrictions and the general rise in demand for consumer goods. The domestic receiver industry accounted for over 40 per cent of the component output and in fact reached its highest ever level of output, producing some 2¾ million television receivers and 1½ million radio receivers during 1959.

Exports too were a satisfactory feature, the total figure being in the region of £22½m for components proper and sound reproducing equipment such as record players.

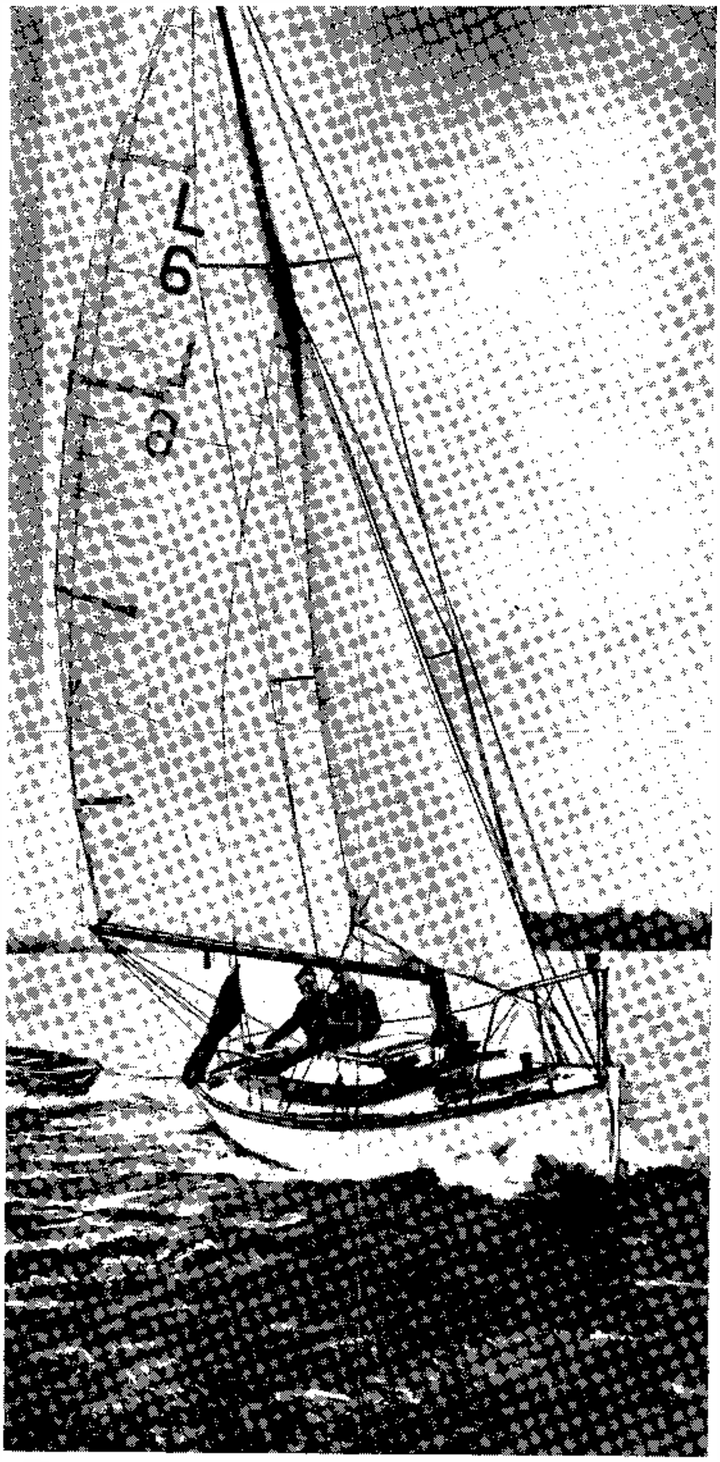
The Federation however, views with some concern the rise in imports which it regards as unprecedented and forecasts that if imports continue at this rate we shall in a few years time be importing more than we export, a position which is almost reached today with regard to valves and cathode ray tubes. These imports come in the main from Holland, West Germany and the U.S.A. and consist of audio products, professional equipment and valves and tubes. Roughly, the imports were just over £5m under each heading. The largest section was valves and tubes (£5.7m) closely followed by audio products comprising tape recorders and disks, gramophone record players and loudspeakers.

Turning to other matters of interest, the Federation in its Report announces that the last of the RECMF's Exhibitions has now been held at Grosvenor House and that from May of next year its show will be a biennial event to be held at Olympia.

This decision has obviously not been an easy one to make and it will be received with mixed feelings, for Grosvenor House in the post war years has become more than an Exhibition.

But because of its very success the Exhibition has outgrown its environment and drastic action has been needed. The acquisition of accommodation at nearby Park Lane House would, it was hoped, relieve the pressure, but it proved unpopular to exhibitor and visitor alike.

The decision to make the Exhibition a biennial event is in itself a sensible one for it will fit in conveniently with the Instrument Industries Exhibition which from this year is also to be staged every other year.



The author's yacht "Wavecrest II" in which the echo-sounding trials were carried out during the winter of 1958-1959.

A Miniature Transistorized Echo-Sounder

By R. N. B. Gatehouse*, A.M.I.E.E.

This article describes the development and the main design features of a very small and robust echo-sounder for use in depths from 2·5ft to 32 fathoms. The instrument is suitable for such purposes as hydrographic surveys for the laying of underwater pipelines and the pilotage of yachts and small fishing vessels. The complete equipment weighs only 5¼lb, has a power consumption of 0·1W and provides an accuracy of ±3 per cent or 6in, whichever is the greater.

(Voir page 394 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 400)

ALTHOUGH the first practical 'Asdic' was constructed as long ago as 1918 (by Paul Langevin in France), it is only during the last five years that echo sounders suitable for use in small craft operating in coastal waters have been produced. The reason for this appears to be that no suitable transducer materials were available to exploit the economic advantage of using very high ultrasonic frequencies where deep soundings are not required. The recent discovery of the piezo-electric ceramics has enabled relatively small and efficient transducers to be constructed which has led to the design of quite compact short-range echo-sounders consuming less than 50W and being of moderate cost.

The principle of echo-sounding is well established. A transmitter sends a train of brief pulses of acoustic energy into the water, a receiver picks up and amplifies the echoes from the sea bed and the time taken by each pulse to do the round trip is measured by means of a time-base.

Sea water is a good medium for the transmission of soundwaves, losses by dissipation being only 0·002dB/m at 20kc/s. Losses rise rapidly with frequency, however, being about 0·1dB/m at 230kc/s[1]. The velocity of sound waves in the sea is about 4 800ft/sec, but varies considerably with temperature and salinity.

Ocean-going ships require soundings in depths up to about 500 fathoms (3 000ft), and must use carrier frequencies in the range 10 to 50kc/s. Trawlers sound and locate fish in depths up to about 100 fathoms, using frequencies of about 100kc/s. For inshore fishing boats and yachts, soundings in depths greater than 20 fathoms are seldom required, and frequencies up to about 200kc/s may be used.

The transducers for deep-sea installations, and most of those for trawlers, use the magneto-striction effect in nickel and are often fitted with separate transducers for transmission and reception. These are sometimes fitted in tanks in the hull, the attenuation in the ship's plating being regarded as a lesser evil than holes through it. The presentation with these bigger sets is nearly always by means of a fast-moving 'electric pen' which in running

* *Brookes & Gatehouse Ltd.*

repeatedly across a slowly-moving strip of special paper, in synchronism with the firing of the transmitter, marks up echoes as dark dots. The sea bed appears as a continuous line, displaced from the left-hand edge of the strip by a distance proportional to depth.

The short and medium-range sets now nearly all use piezo-electric ceramics, such as barium titanate, in their transducers and present echoes as luminous patches on a circular range scale by means of a gas discharge tube spinning on the arm of a motor shaft. This is synchronized with the transmitter, one pulse being fired on each revolution. A few sets display echoes on c.r.t. timebases, as in radar.

The Objective

The development of 'HECTA' was undertaken with a view to producing a 'pocket-sized' echo-sounder with a range of 30 fathoms that might be used in the smallest fishing boats, surveyors' dories and sailing yachts. To make this possible the power consumption was to be limited to 0·1W. The instrument was to be capable of withstanding the severe knocks and wettings to which all small-boat gear is subjected. The hopes for success lay in taking advantage of the very high overall efficiency of transistors as compared with thermionic valves, in abandoning the rotating lamp in favour of automatic range computation and a milliammeter as a means of presentation, and in using a strong, hermetically-sealed case. The use of a meter would bring the important advantages of visibility in sunlight, improved reliability and easy variation of scale sensitivity.

The Development Programme

Calculation showed that the requirements in a pulsed echo-sounder for transmitted power and receiver gain were quite modest in terms of the capabilities of transistor circuits. For example the fraction of the transmitted power which is available in the echo at 175kc/s, using a 60° beam over a muddy bottom 30 fathoms down, is about 2×10^{-11}. Therefore with a peak transmitter power output of 0·5W to a barium titanate transducer, a receiver gain of 80dB would be required to deliver 0·2mW to the range-measuring circuit. The corresponding mean power input would be about 50mW, leaving 50mW over for the receiver and display circuits.

The greater part of the development programme was aimed at finding out the noise characteristics of the sea and the reflectivity of the natural things suspended in it and of the sea-bed itself, so that the practicability of auto-ranging on the bottom echo could be established. (It will be appreciated that with the spinning lamp and c.r.t. displays the operator can, with experience, 'filter' out the unwanted signals.) The transmitting and receiving circuits were developed as part of the trials programme, for the use of miniature, transistorized measuring gear was essential in view of the fact that the 'test bed' was a sailing vessel of only 19ft waterline length carrying no electrical generators.

The Results of the Trials

(a) Sea Noise

Noise generated in the sea[5] has a power which varies inversely as frequency and appears to be due mainly to the movement of shingle and sand on the bottom caused by tidal streams. Breaking seas also produce noise which can be troublesome. Our measurements at 175kc/s indicated a peak noise power of about 10^{-13}W/cm^2/kc/s over shingle in breaking seas in tidal streams of surface speeds 5 knots.

(b) Mid-water Echoes

The sea contains many objects, apart from fish, which scatter sound waves. The physical extent of each of these is generally much less than the diameter of the acoustic beam and so their echo power generally varies inversely as the fourth power of range. The sea-bed echo power varies as the inverse square of range, and so it can be seen that these echoes are to be reckoned with only at the shorter ranges. Our measurements showed that, except in the case of patches of aerated water, the amplitude of suspension echoes very seldom exceeds that of the sea bed echo if the latter were at the same range. Aerated water is found in the wakes of ships and in heavy overfalls, and produces echoes of great intensity. The auto-ranging system will not, however, be at an appreciable disadvantage in comparison with the others under these conditions since the bottom echo is heavily attenuated by scattering in aerated water and soundings are then generally unobtainable with any system.

(c) Wave Echoes

Echoes from wave-crests, due to radiation outside the limits of the main acoustic beam, were troublesome at first, and it was found that care was needed in the design of the transmitter and transducer to minimize radiation from the disk in the radial mode of oscillation.

(d) Bottom Echo Coefficient

The power reflection coefficient of very soft, muddy bottoms was sometimes found to be as low as 0·01. For sandy bottoms the figure was generally 0·15.

Planning the System

Both continuous-wave and pulsed systems are practicable. In a c.w. system information is carried by modulation, and this may be either a.m. or f.m. An experimental f.m. radar system has already been described[3], but so far as is known only the pulsed system has been used in echo-sounders. Pulsed operation confers the advantage of simplicity of design, but for bottom-ranging is inefficient on account of the relatively high values of receiver bandwidth and transmitted power needed for a given accuracy. With c.w. operation, the designer may achieve higher efficiency by taking advantage of the fact that the depth of water below a moving ship changes but slowly with time and may be measured to the desired accuracy in a bandwidth of only a few cycles per second. For example, the transmitter may be modulated by a low-frequency sine-wave and the phase difference between the envelopes of the transmitted and received signals then measured in order to compute range. The narrowing of the receiver bandwidth permits a reduction in transmitted power for a given level of sea noise.

The major difficulties in designing a c.w. system are to provide very high signal attenuation in the direct, local path between transmitter output and receiver input, to ensure that the transmitter frequency does not drift outside the receiver pass-band, and to discriminate against wave- and suspension-echoes. The cost of the resulting echo-sounder might be uneconomically high. For this reason it was decided to adopt a compromise by using long pulses and a commensurately narrow bandwidth for the deeper soundings and in shoal soundings to reduce the receiver gain so that noise becomes unimportant. Then the usual short-pulse wide pass-band arrangement, which is necessary in a pulse system for measuring at the lower limit of range, is acceptable.

The trials results showed that the expected range performance was attainable with the prescribed limit of 0·1W for total power drain. They also showed that the great majority of suspension and wave-echoes would be suppressed if at all times, in each pulse cycle, the receiver gain was only just high enough for the bottom echo to be detected. A system of 'swept gain', such as is used in marine radar sets to discriminate against wave echoes, was developed to meet this requirement. An automatic gain control system, working only on the signal in the range-measuring 'gate' and controlling the mean receiver gain, further reduced the probability of this interference occurring, particularly over sea-beds of high reflectivity.

The second important consideration was that of sea noise. The trials showed that the greatest peak intensity one was likely to experience was about 10^{-13}W/cm^2/kc/s at 175kc/s. It must be arranged that this noise is of considerably less amplitude, by the time it reaches the ranging circuits, than that of the bottom echo. The curves of Fig. 1 show both peak sea noise in a bandwidth of 200c/s and echo power plotted against frequency. 200c/s is the lowest receiver bandwidth that can be tolerated, in consideration of the requirements for range accuracy. The curves indicate if frequencies below 200kc/s are used, the effects of noise may not be serious.

Fixing the Parameters

The following parameters in an echo-sounding system, as in radar, may be varied by the designer.

The carrier frequency.
The transmitted power.
The transmitted pulse length.
The shape and angular width of the radiated beam.
The receiver gain.
The receiver bandwidth.

Let P_t = power transmitted into water during each pulse (W).
P_r = echo power from sea bed incident upon transducer face (W).
d = diameter of transducer face (taking practical case of a cylindrical transducer) (ft).
θ = acoustic beam-width, between first zeros (radians).
h = depth of water below transducer (ft).
K = power reflection coefficient of sea bed.
f = frequency (kc/s).

Then: $$P_r = \frac{36 \, . \, K \, . \, P_t}{h^2 \, . \, \theta^4 \, . \, f^2 \, . \exp\,(4{\cdot}48 \times 10^{-7} \, . \, f^2 \, . \, h).}$$

The derivation of this expression and the approximations used are to be found in the Appendix.

The object is to make P_r as high as practicable in the interest of efficiency and of overcoming sea noise. Let us see how far we can go, bearing in mind the practical limitations:—

(1) In HECTA P_t is limited by considerations of battery power drain at about 0·01W (corresponding to a peak electrical input power of 0·5W and a mean electrical input power of 0·05W).

(2) θ must clearly be made as small as possible, consistent with the need to accommodate rolling and pitching without stabilization. Since the angle of roll is typically three times the angle of pitch, it would be advantageous to provide an elliptically-shaped beam. This was, however, ruled out on account of practical difficulties and θ was fixed at one radian.

(3) f must be as low as possible, consistent with the need to restrict the size of the transducer. It should be remembered that the volume and hence the cost of the transducer material varies as the inverse cube of frequency for a given value of θ.

f therefore remains as the one variable factor, and the curves of Fig. 1 were prepared to assist in deciding upon the best value. Here the value of P_r is plotted against f for a depth of 30 fathoms, with K fixed at 0·1, which is believed to be the lowest value of sea-bed coefficient likely to be found in coastal waters of that depth. Values for f less than 100kc/s were not taken, since transducers then become too large and expensive. The upper limit for f is set by a consideration of the noise power (P_n) in the sea, the density of which is reported[3] to be inversely proportional to frequency. Since, for a fixed value of θ, the transducer aperture area varies as $1/f^2$, it follows that the intercepted noise power varies as $1/f^3$. This function is also plotted in Fig. 1 for a bandwidth of 200c/s. It was found that the ratio $P_r/P_{n(pk)}$ in the auto-ranging system must be greater than about 10 if disturbance of the output is to be avoided. As will be seen in Fig. 1, this sets the upper limit for f at about 200kc/s. A frequency of 175kc/s was chosen as providing a compromise between the requirements for a high value of $P_r/P_{n(pk)}$ on the one hand and low size and cost of transducer on the other.

The pulse power required to operate the range-measuring circuit is about 0·2mW. From Fig. 1 it is seen that the available acoustic echo power at full range is 2×10^{-12}W. The receiving amplifier must therefore have a power gain of about 10^8 or 80dB, a figure that is easily obtainable in practice.

Description of 'HECTA'

HECTA uses 15 transistors and no thermionic valves.

Referring to Fig. 2, it will be seen that a single transducer is used for both transmission and reception. This is a barium titanate disk 23mm in diameter mounted in a tube of 'fibreglass' which contains also a matching transformer.

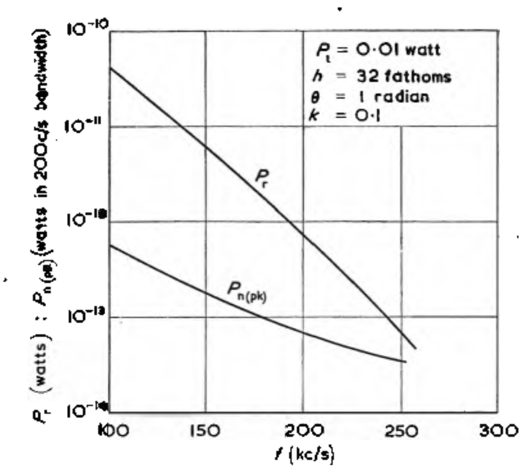


Fig. 1. Curves showing the relationship between
(a) echo power and carrier frequency for a fixed beam angle
(b) peak sea noise power and frequency for a fixed beam angle

The transmitter is an oscillator at 175kc/s, delivering 4V r.m.s. from a very low output impedance. It is modulated by a square-wave pulse generator at a constant rate of 12·5p/s. The mark-to-space ratio is about 1:150 on the short range scale and 1:15 on the long range scale. These scales are 2·5 to 32ft and 2·5 to 32 fathoms respectively. The modulator also controls a monostable relay circuit which has been called the 'guard pulse generator', the purposes of which are to disable the first receiving amplifier while the transducer is energized and to give a 'go' signal to the range-measuring relay at an instant when the remaining energy in the transducer has fallen to a level well below the threshold of detection.

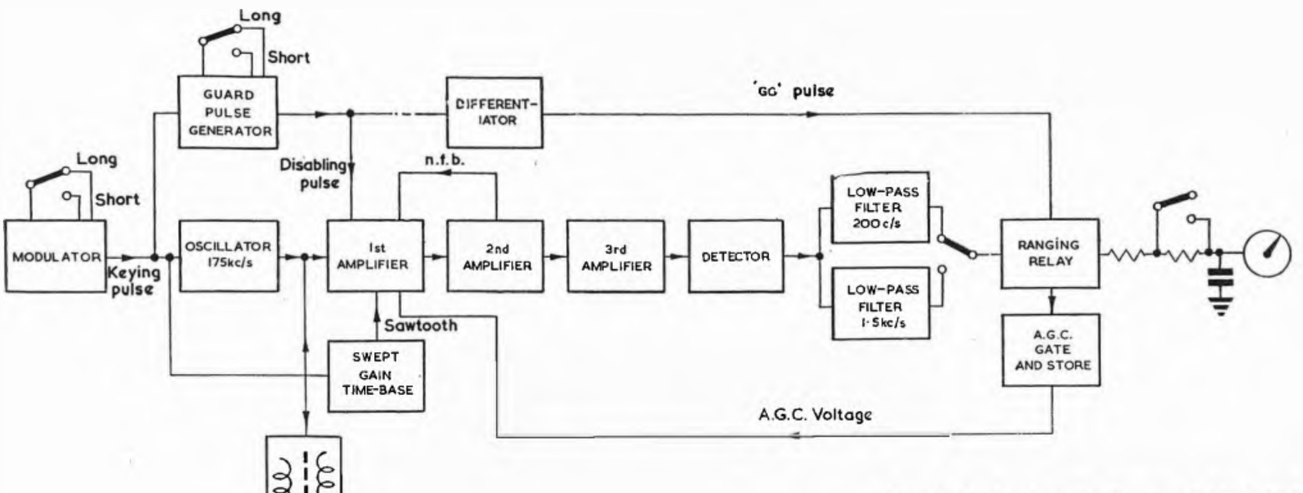


Fig. 2. Block schematic diagram of 'HECTA'. The four switches are ganged and are shown in the 'fathoms' setting

The duration of the guard pulse is changed from 1·1 to 6·7msec as the range switch is changed from feet to fathoms, these periods corresponding to the minimum depth readings on the two scales. When disabled, the first amplifier presents an open-circuit to the transmitting oscillator, and when the oscillator is switched off it, in turn, presents an open-circuit to the first amplifier.

The modulator also initiates a time-base, the purpose of which is to vary the gain of the second amplifier, so providing the 'swept gain' effect mentioned previously.

The output of the detector is fed via a low-pass filter with a cut-off frequency of 200c/s on the fathoms range, and 1 500c/s on the feet range, to the range-measuring relay. This is a bistable circuit which is driven into one of its two states by the leading edge of each transmitted pulse and is reset by the leading edge of the echo. This action is caused to start and stop a constant charging current which flows into a reservoir capacitor. Consequently the charge stored per cycle is proportional to depth. The reservoir discharges through the milliammeter at a rate proportional to stored voltage, and so drives a current through it which, when equilibrium has been attained, is also proportional to depth. When changing range from feet to fathoms the constant charging current is reduced six-fold. The accuracy of the system depends upon the constancy of the pulse repetition frequency, and of the charging current, both of which are controlled by high stability components and made practically independent of transistor leakage currents. The meter accuracy is ± 1 per cent, and an overall accuracy of ± 3 per cent is attainable over the temperature range $-20°C$ to $+50°C$. The meter has a dial 2½in in diameter. The movement travels through 260°, giving a scale of 5ft to the inch on the 'feet' range.

When the on-off switch is set to an intermediate test position the meter monitors the voltage applied by the battery to the range-measuring circuits. This voltage is kept at the correct value to within $\pm 0·5$ per cent by operation of a control knob, notwithstanding the gradual decay of battery voltage with use.

Provision is made for the connexion of an external meter to the instrument, which is plugged into the upper socket on the front panel, so enabling the depth to be observed both by the navigator and the helmsman. For survey work the second meter may be of the pen-on-paper recording type, with clockwork drive. Fig. 3 is a reproduction of a short length of chart from an Elliott recording milliammeter taken while the boat was proceeding southwards at 6 knots across the Lymington Banks in the West Solent. The depth scale is 1ft per division, and undulations of only a few inches are clearly seen. The paper speed was ¼in/min, giving a vertical scale of 100 yards per division. The steep descent into the shipping channel is seen at the top of the chart, followed by a change of scale from feet to fathoms.

The Mechanical Design of 'HECTA' (see Figs. 4 and 5)

The Transducer

This is contained in a tube of 'fibreglass' 3·5in long and 1·2in in diameter, which houses also a transformer that matches the oscillator output to the resonant resistance of the transducer disk. The complete assembly is 'potted' in epoxy resin. A brass top-cap contains synthetic rubber O-rings, fitted into grooves, which create a tight seal when the unit is pushed into its tubular gunmetal housing. This housing is fitted in the bilge of the boat, the tube passing through a 1⅝in diameter hole in the planking. A screw-cap is provided to seal off the top of the tube when the transducer unit is not in place.

The Instrument

This is housed in a hermetically sealed case of silicon-aluminium alloy 8in wide, 4⅞in high and 2⅞in deep. This is protected against corrosion by a film of nylon which is deposited by a fusion process. The meter window is

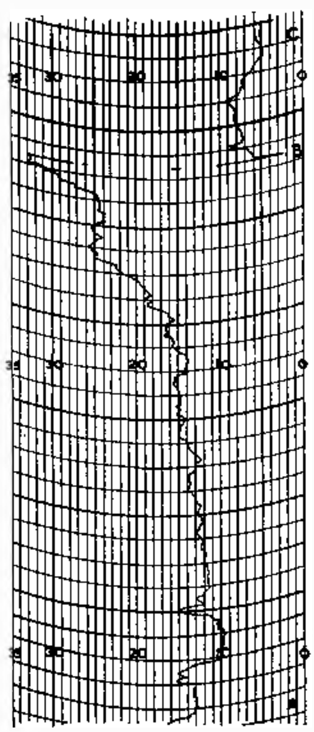


Fig. 3. Trace from the Elliott Recorder in conjunction with 'HECTA'

Horizontal scale is in feet from A to B; fathoms from B to C. Vertical scale is 100yd/division of toughened glass. Its rim, the control shafts and sockets are all fitted with sealing glands, and a removable desiccator with indicator window is provided to prevent internal condensation in case of seal failure. The controls are recessed into the case to protect them from damage.

The battery comprises four Mallory RM.522R hearing-aid cells and is housed in a compartment in the rear cover of the case. This compartment is isolated from the rest of the instrument, so that failure to secure its cover-plate properly can at the worst result only in a wetting of the cells. The instrument is fixed to a bulkhead, when necessary, by means of a back-plate with spring clips which engage with studs on the sides of the case.

The instrument is designed to withstand shock loading up to values of 40g.

The Final Specification

(a) Maximum range: 32 fathoms over mud.
(b) Power intake: 0·1W at maximum range.
(c) Battery supply: 4·8 to 6·0V (four hearing-aid 'A' cells, in instrument case).
(d) Battery endurance: This follows from (b) and (c) and is 150 hours (average) when using Mallory RM.522R cells.
(e) Accuracy: ±3 per cent or 6in, whichever is the greater, over the range of ambient temperature between −20°C and +60°C, subject to correction for extremes of sea temperature and salinity.
(f) Scale of display: 5ft per inch of scale on shoal setting. Scale and pointer are luminized for night use.
(g) 'Repeater': A distant meter of any size may be connected to the instrument, and this may be of the pen-on-paper recording type, clockwork driven, for survey use.
(h) Weight: 5½lb, including transducer.

Appendix

Derivation of an Expression for Echo Power

Assumptions and Approximations:—

(1) Reflection from the sea bed is specular, as has been verified experimentally for mud and sand. Rocks scatter energy in all directions, but their high reflectivity appears to make up for the losses by scattering.

Fig. 4. Rear view of 'HECTA', showing the battery with cover-plate removed. The screw driver control allows the zero of depth measurement to be set to the water-line plane

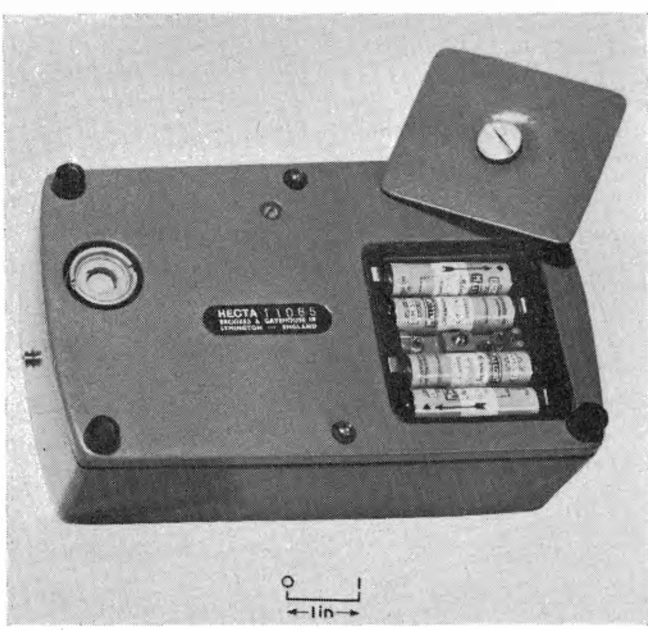


Fig. 5. The complete installation including the transducer housing which is fitted in a hole in the ship's bottom. The transducer is retractable for cleaning when the boat is afloat

(2) The conical beam angle, θ, is taken to be small enough that the cross-sectional area of the beam at a range r may be written as

$$\frac{\pi\,(r\,.\,\theta)^2}{4}$$

The errors introduced by these approximations are probably small in relation to those due to the uncertain knowledge of the range of the "constants" K and A in different localities. The results obtained by calculation, as shown in Fig. 1, are therefore taken only as a guide to design, and a fair margin of extra performance is provided for.

Using the symbols listed on page 338, and in addition letting:—

d = diameter of transducer face
A = sound-wave attenuation in nepers per unit distance at a given frequency, f

the total pulse power reaching the horizontal plane through the transducer, after reflection from the sea-bed, is:—

$K\,.\,P_t\,.\,e^{-4Ah}$, the total distance traversed being $2h$.

The sea-bed reflects the cone of energy as a mirror. Hence the cross sectional area of the cone at the horizontal plane through the transducer is

$$\frac{\pi\,(2h\,.\,\theta)^2}{4} = \pi\,.\,h^2\,.\,\theta^2 \text{ sq. units}$$

$$\therefore P_r = \frac{K\,.\,P_t\,.\,e^{-4Ah}}{\pi\,.\,h^2\,.\,\theta^2}\cdot\frac{\pi\,.\,d^2}{4} \quad \ldots\ldots\ldots \text{(1)}$$

θ is given approximately by $12/f\,.\,d$ radians, where f is in kc/s and d is in feet.

$$\therefore d = (12/f\,.\,\theta) \quad \ldots\ldots\ldots\ldots\ldots \text{(2)}$$

It is known[1] that A varies as f^2 and that at 100kc/s $A = 0{\cdot}369 \times 10^{-4}$ nepers/cm.

Hence $A = 1{\cdot}12 \times 10^{-7}\,.\,f^2$ nepers/ft .... (3)

Substituting for A and d in equation (1):—

$$P_r = \frac{36\,K\,.\,P_t}{h^2\,.\,f^2\,.\,\theta^4 \exp\,(4{\cdot}48 \times 10^{-7}\,f^2\,.\,h)}$$

Acknowledgment

The author is indebted to his colleague, P. Bailward, for his assistance in sea trials and design work and in computing the performance data for Fig. 1.

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A Passive Waveform Shaping Circuit

By M. J. Wright*, B.Sc.

A simple circuit is described which is able to perform a variety of waveform shaping functions. Although no active components are used, sharp transitions in output voltage can be obtained which are delayed in time with respect to the input voltage. Among the applications are: (1) Pulse lengthening, (2) 90° change in phase of a square wave, (3) combined filtering and limiting, (4) mark-to-space modulation of a square wave input.

(Voir page 394 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 400)

THE combination of an inductance together with one or more suitable semiconductor diodes produces a simple circuit capable of performing a number of diverse functions in waveform shaping or pulse delay circuits.

The principles underlying these applications are:—

(a) The current flowing through the inductance rises or decays exponentially towards an asymptote determined by the instantaneous voltage across it.

(b) The diode or diodes give a change in output voltage when the current through them decays to zero or commences to flow in a previously non-conducting diode.

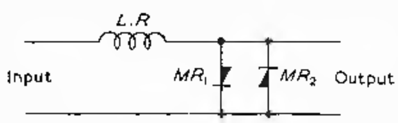


Fig. 1. Circuit for 90° change of phase

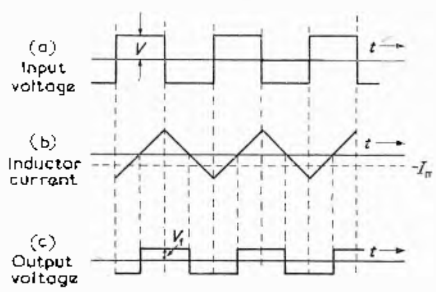


Fig. 2. Waveforms for circuit of Fig. 1 (Change of phase application)

The applications to be discussed are:—

(1) 90° Change in phase of an applied square wave.
(2) Mark-to-space modulation of an applied square wave.
(3) Filtering-limiting of a complex waveform.
(4) Pulse Lengthening.
(5) Pulse Delay.
(6) Pulse Area Discrimination.
(7) Single or Multiple Pulse Generation.

Phase Changing

Consider the application of a square wave of even mark-to-space ratio to the input of the circuit shown in Fig. 1.

Let $\pm V$ = Input voltage amplitude

V_o = Instantaneous output voltage amplitude

V_f = Forward voltage drop of diodes MR_1 and MR_2.

If $V \gg V_f$ the voltage $\pm V$ can be considered to exist across the inductor only since the voltage at the output terminals cannot exceed V_f in magnitude. The current through the inductor therefore rises exponentially towards the asymptote $+V/R$ during one half cycle and decays towards the asymptote $-V/R$ during the opposite half cycle. However, if the time-constant L/R of the inductor is long compared with the period of the input waveform, the rise and fall of current may be considered practically linear. Figs. 2(a) and 2(b) show the time relationship between the input voltage and circuit current for this condition. The current passes through zero one quarter cycle after the input voltage transitions and at these times an abrupt change in output voltage occurs while the current switches from one diode to the other. Thus, subject to the approximations made, the output voltage is a square wave displaced by 90° with respect to the input waveform (Fig. 2(c)). Also, the phase change is constant over the frequency range for which the approximations are valid.

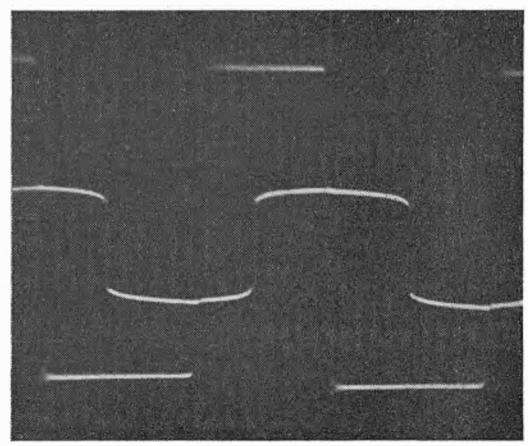


Fig. 3. Input (large amplitude) and output waveforms for phase change circuit

Diodes MR_1 and MR_2 should be low power silicon junction types to give the best output waveshape.

In practice the phase shift will be just less than 90° but, for any given conditions of operation, a suitable capacitance can be added in parallel with the diodes to give the additional delay required for a total of 90°. The additional delay is produced at the expense of output rise-time.

Fig. 3 shows the input and output voltage waveforms for the circuit conditions given below.

L	=	0·3H
R	=	15Ω
MR_1, MR_2	=	Silicon junction diodes
Frequency	=	1kc/s
V	=	±15V
V_o	=	±0·6V

Mark-to-Space Modulation

By feeding a bias current I_m from a subsidiary circuit through diodes MR_1 and MR_2 the mark-to-space ratio of the output voltage is altered. The inductor current remains essentially as before but the transition in output voltage now occurs when the sum of the inductor current and the bias current is zero, i.e. when the current through the inductor is $-I_m$. Reference to Fig. 2(b) will show the

* *Group Research Centre, Joseph Lucas Ltd.*

change in mark-to-space ratio which occurs for any given bias current.

To avoid excessive lengthening of the output voltage rise-time the bias current should be supplied from a high impedance source. Where the modulating signal is a volt-

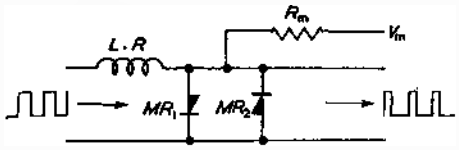


Fig. 4. Mark-to-space modulation circuit

age, as in Fig. 4, it should be of large amplitude so that the resistor R_m may be large. Alternatively, a blocking capacitor can be used at the input and the modulating voltage applied via resistor R_m to the junction between the capacitor and inductor (see Fig. 9). This avoids loading the output terminals with resistor R_m but introduces time-constants because the diode bias current is applied via

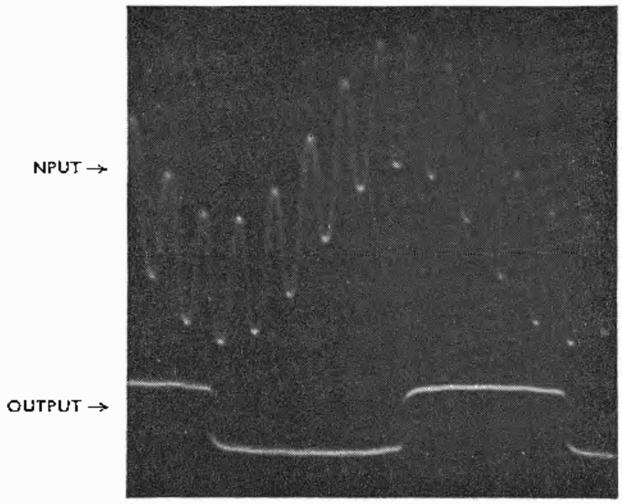


Fig. 5. Filter-limiter circuit waveforms

reactive elements. This arrangement is therefore more suitable where the modulating frequencies are low compared with the carrier frequency.

Filter-Limiter

Most systems of frequency measurement require the input waveform to be either—

(a) An amplitude limited square wave

(b) A pulse train.

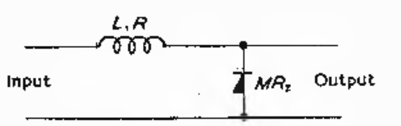


Fig. 6. Pulse lengthener

In the majority of cases the signal of interest is in neither of these forms and filters, amplifiers, limiters and differentiating circuits are commonly used preceding the frequency determining network. The circuit of Fig. 1 can sometimes be employed to advantage in a combined filter-limiter role producing a square wave output of frequency equal to the fundamental of a complex input waveform. The inductor in itself forms a low-pass LR filter, where the input is the voltage across it and the output is the current flowing. The diodes sense the direction of current flow and provide limiting. Fig. 5 is typical of the results obtained.

Where the input waveform is of suitable amplitude, diodes MR_1 and MR_2 may be replaced by a Zener diode to give improved amplitude limiting[1].

Differentiation of the square wave will produce a train of pulses suitable for operating counter circuits in digital frequency measurement systems.

Pulse Lengthener

Consider a rectangular pulse of amplitude $+ V$ applied to an inductor and Zener diode in series (Fig. 6). Assuming that V is greater than the reverse breakdown voltage V_r of the Zener diode the output voltage follows the leading edge of the input pulse until limited by conduction of the Zener diode. Thus, for almost the entire duration of the input pulse, a voltage $(V\text{-}V_r)$ exists across the inductor causing an exponential rise in current. At the end of the input pulse, the voltage $- V_r$ appears across the inductor since the input voltage is zero and the output voltage is still V_r (the inductor prevents a rapid fall of diode current). The inductor current now decays exponentially. It is important to note that this decay is towards the asymptote $- V_r/R$ and not zero. Thus there is a definite rate of change of current as zero is approached. However, when the current reaches zero the diode cuts off and both the output voltage and the voltage across the inductor fall to zero. Fig. 7(b) shows the circuit current for the input voltage shown in Fig. 7(a) applied to the case where the inductor

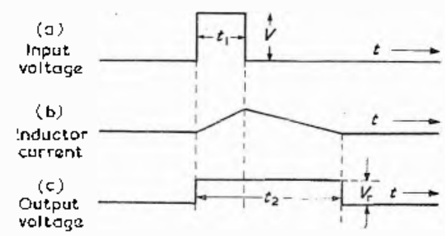


Fig. 7. Waveforms for pulse lengthener

time-constant is large compared with the pulse length t_1. An almost linear rise and fall of current is therefore obtained. The output voltage is shown in Fig. 7(c).

Where linear changes of current can be assumed,

$$(t_2/t_1) = (V/V_r)$$

Any practical inductor will have some self-capacitance which will cause ringing to occur when the Zener diode cuts off. A resistance of suitable value connected in parallel with the Zener diode can be used to give critical damping of the tuned circuit. For best results, sectional winding of the inductor should be used to give minimum self-capacitance.

Fig. 8 shows typical input and output waveforms for the circuit.

Pulse Delay

Applying a bias current to the circuit of Fig. 6 will delay the leading edge of the input pulse and so give pulse delay. The current is best applied via the inductor as shown in Fig. 9.

In the standing condition Zener diode MR_2 is conducting on its forward characteristic, the value of current flowing being given by the equation:

$$I_m = (-V_m/R_m)$$

subject to the approximations

$$V_m \gg V_f \quad R_m \gg R$$

On application of a rectangular pulse of amplitude $+ V$ to the input terminals (Fig. 10(a)) the voltage V appears across the inductor (the voltage drop $R \, . \, I_m$ can usually be ignored). The inductor current rises exponentially toward the positive value V/R. When the inductor current passes through zero the diode is switched into conduction on its reverse characteristic and the output voltage changes from $-V_f$ to $+ V_r$. The inductor current continues to rise, but towards a less positive asymptote. At the end of the input pulse the voltage across the inductor becomes equal to $-(V_f + V_r)$ and the current decays towards the value $-(V_f + V_r)/R$. The diode cuts off when the current becomes zero and the output voltage falls practically to the initial value $-V_f$ (see Fig. 10(c)). Where the pulse period t_1 is short compared with the time-constant L/R these current changes would be linear as shown in Fig. 10(b). The Zener diode current gradually attains its initial value, the time taken depending on the time-constants L/R and CR.

Where the time between pulses is long the circuit reacts to each one as described above. However, if the period between pulses is of the same order or shorter than the time-constants L/R and CR, the leading and trailing edge delays will both be a function of the time between pulses. The circuit is not recommended where the pulses may be close together with a variable spacing.

If, in the circuit of Fig. 9, the diode is replaced by a normal silicon diode the leading edge of an applied pulse is delayed by the circuit but the trailing edge of the output would be time coincident with that of the input.

Pulse Area Discriminator

The leading edge delay circuit may also be used as a pulse area discriminator since no output can be obtained if the inductor current is still negative when the input pulse terminates. Since the rate of change of current in the inductor is proportional to the applied voltage, the total change in current ΔI caused by a voltage V applied for a time t_1 is given by

$$\Delta I = (V \, . \, t_1/L)$$

If, for a given pulse, ΔI is less than the initial bias current I_m in amplitude, the diode remains conducting on its forward characteristic and no pulse appears at the output. Thus

$$V \, . \, t_1 < L \, . \, I_m \text{ no output}$$
$$V \, . \, t_1 > L \, . \, I_m \text{ output obtained}$$

If the pulse amplitude V is fixed the circuit is a pulse length discriminator rather than a pulse area discriminator.

The remarks concerning pulse spacing when the circuit of Fig. 9 is used for pulse delay apply also when it is used as a pulse area discriminator.

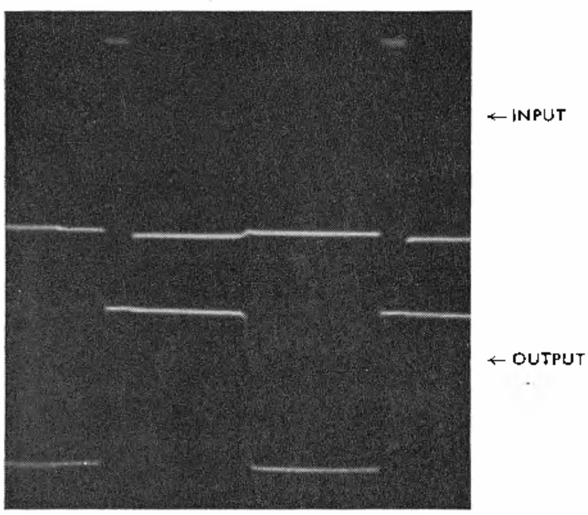


Fig. 8. Pulse lengthener circuit waveforms

Single or Multiple Pulse Generator

On closing switch S in Fig. 11, the Zener diodes MR_{z1}, MR_{z2}, MR_{z3} are brought into conduction on their reverse characteristics and the branch currents increase exponentially towards values I_1, I_2 and I_3. If, after some time, the power supply is removed by opening switch S, diode MR is brought into conduction by the back e.m.f. developed across the inductors. The branch currents now decay exponentially until the Zener diodes cut off.

Fig. 9. Pulse delay circuit

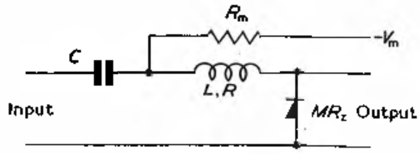
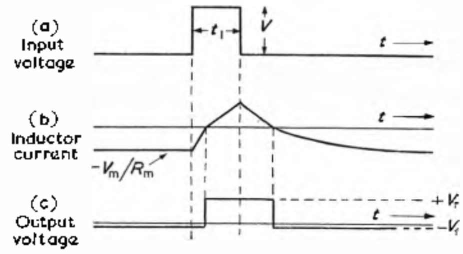


Fig. 10. Waveforms for pulse delay circuit



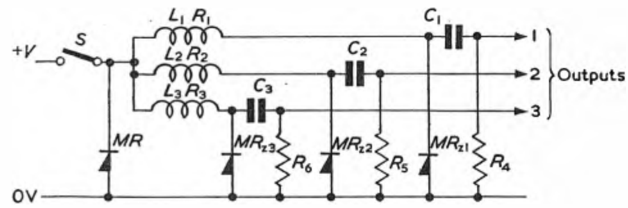


Fig. 11. Pulse generator

Considering the branch containing diode MR_{z1}, the final current after closure of S is given by

$$I_1 = \frac{V - V_{r1}}{R_1} \text{ where } V_{r1} = \text{Reverse breakdown voltage of } MR_{z1}$$

On opening S the current decays with a time-constant L_1/R_1 towards the asymptote

$$I = \frac{-(V_{r1} + V_f)}{R_1} \text{ where } V_f = \text{Forward voltage drop of } MR$$

When the current reaches zero MR_{z1} ceases to conduct and a short differential pulse is generated at output terminal 1 while C_1 discharges through R_4, the inductor and diode MR.

Use of different circuit constants in the branch elements gives a series of time displaced pulses, each at a different output terminal.

To stabilize the time periods against variations in supply voltage diode MR can be a Zener diode and a limiting resistor placed between $+V$ and the switch S.

In place of a d.c. supply and switch S positive rectangular pulses can be applied to the input to give groups of time displaced pulses.

Other Applications

The circuit of Fig. 9 can be used to generate a reasonably sharp rectangular pulse when fed by a pulse of poor shape, while the circuit of Fig. 1 will adjust an uneven mark-to-space rectangular wave input to an even mark-to-space ratio square wave output. Fig. 3 does in fact illustrate this since the input is not of even mark-to-space ratio.

Acknowledgment

The author's thanks are due to the Directors of Joseph Lucas Ltd for permission to publish this article and Messrs. D. W. Morris and P. Smith for helpful discussions.

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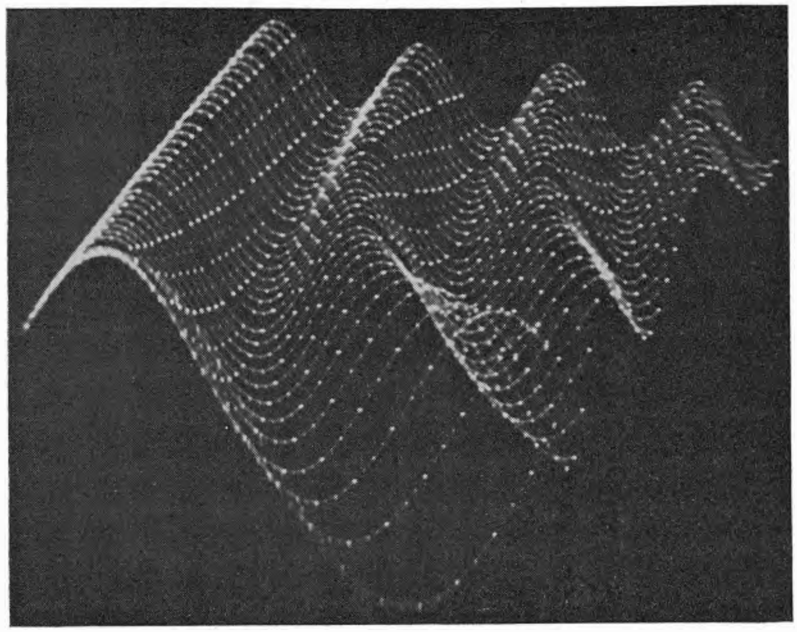
A Simple Multi-Dimensional C.R.T. Display Unit



The above photograph shows a typical three-dimensional display

By D. M. MacKay*, B.Sc., Ph.D.

Functions of two or more variables can be displayed on the screen of a c.r.t. as a two-dimensional projection of the hypersurface they define. The angle of projection may be altered by rotating potentiometers which determine the proportions in which the different input voltages contribute to the X and Y deflexions.

While this is best done with the help of sine-cosine potentiometers having four sliders, and matching amplifiers, a surprisingly satisfactory result can be achieved with much simpler circuits, requiring only two-ganged linear potentiometers and a few carefully chosen resistors. The outputs for a fully stereoscopic display can also be derived at the cost of little additional circuit complexity.

(Voir page 394 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 400)

'PROJECTIVE' methods of displaying functions of more than one variable on a cathode-ray tube were first developed during the last war for radar purposes[1]. In 1949 the writer described in this Journal[2] an adaptation of the same principle, developed for displaying sequentially the solutions of equations in a high-speed analogue differential analyser[3], and the reader is referred to the foregoing and other references[4,5] for some examples of the many other applications which are possible.

In most of these the apparatus described for transforming from three to two dimensions has been fairly complex, comprising several amplifiers and either magslip resolvers or four-gang sine-cosine potentiometers. The present brief article describes a 'utility' version of the apparatus described earlier[2,3], devised to reduce to a portable minimum the equipment needed to give an acceptable three-dimensional display.

Basic Principles

The basic principles of projective display may be recapitulated in terms of Fig. 1. Looking along any axis, Y say, of a three-dimensional reference system, one has a 'plan view' of the other two axes X and Z. If, however, the line of sight is rotated to a position half-way between X and Y axes in the XY plane, the view of the Z axis will be unchanged, but the other direction of the field of view will include some of X and some of Y.

In mathematical terms what one sees is the projection of X and Y axes on the plane normal to the line of sight. The apparent horizontal co-ordinate x' of any point in the XY plane will thus be the difference between the projections of its true co-ordinates x and y on the plane of sight. In general, (Fig. 2), when the X axis has been rotated through an angle θ from the position of Fig. 1(a), i.e., when the Y axis makes an angle θ with the line of sight, the visible co-ordinate x' of a point P (x, y) is given by:

$$x' = x\cos\theta - y\sin\theta \quad \ldots\ldots\ldots\ldots \quad (1)$$

The corresponding 'invisible co-ordinate' y' is given by:

$$y' = x\sin\theta + y\cos\theta \quad \ldots\ldots\ldots\ldots \quad (2)$$

Assume now three voltages, x, y and z, of which z varies with x and y so as to describe a three dimensional pattern. Clearly all one has to do to obtain a view along the y' direction is to mix the X and Y voltages in the proportions $\cos\theta$ and $1 - \sin\theta$ before applying the resultant to the X-plates of the c.r.t. In fact, if a fraction $x\cos\theta$ is applied to one X-plate, and $y\sin\theta$ to the other, the tube will subtract them automatically and give the required x' deflexion to the spot according to equation (1). The result will be a view of the pattern described by x, y and z, from ground level as it were, and at an angle θ to the Y axis.

More generally, the line of sight may be at an angle oblique to all three axes. The rule which has been found for the simple case can easily be extended, and in its general form may be stated thus:

To obtain a projection on a plane in which the axes X', Y' are specified by direction cosines $(l_1m_1n_1)$ and $(l_2m_2n_2)$ relative to the X, Y, Z axes, mix the x, y, z voltages in the proportions of the direction cosines of X' and Y'. In other words make:

$$x' = l_1x + m_1y + n_1z \quad \ldots\ldots\ldots\ldots \quad (3)$$

$$y' = l_2x + m_2y + n_2z \quad \ldots\ldots\ldots\ldots \quad (4)$$

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and if z' is required:

$$z' = l_3x + m_3y + n_3z \quad \ldots\ldots\ldots\ldots \quad (5)$$

where $(l_3m_3n_3)$ are the direction cosines of the line of sight.

Rotation

In everyday life, the two-dimensional projection of the solid world on the retina of the eye is given depth in a number of ways. The tension in the focusing muscles, the convergence of the eyes and the different views seen in binocular vision, and, of course, long experience of familiar objects, all help one to distinguish between a small object near at hand and a large one farther away. Sometimes, however, even these are insufficient, and one has to take the trouble to move one's point of view to see another aspect of the field, from which the ambiguity is resolved. This important facility of relative rotation—so commonly used, for example, in 'parallax' methods of optical measurement—is fortunately very simple to secure in projective displays.

There are many ways in which one may choose to rotate the pattern, but one of the most convenient methods is to carry out the selection of aspect in two independent stages. The first corresponds to the transformation that was considered (equations (1) and (2)), which effectively rotates the pattern about the Z axis. With ganged attenuator-controls to mix continuously-variable proportions of x and y according to equations (1) and (2), any angle of view (θ) can be selected at the turn of a knob.

The second stage of aspect-selection can conveniently correspond to rotation about the X' axis, i.e. choice of the angle of elevation (ϕ) of the line of sight. The principle is the same as before, but in this case z must be combined with the voltage y' which is synthesized in the first operation (equation (2)) so as to form:

$$z' = z \sin\phi + y' \cos\phi \quad \ldots\ldots\ldots\ldots \quad (6)$$

and if desired:

$$y'' = z \cos\phi - y' \sin\phi \quad \ldots\ldots\ldots\ldots \quad (7)$$

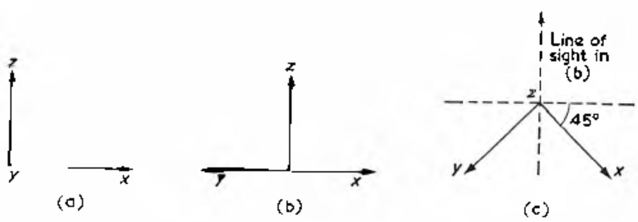


Fig. 1. (a) The view along Z axis
(b) The view at 45° to X axis
(c) The view of (b) along Y axis

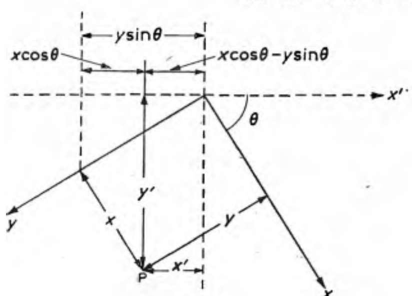


Fig. 2. Rotation of XY plane through angle θ

Here y'' is the 'invisible' co-ordinate normal to the face of the c.r.t. Its use in producing perspective was discussed in the earlier paper[2] and will be mentioned below.

These two facilities can together remove any structural ambiguities present in a given projection, since any surface that remains edge-on under one rotation is rendered visible by the other.

Approximating to a Sine-Cosine Law

As indicated in the paper referred to earlier[2], it is possible to derive an approximation to a sine (or cosine) law by suitably modifying a circuit incorporating a linear potentiometer. Fig. 3 shows the simplest arrangement, in which a

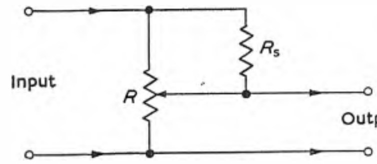


Fig. 3. '3 point' approximation to a sine law

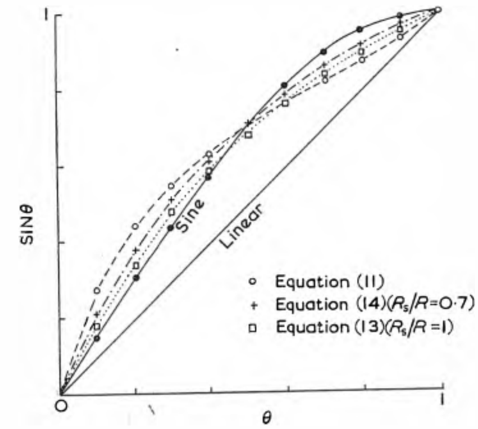


Fig. 4. The performance of circuits in Fig. 3 and Fig. 5 compared with the ideal sine law and the original linear law

resistance R_s links the slider to the live terminal. If the potentiometer is linear, the ratio at the half-way position will be:

$$V_2/V_1 = (2R_s + R)/(4R_s + R) \quad \ldots\ldots\ldots \quad (8)$$

where R is the resistance of the element. Since a full rotation of the potentiometer must turn the projection on the screen through an angle of 90°, it is desirable that half-rotation should give a ratio:

$$V_2/V_1 = \sin 45^\circ = 0{\cdot}707$$

Substituting in equation (8), it is found that R_s should be approximately:

$$R_s = 3/8\,R \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (9)$$

The accuracy of this approximation may be easily calculated. Taking units of $\theta = 90^\circ$, the ratio V_2/V_1 at an angle θ is:

$$V_2/V_1 = \frac{(1 + (1-\theta)\,R/R_s)\theta}{1 + \theta(1-\theta)\,R/R_s} \quad \ldots\ldots\ldots \quad (10)$$

When $R/R_s = 8/3$, this gives:

$$V_2/V_1 = \frac{(11 - 8\theta)\theta}{3 + 8\theta\,(1-\theta)} \quad \ldots\ldots\ldots\ldots \quad (11)$$

The comparison in Fig. 4 shows that the greatest errors occur around $\theta = 0{\cdot}2$ and $0{\cdot}8$, and do not exceed 13 per cent of maximum.

A 'Utility' Scheme

The above arrangement has the advantage that the full input is transferred to the output when $\theta = 1$ (90°). As with the normal sine-cosine potentiometer, however, four sliders must be ganged together to provide a full transformation of the type in equations (1) and (2).

Fig. 5 shows a 'utility' form of circuit which enables two sliders to do the work, at the cost of some overall attenuation. Linear potentiometer elements are again used, but the slider is earthed, and the ends are supplied through resistances R_s chosen so as to make the output again 0·707 of maximum at the halfway point. (R_s must, of course, include the source impedance.)

For a simple resistive feed, the value of R_s is easily found. The output/input ratio V_2/V_1 at the halfway point is:

$$(R/2)/(R_s + R/2) \text{ or } R/(R + 2R_s)$$

At maximum V_2/V_1 is $R/(R + R_s)$

Hence $R + R_s = 0{\cdot}707\,(R + 2R_s)$

or:

$$R_s \simeq 0{\cdot}7R \qquad (12)$$

With the same convention as before the ratio V_2/V_1 may be written in general as:

$$V_2/V_1 = \theta/(\theta + R_s/R) \qquad (13)$$

or with the above value of R_s (equation (12)):

$$V_2/V_1 = \theta/(\theta + 0{\cdot}7) \qquad (14)$$

This function (normalized for comparison) is also shown in Fig. 4, the biggest departure from the sine function being now about 7 per cent of maximum. If somewhat greater attenuation is tolerable, R_s may be set equal to R, enabling the maximum error to be reduced to some 3 or 4 per cent (Fig. 4).

A Complete Display Unit

The circuit of a complete three-dimensional display unit of the simplest possible kind on this 'utility' principle is

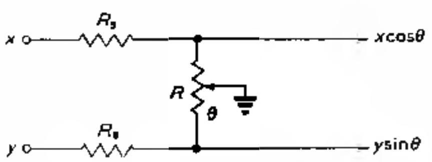


Fig. 5. A 'utility' circuit for approximation to a sine cosine law

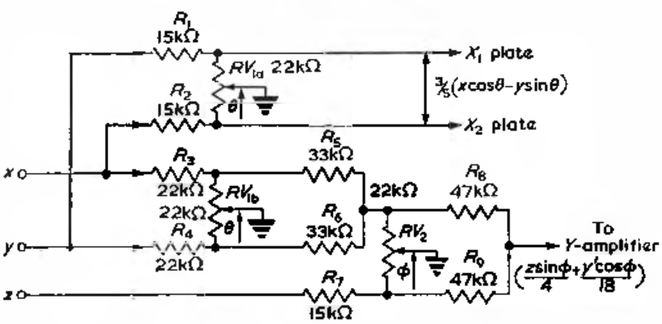


Fig. 6. The simplest three-dimensional display unit

shown in Fig. 6. Subtraction of $x\cos\theta$ and $y\sin\theta$ is achieved directly at the c.r.t. by applying the appropriate voltages to opposite plates. Addition is performed through mixing resistors such as R_5 and R_6, which place an additional load on the other two sine-cosine units. These resistors and the input resistors (such as R_3 and R_4) must therefore be chosen so that in conjunction they present an effective series input resistance of about $0{\cdot}7\,R$ both to the supplying and to the supplied potentiometer. The values shown in Fig. 6 are not accurately optimal, but are preferred values which give an acceptable performance. Altering the setting of one potentiometer will in any case slightly change the effective series input resistance to the other, so that greater precision is normally pointless.

The operation of the circuit should now be self-explanatory. Ganged potentiometers RV_{1a} and RV_{1b} determine the first rotation (θ) about the Z-axis. The output to the X-plates is approximately $(3/5)(x\cos\theta - y\sin\theta)$. A voltage proportional to y' (equation (2)) is supplied through R_5 and R_6 to RV_2, which also receives the input z through R_7. The output from R_8 and R_9 is proportional to $z\sin\phi + y'\cos\phi$ (equation (7)) and this is supplied (after amplification) to the Y-deflecting system. There is an overall attenuation by

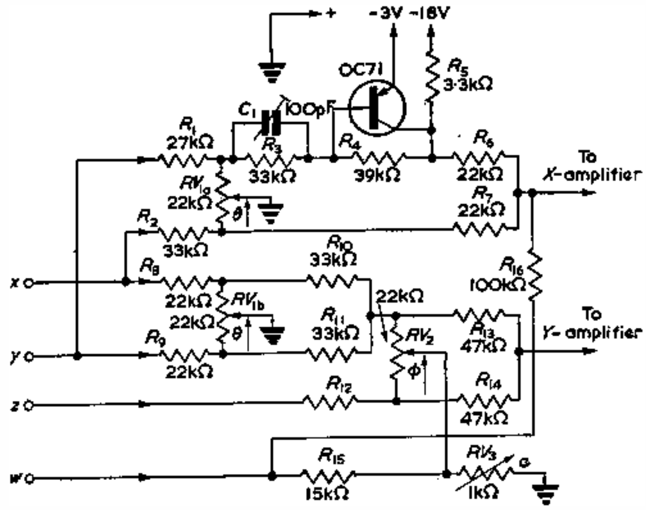


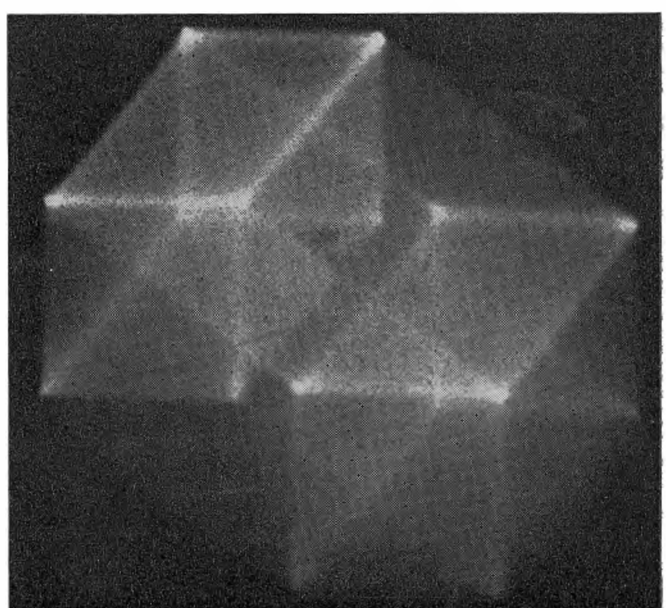
Fig. 7. Display unit with subtracting circuit and provision for a fourth input

a factor of about 1/18 from the x and y terminals, and about 1/4 from the z terminal, to the above output.

Equation (6) here represents a more practical form of output than equation (7), since the signals from RV_2 are too small for direct subtraction to be possible at the Y-plates, and addition is readily performed in resistors. The difference is only in the choice of projective quadrant, and the resulting picture is just as satisfactory as if the orthogonal component of equation (7) had been used. It will be noted that the scale of z is different from that of x and y.

In some oscilloscopes the X-plates may not be accessible, or their sensitivity may be inadequate to allow use of the simple arrangement of Fig. 6. In the circuit of Fig. 7 a transistorized subtracting stage has therefore been incorporated in the output from RV_{1a}. This supplies an output of approximately $(1/6)\,(x\cos\theta - y\sin\theta)$ to the X-deflecting system, the maximum undistorted output with the components shown being of the order of $\pm 3{\cdot}5$V, with inputs of ± 20V to the x and y terminals. The capacitor C_1 is adjusted to compensate for phase shifts due to the self-capacitance of the transistor.

Fig. 8. A four-dimensional "tessaract" photographed from the screen of a multi-dimensional c.r.t. display



Handling Four (or more) Dimensions

The circuit of Fig. 7 has a further feature, which enables yet another 'dimension' to be added to the display. It is in fact the most rudimentary facility possible of its kind, and is described here only in case the reader might enjoy contemplating the projections of a four-dimensional object at little cost. A more businesslike approach to the problem will be found in one of the papers cited[3].

Basically all that is required is a means of displacing the whole of a three-dimensional picture in a new direction, projectively independent of those displayed, and hence logically on the same footing. In Fig. 7 this is partially provided by the network R_{15}, R_{16}, RV_3, which

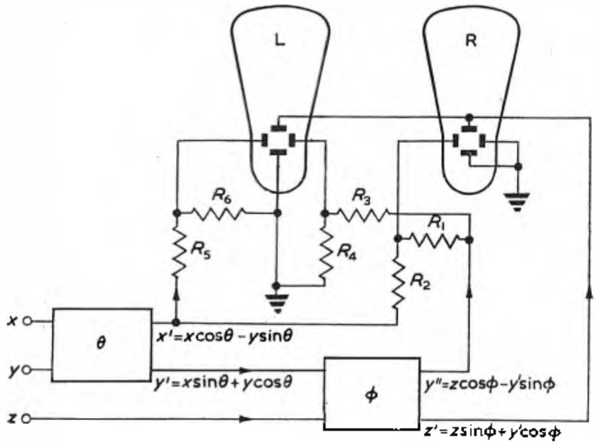


Fig. 9. Principle of stereoscopic display

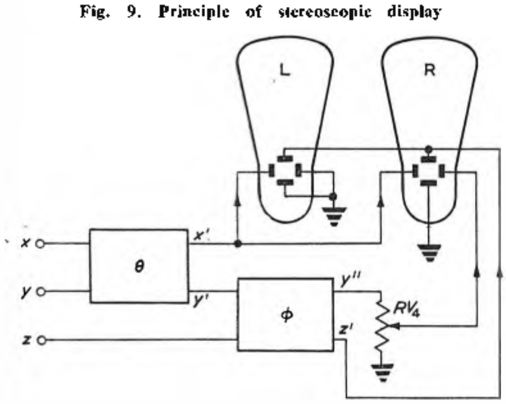


Fig. 10. Simplified approximately stereoscopic display

enables the voltage w to be mixed in controllable proportions with the inputs to X and Y amplifiers.

An impression of the result may be gained from Fig. 8, which shows a four-dimensional 'tessaract' (the four-dimensional analogue of a cube) formed by sweeping all four inputs asynchronously between fixed voltage-limits. Rotation of RV_1 now produces a pseudo-rotation of the figure about a plane (i.e., it leaves a plane invariant). Some conceptually elusive facts about a four-space—e.g. that it possesses planes which meet only in a point—are readily illustrated with this simple device.

Stereoscopic Display

The 'solidity' of a projective display is enhanced by nothing as vividly as by rotation of the controls. Various other aids, such as the modulation of brightness and sensitivity as a function of the depth-co-ordinate, have been discussed before[3]. It may be worthwhile in conclusion to refer again to the possibility of a fully stereoscopic display, as (apart from the additional tube) this requires remarkably few extra components. In stereoscopic vision each eye sees a

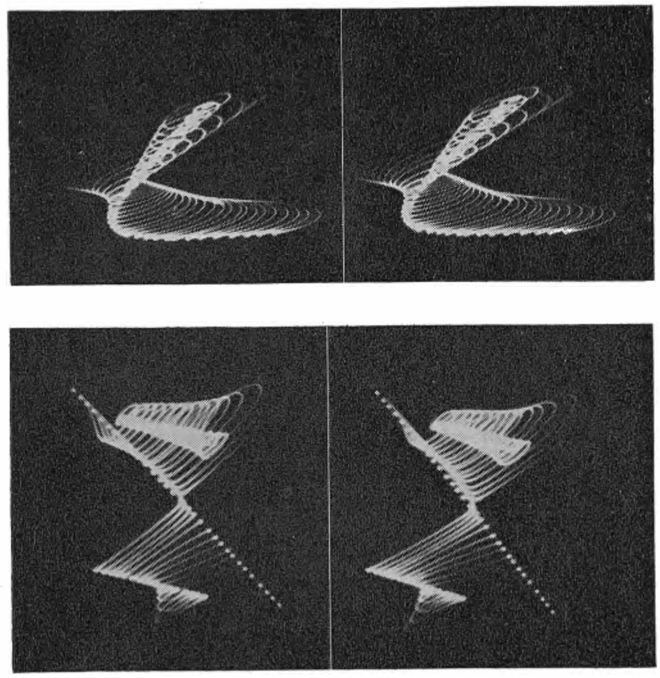


Fig. 11. Stereoscopic pairs from a three-dimensional display

projection at a slightly different angle. In the notation of equations (1) to (7), the x' co-ordinate seen by the left eye contains a small mixture of the depth-co-ordinate y'', while that seen by the right contains the same proportion of $-y''$.

It follows that with a single transforming unit of the type in Fig 6 or Fig. 7, a stereoscopic display can be achieved by using two tubes, one viewed by each eye, and simply adding to the X-deflecting voltage on each tube a small signal proportional (in this case) to $\pm(z\cos\phi - y'\sin\phi)$. This is readily secured by ganging a second potentiometer to RV_2, and feeding suitable fractions of its outputs $z\cos\phi$ and $y'\sin\phi$ either to opposing X-plates or to the corresponding subtracting circuit.

Fig. 9 illustrates the principle in a block diagram, where the rectangles marked θ and ϕ comprise two-gang transformation circuits of the general type shown in Figs. 6 and 7. The mixing resistors R_1 to R_4 are chosen to allow a small fraction of y'' to reach opposing X-plates in the two tubes. R_5 and R_6 enable the attenuation of x'' due to R_1 and R_2 to be matched.

If slight imperfection is tolerable, an even simpler arrangement is possible (Fig. 10) in which only one tube receives the input from y''; the degree of binocular disparity (stereoscopic depth) is then adjustable by means of RV_4.

Sample 'stereoscopic pairs' of photographs from a three-dimensional display (prepared by Dr. M. E. Fisher of King's College) are shown in Fig. 11. They may be viewed in any conventional stereoscope, or directly (after some practice) by allowing each eye to fixate the appropriate picture. To see such apparently 'solid' objects rotate in depth, as θ and ϕ are altered, however, offers a satisfaction all its own.

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A Graphical Analysis of the Blocking Oscillator

(Part 2)

By D. Chambers*, B.Sc.

(Voir page 326 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 333)

The Valve

Positive Grid Operation

The blocking oscillator is a very economical type of circuit for generating relatively high power pulses from small receiving type valves, but its use is sometimes restricted because of the variations in pulse shape which may occur when the valve is changed. In general, the types of valve employed in blocking oscillator circuits were originally designed for negative grid operation, and the characteristics in the positive grid region are therefore not controlled by the manufacturer. This explains why valves having identical characteristics under normal operating conditions will not necessarily produce identical pulse shapes in a blocking oscillator circuit. Experience has shown, however, that variations in pulse amplitude and duration can be minimized if the operating conditions of the valve are kept within reasonable limits. When the output pulse width is required to be independent of valve characteristics, a delay line must be employed to control the width of the pulse.

The tolerances to be expected in valve characteristics at low values of positive grid voltage (0V to +30V) are related to the manufacturers allowable valve spread in the negative grid region. At high positive grid voltages (greater than +75V), much wider variations occur, due primarily, to differences in peak cathode emission. Although high values of peak current may be a desirable requirement, the highest value of grid-cathode voltage that can safely be used is limited by cathode emission and grid dissipation and depends upon the valve type and the duty ratio. As a general rule, in low duty applications (less than 0·001) the peak value of grid voltage during the pulse top should be kept below +75V. For higher duty operation, correspondingly lower peak values of grid voltage should be used. The required operating conditions of the valve can be adjusted if necessary by a suitable choice of transformer turns ratio or by the inclusion of a series resistor in the grid, anode or cathode circuits.

Ruggedized valves which have the same negative grid characteristics as their commercial counterparts often give much lower peak cathode currents under positive grid operation. This may be due to a change in cathode material or a lower operating temperature of the cathode in the ruggedized types.

Excess and Peak Currents

The most important characteristic of valves used in blocking oscillator circuits is the value of excess current available during the pulse. The higher this value, the longer the pulse or the greater the useful power than can be supplied to a load connected across a winding on the transformer. In Fig. 12, the variations in excess current are shown for the cases when different valves replace the ½ 6J6 in the original circuit, and Fig. 10 shows the effect on the pulse duration and shape.

It will be noted from Fig. 11 that a peak cathode current greater than 500mA can be drawn from general purpose valves such as ½ 12AT7. Currents of this magnitude do not normally damage the cathode so long as the mean current is kept well below the manufacturer's rating. During valve life, the peak cathode emission falls off considerably, and it is recommended that allowance is made for this deterioration by limiting the peak cathode current during the pulse to less than 25 per cent of the peak cathode emission available from a new valve.

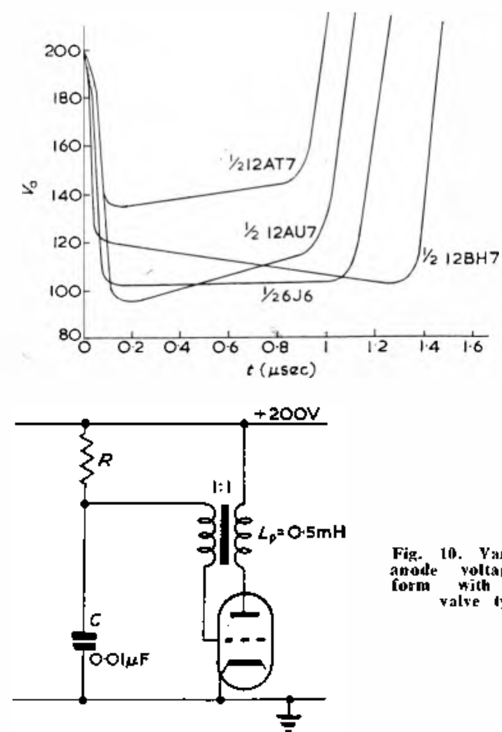


Fig. 10. Variation of anode voltage waveform with different valve types

Dissipation

Mean dissipation at the anode or grid is a product of pulse dissipation and the duty ratio. The pulse dissipation can be taken to be the product of the average current and average voltage at the electrode during the pulse top. The permissible grid dissipation is not normally given in data for receiving type valves, and its value depends primarily on the cooling arrangements and surface material of the grid wires. As a general rule, in low power blocking oscillator circuits the mean grid dissipation should be kept below 0·1W. Valves with fine grid wires require a smaller rating factor.

Choice of Valve Type

The choice of valve for a particular application depends primarily on the peak output power required in the load circuit. For the generation of low power pulses, the choice will be influenced by standardization of types, heater consumption, availability or price. General purpose valves such as the 12AT7, 12AU7 or triode strapped 6AM6 are satisfactory for peak pulse powers up to 15W. Pulses of higher power can be obtained from valves with bigger cathodes such as the 12BH7. When very fast rise-times

* Formerly Coss or Radar & Electronics Ltd.

are required, the most suitable types are those giving large excess currents and low values of grid current during the major part of the rise, for example, the EL81. Some of the best valves for high power applications are types originally designed for series stabilizer operation.

In general, optimum performance in blocking oscillator circuits is limited to valves having the following characteristics: (1) Good cathode emission. (2) Medium μ (15-40). (3) Moderately high g_m (4-10mA/V). (4) High value of i_a at zero bias and (5) low knee voltage on the anode characteristic in the positive grid region. Some of the newer special purpose valves, such as American types 5687, 5965 and 6463, have particularly favourable characteristics.

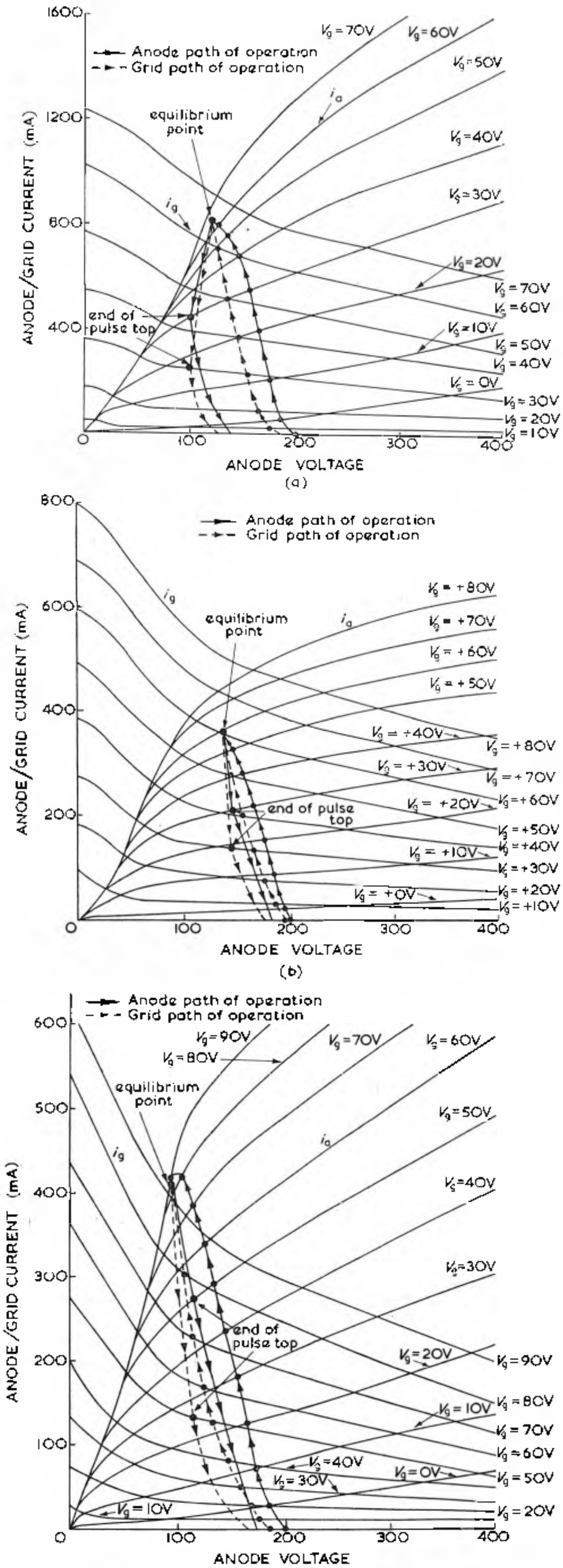


Fig. 11. Positive grid characteristics
(a) ½12BH7; (b) Average ½12AT7; (c) ½12AU7

Effect of Valve Type on the Pulse Shape

It was shown in Part 1 that the amplitude, slope and duration of a blocking oscillator pulse are controlled by a combination of three factors; the transformer pulse inductance, the value of the blocking capacitor, and the positive grid characteristics of the valve. Changing any one factor will affect the shape of the pulse, and in Fig. 10 waveforms have been constructed

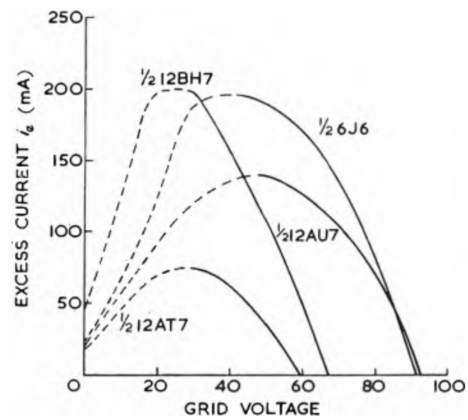


Fig. 12. Excess current during pulse top for different valve types

from the characteristic curves for the cases when the ½ 6J6 valve used in the original circuit (L_p = 0·5mH, C = 0·01μF) is replaced by ½ 12BH7, ½ 12AT7 or ½ 12AU7. The method of constructing the waveforms in these cases is slightly different from the method previously described because the values of L_p and C are already known and the anode waveforms will not necessarily be flat-topped. The equilibrium point is located on the valve curves as before, but the end of the pulse top cannot be found directly, and the following alternative method has been used to complete the paths of operation in Figs. 11(a), 11(b) and 11(c).

The approximate terminating point is first located by dropping a perpendicular from the equilibrium point and noting the grid voltage where maximum excess current occurs. Since the value of the transformer inductance and the approximate voltage across the primary are known, the time taken for the primary current to reach the value of maximum excess current can be determined. The average value of grid current during the pulse top can be estimated from the valve curves, and the voltage drop across the blocking capacitor C in this time is then known. If this voltage across C is greater than the grid voltage drop (between the equilibrium point and the terminating point) the terminating point occurs to the left of the perpendicular line at a lower value of anode voltage. This fall in anode voltage during the pulse top appears at the grid as a rising voltage which opposes the voltage built up across C and, by a little trial and error, a point can be found where the

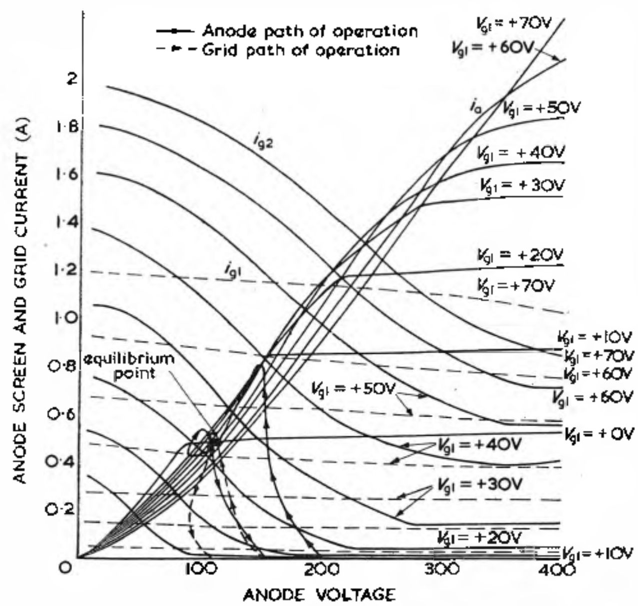


Fig. 13. Characteristic curves of average EL81 CV2721 positive grid region
$V_{g_3} = 0 \; V_{g_2} = +200V$

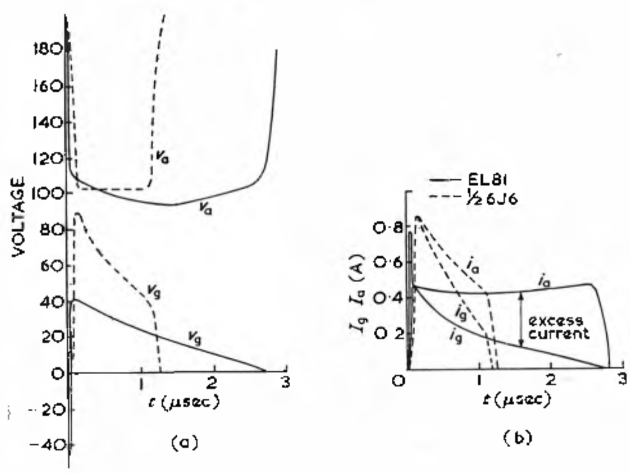


Fig. 14 (a) Voltage waveforms when EL81 replaces ½6J6
(b) Current waveforms when EL81 replaces ½6J6

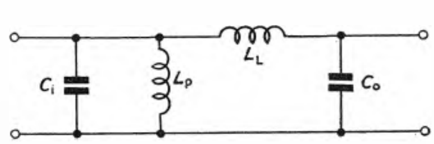


Fig. 15. Simplified equivalent circuit of transformer

circuit conditions are satisfied. If the voltage developed across C is less than the grid voltage drop, the terminating point is located to the right of the perpendicular. The approximate time of rise is established by taking the average value of excess current during the rise-time and substituting in $t_r = (C/i_{e(av)}) \,.\, V$, where V is the anode voltage drop.

Circuit Operation with Pentodes

The mode of operation has so far been described for a blocking oscillator employing a triode valve, but some types of pentode are more suitable in applications where high anode voltages must be used or when flat topped current pulses are required. Suitable pentodes include the EL81, EL360, CV2179, and CV2231. The operation of the circuit is influenced by the sharp knee and the bunching of the anode characteristics in the positive grid region (see Fig 13). At normal operating anode voltages, anode current is a maximum at relatively low values of grid voltage ($+10$ to $+30$V) and decreases slightly if the grid voltage is further increased. The pulse top tends naturally to be flat since the anode current is practically constant over a wide range of positive grid voltages, and the valve has a low anode impedance under these operating conditions. During the pulse top, the fall in grid voltage due to the charging of C causes both the excess current and the anode current to increase, and the point of maximum excess current occurs at a lower grid voltage than with triodes. It is therefore more efficient to operate pentodes at relatively low values of positive grid voltage in order to save power that would otherwise be wasted in the screen and grid circuits. When the anode is returned to a high voltage supply, a resistor should be connected in series with the grid to limit the peak grid voltage swing. Alternatively, a transformer with a large step down ratio can be used, but the circuit is then more difficult to trigger and the rise- and fall-times of the pulse are increased.

Figs. 13 and 14 show respectively the paths of operation and the waveforms when an EL81 replaces the 6J6 used in the original circuit. The construction of the pulse top waveform cannot be simplified in this case since the bunching of the characteristic curves makes the paths of operation unpredictable. In this circuit, $L_p = 0{\cdot}5$mH and $C = 0{\cdot}01\mu$F and the rate of increase in magnetizing current and the charging rate of C can be found at each point using values of $v_{t(av)}$ and $i_{g(av)}$ taken from the characteristic curves. For instance, during the first microsecond of the pulse top the average voltage across the transformer $v_{t(av)} = 90$V, and the magnetizing current therefore increases by 0·18A. This current is supplied by the excess current, which is equivalent to the drop in grid current (the anode current remains virtually constant during the increment). The value of grid current at the equilibrium point is 0·48A and the average value of grid current during the first microsecond is $(0{\cdot}48 - (0{\cdot}18/2))$A. From Fig. 13, the fall in grid voltage to give a 0·18A change in grid current is about 11V, but C charges by 39V, so the voltage pulse across the transformer must change by approximately (39-11)V. This drop in anode voltage reduces the available excess current to less than 0·18A, and v_g must in fact drop by more than the estimated 11V in order to give the required value of i_e. A little trial and error determines that the anode and grid voltage after 1µsec are 93V and 25V respectively. The step by step method has been used to construct the remainder of the pulse waveforms shown in Fig. 14. It will be seen that the anode current remains relatively constant during the pulse and that the pulse width is three times longer than with the ½ 6J6 in the original circuit.

The Transformer

The complete design of the transformer involves a large number of factors, many of which are based on experience, and in this article it is only possible to deal with those factors which are of special importance in blocking oscillator applications, namely, the pulse inductance L_p, the leakage inductance L_L and the effective shunt capacitance.

The simplified equivalent circuit of the transformer is represented in Fig. 15. C_i and C_o comprise respectively the shunt capacitance of the primary and secondary windings to the core and include the effective capacitance between the primary and secondary windings. Although this interwinding capacitance is smoothly distributed along the winding, for the purpose of the analysis it has been lumped with the shunt capacitances into two parts (C_i and C_o). This simplification does not normally introduce serious errors into the result, and in many cases when L_L is small, the circuit can be further simplified by putting the entire interwinding capacitance across the primary winding.

Pulse Inductance and Permeability

The pulse inductance L_p is the most important factor because it determines the magnetizing current flowing in the primary winding at the end of the pulse top ($E = L_p(\delta i/\delta t)$, where E is the pulse amplitude for δtsec). The required value of L_p can be obtained by applying the expression:

$$L_p = \frac{4\pi N^2 A K \mu_p}{l} \,.\, 10^{-9} H$$

where N = number of primary turns
A = cross-sectional area (cm^2)
l = mean magnetic path length (cm)
K = stacking factor
and μ_p = pulse permeability of the core material.

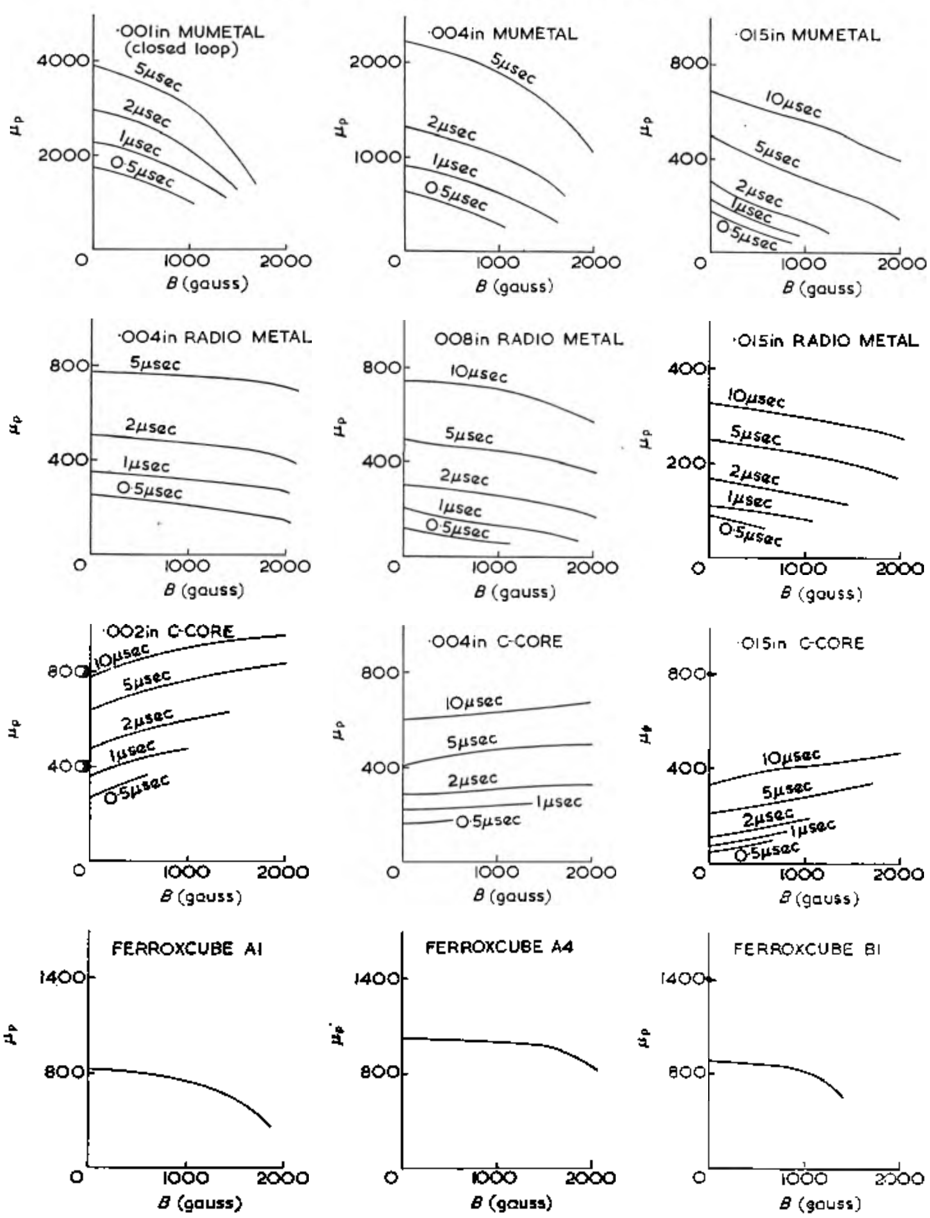
Pulse permeability may be defined as the ratio of the flux density change to the total magnetizing force required. Values of μ_p are not generally available, and those shown in Fig. 16 have been determined by the method illustrated in Fig. 17, using typical core assemblies. It will be seen that μ_p is not constant for a particular core material but varies widely with lamination thicknesses and to a certain extent with the duration of the applied pulse. A full description of the magnetization process in the core of a pulse transformer is outside the scope of this article, but the essential principles will be mentioned.

When a voltage pulse is applied to a transformer there is a change of flux in the core. The magnetizing force causing the change may be split into two components; one supplies the useful flux while the other component provides the energy for eddy currents and other losses which increase with the rate of change of flux in the material. When a series of rectangular pulses is applied to the transformer, the behaviour of the core and its operating point on the d.c. B-H loop depend on its state of magnetization at the onset of each pulse. In most blocking oscillator applications, the applied magnetizing force falls to zero between pulses. The core material settles down to a level of magnetization where the gain of flux during each applied pulse is equal to the loss of flux between pulses. Fig. 18 shows the build up of flux on the B-H curve when pulses are applied to an initially unmagnetized core. The pulse permeability μ_p is given by $\delta B/\delta H$. The actual operating region is a minor B-H loop and its location on the main loop is dependent upon the remanence of the core material and the air-gaps. The value of B at the start of each pulse is not easy to establish, being dependent upon air-gaps which are difficult to control in the assembly of the core. The flux swing δB is determined from the expression:

$$\delta B = (Et/NKA) \,.\, 10^8$$

where E = voltage across the primary winding
t = pulse duration
N = number of turns
K = stacking factor
and A = cross-sectional area of the core.

Fig. 16. Average values of μ_p for various cores



The available flux swings for typical core assemblies can be taken to be 1 000 gauss for Mumetal and Ferroxcube A, 2 000 gauss for Radiometal, and 4 000 gauss for orientated silicon steel in low duty ratio applications (less than 0·01). When the duty ratio is higher, it may be necessary to lower the above values of δB or apply a bias to the core. The values apply in most blocking oscillator circuits since the core is normally operated below saturation.

The inclusion of air-gaps reduces the pulse permeability but increases the available flux swing during each pulse and hence delays the start of core saturation. In circuits where core saturation has been allowed to occur before

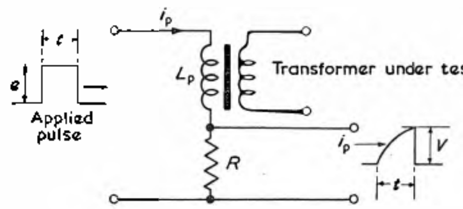


Fig. 17. Method of determining μ_p

$H = (4\pi NV/10\,R)$ *oersteds* $B = (Et/KNA)\,.\,10^8$ *gauss*
N = *numbers of turns on primary winding*
l = *mean magnetic path length (cm)*
t = *duration of applied pulse (sec)*
A = *cross sectional area of core (cm)*
K = *stacking factor*

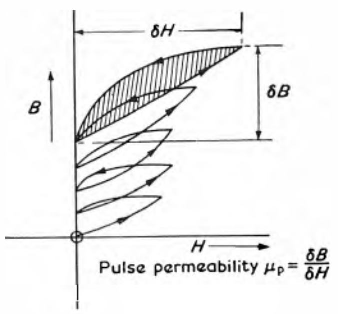


Fig. 18. Build-up of flux due to application of pulses to winding

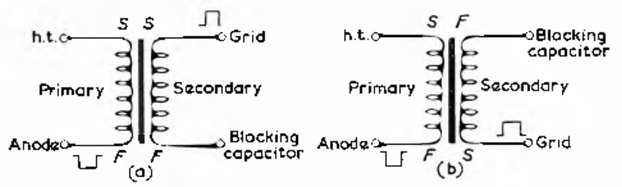


Fig. 19 (a) Outputs taken from the same end of winding. Windings have same angle of rotation

(b) Output taken from different ends of the winding. Windings have opposite angle of rotation

the maximum value of excess current is reached in the valve, an increase in the air-gap reduces L_p, but paradoxially increases the duration of the pulse.

The choice of core material is often a compromise between performance and cost. For sharply rising pulses, thin laminations have the highest values of pulse permeability and give the best results, but transformers using these cores are most expensive to produce. Ferroxcube A4 or 0·004in Mumetal are generally satisfactory for short pulse applications, while Radiometal and orientated silicon steel are more suitable for longer pulses where larger flux swings occur.

Winding Arrangement

The primary and secondary windings should be single layer and of suitable gauge to fill the winding length, and should be separated by sufficient insulation to prevent breakdown between windings during the overswing. Reducing inter-winding spacing reduces the leakage inductance but increases the effective shunt capacitance of the transformer. With triodes, where the rate of increase of grid current can be very high during the leading edge, it is usually advantageous to decrease the leakage inductance at the expense of an increase in capacitance. However, when pentodes are used, the greater part of the grid rise takes place with zero grid current, and it is often worth while to reduce the effective capacitance at the expense of leakage inductance by increasing the space between the windings or by winding the primary and secondary side by side.

The effective leakage inductance L_L is not strictly independent of frequency, but for design and comparison purposes, it is usually sufficiently accurate to take the value determined in the normal manner by measuring the inductance of the primary winding at audio frequency when the secondary is short-circuited. A normal value for L_L is $0{\cdot}005\ L_p$. The effect of L_L on the circuit operation is complex and will be dealt with separately.

The effective capacitance between the windings is dependent on the winding arrangement. Two arrangements are possible for blocking oscillator transformers having single layer windings with phase reversal, and these are shown in Fig. 19 for a 1 : 1 turns ratio. When the windings are in the same direction, as in (a), the effective shunt capacitance is the same as the measured capacitance since there is a constant voltage gradient between windings along the whole winding length. When the primary and secondary windings are earthy at the same end (Fig. 19(b)), the effective capacitance 4/3 of the measured value and this arrangement therefore gives a slower rise-time to the pulse.

The resistance of the windings is usually small (less than 10Ω) compared with the valve impedances and may normally be neglected in the analysis. However, if thin wire is used with a large number of turns, the resistance should be taken into consideration when drawing the paths of operation and in calculating the value of C.

Design Procedure

The following procedure is suggested for the electrical design of the transformer.

(1) Choose the core material, size and thickness of laminations according to the pulse permeability and flux swing required.

(2) Calculate the approximate number of primary turns required $\left(N = \left(\frac{L_p l\,.\,10^9}{4\pi A K \mu_p}\right)^{\frac{1}{2}} \right)$ using an average value of μ_p from Fig. 16.

(3) Determine the flux swing of the core ($\delta B = (Et\,.\,10^8/NK\ A)$) to obtain a more accurate value of μ_p and also make sure that the core will not saturate.

(4) Re-calculate the number of turns using the new value of μ_p.

(5) Determine the gauge of wire necessary to just fill the winding length.

As an example, a transformer will be designed to satisfy the requirements of the circuit previously considered when dealing with the construction of the waveforms. In this case $E = 98$V, $t = 1\mu$sec and the required value of $L_p = 0{\cdot}5$mH.

Ferroxcube would be the most economical core material for this pulse duration, but owing to the low losses of this material, an external damping resistor would be required across one of the transformer windings to prevent re-triggering of the circuit by the overswing voltage, and this resistor would modify the values of L_p and C previously calculated to produce a 1μsec pulse. A suitable alternative core can be made up of 0·004in Mumetal, and the losses in this material would normally be sufficient to prevent re-triggering of the circuit without an external damping resistor. Using a 0·25in stack of Inter-Services shape 450

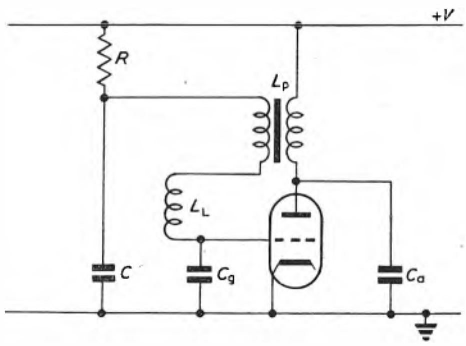


Fig. 20. Simplified circuit with leakage inductance included

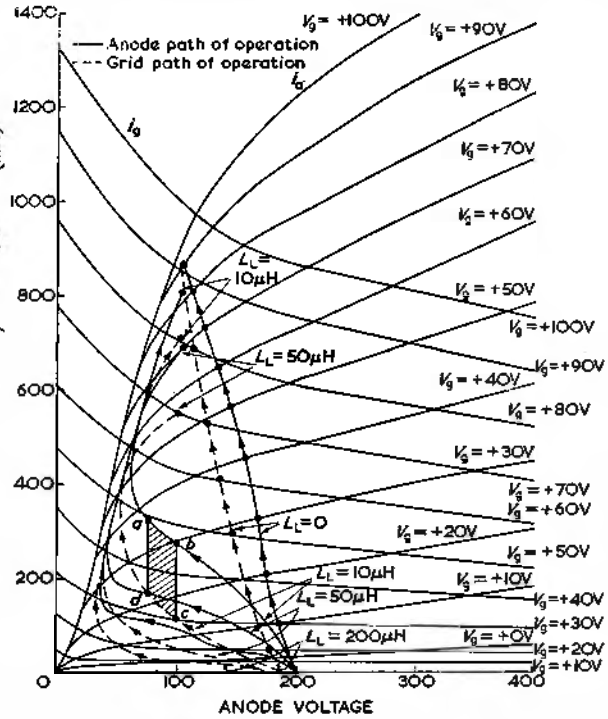


Fig. 21. Positive grid characteristic for average 6J6
Effect of L_L on paths of operation during leading edge

lamination, the constants of the core are $A = 0{\cdot}403\text{cm}^2$, $l = 5{\cdot}06\text{cm}$, $K = 0{\cdot}8$. Fig. 16 shows that μ_p lies between 500 and 900 for a 1μsec pulse, and a value of 700 will be used to determine the approximate number of turns.

$$N = \left(\frac{L_p l\, 10^9}{4\pi A K \mu_p}\right)^{\frac{1}{2}} = 30 \text{ turns.}$$

The flux density swing is:

$$\delta B = \frac{Et\,.\,10^8}{NKA} = 1010 \text{ gauss.}$$

The new value of μ_p is then 600 and the correct number of turns is 32. 32 s.w.g. enamelled wire would comfortably fill the 7/16in winding length.

Leakage Inductance

In the construction of the waveforms considered earlier, it was assumed that the transformer leakage inductance was zero and that for a 1 : 1 turns ratio, a change of voltage at the anode of the blocking oscillator valve produced instantaneously an equal and opposite change at the grid. The effective transformer interwinding capacitance and the stray capacitances were lumped together and the paths of operation and the pulse shapes were constructed on

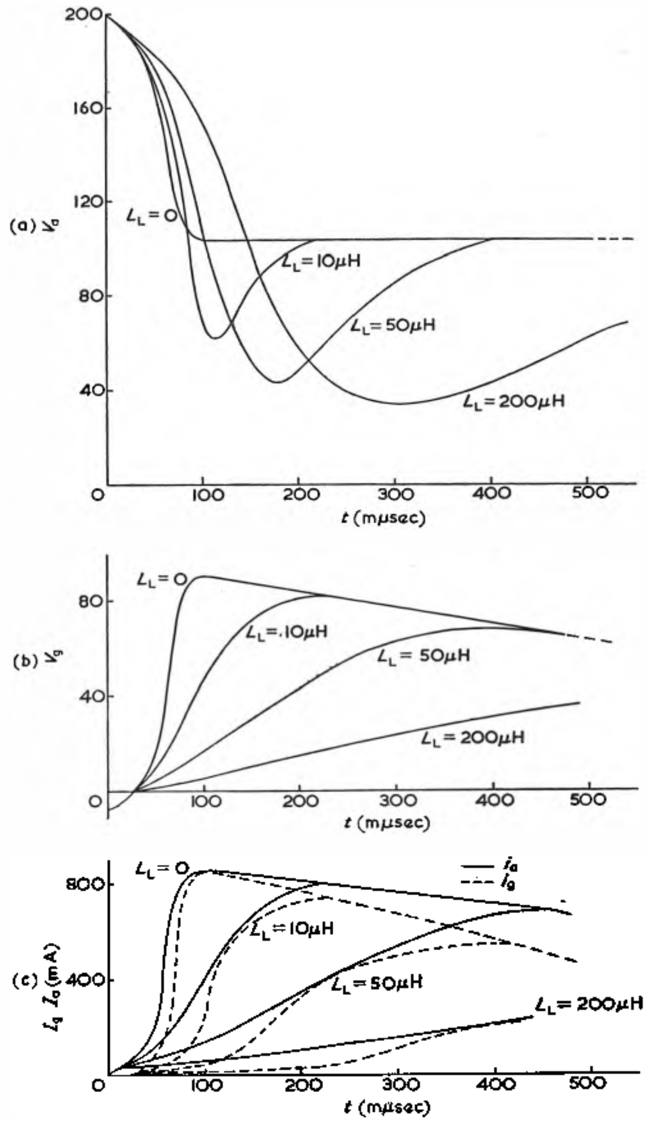


Fig. 22 (a) Effect of L_L on leading edge of anode voltage waveform
(b) Effect of L_L on leading edge of grid voltage waveform
(c) Effect of L_L on leading edge of anode and grid current waveforms

the assumption that this total capacitance appeared between anode and cathode. The presence of leakage inductance however, modifies the circuit operation during the leading and trailing edges of the pulse by partially isolating the grid circuit load from the anode circuit and thereby reducing the rates of change of grid current.

Fig. 20 shows the simplified circuit of the blocking oscillator when leakage inductance L_L is included. The transformer shunt capacitance has been split into two parts and lumped with C_a and C_g.

The operation of this circuit during the leading edge is difficult to analyse quantitatively but it is possible to plot the paths of operation by the step by step method. This involves taking increments of anode and grid voltage, and finding by trial and error, the points on the characteristic curves which satisfy the circuit conditions. These conditions require that for each increment, the time taken for the mean excess current to charge C_a is equal to the time taken for the corresponding increase in grid current through L_L.

In Fig. 21 the paths of operation are drawn for the cases when $L_L=0$, 10μH, 50μH and 200μH, and the method of construction will be illustrated for a typical increment *a b c d*. The mean excess current during this increment is 150mA and since this current flows into the shunt capacitance at the anode (50pF), the time taken for the anode to drop from 100V to 75V is $\delta t = C_a \, \delta v_a / i_e = 8{\cdot}3m\mu$sec.

At the start of the increment $(b-c)$ the anode has dropped 100V from its initial value while the grid voltage has increased by only 37V, so that the voltage across L_L must be 63V. As the anode falls a further 25V, the grid rises by 8V giving an additional 17V across L_L. The mean voltage during the increment is then 71·5V. The characteristic curves show that the grid current (i.e. – the current in L_L) increases by 60mA and, from the expression $E=L_L \, \delta i/\delta t$, the time taken for this rise is 8·4 mμsec.

Since the time of rise of grid current during the increment is virtually the same as the fall time of the anode voltage, the points on the curves have been correctly chosen. If the times had differed however, other values of grid voltage would be tried until the conditions were satisfied.

The waveforms in Fig. 22 have been drawn by the step by step method. It will be seen that the presence of leakage inductance slows down the rates of change of anode and grid current and introduces an overswing into the anode voltage waveform. The grid voltage rate of rise is considerably reduced relative to the rate at the anode, causing C_g to have less effect on the circuit operation than C_a.

In many practical applications, it is useful to be able to estimate the effect of leakage inductance on the rates of rise of anode and grid current without having to construct the waveforms by the laborious step by step method. If an average value for the grid resistance $r_{g(av)}$ is derived from the valve curves, the rise in grid current during the leading edge is given approximately by:

$$i_g = I_{max}\left[1 - \exp\left(\frac{-r_g t}{L_L}\right)\right]$$

where I_{max} is the final current at the equilibrium point, taken in this approximation to be the same value as for the case when $L_L=0$. It will be noted that this expression gives an instantaneous rise in grid current when $L_L=0$ whereas the actual rise time has been previously shown to be, $t_r = C_a V / i_{e(av)}$. A closer approximation is therefore obtained if this value of rise-time is added to the grid current rise derived from the above expression for i_g.

With a well designed and carefully manufactured transformer, the ratio of L_L to L_p is less than 0·005 : 1, and the effect of the leakage inductance on the paths of operation and the leading edge of the pulse can usually be ignored.

(To be continued)

The Simulation of Transfer Functions on an Analogue Computer

By F. C. Harbert*, B.A.

This article describes a one-amplifier circuit for performing the operation of double integration and a method of synthesizing polynomial transfer functions on an analogue computer.

(Voir page 394 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 400)

MOST users of analogue computers are familiar with the second order circuit in Fig. 1 which has the transfer function

$$\frac{V_0(s)}{V_1(s)} = -(R_3/2R_1)\frac{1}{1 + (R_1Cs/2) + \dfrac{R_1R_3C^2s^2}{4}} \quad \ldots\ldots (1)$$

If R_3 is removed from the circuit (i.e. $R_3 \to \infty$) the transfer function reduces to

$$\frac{2}{R_1^2C^2s^2} \quad \ldots\ldots\ldots\ldots\ldots\ldots (2)$$

Thus one amplifier can be used to perform double integration; an operation which usually requires two amplifiers, or three if the sign of the output is taken into account.

By connecting output to input the double integral circuit can be used as a simple one amplifier oscillator. Referring again to Fig. 1 with R_3 omitted it can be shown that the transfer function between V_2 and the input is given by

$$\frac{V_2(s)}{V_1(s)} = \frac{1}{R_1 C s (2 + Ts)} \quad \ldots\ldots\ldots\ldots (3)$$

It would appear at first sight that V_2 could not be used as a signal due to circuit loading but as the grid of a computing amplifier is at virtual earth, the resistor $(R/4)$ can be utilized as the input to any additional amplifier.

It can thus be seen that in many problems the double integral circuit can, with ingenuity and skill, lead to a great saving in the number of amplifiers required. For this reason its use can greatly increase the capacity of small analogue computers.

Simulation of High Order Transfer Functions

A basic approach to the simulation of transfer func-

* *Louis Newmark Ltd.*

tions on an analogue computer is to convert the expression to its differential equation form and set up the computer to solve this differential equation. An alternative method of synthesizing transfer functions makes use of the relationship

$$\frac{KG(s)}{1 + KG(s)} \quad \ldots\ldots\ (4)$$

where $KG(s)$ is the transfer function of the forward path of a closed loop. This method has the advantage that in general

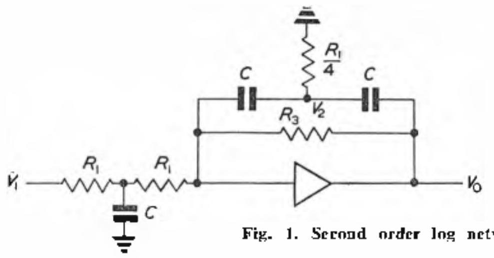


Fig. 1. Second order lag network

fewer amplifiers are required. To illustrate the method consider first the transfer function,

$$\frac{V_o(s)}{V_1(s)} = \frac{1}{1 + s + s^2 + s^3} \quad \ldots\ldots\ldots\ (5)$$

Putting $Q(s) = s + s^2 + s^3$,

$$\frac{V_o(s)}{V_1(s)} = \frac{1}{1 + Q(s)} = \frac{1/Q(s)}{1 + 1/Q(s)} \quad \ldots\ldots\ (6)$$

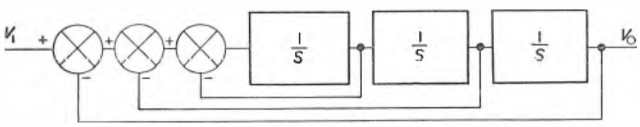


Fig. 2. Feedback loops required to simulate the transfer function $\frac{1}{1 + s + s^2 + s^3}$

From equation (6), it is seen that $1/Q(s)$ is represented by $KG(s)$ in equation (4). If a 'block' circuit could be designed to produce $1/Q(s)$, feedback could be added around it to produce the required results. Thus the problem is to generate $1/Q(s)$.

Now

$$\frac{1}{Q(s)} = \frac{1}{s + s^2 + s^3} = \frac{1}{s(1 + s + s^2)} \quad \ldots\ldots\ldots\ (7)$$

$1/s$ can be produced by using an integrator. Thus the problem reduces to generating

$$Q_1(s) = 1 + s + s^2 \quad \ldots\ldots\ldots\ (8)$$

The process is now started again and repetition of the method produces the series of feedback loops shown in Fig. 2.

In the above example all coefficients have been made unity in the transfer function to simplify the working. However, the method is generally applicable.

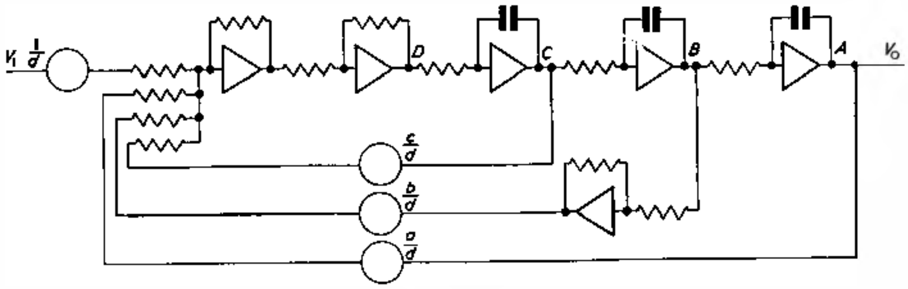


Fig. 3. Computer arrangement for the simulation of the transfer function $\frac{1}{a + bs + cs^2 + ds^3}$

The transfer function

$$\frac{1}{a + bs + cs^2 + ds^3} \quad \ldots\ldots\ldots\ (9)$$

can, by the above method, be simulated with the computer arrangement in Fig. (3).

Consider now the transfer function

$$\frac{V_o}{V_1} = \frac{e + fs + gs^2 + hs^3}{a + bs + cs^2 + ds^3} \quad \ldots\ldots\ (10)$$

$$V_o = \frac{e}{a + bs + cs^2 + ds^3}V_1 + \frac{fs}{a + bs + cs^2 + ds^3}V_1 + \frac{gs^2}{a + bs + cs^2 + ds^3}V_1 + \frac{hs^3}{a + bs + cs^2 + ds^3}V_1 \quad \ldots\ (11)$$

Referring to Fig. 3 the signal at point A referred to the input is $\frac{-1}{a + bs + cs^2 + ds^3}V_1$.

At point B (the input end of the third integrator) the signal is $\frac{+s}{a + bs + cs^2 + ds^3}V_1$; at C $\frac{-s^2}{a + bs + cs^2 + ds^3}V_1$; at D $\frac{+s^3}{a + bs + cs^2 + ds^3}V_1$. Thus the terms in equation (11) can readily be obtained and the complete computer diagram is shown in Fig. 4.

The method is applicable of course to transfer functions of any order and when some coefficients are zero the number of amplifiers used can be reduced by employing the double integrating circuit.

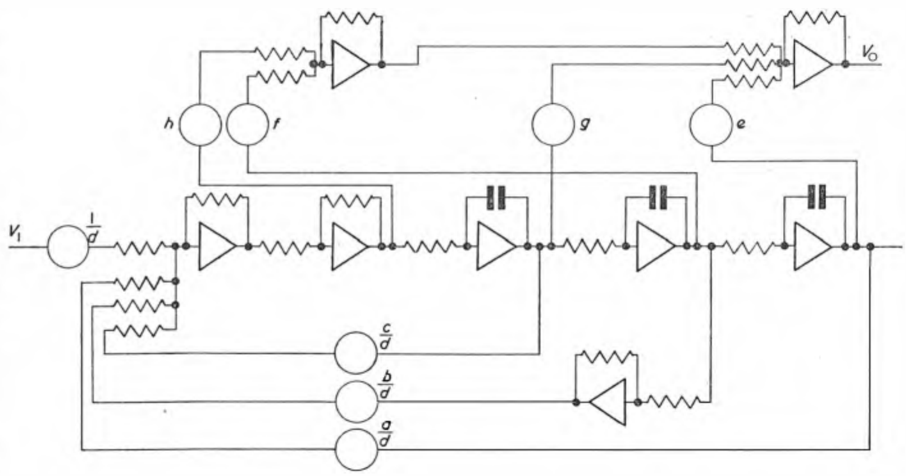


Fig. 4. Computer circuit for the simulation of the transfer function $\frac{V_o}{V_1} = \frac{e + fs + gs^2 + hs^3}{a + bs + cs^2 + ds^3}$

Henri de France Colour Television System*

IN known systems of colour television, when cutting down the chrominance signals the horizontal colour definition is lowered to an extent which is not perceptible to the eye. But vertical definition is practically unchanged. This fact seems rather abnormal when one considers that in all Standards, bandwidth and number of lines are related to achieve balance between the horizontal and vertical definition.

The first important aspect of the 'Henri de France' system takes advantage of this observation; it is of little interest to supply the eye with colour information which is denser in the vertical direction than in the horizontal; having reduced the horizontal colour definition it is also possible, without disadvantage, to supply the new chrominance information in every other line only.

The application of this first idea gives the system its second, and no less important, characteristic. It is clear that the simultaneous transmission of three sets of information over a given frequency band raises greater risks of interaction than if only two signals were transmitted simultaneously over the same frequency band. Sequential transmission of chrominance information, the means employed in the Henri de France system for harmonizing the vertical and horizontal colour definition, makes it possible to transmit only two sets of information simultaneously.

At the transmitter, luminance information is transmitted over a wide band for all lines, in accordance with the usual rules for black and white television, but each of the complementary chrominance signals is alternately transmitted, over a narrow band, only every other line.

The chrominance information which is not transmitted is little different from that corresponding to the preceding line, this information having already been transmitted. At the receiver, it is considered as being identical, the eye being incapable of appreciating the difference. This therefore provides three simultaneous signals while only two have actually been transmitted. For a line, the three signals are luminance, the chrominance information transmitted at that instant, and the chrominance information transmitted at the preceding line. The latter signal is then available, due to a delay system which causes a delay of one line. At any instant, the chrominance signal transmitted follows two paths. One takes it directly to the decoding matrix before the receiver cathode-ray tube; the other makes it available for the following line.

At the transmitter, the two chrominance signals, multiplexed in time, are frequency multiplexed with the luminance signal so that it may be transmitted at the same time as the latter. The luminance information keeps its normal place in the frequency spectrum. A sub-carrier allows these two chrominance signals to be transposed. It is situated at the end of the video spectrum with which it is interlaced. It can be transmitted with symmetrical or asymmetrical sidebands. In order to reduce its effect on the luminance signal its frequency is chosen as an odd multiple of half the line frequency.

The arrangement obtained in this way is compatible. The black and white picture can be extracted from the composite signal by a simple filter. Conversely, the principle of constant luminance being observed, a colour receiver gives a faithful reproduction of a black and white picture.

Lastly, stress should be laid on the fact that the chrominance sub-carrier is at all times modulated only by a single signal. This secures all the advantages of a single modulation. For example, if the sub-carrier is amplitude-modulated, there is no need to be concerned about spurious phase modulation. If the sub-carrier is phase- or frequency-modulated, then the behaviour of the amplitude of the sub-carrier is of no significance.

In short, it can be said that the Henri de France system is characterized by:

(a) Sequential transmission of chrominance information lines.

(b) Simultaneous transmission of two signals instead of three.

(c) Single signal modulation of the chrominance sub-carrier.

(d) Storage at the receiver of chrominance information from line to line.

Transmission Encoder

Fig. 1 shows the arrangement of the encoder.

This unit processes the composite signal E_t' from the camera voltages E_G' E_R' E_B' and also the synchronizing and sub-carrier frequency voltages which it produces itself.

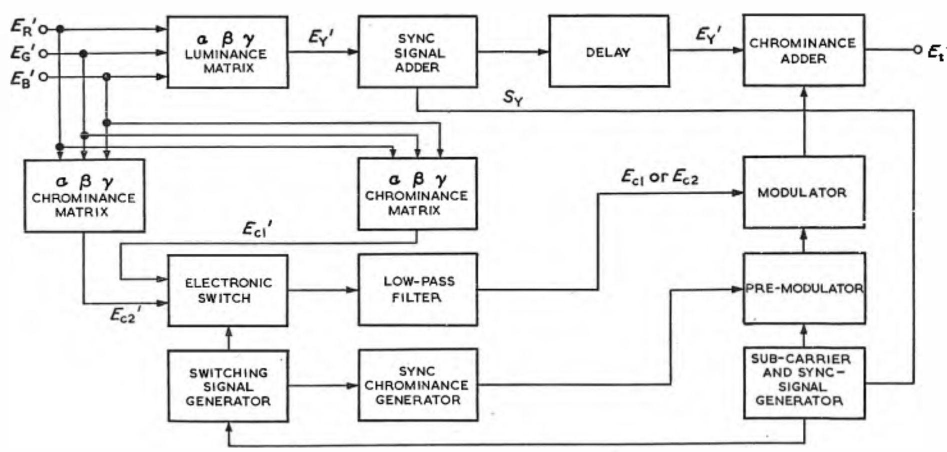


Fig. 1. Block diagram of encoder

The three signals, E_G' E_R' E_B', issuing from some analyzing device are applied to a matrix $\alpha\beta\gamma$ supplying the luminance signal by linear combination of the three signals. After the synchronizing signals have been added, the signal E_Y' is applied to a delay line whose function is to phase together the luminance and chrominance signals, and then to a chrominance mixer which also superimposes the modulated sub-carrier.

* *Compiled from information supplied by Cie Française de Télévision.*

Two other matrices $\alpha'\beta'\gamma'$ and $\alpha''\beta''\gamma''$ produce wide-band chrominance signals E_{c1}' and E_{c2}' which are applied to an electronic switch controlled at line frequency by the synchronization generator. A filter follows the output of the electronic switch and reduces the bandwidth of the chrominance signals. These narrow-band sequential signals then drive the sub-carrier modulator. This sub-carrier is not sent continuously to the modulator to be modulated by the chrominance signals. It is pre-

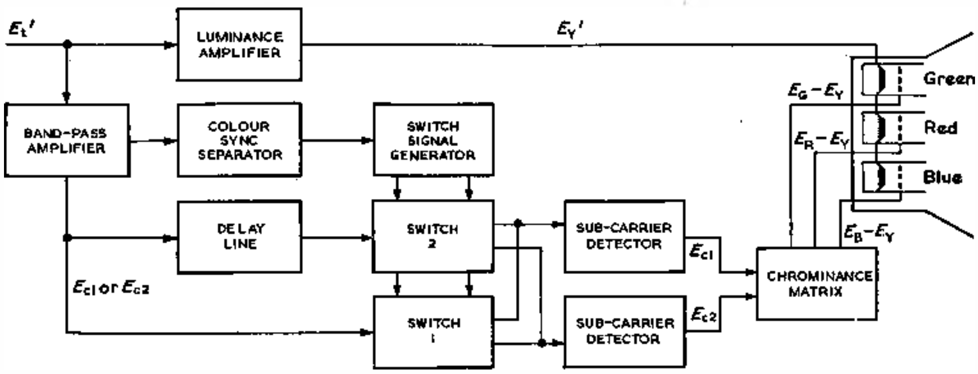


Fig. 2. Block diagram of decoder

modulated by signals issuing from the colour synchronizing generator and is used to discriminate between the two chrominance signals. This pre-modulation is in fact a blocking of the sub-carrier which occurs at the same instant as the line synchronization signals of the black and white picture. The sub-carrier blanking signals are rather wider than the line synchronizing signals of the black and white picture and overlay the blanking levels proper to that picture. The sub-carrier suppression pulses, syn, vary in width on alternate lines. This allows the two chrominance signals to be separated.

Receiver Decoder

Referring to Fig. 2, the composite signal E_t' passed to the decoder input is first subdivided, by filtering, into two signals, the luminance E_y' and the chrominance E_{c1} or E_{c2} transposed by the sub-carrier. The use of the two chrominance signals E_{c1} and E_{c2} will now be examined in detail. The first operation to carry out is their identification, as either are liable to appear at any instant; this is the role of the pulses S_{yc1} and S_{yc2} which control the switching generator at the right instant. After identification, two sequential signals must be converted to two simultaneous signals. In the exposition of the general principles it was seen that the eye is insensitive to amplitude variations of E_{c1} or E_{c2} from one line to the next. In order to effect the sequential-to-simultaneous conversion, systematic use is made of the chrominance information E_{c1} or E_{c2} available at a given moment and the chrominance information E_{c2} or E_{c1} transmitted at the preceding line. The sequential signals are transmitted over two channels, one direct and the other with a transmission delay equal to the duration of one line. At the output of these two channels two signals are simultaneously available E_{c1} and E_{c2}. One output provides in succession E_{c1} E_{c2} E_{c1} E_{c2} . . . etc . . . and the other E_{c2} E_{c1} E_{c2} E_{c1} . . . etc. All that is needed is to switch these two outputs in synchronism, at line speed, to two new channels in order to obtain two simultaneous and continuous voltages E_{c1} and E_{c2}. These operations are effected by switches 1 and 2.

The twin outputs of switches 1 and 2 feed two detectors which rectify the sub-carrier envelope, that is to say they reproduce the true signals E_{c1} and E_{c2} and not transposed signals.

A chrominance matrix then effects the transformation of (E_{c1}, E_{c2}) into ($E_R - E_Y$, $E_G - E_Y$, $E_B - E_Y$). It will be noted that E_Y contains only the low-frequency components E_Y of the chrominance signal E_Y'. Finally, these three voltages $E_R - E_Y$, $E_G - E_Y$ and $E_B - E_Y$ are applied to each grid of a three-colour tube with three guns whose three cathodes are fed in parallel by E_y'.

Decoder Delay Unit

For a 625 line system a delay of 46μsec is necessary in converting the line sequential chrominance signals to continuous form. This problem has required special attention and, after investigating the performance and economics of units employing delay cable, preference is now given to a quartz delay system of the multiple reflexion type[1,2]. The characteristics of this are as follows.

The transducers are barium titanate ferro-electric cells using the propagation of longitudinal sinusoidal waves in a solid medium of fused silica.

These lines have the following characteristics:

Delay	:	64μsec
Pass-band at 3dB about the 4·1 Mc/s sub-carrier	:	1·1Mc/s
Spurious echos	:	28dB below the useful signal
Attenuation	:	29dB
Temperature co-efficient	:	less than 10^{-6}/°C
Dimensions: height	:	25mm
diameter	:	75mm

Further work is in hand for improving certain characteristics and for further reducing the cost. But it is already clear, judging from these latest results, that the problem of the delay line has become a secondary one, both from the technical and economic point of view.

Tests on the Effects of Distortion and Spurious Signals

Experiments have shown that, as expected, the Henri de France Colour System is virtually immune to differential phase distortion of the type frequently encountered in f.m. links.

Differential gain should be constant to within about 1dB to avoid hue distortion. Also, to avoid colour saturation distortion a tolerance of 1dB is placed on the luminance-chrominance gain ratio.

For uniformly distributed noise, the threshold of noise visibility on the colour picture is found to be about 10dB below that of the monochrome picture. In the case of an interfering signal at sub-carrier/frequency the visibility threshold of the intereference pattern occurs at a level 13dB below that of the monochrome picture.

Reducing the Visibility of the Sub-Carrier

Line sequential transmissions of the chrominance signals can result in a coarse sub-carrier dot-pattern if there happens to be a significant difference in the amplitude of the two signals.

This defect can be reduced by reversing the sequence of the signals at the end of each field. The resultant dot-pattern then has a downward diagonal crawl. A further improvement can be effected by periodically reversing the direction of the crawl; this is done by omitting to reverse the colour sequence in alternate groups of four fields but instead reversing the phase of the sub-carrier. Eight fields are then required to establish the behaviour of the dot-pattern.

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Computer for Air Traffic Control

One of the first computers to be used for air traffic control experiments in this country is to go into operation at the Scottish air traffic control centre at Redbrae, Prestwick, early in the spring of next year.

The computer, known as APOLLO, is a high-speed, transistorized, parallel machine, which is at present under construction at the computer research establishment of Ferranti Ltd at Bracknell, Berkshire. It will be used for processing flight information, but initially this will be done only on an experimental basis.

The main problem facing civil air transport today is the ever-increasing number of piston-engined aircraft flying on the Atlantic routes and the rapid introduction of turbo-jet aircraft requiring different control techniques, both factors combining to aggravate potential hazards. This brings with it the attendant problem that as the numbers of aircraft increase, so does the volume of work for controllers.

With this in mind the Ministry of Aviation approached Ferranti Ltd to design a computer which could be used for experimentation in order to find a solution to the more efficient use of airspace by investigating new control techniques and the possibility of using smaller separation standards. It was also felt that the use of the computer could ease the manpower problem to some extent as it would free the controller from a number of time-consuming routine tasks.

Aircraft flying the North Atlantic are subject to rigid traffic regulations. Whether flying eastbound or westbound, each aircraft must conform to agreed separation standards, which are 30 minutes longitudinal flying time between aircraft, 2° of latitude and 1 000 or 2 000ft in height, according to the flight level.

To prevent possible collisions, as there may be as many as 100 aircraft in the sky at any one time, it is the oceanic controllers' duty to be fully aware of traffic conditions and to re-route aircraft, should it be necessary, to avoid infringement of separation standards.

For this purpose, careful attention is paid to pre-flight planning and the present practice is for pilots to submit a flight plan some hours before the proposed departure time. Flight progress strips are then prepared for use on the controllers' flight progress boards. These strips contain such information as the aircraft identification number, aircraft type, point of departure, departure time, destination, cruising speed, and altitude, longitude and estimated time of arrival at specified meridians. Radio contact is also maintained between the aircraft in flight and the control centre, so that position reports giving time, altitude and latitude at the meridian just crossed and estimates for the next meridian, can be verbally reported by the pilot. In this way the controller is able to keep a detailed record of the route and position of each aircraft, in the form of flight progress strips, which enables him to make control decisions.

The whole process of sorting out information, amending flight progress strips and guarding against infringements of the regulations is carried out manually by the controller.

The new computer, which has taken two years to develop, will print out flight progress strips relating to each aircraft and provide up-to-the-minute information on air traffic conditions. It will also carry out calculations to see whether aircraft are in conflict and, if it is found that separation standards are being infringed, details of this will be printed out for the benefit of the controller who retains the basic responsibility of re-routing the aircraft. The complete situation may also be shown pictorially to the controller.

The computer which works in real time consists basically of three cabinets, each 6ft high, 22in wide and 19in deep. Positioned alongside the computer is a table upon which is mounted a number of Creed 25 tape punches and in front of the computer is the control desk upon which is mounted a Ferranti TR5 tape reader and a Creed 75 printer.

The left hand cabinet contains racks of printed packages which carry out the arithmetic and control operations of the computer, next to that is the cabinet containing the programme and data store as well as the power supplies. The third cabinet contains input/output logic for control experiments. A fourth cabinet on the extreme right contains electronic equipment, developed by the Ministry of Aviation in conjunction with R.R.E. for cathode-ray tube displays.

For processing flight information, the computer receives data from the local operators by way of four Creed printers with keyboards. Under the new system, the operators will receive the same information as the controller now receives. This information will be sorted out in conjunction with a controller and only relevant information will be passed to the buffer store of the computer. At the same time, a record of what has been typed appears on the printer associated with the keyboard, and if the operator is satisfied that everything is in order, he then presses a key on the keyboard indicating to the computer to transfer the information into the main file.

Depending on the message, certain stored programmes are brought into action and data is processed and checked for conflict. If the information concerns a flight plan it is stored until the departure time is received. This information is then processed and flight progress strips are printed out automatically on Creed printers. This is done according to a time criterion, which enables the strips to be entered on a board in readiness for the entrance of an aircraft into the controlled area.

Another feature of the machine is that the programme can be interrupted at any time and a wide variety of special sub-routines inserted into the programme automatically. This interrupt facility, which is carried out on a priority basis, enables the computer to deal with data-processing functions which originate with little or no advance notice, a particularly important factor in real time applications such as air traffic control.

It also incorporates a time-sharing facility which ensures that it can be always fully engaged even while peripheral transfers are taking place. Although operational programmes have priority, the computer can also be used for carrying out routine calculations and developing programmes.

The cathode-ray tube visual displays are employed as another means of showing numerical data stored inside the computer memory to the controller, and they may also be used to show plan position information.

For real-time operation, a time register containing a number representing G.M.T. is fed by pulses from a master clock, but for non-real time applications the register can be fed from an independent variable oscillator.

The Design of
OPTICAL DIGITAL INSTRUMENTS

By I. R. Young*, B.Sc.

Various design considerations for optical digital instruments are first discussed with emphasis placed on the importance of readings being at once available if demanded, as this is desirable in time sharing systems. Some low resolution instruments are then described with notes on the use of the data obtained from them. Finally, the problems associated with the design of very high resolution units are briefly described.

(Voir page 394 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 400)

OPTICAL digital measuring systems are capable of extremely high resolutions, probably greater than with any other method. This article discusses some problems associated with the design of digital instruments using optical techniques.

There are two main approaches; one using coded scales[1,2], and the other diffraction gratings[3,4]. It is possible to use one or other of these in various forms, or in conjunction with other types of scale[5], but these do not constitute fundamental types.

In arguments about the relative merits of coded and counting systems, the most common basis for attacks on the latter is the danger of missing counts, and consequently, of changes in the origin of the measurement. This can be minimized electrically, and a technique for incorporating check marks on the scale will be described, but it is more difficult to see a solution for another major disadvantage. This is the fact that whereas the outputs from a number of coded commutators can be fed, using time sharing, through one output system, each fringe detecting unit must have a counter permanently associated with it, in addition, probably, to output equipment which can then be common to a number of counters. Primarily for this reason, the bulk of this article is devoted to coded scale considerations.

Some Design Pointers

The design of an optical coded commutator can vary enormously with the use to be made of the output signals from it. A digitizer, the output from which is to be continuously displayed, can be very different in structure from one which is being sampled intermittently for short periods. In both cases, the idea of the accessibility of information is important. If a reading must be obtained in an arbitrarily short time after a demand for it has been made, the scale must contain all the information which is needed and the detectors must record it all at once. If the time is very short, flying spot systems in which a light spot is traversed across the scale, and detected by a single cell, may be impracticable. Similarly, code arrangements in which, for example, the elimination of ambiguity depends on one or more circuits being set from the scale before another part can be read, may also be unusable.

The requirements for high accessibility usually eliminates the easiest method of achieving very high digital resolutions in which the scale consists of fine lines (which are 'numbered' by a coarse code) accurately spaced at a convenient but relatively large distance. The line centres are determined by a photo-electric microscope[6,7] or similar device, and the interval between the lines is resolved by a micrometer with appropriate coded scales attached to it. Thus, a linear resolution of 0·0001in can be conveniently obtained using a scale with lines 0·0005in wide positioned 0·1000in apart centre to centre. The micrometer must then have a resolution of 1 in 1 000. It is obvious, however, that, as the micrometer can only be read when a line centre is being detected, it is impossible, in general, to obtain a reading within perhaps 10 to 50msec of a demand for one being made.

Requirements for high accessibility also condition the type of code to be used. There are three fundamental methods of avoiding ambiguity—the guard ring technique in which a track of the code is used to prevent readings being taken in places where tracks are changing; the choosing system in which pairs of detectors are used on all tracks except the least significant and the one of each pair to be read determined from it; and the use of a cyclic permuting code in which only one track changes at any one place. The first has obviously a very low accessibility as over a finite fraction of the disk no reading at all can be obtained: the second can be read quickly but involves duplication of detectors and often amplifiers (though if time is available it is possible to select which cell of each pair is to be read and use one amplifier. This involves a delay between the interrogation of the reference cell and the other detectors.) Particularly in small digitizers the main difficulty with this technique is the close packing needed for the cells and, for this reason, unless otherwise stated, only cyclic coded systems will be considered.

Design can also be very much affected by the frequency with which readings are taken, and by the duration of any one interrogation—these two are, in general, completely independent. From a theoretical point of view, the latter in particular, can affect the resolution of which a given configuration of lamps and detectors is capable. From a practical point of view, these parameters are important mainly in deciding the type of lamp to be used. In most applications it seems that the interval between the readings is long compared with the time of an interrogation. This is an argument for the use of cold-cathode discharge lamps, in which the intervals between readings are not simply a waste of lamp life. Obviously a system in which tungsten filament lamps must be switched on before a reading may be taken is one with very low accessibility.

Another problem is the way in which the outputs from the cells are to be specified. In principle it is sufficient that the signal contrast between the 'light on' and 'light off' conditions should be sufficiently great, though it is obviously desirable that the minimum 'on' signal should be large enough to avoid much amplification—in any system the contrast is ideally large between the greatest 'light off' signal and smallest 'light on' one (which may be from another part of the device altogether).

* *Evershed and Vignoles Ltd (formerly with Hilger & Watts Ltd)*

Reading Techniques

There are three primary techniques for reading coded commutator scales optically, of which the simplest uses a number of photo-detectors and one or more d.c. lamp sources—this type of approach is, of course, obligatory with fringe counting systems.

Secondly, the arrangement of the cells can be like that in the first case but they are illuminated by an a.c. source (giving a pulsed or flashed light, or a continuously modulated one.) Thirdly, a single detector may be used with a light source which illuminates the code tracks in series. The work to be described here is all concerned with the first and second approaches.

The two methods of reading optical digitizers have been hinted at—that in which the output is continuously monitored, and that in which it is sampled. The latter is probably the more generally useful as readings can be taken very quickly and, as repetition rates are often low, the amplifiers, decoders, and so on can then be used for other

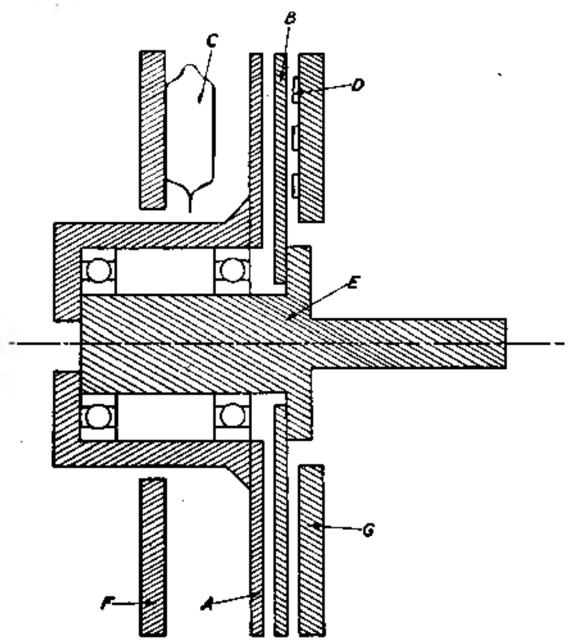


Fig. 1. Schematic drawing of a small optical digitizer

digitizers in the intervals between interrogations. This method lends itself to units using a flashed light source.

With a modulated light continuous outputs can be obtained by demodulating the signals from the digitizer, but this arrangement limits the speed at which the shaft can be turned as it is necessary to allow 5 to 10 cycles of light to establish the presence or absence of an output from a cell. As moving mechanical parts and large optical systems are undesirable in a digitizer, modulation frequencies are necessarily fairly low, and the shaft speed is limited to that in which the order of only 1 000 digits per second pass the detectors.

Systems using d.c. light can give either type of output (sampling being achieved by pulsing the cells) but do depend on having stable and uniform detectors.

Low Resolution Digitizers

It is convenient to divide the instruments to be discussed into two classes—low resolution and high resolution—the division being made by whether or not the unit has a projection optical system to image the scale on a graticule. In practice, on this argument, units in the high resolution class have a digit size less than about 0·003in.

Two types of detectors are used in the low resolution instruments—silicon photo-voltaic and Hilger Schwarz photo-conductive cells. The approaches needed in using these very dissimilar types differ considerably in philosophy and operation and they will, therefore, be discussed with little reference to each other.

Schwarz cells have a very high sensitivity, but their variations in cut-off frequency (anywhere between 50 and 300c/s) dark resistance and temperature coefficient of sensitivity makes it very difficult to use them with d.c. light. With these cells there is, in general, little difficulty in getting enough light to produce a given signal amplitude—the problem is to obtain a sufficiently short time of read-out to enable a reasonable speed of shaft rotation to be obtained. (If a flashed source is used with some sort of storage of the parallel outputs from the cells, the maximum speed of shaft rotation which allows a completely unambiguous output to be obtained is that at which one digit only of the scale can be seen by the cells during the time in which light is present.) In practice, best results are obtained using neon filled discharge lamps. Assuming that the time-constant of the cells is much longer than the time duration of the flash (almost invariably so), it can be shown that the maximum signal amplitude is obtainable when the ratio between the flash length and the interval between flashes is about 1 : 5. This is borne out in practice, though, as a rule, single flashes separated by intervals of 100 or more times their duration are required. Two types of lamps have been used—special neon flash tubes, giving a minimum flash duration of about 0·25msec and neon indicator tubes, with which the shortest practicable flash is about 1msec (both figures for 0·005in digit width.) The latter are of much less peak brilliance than the former, but are very much smaller, and can, therefore, be used in a more compact instrument.

Where size is of the essence, and, unless there are excellent reasons, a low resolution instrument must be small, normal projection optical systems tend to be too large. For this reason, such instruments use one or more slit systems in conjunction with the disk to define the area of code which can be seen by the detectors. A schematic drawing of a simple digitizer (using neon indicator bulbs) is shown in Fig. 1. An opaque disk (A) with slits in it, one to each cell, is mounted close to the code disk (B). The three lamps (C) are arranged so that each illuminates four cells (D). The code is staggered on the disk so that only every third track is illuminated by any one lamp.

The main limit on resolution with this method is imposed by the finite gap between the code and slit disks, which has the affect of spreading the apparent size of transparent segments of the disk. Assuming that the gap between the circles is d; the width of a digit X; of a reading slit Y; and the angular width of the source α, the apparent edge of a digit will be advanced by up to $2/5$ $(Y + 2d \tan \alpha/2)$ approximately where the track concerned is changing from opaque to transparent, allowing for a four to one variation in sensitivity about the mean value at which the cells nominally trigger the associated equipment. There is thus a danger of missing digits if $X < 4/5\ (Y + 2d \tan \alpha/2)$ as cells which change consecutively may have the maximum difference in sensitivity —that is, in practice in a 1 000 division 3in diameter digitizer if $d > 0{\cdot}004$in.

Crosstalk, due to light passing through one slit and code track, reaching a cell corresponding to another, can cause a similar effect. Both can be minimized by using plates with collimating slits in them.

With an optical digitizer, regardless of type, the most likely source of failure is the illumination system. This is particularly true of arrangements in which more than one lamp is used, as a fault tends to be more difficult to detect. When binary coded decimal, or other, codes with some redundancy are used, complete lamp failure is easily recognized by ensuring that the code combination in which no cell is illuminated does not appear in the pattern. With binary codes it is necessary to have an extra, wholly transparent track on the disk with an additional detector. Where a single lamp is used, for example, the flash tube used in one digitizer based on this discussion, the problem is relatively simple. With multiple lamps, series wiring is to be preferred (in the digitizer using neon indicators, the lamps are usually in series, even though their performance, in terms of life, may be impaired), but is unsatisfactory if many lamps—one for each cell, perhaps—are used.

From a purely practical point of view, the neon-indicator type of digitizer has been made as simple as possible so that it consists essentially of four plates on the same axis. One of these, the coded disk, is accurately centred on a

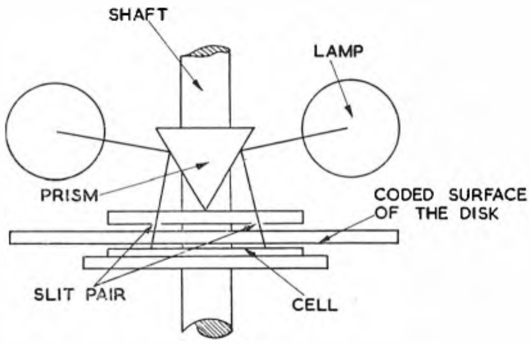


Fig. 2. Lamp and slit arrangement for an optical choosing system

shoulder on the shaft and cemented with Araldite (*E*, Fig. 1). The slit disk is similarly mounted on the bearing assembly of the instrument; with, immediately behind it, a printed circuit (*F*), carrying the three lamps. On the other side of the code (with the shaft passing through its centre) is another printed circuit plate (*G*) with the cells, in the form of small mica pellicles, stuck on and contacted to it.

Multi-Turn Units

Mention has already been made of the difficulty of using the method of avoiding ambiguity in which pairs of detectors are positioned on each track. In addition to the problem of packing in the cells, extra amplifiers or low level switching circuits are needed. If it is essential to adopt this technique, there is much to be said for an approach[8] in which pairs of lamps—and single cells—are used, as long as each pair of lamps can illuminate a number of cells. This arrangement is valuable in reading the coarser sections of multi-turn digitizers, as under these circumstances cyclic coding is impossible.

In practice, to allow for as much error as possible in the gearing, the two reading positions on the least significant track of the coarse scale ought to be spaced by half a digit width. The limitation on making full use of the latitude available for back-lash is generally set by the size of the cell where, because of space considerations, it is undesirable to use lenses or piped light units to collect light on to it. Another condition for a satisfactory system is that the cell shall have roughly uniform sensitivity over that part of it which is used.

A practical arrangement is shown in Fig. 2 in which two neon indicator lamps are used in conjunction with Schwarz cells. Each cell can be illuminated through one of a pair of slits by a lamp which is selected by reference to the fine part of the system. In practice, it is convenient to use one pair of lamps for only four cells—but there is still a major saving in associated electronic circuits even with as few as this.

This technique is, however, only of value with lamps that can be very rapidly switched—i.e. in a time which ideally is small compared with the response time of the detectors. Systems in which filament lamps are used, unless they are very slow, must use the standard technique of pairs of detectors.

D.C. Systems

The simplest form of optical digitizer is that in which a d.c. output of sufficient power to operate a simple decoding element such as a relay is obtained directly from the cells. Photo-conductive cells like the Schwarz cell, the photo-transistor and various photo-diodes meet this condition—but suffer, more or less, from a lack of uniformity and from thermal instability. Even so, they would be very attractive if the most profitable method of using such digitizers were to feed them individually into relay handling equipment. However, as in the majority of cases it is usual to feed the outputs from a number of digitizers into the same display mechanism, more sophisticated techniques are justified if they enable digitizer switching to be accomplished earlier in the system, more quickly, and with less equipment.

For these reasons, and because the cells are relatively uniform in performance, silicon photo-voltiac cells are used in the rest of the equipment to be discussed. In one type of instrument, there is enough output from the cells to switch a transistor directly, but in most the current from the detectors is so small that the signals can only be conveniently handled by a.c. amplifiers. However, the general principles for handling the output from the instruments are very similar in the two approaches.

Three cases may be distinguished:

(a) D.C. output

(b) Parallel sampling

(c) Serial sampling

In the design in which the cells supply sufficient current to switch transistors directly (the specified minimum is $20\mu A$ into $1k\Omega$ for the light "on" condition), case (a) is simple. With the other instruments, the specified minimum light "on" signal can be as small as $0{\cdot}5\mu A$ into $1k\Omega$ (for a high resolution instrument). In other terms, the output minimum is 50mV with the cell open-circuited. As a high resolution digitizer may have 30 output channels, it is undesirable to use d.c. amplifiers, particularly as it is essential that all the outputs should be referred to earth. Further, transistor amplifiers are preferable as there is little point in using reliable cells and making every effort to use the minimum number of lamps—and then following the digitizer with a system in which the chance of failure is much higher. For this reason, the output of the cells is chopped and amplified before being rectified to give the desired signal form. This technique suffers from the same inefficiency in terms of speed as that in which modulated light is used, and merits little attention as it is potentially much less useful than the sampling ones.

The techniques used in cases (b) and (c) are very similar.

In all the methods, the chopping mechanism is a silicon transistor. In systems where the current is relatively high this is used simply as a switch, but with small signal devices it is used to discharge a capacitor which has been charged previously by the cell.

Various circuits are shown in Fig. 3 for use with digitizers from which the higher currents are obtained. Detailed design considerations for the circuits will not be discussed here as the emphasis is on the instruments and the general techniques of use which condition their design. Fig. 3(a) shows the fundamental circuit, feeding into either a bistable Eccles-Jordan circuit or an avalanche diode (which can operate a medium-sized relay if desired). Other outlets are obviously possible. Fig. 3(b) shows a bank of amplifiers fed

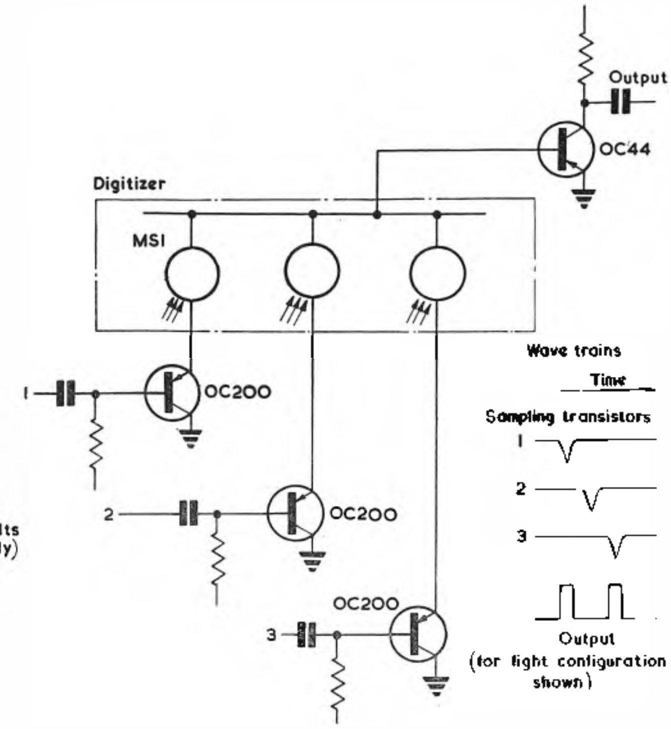
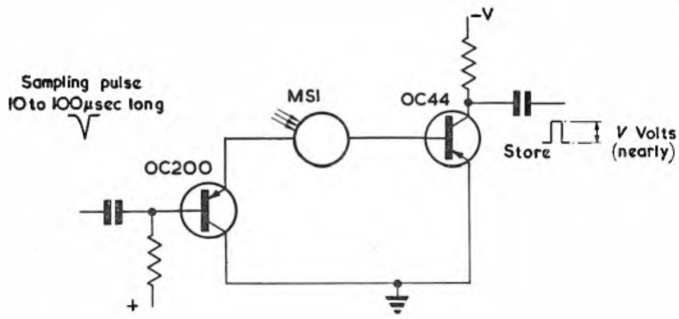


Fig. 4. Serial sampling of digitizer cells



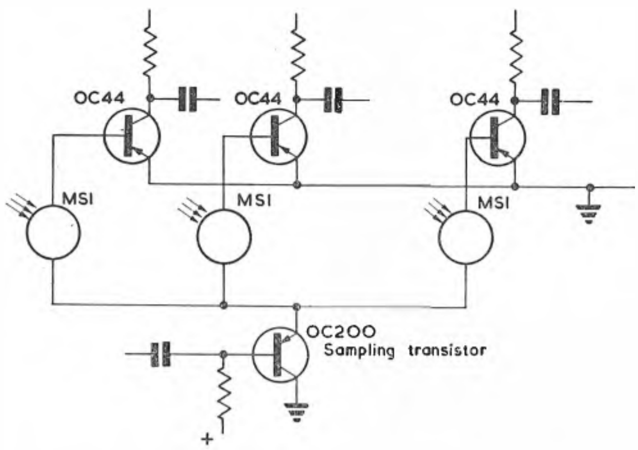
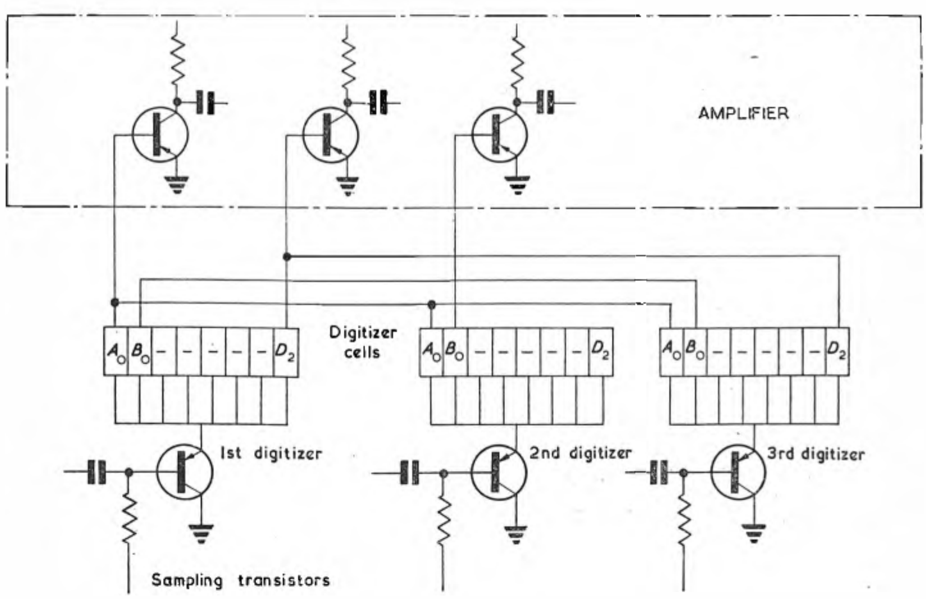


Fig. 3. Parallel sampling for digitizer cells

from the cells of a digitizer in which all are operated by one chopping transistor; Fig. 3(c) shows several digitizers commoned into the same amplifier shared by them in time by selecting the appropriate chopping transistors. Fig. 4 shows an arrangement in which several cells from the same digitizer are fed into a common amplifier, but each having its own chopper. By cycling through these in turn, a serial output can be obtained, which is at once suitable for computation. This arrangement is more economical in components than that in which the cells are sampled in parallel (Fig. 3(b)) and the signals fed into a shift register.

In practice it is convenient to design sub-units, and for this reason four cells only, corresponding to the output from one decade of a binary decimal scale, are fed into an amplifier. Similarly, it is convenient to arrange to feed only four digitizers into one amplifier.

An example of the use of these ideas is the coincidence circuit, shown in Fig. 5, designed to give a signal when the output of a digitizer and a switched-in number are identical. Here a two-stage counter is used to derive four pulses for the four functions of sampling the digitizer and pre-set number, providing a signal when coincidence occurs and clearing the system ready for the next reading. The pulses are selected by gates connected to the counter transistor collectors. The first one is used to sample the digitizer; the second interrogates the switches in which the selected number is coded; the



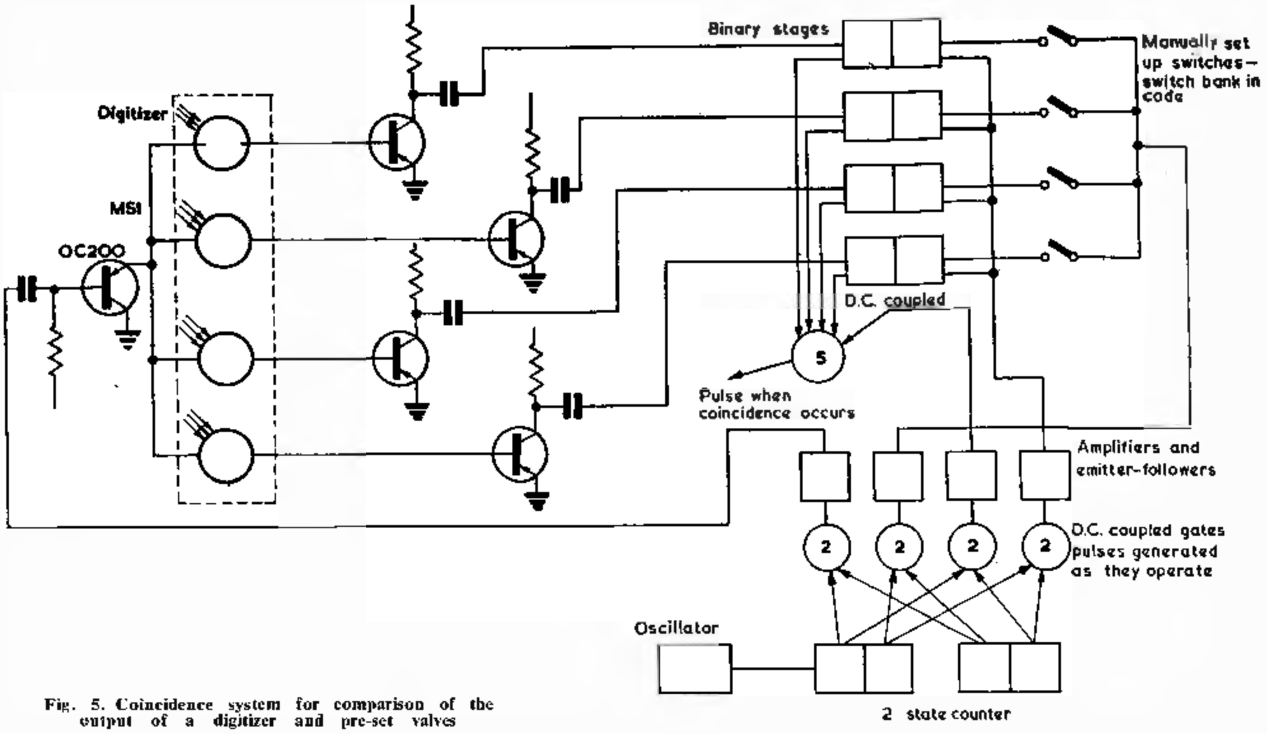


Fig. 5. Coincidence system for comparison of the output of a digitizer and pre-set valves

third is fed to the output gate and only appears at the output at coincidence, while the fourth returns to the system as its original state ready for next sample. Obviously, only if a bistable circuit receives a pulse from neither the digitizer nor the switches, or alternatively from both, will it be in its original state when the third pulse arrives. Only if all are in this state will the third pulse pass through the five-way gate.

With digitizers from which the current output is very low so that a more refined chopping and amplifying technique is required, it is not possible to chop the outputs from all the cells of a digitizer with a single transistor. In this case the chopping transistor is in parallel with the cell, coupled via a capacitor to the amplifier (Fig. 6)[9]. When the transistor is switched to the conducting state the capacitor is discharged and a signal pulse produced. The maximum repetition frequency with which the cells may be sampled falls with their output signals but at the minimum signal allowed (50mV with the cells open-circuited), this frequency is about 10kc/s. Sample pulse widths of about 5μsec are normally used. Two stages of amplification using Mullard OC44 transistors give 2V of output signal across 2·7kΩ from the standard open-circuit input of 50mV.

It is again possible to feed several digitizers into the same amplifier, though each cell must, of course, have its own chopping transistor. As in the previous arrangement the cells can be sampled in turn to provide a serial output.

Though several types of instrument have been made which give the low current output which has to be handled by the multi-stage amplifier described above, only one gives the 20μA minimum light on signal needed for the other system. This is a relatively crude design using a 4·5in diameter disk with tracks pitched 0·060in centre to centre. The cells are mounted in groups of four behind individually adjustable slit plates. Though it is very undesirable in principle, one lamp is used for each cell (12V, 0·1A aircraft panel bulbs). This is because the form of the coded disk is such that it is difficult to obtain a single bulb which will cover all the four cells in a group.

A better arrangement is that used in a digitizer based on the same disk as that used for the small Schwarz cell units. In this device the filaments of three festoon lamps are imaged on the slits in the mask disk. The outputs from the cells are very much lower in terms of current than those of the previously mentioned type. Another design uses a scale wrapped round a transparent drum[10]. With a standard linear coded scale any desired angular range of measurement less than 350° can be obtained by simply choosing the diameter of the drum correctly.

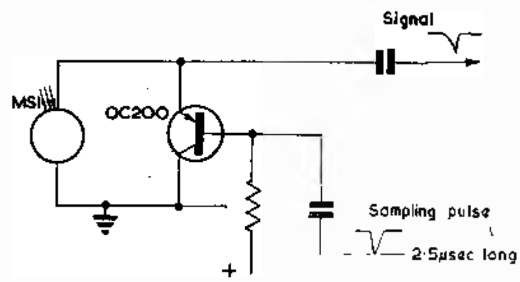


Fig. 6. Sampling arrangement when cell signals are small

An Approach to High Resolution Digitizers

The foregoing description has been concentrated on low resolution equipment with the emphasis placed on the philosophy of reading the instruments. The codes used have, as was indicated, been assumed to be standard cyclic binary and binary coded decimal arrangements. In this section, a few remarks will be made about the choice of scales for high resolution digitizers, though little about their physical realization.

If a standard cyclic progressive binary code were used in a digitizer having a resolution of 1 in 2^{17} this would involve an edge accuracy of, at worst, $\pm$0·00005in for each of the 131 072 edges together with (in practice) no pinhole or blemish on the disk greater in area than 5×10^{-7}in^2

using a code outside diameter of 5in. Further, with these tolerances all the edges must be really well defined, which means in practice using a medium for the pattern with a resolution of about 4 000 to 5 000 lines/mm. These requirements, cumulatively, mean that even high resolution photographic emulsions are very nearly impossible to use. If silicon cells are used, too, the upper temperature limit at which the digitizer is potentially capable of operating may well damage an emulsion.

Trouble due to blemishes and incorrectly formed edges can be minimized by averaging over a number of digits. On tracks in which the pattern repeats itself in a small physical distance, (e.g. 2^0, 2^1 tracks of a cyclic progressive binary code), it is obviously possible to arrange a series of slits with the same period (or an appropriately magnified one if an optical system is used) as the code and read them all with a single cell. This is, however, impracticable with the most significant tracks of a cyclic disk, and both the reliability and signal size are much less (as no slit should exceed the width of a digit of the scale) in the channels associated with them.

The most direct approach, therefore, to this problem is to use diffraction gratings and count Moiré fringes. It is easy to obtain a great deal of averaging—so much so that the existence of any given ruling is irrelevant. The disadvantages have already been indicated, though some of these can be overcome to a great extent by associating with the grating a series of reference marks placed accurately so that a known number of counts occurs between each. Assuming then that the count starts from zero at one such mark, and that, for example, there are 100 counts between each mark, the desired state of the stages representing the last two decades is known at each mark. As long as the number of errors does not exceed 99 (in this case; $(2^n - 1)$ excess, or 2^n missed counts in the case of a binary system with an interval of 2^n digits between marks), the count can be corrected. Even if the check marks do not occur at a count of zero, correction, though relatively complex, is possible.

Derivation of the check signals is complicated by the need to obtain check signals of a comparable amplitude to

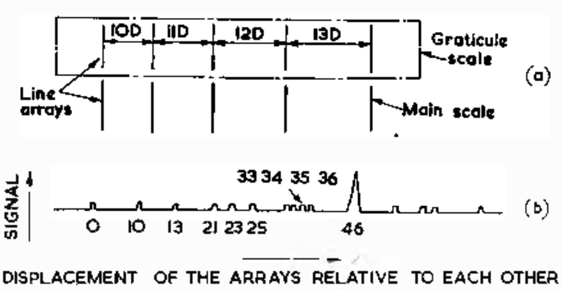


Fig. 7. System for deriving check pulses

those from the detectors which are operating off the grating. One approach to this is indicated in Fig. 7(a) in which the two arrays of lines are to be taken as moving over each other. If the spacing of the lines in the groups is suitable, a wave train of the type shown in Fig. 7(b) is obtained in which there is one large unambiguous signal. The ratio between this and its neighbours depends on the number of lines in the groups and their spacing.

The width of the lines has to be less than that of the smallest digit of the scale—that is, a quarter of the pitch of the grating (from which four counts can be derived). This approach is limited primarily by diffraction as the gap between the circle and the graticule with the array of lines on it quickly becomes much greater than the width of one of the lines as the resolution of the grating is increased. The useful limit of line width is about 0·0005in.

Another variation is to use these check marks to trigger the detectors of a coarse coded scale which numbers them. The scale need not be cyclic as it is simple to ensure that the code edges are kept well away from the positions at which the check marks occur. This gives a relatively simple system, but one with a low accessibility, and one in which the check system can be a serious limitation on speed

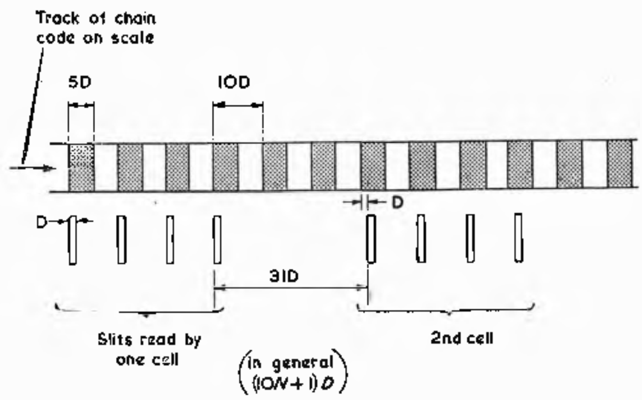


Fig. 8. Chain code with multiplicity of reading slits

of operation at high resolutions as a simple design of photoelectric microscope is needed.

Following the above principles of scale design, a reasonable approach is to use a chain code arrangement[11] (Fig. 8) in which a number of detectors view the same track of a coded scale through slits spaced at a distance apart equal to a whole number of pitches of the scale plus a fraction equal to $\frac{1}{2 \times \text{number of detectors}}$. Each detecting system can consist of a number of slits, as described elsewhere[11], and some measure of averaging obtained. This method is suitable for only one or two decades (of a binary coded decimal scale) as, first, the interval between consecutive slits of the same detector group becomes too great, and, second, the number of detectors needed for either the cyclic or choosing systems of avoiding ambiguity is considerable (10 per decade).

In any case, as this type of scale can in practice only be made by photographic means—there are always several tracks with different pitches—the requirements for high edge definition is still difficult to meet, though the problem of blemishes on the scale is minimized. This type of scale has been made to read with a resolution 0·0002in but only with very critical adjustment of amplifier triggering levels, as, due to the degradation of digit edges, the light reaching any one cell changes in intensity during a movement of greater than one digit, and variations in cell sensitivity become, thereby, very apparent.

The simplest methods of achieving very high resolutions are those based on an idea evolved at the Royal Aeronautical Establishments[12] in which the scale consists primarily of lines equal in width, ideally, to that of a digit and accurately spaced at a distance of a hundred or more digits apart. The lines are imaged by a lens system on a coded graticule, the length of which is equal to the line interval times the optical magnification. The lines may be engraved by a cutter, and can therefore be of very high definition and an adequately blemish-free background obtained without too much difficulty. A straightforward choosing system is necessary to 'number' the lines and, as it is, in general, impracticable to maintain the accuracy of the ruled scale in this section, pairs of detectors, chosen from the high precision scale in the conventional way, must be used to read its tracks.

The system described by Ballantyne[12] contains an optical scanning device and is not, therefore, of very high accessibility nor of great flexibility as the output must be in serial form. A purely experimental device, with a single revolution resolution of 1 in 2^{17} (131072) in an effective disk diameter of 4½in has been made, which has all the flexibility associated with the use of silicon photocells in parallel, except that, as is usual with the technique used in it, part of the output has to be obtained by selection from pairs of cell signals. This need not limit the accessibility of the disk information, but may result in a delay during the manipulation of the output. Duplication of amplifiers is, however, then obligatory, and, as all signals are inevitably small, though the desired signal to background ratio is still comfortably high (4 or 5 to 1 is a typical figure), this can involve a considerable amount of equipment.

Conclusion

The design of all the instruments which have been described has been conditioned by the idea that their main use will be in systems in which they are interrogated for short intervals rather than read continuously. By this means the often extensive information handling equipment needed can be used to best advantage, and at the same time kept to a minimum. This aspect is, of course, less important with high resolution instruments, as these units would certainly be used for measurements which could be done no other way with true digital equipment and would generally be part of a considerable system.

In the case of low resolution units, most of the functions they perform can be done using electronic analogue to digital convertors. However, if it is unnecessary to convert electrical signals into shaft rotations in order to obtain a digital reading, it is also unwise to turn shaft rotations or linear movements into voltages in order to measure them. It is desirable, then, that the output of the two types of transducer should be compatible to allow them to be used together in systems. The Schwarz cell type of instrument is unsatisfactory in this respect, because of the relatively long sampling times, and rather indeterminate shape of the signals. A digitizer containing silicon cells can, however, be sampled, either in series or parallel, using the same clock oscillator that is used with the electronic convertor. Consecutive readings, one from a shaft digitizer, the other from an electronic one, can then be passed to the handling equipment in exactly the same form.

In the range of instruments, described—obviously the ideas can be applied in a very wide range of forms, both linear and circular—all the designs have been conditioned by the concept of accessibility. It is for this reason that systems such as those using micrometers, the check mark system described previously here, and the simple ambiguity avoiding system using a track to prevent readings being taken near code changeovers, have been avoided. Ultimately, a very fast frequency of response from the cells is necessary to obtain high accessibility—and in this respect Schwarz cells are unsatisfactory, though they allow a very simple instrument design.

The problem of detecting failures—except the most common one of a lamp breakdown—is difficult. In these units error checking codes have not been used—as they involve extra tracks (and so space on the disk) and cells and complicate the information handling equipment very considerably. On a straight-line scale, or one which is not complete through 360°, the problem is simple as a region in which, for example, all the cells should receive light can be incorporated, and occasional excursions made to the region for test. Otherwise, it is necessary to run the digitizer through its range of measurement checking the behaviour of each cell. As the subject is so vast, the problems of error determination will not be discussed in detail in this article.

Acknowledgements

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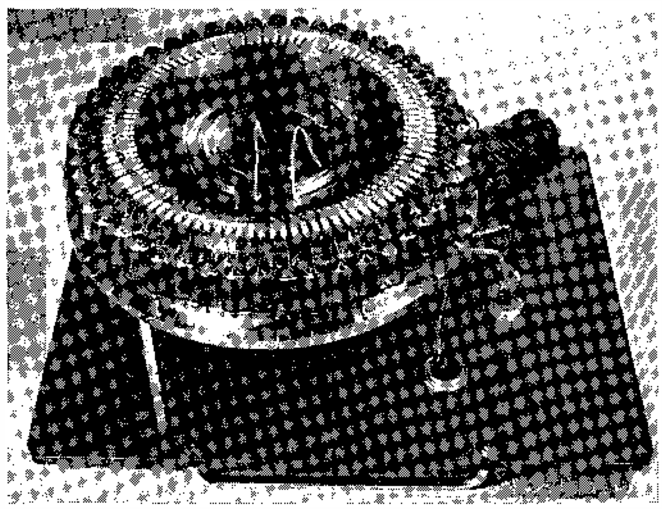
A Distance-Velocity Lag Simulator

This instrument, designed by Elliott Automation Ltd., is intended for use in conjunction with analogue computing equipment for the representation of large transit time effects which occur in a number of industrial processes, usually associated with the flow of fluids in ducts. The transport time provided by the instrument can be varied between 2 and 60 seconds, and is inversely proportional to the magnitude of an electrical input signal, so that it may be either preselected to a constant value, or caused to vary in relation to other plant parameters represented in the computer. The input and output signals corresponding to the transported information may be in the range of ±100V and between d.c. and the cut-off frequency which is dependent upon the transport time.

The instruments consist basically of a 100 position motor driven switch having two wipers, one of which is connected to the output of a "writing" amplifier, and the other to the input of a "reading" amplifier. The motor, whose speed is servo controlled, drives the switch continuously in one direction, through a reduction drive, the time taken for one complete revolution being equal to the distance-velocity lag. To each of the 100 fixed contacts is connected a 22,000 pF polystyrene capacitor, which is successively charged to the "writing" amplifier output level as the switch rotates.



The Effects of Encapsulation on Electronic Components

By F. P. Campbell*, B.Sc.

Potting of electronic circuits has long been known to affect their performance. A survey is made of the effects of encapsulation on commonly used components.

Valves may be encapsulated safely provided that the correct techniques are used to remove self-generated heat and to protect the glass envelope. Little difficulty is encountered with transistors and capacitors. The effects of potting transformers and chokes are described. Wirewound resistors are not suitable for casting in rigid resins when heavily dissipating.

The potting of carbon and cracked carbon resistors may affect various manufacturer's products differently; an experiment is described. The causes of component change are summarized.

(Voir page 394 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 400)

IT HAS long been recognized that the encapsulation of electronic assemblies causes changes to take place in their electrical parameters. For example, when the first cold-curing pourable resins became available in the late forties and early fifties, it was believed that components would be most affected by the increased permittivity and loss angle of the resins compared with those of air. Early confirmation of the reduction in Q of coils, particularly at the higher frequencies, was soon available[1]. Nevertheless, there was a feeling that potting would solve most reliability problems.

Since those days many manufacturers have used potting as a constructional technique on a large scale. Some have subsequently dropped it in favour of a return to more conventional construction. Others have restricted the use of potting to some parts only of their equipments. Finally, there are those who believe that potting techniques, when correctly applied, bring about an overall improvement in the performance of equipments under severe environmental conditions, and who therefore continue to pot wherever possible.

Why have these changes in outlook occurred? The answer lies in the increasing knowledge of the behaviour of potted components which has been gained in manufacture and test. This article is an attempt to summarize that knowledge.

General

Each type of component is treated under a separate heading. The potting effects discussed are mainly those of the polyester and epoxy resins, since these are believed to be the most widely used in this country. Mention is made, however, of silicone rubber and polysulphide rubber encapsulation where the effects are known. Dip-coating and foaming techniques are not yet sufficiently widespread for their effects to have been assessed. Emphasis is laid on the potting of carbon and cracked carbon resistors in polyester and epoxy resins. This is partly because these components are among those which are most markedly affected by encapsulation, and partly because more evidence is available.

It is inevitable that the means of overcoming potting troubles should be discussed in the same context as the effects of potting on components. It is the opinion of the author that with intelligent design the injurious effects of encapsulation can generally be rendered harmless.

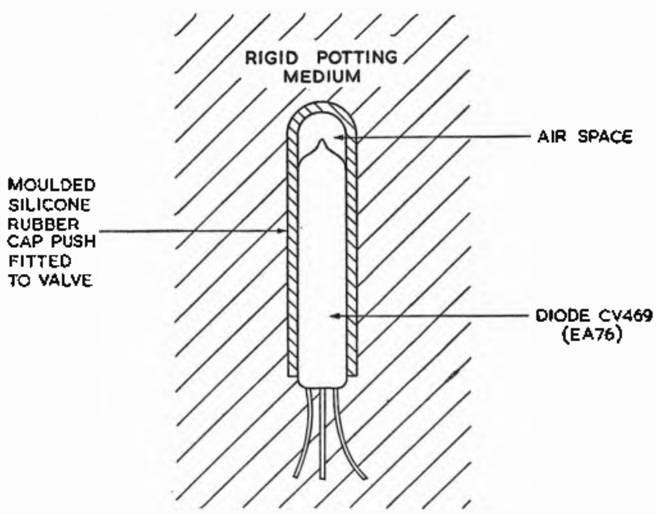


Fig. 1. Potted subminiature diode

Valves

Valves are encapsulated for two principal reasons:—

(1) To provide a more rigid mounting than that offered by many types of holder.

(2) To bring up the envelope diameter to a closer tolerance than that achieved by the glass-worker.

The main difficulties are the removal of heat to ensure that the valve and potting medium do not exceed their maximum temperatures, and the effects of differential expansion and shrinkage.

Since the expansion coefficients of polyester and epoxy resins, even when heavily loaded with quartz or mica flour, are some four times that of glass it is likely that the use of a rigid resin will result in breakage of the envelope at extremes of temperature. Attention has therefore been concentrated on cold-curing silicone rubber

* *English Electric Aviation Ltd.*

as a potting medium. It is better than the various epoxies by virtue of its high maximum working temperature (~250°C) and its excellent electrical properties. Some ways in which it has been employed are shown in Figs. 1 and 2.

The Fig. 1 method was used when it was required to include subminiature diodes CV469 together with other components in a block of polyester resin. Since the compressibility of silicone rubber is low it is necessary to leave an air-space inside the cap to allow the rubber to flow. This was found to be quite successful so long as the temperature rise at the envelope was not too great.

In Fig. 2 is shown a technique for mounting subminiature valves which has been used both in America and in the U.K. where high vibrational stresses were encountered. A silicone rubber envelope is injection moulded around the valve. This is now treated as a sub-assembly which is finally pushed into a hole in a light-alloy block. In a typical case the temperature rise at

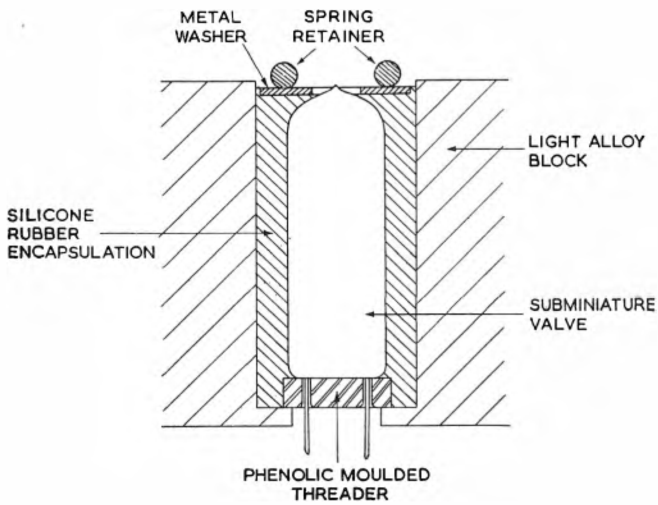


Fig. 2. Encapsulated subminiature valve

the glass envelope was only 23°C, compared with 82°C for a plain valve suspended in air[2]. The technique of Fig. 2 has also been applied successfully to miniature valves. Some curves of temperature rise against time are shown in Fig. 3. Here the differences between the encapsulated and clip-mounted envelopes are seen sometimes to be positive and sometimes negative. The rubber tends to even up the envelope temperatures.

Semiconductors

Germanium and silicon devices have been potted for a number of years with few failures. It is important that, in the case of germanium semiconductors, the temperature during the exotherm does not exceed about 70°C and that the curing temperature be limited accordingly.

It is common to fit semiconductors with a synthetic rubber sleeve in order, so that argument runs, to minimize stressing of the component when the potting resin gells and contracts. Why this should be effective when both ends of the component are left uncovered is not clear, but there is evidence of failures due to glass breakages when the sleeve is not fitted. The sleeve is useful in preventing removal of the paint from some types of transistor, so minimizing photo-electric effects.

The usual precautions must be taken to ensure that self-generated heat is removed from the block.

An interesting difficulty can arise when transistors mounted only by their leads are potted. Since the leads are very thin and the component is light and held only at one end the pouring or injection of the resin can move the transistor so as to cause short-circuits to neighbouring components and between the lead-out wires.

In general, however, the characteristics of the semiconductor are not affected by encapsulation.

Components Sealed in Containers

The potting of sealed relays, switches, potentiometers, etc., seldom affects the performance of the component. The sealing method should be scrutinized for compatability with the resin chosen (e.g. a wax seal may be melted by the exotherm).

When a sealed component is large in relation to the potted block there is a danger that stress concentrations may be set up when the resin gells. Fig. 4 shows a potted

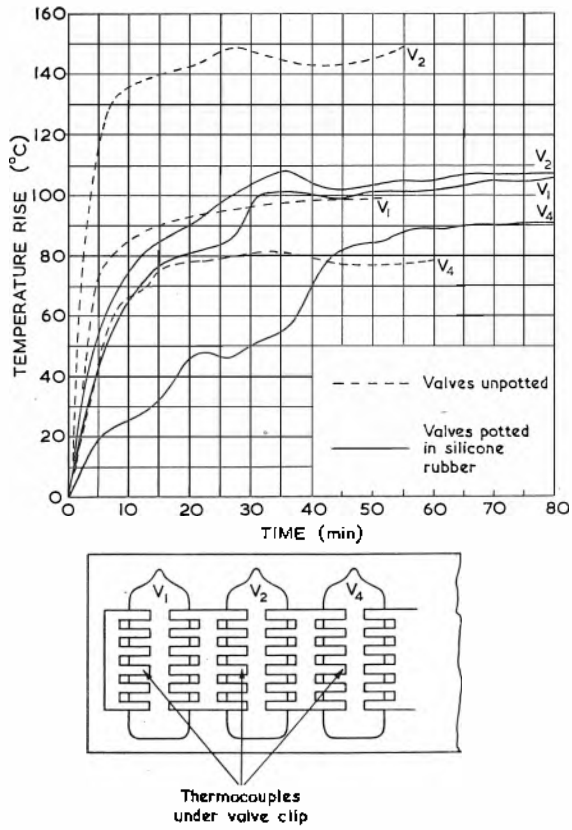


Fig. 3. Temperature rise curves for miniature valves

unit containing a large capacitor and a delay line. The block has cracked due to insufficient wall-thickness of resin and the outlines of the large components are clearly seen. As well as damaging unsealed components such cracks become moisture traps in highly humid atmospheres.

Capacitors

Provided that the capacitor is fully sealed when it is potted there is seldom any effect on its value or rating. This applies not only to the metal-cased types but also to the sealed-in-Bakelite and sealed-in-resin varieties. The results of ageing and climatic cycles on various types of capacitor potted in Araldite D were quoted in an M.O.A. Report[3]. In no case did the deviation appear to be greater

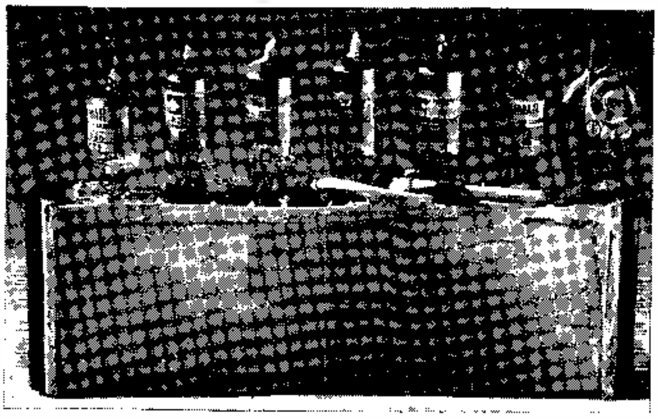


Fig. 4. Potted block containing large components

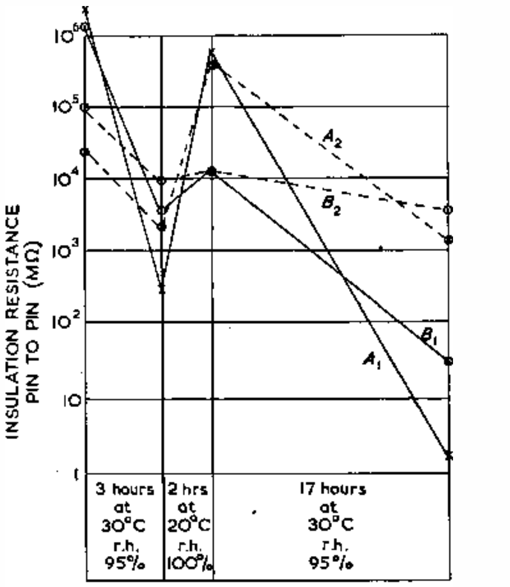


Fig. 5. Insulation resistance of mated connectors

A_1 : *phenolic moulded connector*
A_2 : *as A_1 with terminations encapsulated in polysulphide rubber*
B_1 : *polyethylene moulded connector*
B_2 : *as B_1 with terminations encapsulated in polysulphide rubber*

than that expected of an unpotted component, and all were within RCS limits for drift by a large factor.

In order to save space it is possible to encapsulate capacitors which have been supplied without outer containers. This procedure is usually satisfactory, provided that the component has been stored in a perfectly dry atmosphere up to the instant of potting. It is seldom easy to achieve this in practice.

Connectors

Connectors are encapsulated to provide a seal against moisture and to prevent undue movement of the lead-out wires. A medium often used is polysulphide rubber. This has excellent electrical properties and good adhesion to most metals, so that a seal is formed at the shroud. Some measurements of insulation resistance under humid conditions are shown in Fig. 5. Even with a polythene plug and socket (to which this encapsulant does not bond) an improvement in insulation resistance of several orders is obtained.

Transformers and Inductors

The effects of potting may be divided into those affecting the core and those affecting the windings.

The core may be affected in three ways—change of magnetizing current due to alteration of the effective gap, change of permeability resulting from suppression of the magnetostrictive movement, and damage due to differential expansion.

Taking the first effect, it is found that if the gap is uneven it is possible for resin to creep between the core faces and, on subsequent temperature cycling, to alter the magnetic reluctance of the gap. This is particularly the case in cores where a high inductance is obtained by having face-to-face contact without any separating shim to control the gap. It affects both iron and ferrite materials.

The second type of core defect is observed in nickel-iron cores such as Radio-metal and Mumetal. Their change of length with change of magnetization is prevented by a rigid potting medium. As a result the magnetization current rises. The effect is dependent upon temperature, the magnetizing current being inversely proportional to temperature[4].

The third type of trouble is particularly severe on ferrite cores. There is generally good adhesion between the resin and the core, and the expansion coefficients are different enough to cause fracturing of the ferrite. All these difficulties have been largely overcome by the use of a pre-potting core encapsulant.

The effects of potting on windings usually arise from strains set up in the block as it gells. Fine wires can be broken or displaced, even if they have previously been vacuum impregnated. Pre-impregnation, however, will usually mimimize this effect.

The effect on the Q of coils is not usually noticeable at power and audio frequencies. At radio frequency the power factor of the resin can make a marked difference. Some typical figures after 7 days are[1]:—

5 per cent	reduction in	Q at	30Mc/s.
17 ,, ,,	,, ,,	Q ,,	45Mc/s.
36 ,, ,,	,, ,,	Q ,,	60Mc/s.

Fig. 6. Wirewound resistor—longitudinal split

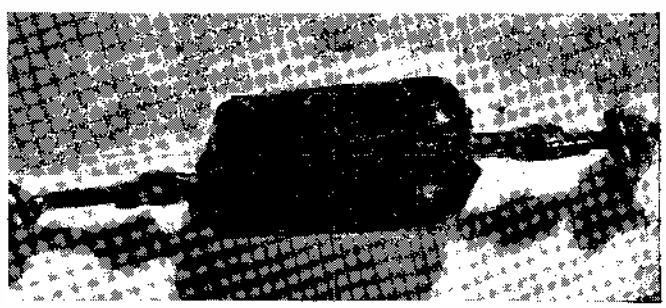
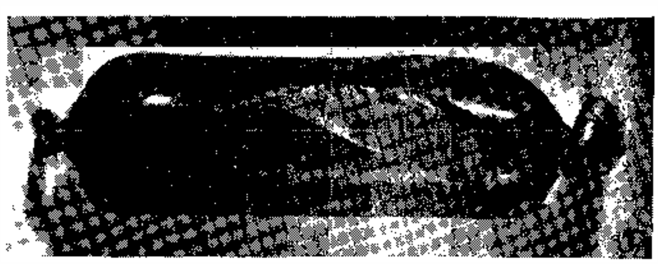


Fig. 7. Wirewound resistor—rupture of vitreous enamel



For a 45Mc/s coil potted in polyester the Q fell by 18 per cent in four weeks.

The dependence of loss angle on time arises from the fact that the cure continues for days or weeks after gelation. This could be an embarrassment in production testing, but it is seldom met with in practice since video frequency circuits are not often potted in resins because of their high permittivities (approximately 3·5).

Wirewound Resistors

It is not considered good practice to pot wirewound resistors.

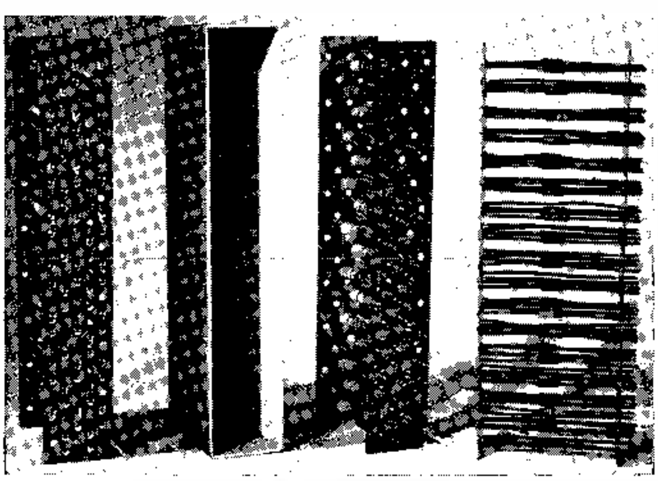
Because of their high surface temperature—as much as 400°C when fully loaded—differential expansion effects will be very severe. That type of resistor which consists of a hollow ceramic tube with open ends is particularly prone to failure. Resin enters the core and since its expansion coefficient is some six to eight times that of the ceramic the latter is usually cracked, so fracturing the fine-wire resistive element.

Where wirewound resistors are used merely as stable circuit elements and do not dissipate appreciable heat they can, of course, be satisfactorily potted, the most suitable types being those with closed ends.

Where it is essential to include heavily dissipating wirewound resistors in a block of rigid resin a protective sleeve of silicone rubber has been found useful in minimizing failures. This is because it breaks the excellent bond which occurs between the vitreous enamel coating of the components and some resins, especially the polyesters. If the component is not sleeved expansion causes the enamel to craze or fracture, and the resistance wire breaks. Photographs of resistors which have so failed are shown in Figs. 6, 7 and 8.

Fig. 8. Wirewound resistors—rupture of vitreous enamel and circumferential cracking

Fig. 9. Resistor potting experiment—form of construction



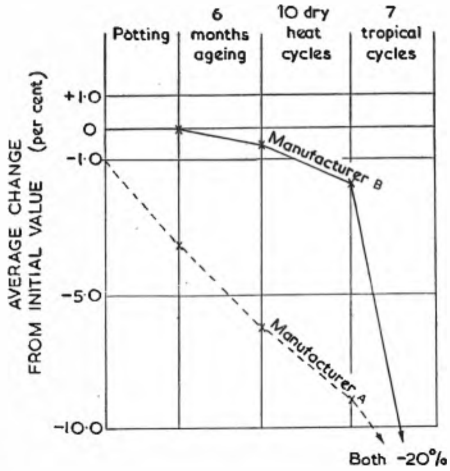


Fig. 10. Temperature cycling—RC7K resistors potted in polyester resin

The following figures may be of interest. In a block of Araldite D with quartz filler, six 6W (R.C.S. rating) resistors were cast. Temperature rises at three points were measured with full dissipation in all resistors:—

	Before potting	*After potting*	*Difference*
Surface of an outer resistor	218°C	103°C	−115°C
Resin ¼in from two resistors	82°C	137°C	+55°C
Surface of an inner resistor	240°C	159°C	−81°C

The very considerable moderating effect of the resin is seen.

Carbon and Cracked-Carbon Resistors

One way to approach this large subject is to describe an experiment carried out by English Electric Aviation Limited with considerable assistance from the RRE Components Division.

Fourteen blocks each of 40 resistors, mounted as in Fig. 9 were potted in quartz-filled polyester resin. Six blocks contained Style RC3N resistors of various values, i.e. a total of 240 resistors. Four blocks contained equal numbers of Style RC7K resistors by manufacturers *A* and *B*. Four blocks contained equal numbers of style RC2 resistors by manufacturers *C* and *D*. In addition, two blocks were cast in Araldite, each containing RC3N resistors. The main purpose of the experiment was to observe the behaviour of resistors, particularly RC3N's, when potted in polyester and to compare their behaviour in Araldite.

All resistance values were measured to four significant figures immediately before and immediately after potting. Eight blocks, representing all resistor types, were stored for six months and then subjected to ten dry-heat cycles between 20°C and 85°C. Measurements were taken at various stages.

Six blocks underwent tropical exposure cycling. Resistance measurements were made at the end of each cycle. The two remaining blocks did not survive potting.

An analysis of these results, totalling some 4 000 measurements, was made. The blocks were opened up and examined. With the addition of the failures data gathered over two years by the English Electric Guided Weapons Reliability Group, it is possible to make the following statements.

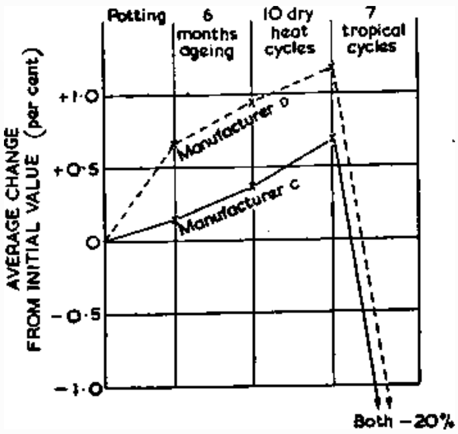


Fig. 11. Temperature cycling RC2D and RC2E resistors potted in polyester resin

circuits using Grade 2 resistors these changes would not be critical.

No clear explanation of these effects is advanced. There are two main possibilities—absorption of moisture and stressing of the terminations. In support of the latter it is known that the two manufacturers of these resistors use different means to secure the terminations internally.

On the other hand, Church[5] shows that moisture is the principal cause of drift in this type of resistor. The explanation of the large changes seen in these tests may be that moisture is alternately absorbed and driven off. Support is lent to this point by the description of a test on ten RC7K resistors[3]. Here they were encapsulated in epoxy instead of polyester resin as above. The epoxies generally have a lower moisture permeability than the polyesters. The mean change of resistance after 14 climatic cycles was only +0·17 per cent. The number of components tested was small however, and not too much significance should be attached to this result.

It remains to describe a peculiarity observed in the main series of tests. One of the four blocks containing RC7K's was found after potting to have its resistors faulty, all the resistors by manufacturer *B* becoming open-

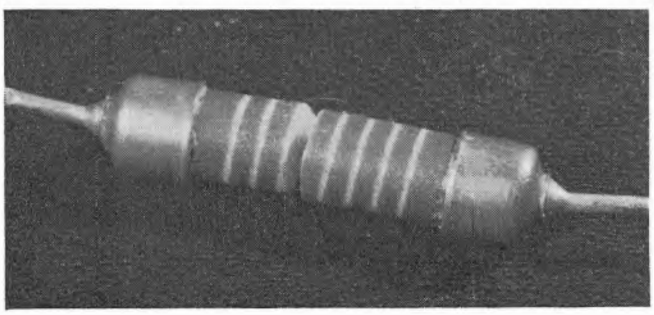


Fig. 12. RC2D resistor—fracture

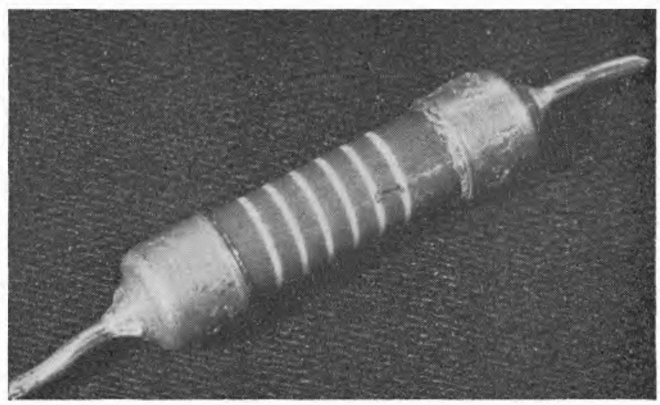


Fig. 13. RC2 resistor—arcing of longitudinal crack

circuit and all *A*'s falling to between 1 per cent and 10 per cent of their values before potting. The resin was dissolved away and the resistors were found to have regained their correct values. It is postulated that an incorrect mix must have been used in this case and that it failed to gel. The chemical effect of the resin on the varnish coatings of the resistors could account for the partial short-circuits at least, but not for the open circuits.

RC7K Resistors

These are carbon resistors having an insulating phenolic casing. They are rated at ½W by the R.C.S.C.

In Fig. 10 are shown the results of the tests. These indicate that:—

(1) There are significant differences in behaviour between resistors by the two manufacturers.

(2) The immediate effect of potting is considerable in the case of manufacturer *A*.

(3) The average effect of ageing on resistors by manufacturer *B* is not great. The figures for average drift do not, however, show the behaviour differences of individual components. Some tended to return to their initial values, while others continued to drift.

(4) The effect of dry-heat cycling is to accelerate drift.

Some resistance values were much less stable than the aggregate. For example, A's 100Ω nominal resistors decreased by up to 30 per cent during potting. Some values on the other hand were very stable. In most

RC2D and RC2E Resistors

These are Grade 1 cracked-carbon non-insulated resistors of ¼ and ½W RCS ratings.

Fig. 11 shows the test results. The picture here is a very different one;

(1) There is no significant difference between components by the two manufacturers.

(2) The effects of potting, ageing and dry-heat cycling are not great. Most of the resistors kept well within the RCL limits for drift. There were nevertheless a few exceptions. For example all the 100kΩ resistors by manufacturer *D* had higher drift rates than the mean, both during potting and subsequently.

(3) There was one catastrophic failure out of 120 resistors tested. This was a 4·7kΩ RC2E which went open-circuit at the second dry-heat cycle.

There is again no theory to account for the changes which did occur. The fact that these components are not

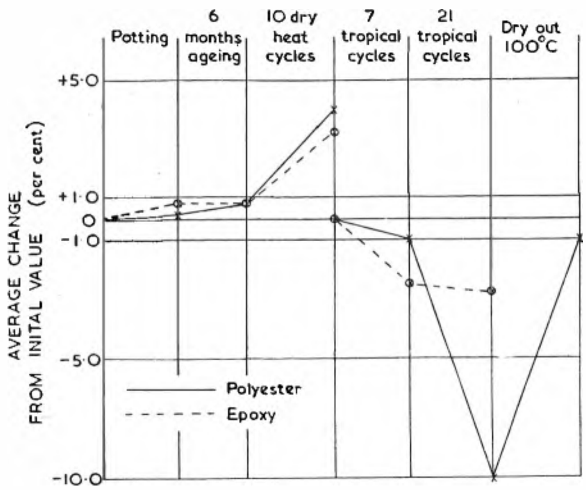


Fig. 14. Temperature cycling -RC3N resistors potted in polyester and epoxy resins

Fig. 15. Insulated cracked carbon 820k resistor—blotched carbon film

insulated and are only protected from the resin by a thin varnish coating is very interesting but it does not really help one to formulate a theory.

Failures in production equipments have been low. The open-circuit component found in the tests was caused by the resin cracking at that point. A photograph of a similar case is shown in Fig. 12. The resistor has sheared across the centre. Another production defect is shown in Fig. 13. Here the surface of the resistor was cracked longitudinally, and arcing between spirals has taken place at that point. Again the original trouble was a block which cracked at a weak point.

RC3N Resistors

These are Grade 1 cracked-carbon resistors of ¼W R.C.S. rating, having a ceramic insulating sleeve which is end-sealed with a cement.

The behaviour of some 320 components is shown at Fig. 14. These resistors are seen to be much less stable than the RC2 styles. In view of the protection afforded by the ceramic sleeve the RC3N's might have been expected to be better. The higher drift rates are accounted for by a few critical values which changed by large amounts when temperature cycled. Thus, 27kΩ resistors both in epoxy and polyester increased by amounts up to 35 per cent. 8·2kΩ resistors were nearly as bad. Yet some higher values, e.g. 1MΩ, were extremely stable during the series of tests. The seriousness to circuit designers of these changes is obvious.

The difference between polyester and epoxy is only slight so far as the dry-heat cycling is concerned, but the effect of tropical cycling becomes very marked after seven cycles. The moisture absorption of the polyester is adversely affecting the resistance values.

Three catastrophic failures occurred, one during potting and two during dry-heat cycling, that is, about 1 per cent of the total.

To account for these failures and for the gross changes in value of certain resistors it has been suggested that the high temperatures experienced during the exotherm and subsequent cycling had caused exudation of the end-seals, so giving rise to voids and stressed regions in the block.

Resistors which have failed in potted blocks often show some lifting of the carbon spiral, as seen in Fig. 15. It is by no means clear how this can be caused by local stressing of the block. The effect is most marked when the width of the spiral resistance track is least, as one might expect.

Conclusions

To summarize, the causes of component change are:—

(1) Stressing caused by:

(a) Shrinkage on to component when resin gells.

(b) Fracture of resin block due to faulty potting, poor layout or extremes of temperature.

(c) Differential thermal expansion.

(2) Temperature effects causing deterioration of components.

(3) Moisture absorption into block.

(4) Effect of increased permittivity and loss angle.

In reading this sad survey of failures and defects one may come to feel that encapsulation is a dangerous technique. It must not be forgotten, however, that the failures are few in relation to the number of components encapsulated. With the increasing knowledge of the effects of the potting media the number of defects will fall, and a constructional technique which has already brought great benefits where high environmental stresses are encountered will come to be more widely used.

Acknowledgment

Acknowledgments are made to the author's colleagues, particularly Mr. W. T. Truscott, for help and information; to the Chief Engineer of the Guided Weapons Division, English Electric Aviation Ltd., for permission to publish this article; to the Royal Radar Establishment, Gt. Malvern, for permission to quote from Ministry of Aviation reports; and to the Electrical Research Association.

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A Stabilized and Programme-Controlled Power Supply Using Positive Feedback

By M. Pacák*

A stabilized power supply for a mass-spectrometer magnet is described. The output is continuously variable from 3 to 800V or it may be programme controlled by means of an RC device. Positive feedback is used in the amplifier of the stabilizer.

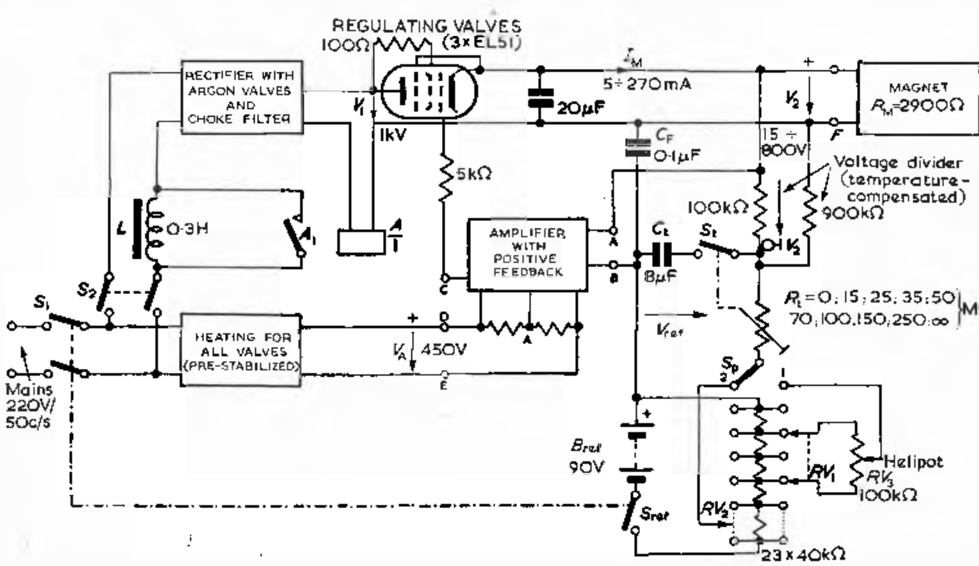
(Voir page 394 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 400)

FOR the main-magnet circuit of a Nier-type mass spectrometer a stabilized supply was required with the facility for controlling the current value over a wide range. It was necessary to be able to control the current easily and precisely manually, or automatically with the change of the current between two preselected limits disposed anywhere within the range. The automatic control had to be continuous, nearly linear and sufficiently slow for the direct ink-recording of the mass spectra. The temperature change of the magnet coil resistance is irrelevant in this case. Therefore the supply might be realized as a programme-controlled voltage stabilizer. The current value had to be adjustable between 5 and 270mA. The magnet resistance was 2·9kΩ, hence the voltage range required was 15 to 800V. This wide range together with the high degree of stabilization and the automatic control required represent rather strict demands on the instrument performance.

In the circuit suggested (Fig. 1) the same regulating system is used both for the stabilizing and the programming function. A part of the output voltage V_2 is compared with the reference voltage V_{ref}: the difference is amplified and it controls the regulating tubes in a well-known manner. V_{ref} is supplied by the battery B_{ref} shunted by a step voltage divider with twin multi-way switch RV_1 and RV_2. In position 1 of the switch S_p an arbitrary value of V_{ref} can be set from 0 to 90V. As $V_2 \simeq 10V_{ref}$, any value of V_2 between 3V and the upper limit can be set. In position 2 of the switch S_p one of 24 equally disposed discrete values of V_2 can be selected as the second limit for the programme-control action. If S_p is switched over from one position to another, V_{ref} and therefore V_2 and I_M change from one limit to another and the limits are defined by the positions of RV_1 and RV_3 and RV_2. This change is exponential, lasting some tenths of a second, if the switch S_t is open and if the step resistor R_t is set to zero. In another case the capacitor C_t and the resistor R_t constitute a delay circuit with the time-constant, which may be adjusted between 120 and 2 000sec. Thus the automatic control is realized obeying the exponential law. For $R_t \to \infty$ the capacitor C_t loses its charge through its own dielectric resistance and the value of V_2 changes then quite slowly, the natural time-constant of C_t being of the order 10^4sec. In such a way the scanning of a peak can virtually be stopped for some seconds, if it is appropriate for the purpose of recording. It is important to realize that the delay circuit acts only in programme-control and does not delay the transfer of variations to the amplifier input. Hence the proper stabilizing function of the circuit correcting for the deviations caused by the disturbing effects is not delayed.

The amplifier (Fig. 2) contains two pentode stages in cascade; their symmetrical arrangement results in a better stability of the d.c. circuit with respect to heating and auxiliary voltage (V_A) variations, etc. The voltage amplification, with the positive feedback switched out, including the input divider is approx. 200. The first stage of the amplifier involves the positive feedback, which makes it possible to obtain the qualities described previously[1], especially the possibility of compensation to obtain zero deviation in the static output and zero output resistance for lower frequencies. This is the second main advantage of the circuit proposed. The third one is the compensation for the V_A variations which is achieved by the bridge arrangement of the circuit supplying the amplifier and is finally adjusted by the potentiometer RV_4. This method, suggested by the author[2], practically eliminates the necessity for stabilization of the auxiliary voltage which supplies the amplifier even in high-grade stabilizers. The following features are also incorporated in the circuit. The impedance of the choke L (Fig. 1) is automatically switched in the primary lead of the main transformer, if the current I_M goes below 70mA, and it is short-circuited, if I_M rises above 130mA; thus the excessive rise of the rectifier output voltage V_1 is eliminated. A cathode-follower is used for

Fig. 1. General arrangement of the programme-controlled stabilizer



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the output of the amplifier (Fig. 2), thus separating the sensitive high-resistance output of the second stage from the disturbing effects of the grid currents which frequently occur in regulating pentodes. The capacitors C_1, C_2, C_3 (Fig. 2) secure the dynamic stability of the system in a rather compressed and incompletely shielded amplifier. These capacitors are placed so that they do not raise the response-time substantially. Some special features omitted from the diagrams for clarity are as follows: A quick-return push-button returning V_3 to the beginning of the scanning interval; temperature equalization of the input voltage divider which shortens the warm-up period to about 10min and reduces the influence of the fast temperature variations (for a description of this see Pacák[3]) a demagnetizing device, which makes it possible to introduce slow decaying oscillations of V_2, if it is necessary to reduce the remanent magnetization in the magnet iron core; a measuring circuit for a quick checking of the performance.

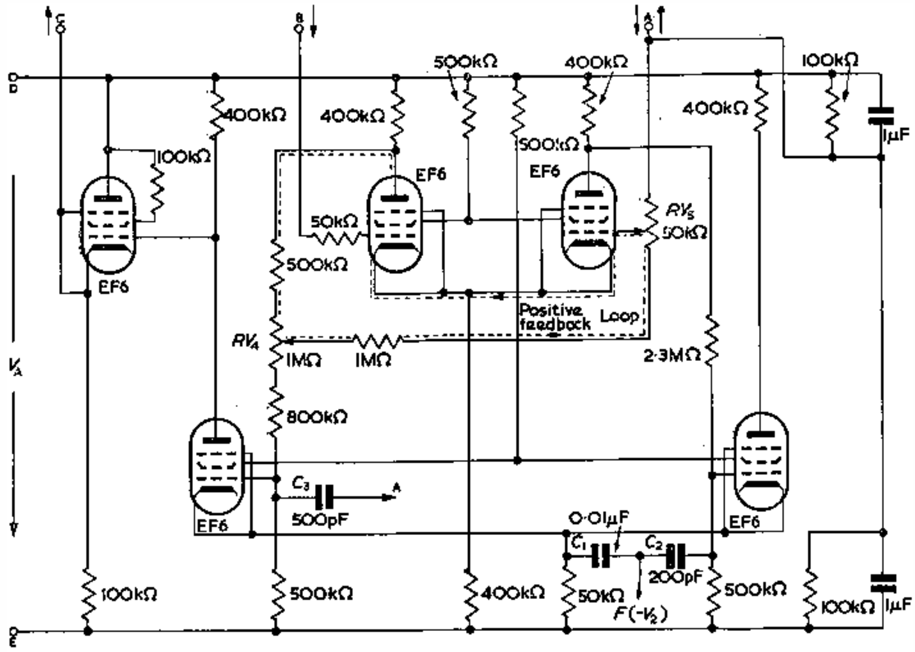

Fig. 2. The amplifier

The performance of the instrument has been investigated by well-known methods both in the laboratory and in practical use. They are as follows: The output voltage is adjustable manually or automatically between 3 to 800V (load 2·5 to 10kΩ). Short-term output voltage variations within the range of 10 to 800V less than 20mV including the effect of mains voltage variations of ± 7 per cent. Long-term stability after 1 hour warm-up period better than 100mV in one hour. Ripple less than 5mV. D.C. output resistance within ±0·1Ω.

The brief description given does not include the complete and detailed analysis of the circuit features and performance. It shows, however, the practical applicability of the principles and methods, a review of which was published previously[1].

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Semiconducting Indium Compounds *

IN the course of a recent review, Wright[1] has listed numerous semiconductors, some of which have been known for a very long time and some of which have only recently been investigated for their fundamental properties and potential applications. Among the latter group of materials, semiconducting indium compounds occupy a prominent place. Their applications cover a wide range of technological practice and there is every reason to believe that they will be further developed as even purer specimens become available and as their remarkable properties become better known. In this connexion the compounds of most interest are indium antimonide, indium arsenide and indium phosphide, of which the first is at present the most important. In principle, indium nitride is also a member of this series, but little seems to be known about the character of this compound. Research has, however, been carried out on indium-selenium[2] and indium bismuth[3] and also on a great variety of tertiary systems as, for instance, indium in arsenic-phosphorus[4], gallium-antimony[5], antimony-germanium, antimony-silicon, antimony-tin, antimony-lead, copper sulphur, silver-selenium[6], copper selenium[6,7], silver-tellurium[7], copper-tellurium[7], and arsenic-gallium. Some of the earlier investigations may have been motivated by the desire to find alternative materials for germanium and silicon for transistor applications. Although, in this sense, the efforts have not been rewarded with practical success, the researches have revealed other properties which are interesting and valuable on their own account.

* *Communication from the Laboratories of Mining and Chemical Products Ltd.*

Preparation and Purity

As a rule, semiconducting indium compounds are prepared by melting stoichiometric proportions of the constituents together in vacuum or in an inert atmosphere. This process is simplest in the case of InSb because the vapour pressure of Sb, even at high temperatures, is relatively low. In other cases, dangerously high pressures may arise, unless suitable precautions are taken to ensure that most of the reaction takes place at lower temperatures before the melting point of the resulting compound is reached. Crystals or, more usually, multi-crystalline ingots produced in this way can be subsequently purified and recrystalized by zone refining. This process is greatly aided by the fact that the constituent elements themselves are only slightly soluble in the compound. Repeated zone melting, followed by cropping of the impure ends therefore results in a material which has a low content of foreign impurities and which is almost free from non-stoichiometric excesses. The efficiency of zone refining differs for the various impurities encountered. It depends on the ratio of the solubilities of the impurity in the liquid and in the solid material respectively. Reliable information is now available for Cu, Ag, Au, Zn, Cd, Ga, Sn, Tl, Se, Te, P, As, Fe and Ni, which gives a measure of the care with which these matters have been investigated[8]. In this way InSb can be prepared with an impurity content of 1 part in 10^5 without undue difficulties and indeed with still smaller impurity contents when special care is taken. To achieve purities of this order, the constituent elements must be separately refined beforehand. In the case of indium, this is done by vacuum baking; antimony is zone refined and arsenic purified by a sublimation process. The compounds can now be obtained in the form of single crystals for research purposes, prepared either by pulling a seed from the melt or by seeding a directionally cooled horizontal boat. Care must always be taken to ensure accurate temperature control[9].

InAs dissociates at the melting point because arsenic has a high vapour pressure. The procedure is therefore a little more complicated and involves maintenance of the entire charge at a high temperature during zone refining. Using the best available techniques, material with a residual impurity content less than 1 part in 10^6 has been made. A description of the process has been given by Gremmelmaier[10]. In many important cases, the degree of purity with which one is concerned in these materials is beyond the scope of spectroscopic analysis[11]. The purity assessment must therefore be made on the basis of the electrical properties themselves.

Binding Energies

A glance at the periodic table shows that InP would be expected to have properties which are very roughly comparable with those of Ge, and InSb properties which correspond to those of (grey) In. InAs should occupy an intermediate position and this is confirmed by experience.

InSb, InAs and InP all crystallize with a zincblende structure. Their binding is primarily covalent, but an appreciable ionic component is also effective. As a result of the ionic binding component, these Group III-V compounds have higher binding energies than their corresponding Group IV analogues. As an immediate consequence, they have higher melting points. The binding energy and the melting point are also related to the crystal structure, and it has actually been shown[12] that, to a good approximation, the melting point is a linear function of $1/a^2$ when a is the lattice parameter. The energy relations of electrons in these crystals can, as usual, be described by a band structure. One of the important features of this band structure is the so-called 'band gap', i.e. the minimum energy required to make a valence electron mobile within the crystal. The high binding energy gives rise to relatively wide band gaps. From the point of view of semiconductor applications, this is a desirable feature, since the band gap is perhaps the most important single factor which determines the maximum temperature at which these materials can be effectively used. On the other hand, materials in which the band gap is very low (e.g. as it is in InSb) are generally less sensitive to the presence of impurities than others. Table 1 provides a summary of these relations. In particular, InP has a band gap which is greater than that of Ge by a potentially useful margin. It could therefore be used at substantially higher temperatures. Unfortunately, the material has not yet been prepared with the degree of purity and crystal perfection that would make more detailed comparisons with Ge meaningful. In some cases, as indicated in Table 1, there remains a measure of uncertainty as to the precise values of the parameters quoted.

Electronic Mobilities

As is well known, the electrical conductivity of a material is determined partly by the concentration of free charge carriers and partly by their mobility. In semiconductors, the carrier concentration is usually (but not under all conditions) dependent on foreign impurities or else on departures from the stoichiometric composition. This applies particularly when the band gap is large (InP), since this makes the thermal 'activation' of resident valence electrons relatively difficult. At the other extreme is InSb which has the smallest band gap. Impurities still control the carrier concentrations at low temperatures, but at high temperatures nearly all the carriers which are free to move arise from the lattice itself. The effect of impurities is then primarily to lower the mobility by acting as scattering centres.

As Table 1 shows, the three principal compounds are all characterized by high mobilities. The values quoted under μ_n and μ_p refer respectively to the mobilities of free electrons and of 'missing electrons' in the valence band or 'positive holes' as they are called. μ_n is particularly high for InSb and, indeed, higher than any mobility known before the first experiments on this material. The hole mobility is always smaller than the electron mobility, and the practical utility of a material (e.g. for galvanomagnetic applications) is therefore governed by μ_n and its temperature dependence.

Indium Phosphide

By virtue of its large band gap, InP is in a somewhat special position. It is a material of much higher resistivity and has been reported to show transistor action[14]. The specimens so far available have not approached modern germanium as far as purity and crystalline order is concerned. Thus, although substantial power gains have been recorded with point contact assemblies, the transistor action in InP has not yet reached the level of efficiency with which one is familiar in Ge. The traditional transistor materials cannot as yet be replaced, but InP and similar compounds clearly merit careful attention from this point of view. In contrast to this, transistor action in InSb and InAs at room temperature is not conceivable since valence electrons activated across the much smaller

TABLE 1.

	MELTING POINT (°C.)	BAND GAP AT 300° K(eV)	μ_n	μ_p
InSb	523	0·18	78 000	800
Sn	232	0·1	~3 000	—
InAs	942 } 936 }	0·36	~33 000	500
InP	1070	1·28	5 000	~150[12]
Ge	958	0·66	3 900	1900
In_2Se_3	890	1·2	30	—
InSe	660	1·05	900	—
GaSb InSe	?	0·59	—	~330

band gap would mask and swamp any modulation of carrier concentration which might be induced by electrical means. Metallic point contacts on the surface of InP exhibit a high degree of rectification. Current ratios of the order of $10^5:1$ have been quoted. Such point contacts are also highly photo-sensitive. There are possibilities for the application of InP in solar energy conversion and the material can also be used as a detector of thermal neutrons. For this purpose, pn junctions can be prepared by means of a diffusion process, and used as described by Gremmelmaier and Welker[15].

The behaviour of this material is reasonably well understood, but a number of interesting problems remain. One is associated with an unusual form of electrical breakdown observed at low temperatures[16]. At 4·2°K such a breakdown can occur at fields as low as 9·1V/cm. It is completely reversible. The effect has not yet been completely explained but is believed to be associated with conduction processes in which charge carriers move directly from one impurity centre to another. (Impurity band conduction.)

Tertiary Compounds

The most reliable and extensive information is probably that available for compounds which can be represented by the general formulae[4,17] $InAs_y\ P_{1-y}$ where $0 \leqslant y \leqslant 1$. These form a continuous series of mixed crystals and can thus be systematically studied. Such work has shown that the band gap is, in this case, a linear function of composition and that the high mobility ratio μ_n/μ_p tends to be maintained as the composition changes gradually from InAs to InP, even though the mobilities themselves become smaller. The mixed crystals can be made n-type or p-type by suitable additives in the same way as the binary compounds InAs and InP. A wide range of solid solutions also exists between GeSb and InSb[5,18,12]. Experiments of this kind demonstrate a promising measure of control and 'designability' of these semiconducting materials. Further information on their preparation and properties has been given by Austin *et al.*[13]

Applications of Indium Antimonide and Indium Arsenide

The applications of InSb depend directly on two principal features: the small band gap and the high electron mobility. As a result of the small band gap, the material is capable of showing photo-conduction which extends to 7·5 microns. The outcome is a sensitive radiation detector which also happens to have a low noise level and a response time of less than 1μsec. It finds application in infra-red spectroscopy and low temperature pyrometry. By the addition of lead, tellurium or tin the optical absorption edge can be shifted to a shorter wavelength (larger band gap) and a variety of materials suitable for the preparation of infra-red filters can thus be obtained.

As indicated above, the large electron mobility has some special consequences, particularly as far as the galvano-magnetic properties are concerned. When placed in a magnetic field which is at right-angles to the current, specimens of InSb develop a relatively large transverse voltage which is known as the Hall voltage. Some current can be drawn from this voltage source. The device is then called a Hall generator[19]. Its efficiency (for an n-type specimen) is proportional to $(\mu_n H)^2$, where H is the magnetic field. In practice this amounts to approximately 16 per cent in InSb, which is about 160 times as high as the values calculated for germanium[20]. For InAs, the calculated value is 7·8 per cent. The Hall voltage is proportional both to the longitudinal specimen current and the magnetic field which, in turn, can be controlled by an independent external current. It is thus a mixing, modulating and multiplying device. Indeed, by a magnetic field arrangement of suitable geometry and by providing for a sufficient amount of positive feedback, a power amplification effect can be achieved. An oscillator can also be constructed in this way. Since only low fields are required, the Hall effect itself can be used for sensitive magnetic field measurements. The same object can be achieved by measuring the change of resistivity which occurs when a specimen is placed in a transverse magnetic field. (Magneto-resistance.) The effect can be most sensitively demonstrated by means of a so-called Corbino disk. This is a semiconducting specimen of circular shape with one electrode at the centre and another on the entire periphery. A transverse magnetic field is applied at right-angles to the area of the disk. In the absence of the field, the carrier mobility has its normal value and the current flows radially along the shortest path. When the field is applied, the effective mobility is smaller and the path length of the current is greatly increased by the magnetic deflexion. These two effects reinforce one another in producing a substantial change of resistance. For a field of 10kG it could amount to a factor of 18, and perhaps more. Such techniques have been applied to the measurement of magnetic fields as low as 10^{-5}G. Microphones, strain gauges, thickness gauges, etc., can be based on this principle. Magneto-sensitive components can be used in a variety of control circuits operated at normal temperatures. (At very low temperatures, the magneto-resistive effect in InSb becomes negative, an observation which is not yet satisfactorily explained.) The multiplication effect mentioned above can be utilized for the construction of wattmeters and various forms of modulators. Devices of this general kind have found application in fields as far apart as railway signalling and the industrial counting of ferromagnetic objects. There are also piezo-resistive properties (change of resistivity with pressure) which may prove to be of practical interest[21]. InSb (and, indeed, also InAs and InP) can be used for the preparation of pn junctions and these can act as photo-sensitive elements.

Practical devices should have a high degree of stability as well as a high sensitivity. Since InSb has the highest electron mobility, devices based on this material are the most sensitive. On the other hand, they are much more temperature dependent than comparable devices based on InAs. For various applications, the stability requirements may be sufficiently important to favour InAs, despite the loss of efficiency involved. It is also a considerably harder material which may have practical advantages.

Numerous applications which are now known to be possible in principle still remain to be fully exploited in practice and, as this short survey indicates, these materials are of great versatility, substantial technological promise and scientific interest.

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The Instruments, Electronics and Automation Exhibition

A description, compiled from information supplied by the manufacturers, of some of the exhibits at the Instruments, Electronics and Automation Exhibition held at Olympia, London from 23 to 28 May.

(Voir page 389 pour la traduction en français: Deutsche Übersetzung Seite 395)

ALMA COMPONENTS LTD

55 Holloway Road, London N.19

WIREWOUND RESISTORS

A number of new types of wirewound resistors were shown, including the following:

Type Y

A new range of resistors on American sized bobbins to provide interchangeability. Rated at $\frac{1}{4}$, $\frac{1}{3}$ and $\frac{1}{2}$W and set to 0·1 per cent, these resistors are made to a high standard and have temperature coefficients guaranteed better than 0·002 per cent/°C.

Type A4

The first of the type A range, this resistor is fully hermetically sealed and tropicalized and is being submitted for Type Approval to DEF.5113. It will operate over the range −55°C to +120°C, and is available from 100Ω to 400kΩ set to 0·1 per cent.

Type C

A compound wound resistor with a temperature coefficient guaranteed to be zero ±4 parts per million per °C over the range 0°C to 100°C. Wound on Alma standard bobbins type S, these resistors are rated at 1W (Type C1), $\frac{1}{2}$W (Type C2) and $\frac{1}{4}$W (Type C4). Tolerances to ±0·1 per cent.

EE 22 751 *for further details*

ASSOCIATED ELECTRICAL INDUSTRIES (Woolwich) LTD

155 Charing Cross Road, London W.C.2

TRANSISTORS

Among the devices recently introduced for industrial applications are four power transistors, two of which have peak current ratings of 10A and collector-base voltage ratings of 80V and 100V (types XC155 and XC156). The other two power transistors have peak current ratings of 3A, with collector-base voltage ratings of 40V and 60V (types XC141 and XC142).

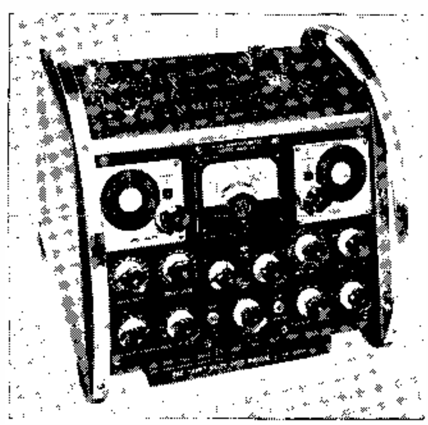
A range of three drift transistors has been released for high speed switching applications. The minimum frequency (f_1) at which the current gain is unity for these transistors (XA141, XA142 and XA143) is 20Mc/s, 40Mc/s and 60Mc/s respectively. The exhibit also contained two switching transistors of the Mesa construction, with minimum f_1's of 25Mc/s and 35Mc/s (types XA161 and 162) offering a low saturation resistance and absolute maximum collector dissipation of 75mW at 55°C and an absolute maximum collector-base voltage of 13V.

For low speed high voltage switching, the XB121 with a collector-base voltage of 105V is of considerable interest.

A range of three germanium npn alloy devices for complementary switching circuits was also included. These are types XA701, 702 and 703, with minimum f_c's of 3Mc/s, 5Mc/s and 10Mc/s and with a collector dissipation of 70mW at 55°C.

Drift transistors with r.f. amplications up to 100Mc/s were shown, as well as the more conventional and established range of pnp alloy devices for audio and r.f. applications, some in alternative encapsulation particularly interesting to printed circuit users.

EE 22 752 *for further details*



AVO LTD

Avocet House, 92-96 Vauxhall Bridge Road, London S.W.1.

VALVE CHARACTERISTIC METER

(Illustrated above)

An improved valve characteristic meter—the Mark 4, has recently been introduced.

While it retains all the facilities existing on earlier models, the following additions have been incorporated:

(a) A considerably increased facility for the correct setting of negative grid potentials, particularly at low values.

(b) More accurate and concise methods for determining grid current, in particular when the valve under test suffers from reverse or primary emission.

(c) Additional anode voltages for testing the recently introduced low anode voltage valves.

(d) Improved regulation of the anode and screen voltage supply to reduce reflected impedance problems.

(e) Overload protection for the movements.

Anode voltages between 12·6V and 400V in steps, screen voltages 12·6 to 300V in steps; grid voltage continuously variable from 0 to 100V. Anode current from 2·5mA f.s.d. to 100mA f.s.d. on the meter, screen currents over the same range; grid current 0 to 100μA filament-heater voltages between 0·625 and 117V; mutual conductance from 0·1 to 60mA/V; Go/No-Go facilities are also available on the instrument.

Diodes and rectifiers can be tested under reservoir load conditions between 1mA and 180mA.

EE 22 753 *for further details*

B & K LABORATORIES LTD

4 Tilney Street, Park Lane, London W.1

TAPE LOOP MAGAZINE

This is claimed to be the smallest magazine ever designed to handle continuous tape loops up to 120ft in length. It provides simple conversion from laboratory reel-to-reel recorder/reproducer to a continuous loop device. This magazine contains easily threaded rollers for accommodating $\frac{1}{4}$in, $\frac{1}{2}$in or 1in tape up to 120ft in length.

The tape path is extremely simple and lends itself to quick threading. Long loops can be quickly interchanged and all standard tape speeds from 1$\frac{7}{8}$in/sec to 60in/sec can be selected instantly.

The magazine can be removed and replaced by a regular 10$\frac{1}{2}$in reel-to-reel magazine on the P.S.200 recorder.

This new loop magazine enables the P.S.200 recorder/reproducer to be used for delay applications and spectrum analyses of large samples of data and can accommodate up to 14 channels of information. Such applications have in the past required large rack mounted units with cumbersome suspended idlers or complex interwoven tape paths. With the new magazine, users can start with either a reel-to-reel or loop instrument and later expand the range of applications.

EE 22 754 *for further details*

INK RECORDING INSTRUMENT

The Oscillomink 230B is an ink recording instrument capable of recording up to 1kc/s, with variable sensitivity from 0·2mm/mV to 4mm/100V. The direct writing system is based on a new principle whereby an extremely fine jet of writing fluid is impelled at high speed out of a nozzle, fine as a hair, placed in the driving centre of the galvanometer. Since the free jet is not an integral part of the moment of inertia of the system, a high resonant frequency, about 650c/s, is made possible. The base line can be moved to a suitable position on the paper without influencing the linearity. When recording with several channels the traces can, without interference, cross or be recorded with a common base line. The recorder is also provided with individual adjustment of the line width. Paper speeds are 5, 10, 20, 50, 100 and 200mm/sec adjustable during operation. Accuracy better than 1 per cent.

Separate attachments are available for paper speeds of 0·5, 1 or 2m/sec, locally or remotely controlled.

This instrument is provided with twin push-pull d.c. amplifiers with coarse and fine adjustment of the sensitivity, trimmable phase and amplitude correction, good discrimination between symmetrical and assymetrical signals, and a carrier frequency modulated final stage with expansion coupling. Other units available include 8 and 12 channel recorders and twin amplifier units.

EE 22 755 *for further details*

R. H. COLE (Overseas) LTD

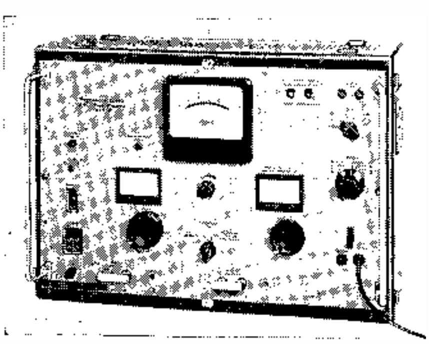
2 Caxton Street, London S.W.1

CARRIER FREQUENCY LEVEL MEASUREMENT

(Illustrated below and next column)

This equipment, which is manufactured by Siemens & Halske, consists of a level oscillator and a selective level meter. It enables selective measurements of level, attenuation and gain to be made at frequencies between 30kc/s and 15Mc/s. This spectrum includes, for example, multi-channel carrier-frequency telephone systems using coaxial cable or radio links, the CCITT basic group (60 to 108kc/s, 12 voice channels), the CCITT basic supergroup (312 to 552kc/s, 60 voice channels), the CCITT basic mastergroup (812 to 2044kc/s, 300 voice channels) and the CCITT basic supermastergroup (8·516 to 12·388Mc/s, 900 voice channels).

A particular merit of this equipment is the high accuracy of the generator frequency and the tuning arrangement of the detector; it is attained by a crystal controlled electronic frequency lock-in system. At any point of the overall frequency range, which is traversed in one sweep of the dial, an indicating lamp lights whenever the frequency has locked in on the frequency corresponding to the scale division to be set (every

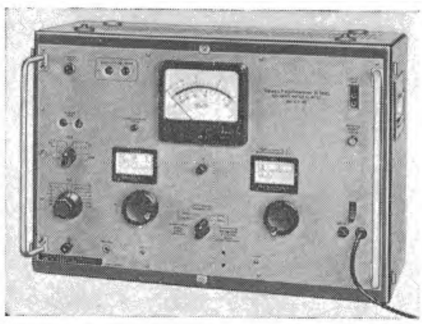


100kc/s). This considerably speeds up the setting of frequencies. Frequency values between the scale divisions can be set to within about 200c/s at any point of the overall range by means of an additional incremental tuning control.

Because of the automatic tuning facility for the main and incremental frequency controls of the level meter, all measurements, where generator and detector need not be physically separated, can be carried out from the generator as simply and rapidly as wideband measurements.

The wide measuring range and high selectivity combined with the high image frequency rejection and low residual harmonic distortion allow measurements of crosstalk and distortion attenuation; the electronic frequency lock-in feature further adds to the ease of such measurements. The stability with time of the settings is of value for continuous supervision purposes.

For lower demands on frequency accuracy equipments are available without the electronic frequency lock-in feature with either neper or decibel calibration.



The level oscillator can be amplitude-modulated at frequencies of 0 to 100kc/s by an external source.

Since the level meter responds to r.m.s. values, it can be used for measurements of random noise, e.g. in conjunction with the random-noise generator type Rel 3 W 432.

EE 22 756 *for further details*

EKCO ELECTRONICS LTD

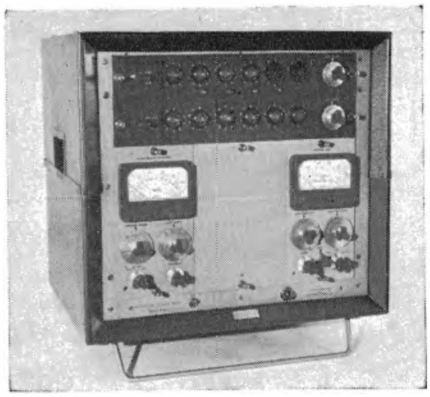
Southend-on-Sea, Essex

AUTOMATIC SCALER

(Illustrated above right)

The new Ekco type N 610 is a combined automatic scaler and timer, incorporating a stabilized high-voltage supply, a high-gain input amplifier and a pulse height analyser. A ratemeter facility is provided, with push-button selection of 1 000, 10 000 and 100 000 counts/sec ranges.

The scaler may be stopped automatically at predetermined counts up to 1 000 000 and at predetermined times up to 100 000sec. The timer can alternatively be used as a second scaler with its own pulse height discriminator, the timing system being referenced to the mains frequency, although provision is



also made for using an external frequency standard. Input sensitivity is 5mV negative-going and 5V positive-going. The switch-selected amplifier gain is 25, 50, 100, 250, 500 or 1 000 and there is a stabilized high-voltage supply of 250 to 2 000V in two ranges.

EE 22 757 *for further details*

ENGLISH ELECTRIC CO. LTD

Marconi House, Strand, London W.C.2

TRANSISTORIZED INTER-TRIPPING UNIT

(Illustrated below)

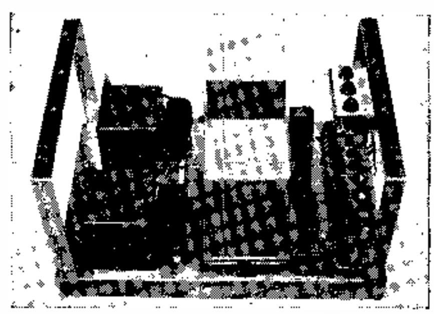
This fully transistorized inter-tripping system costs about half as much as conventional equipment and, working on the frequency shift principle, is designed as a protection link in power networks.

The unit, containing two protective circuits, measures only 20in × 12in × 12in, which is smaller than any existing equipment.

The transistorized logic circuits give reliable service, a low power consumption and a fast operating time. On a normal power line the time taken to relay a signal for tripping a circuit will not exceed 55msec and under the worst conditions 85msec.

A typical application is where a step-down transformer is connected to the remote end of a 33kV line.

Here, a fault on the low voltage side of the transformer will not necessarily be detected by the protective relay at the supply end of the feeder and under this condition problems arise in tripping the circuit-breaker at the supply end. This is particularly true if the fault level on the l.v. side of the transformer is restricted by an earthing resistor. This difficulty is overcome by the use of inter-tripping equipment initiated from the l.v. side of the transformer.



Each terminal unit on this new equipment, known as type SL54, has its own transmitter and receiver. Under healthy conditions a continuous signal on a given frequency is transmitted over the interconnecting G.P.O. lines to the receiver at the other end of the line. When a fault occurs this frequency is shifted and the change, detected by the appropriate receiver, causes operation of a relay which in turn initiates whatever action is required.

EE 22 758 *for further details*

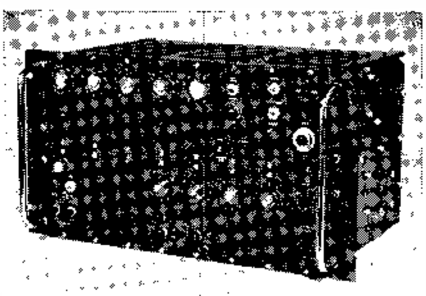
ERICSSON TELEPHONES LTD

Beeston, Nottingham

TACHOMETER

(Illustrated below)

Constructed from E.T.L. sub-units, this instrument uses a reference frequency of 100c/s to measure from a lower limit, governed only by the efficiency of the transducer employed, up to a maximum of 1 200 000rev/min. The 100c/s may be generated either internally or externally.



There are two input channels to cater for pulse amplitudes between 0·1V and 1V or 1V and 100V. Both channels accept either sinusoidal or positive going pulses of greater than 4μsec duration.

The period of count and the length of time of display may each be adjusted between 1sec and 10sec, count and display times being complementary to a total of 11.

EE 22 759 *for further details*

SHAFT DRIVEN TRANSDUCER

(Illustrated below)

The Rotapulse is a shaft driven, photo-electric transducer, which converts the angular velocity of its shaft into a known number of electrical pulses per revolution.

Rotation of the shaft causes the interference pattern produced by light shining through transmission gratings to cycle, the cycling being detected by phototransistors mounted as diametrically opposed pairs. Since each pair is thus subjected to anti-phase signals, both together form a differential device producing the maximum number of pulses. The standard instrument produces 1 000 pulses per revolution, but instruments with other maxima can be provided.

As demonstrated, a Rotapulse is so mounted that its shaft projects through the centre of a disk calibrated in 1 000 units. Spinning the pointer attached to the shaft causes the number of pulses produced by the Rotapulse to be indicated on a bi-directional counter constructed from transistorized counting units, Type 116A.

EE 22 760 *for further details*

LOAD CELLS

(Illustrated below)

Produced to meet the growing need for weighing devices which are fast, accurate, and robust, the Nobel load cell

is designed to give extreme stability against temperature effects, long term drift of the cell output, and creep with sustained loading. These characteristics are unaffected by use in corrosive atmospheres, and the load cell housing is constructed to carry side loads comparable with the maximum rated load of the cell.

When several Nobel load cells are used in a multiweighing installation, a single instrument can be used to indicate the output of any or all of the cells. An accuracy better than ±0·25 per cent is normal, and this accuracy is maintained over long periods. Five standard sizes of cell covering all weights up to 50 000lb have been produced, and were exhibited as a static display.

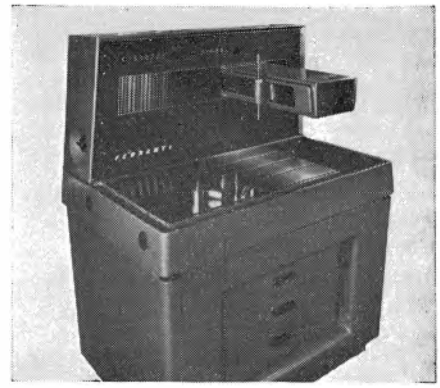
EE 22 761 *for further details*

FERRANTI LTD

Hollinwood, Lancashire

CO-ORDINATE INSPECTION MACHINE

(Illustrated above right)



This co-ordinate inspection machine which can help to speed up the work of inspection departments as it can check the proportions of an object to an accuracy of 0·001in in half the time taken by orthodox methods of inspection such as engineer's dial gauges, etc.

An inspection probe with interchangeable tips, suitably shaped, can be placed at any point on a machined part. This probe is supported from a rigid cantilever beam projecting over the work table to which the item to be inspected is clamped.

Electro-optical sensing units are fitted to the x and y axes of the machine. Motion along either axis causes the Moiré fringe pattern generated by the associated diffraction gratings to traverse photocells, where changes in light intensity are converted into their electrical counterpart. Transistor bi-directional decade counters—one for each axis—record the cyclic variation.

Measurement information is displayed continuously on the two five-figure electronic counter banks, the position of the probe at any point in its x and y travel can be read-off directly to an accuracy of 0·001in.

EE 22 762 *for further details*

THE GENERAL ELECTRIC CO. LTD

Magnet House, Kingsway, London W.C.2

DECIMAL READ-OUT MICROMETER

(Illustrated next column)

A demonstration showed how digitizers can be geared together to obtain high resolution. Four size 11 binary coded decimal digitizers are used, one for each decimal digit, giving a numerical display of the micrometer reading in ten-thousandths of an inch. All circuits are transistorized.

The gearing between digitizers is not critical, since they have additional windings which, by means of an associated circuit, are arranged to prevent ambiguity. This is done by using the most significant digit of the finer of each pair of digitizers to switch the excitation of the coarse digitizer from one winding to another.

The units are excited by means of square waves generated by a transistor multivibrator. The resulting outputs from the digitizers are amplified, phase detected, stored in bistable circuits, and then decoded by means of relays which drive the numerical display.

As shown, each digitizer has a separate block of output circuits associated with it. In many applications it is possible to excite the units sequentially and use a single block of output circuits on a time

sharing basis. This can lead to a considerable economy in equipment, especially in systems containing a large number of digitizers.

EE 22 763 *for further details*

VIBRATOR DRIVE EQUIPMENT

This is designed to provide all the functions necessary for the running of a VEM 5·2 or similar vibrator in a form suitable for use by factory labour.

The vibrator is operated in the frequency range 30c/s to 10kc/s by a low distortion amplifier with a nominal output of 2·5kW. The acceleration level of the vibrator table may be controlled within close limits by a rapid-response *g* control unit operating on any one of three accelerometers, with independent level indication on a meter scaled logarithmically from 2 to 100*g*.

An external signal source, or one of two internal sources may be used. The internal motor driven sweep oscillator is controlled entirely from the front panel and may be swept between any preset limits in the 30c/s to 10kc/s range, at sweep speeds between 4 and 140sec/octave. A manual control may be used to set any required single frequency and the frequency, once set, may be held indefinitely.

The internal tape recorder is a twin-track, automatically reversing machine providing half-hour continuous play back, repeated indefinitely with a reversing time of less than five seconds.

EE 22 764 *for further details*

GENERAL PRECISION SYSTEMS LTD

Aylesbury, Buckinghamshire

ANALOGUE COMPUTER COMPONENTS

Included in a range of d.c. analogue computer components are the following:

Function Generators

This electronic function generator, of the diode switched type, is arranged to accept an input of up to ±100V d.c. Its output is also up to ±100V d.c. and may be adjusted to be any desired function of the input voltage. The function is approximated by a series of straight lines. This is done by splitting the range into 10 sections separated by 9 'break points' which may be set at any value of input voltage. Each of the sections has a straight line law, the slope of which is variable from zero to 10 in 1. All setting controls are so nearly independent of each other that by setting each one once only, an overall accuracy of ±2 per cent is attained. The bandwidth is 2kc/s. The unit is complete with its own internal amplifiers.

Functions may also be generated by the use of the tapped resistance cards fitted to the servo-multiplier units.

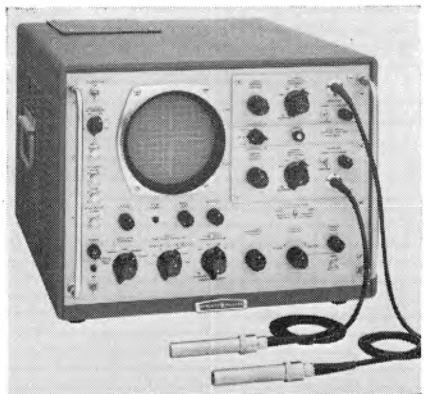
Servo-multipliers

This low speed servo-multiplier consists of a velodyne motor generator driving three potentiometers. One potentiometer is used for motor control and a switch enables this to be disconnected when integration is required. A two-speed gearbox is standard and alternative gearboxes may be fitted, giving a possible variation of integration range from $\frac{1}{2}$ cycle per second to 1 cycle in 8 hours.

The other two potentiometers have 9 tappings so that by suitable resistive loading various functions may be approximated. Two additional potentiometers may be added.

A high speed servo-multiplier is also available using rotary potentiometers driven by an MES servo motor. A speed of 10c/s full scale is attained and either three or six potentiometers may be fitted. In each case one is used for motor control, one has 9 tappings, and the rest are centre-tapped.

EE 22 765 *for further details*



LIVINGSTON LABORATORIES LTD

Retcar Street, London N.19

SAMPLING OSCILLOSCOPE

(Illustrated above)

The Hewlett-Packard oscilloscope type 185A together with the model 187A dual trace amplifier provide a system for viewing fast-pulse phenomena requiring bandwidths of up to 500Mc/s. Successive samples of the pulse under test are taken at slightly later times during each recurrence. The samples are then reconstructed on a long time-base, giving a system with an effective rise-time of 0·7mμsec.

The number of samples per trace can be selected by a control on the front panel and comprehensive triggering and sweep expansion facilities enable leading and trailing edges of pulses to be examined in detail.

In addition, accurate measurements of rise-time, duration or spacing can be made by using the second channel for a simultaneous display of the time calibrator output and an output is provided for operating an X-Y plotter so that permanent records can be made of high speed phenomena.

Each channel has independent sensitivity with calibrated ranges from 10mV/cm to 200mV/cm accurate to ±5 per cent and has an input impedance of approximately 100kΩ shunted by 2pF.

EE 22 766 *for further details*

OSCILLOSCOPE CAMERA

The Hewlett-Packard type 196A oscilloscope camera with its Polaroid Land back, enables a positive photographic record of an oscilloscope trace to be obtained in a matter of a minute.

A viewing hood permits observation of the trace during exposure and the special f/1·9 lens provides an image of a full 10cm graticule width, with no appreciable degrading of the c.r.t. resolution. The object-to-image size ratio is 1 to 0·9.

Problems of light leakage have been eliminated by using a standard camera bellows and a facility is provided for moving the lens through 11 detented positions to assist in carrying out multiple exposure work.

EE 22 767 *for further details*

MARCONI INSTRUMENTS LTD

St. Albans, Hertfordshire

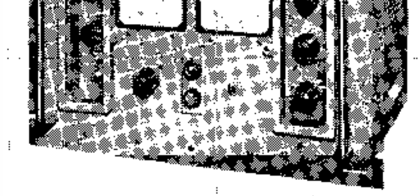
RATEMETER

(Illustrated below)

This instrument (type TF1292) gives an indication by meter of the count rate or integrated count of nuclear particle and radiation detectors. Count-rate ranges of 30, 100, 300 and 1 000 per second full-scale are provided; with these, a time-constant of 1, 5 or 20sec can be selected. The integrated-count ranges are 1 000, 3 000, 10 000 and 30 000 counts full-scale.

Two adjustable e.h.t. outputs are provided: one for a particle or radiation counter and the other for a precipitation chamber. Also available are an h.t. and l.t. supply for an external pre-amplifier.

The instrument is suitable for many radiation measurement applications in industrial processes and in laboratories. One particular use is in a system monitoring the coolant gas in a nuclear power station, thereby detecting faulty fuel elements. Employed for this purpose, it measures the output of a scintillation counter that reacts to radioactive particles in the gas; the particles, being ionized, are precipitated on to a wire fed with an



attracting e.h.t. voltage provided by the ratemeter.

There is provision for remote control and indication, and operation of a warning circuit at any predetermined meter indication, or in the event of failure in the power supplies. In addition to the panel-meter indication of count-rate or count, there are two proportional outputs for external use giving 50V and 100mV at full-scale; the lower is particularly intended for a recorder.

EE 22 768 *for further details*

ANALOGUE-DIGITAL CONVERTOR

(Illustrated above)

The TF 1325 is a fully transistorized analogue to digital convertor for use in feeding to indicating, recording or processing equipment, the value of an applied d.c. potential from 0 to 0·999V. Binary coded (+10/0V) decimal output of three digits provides an accuracy of ±1mV. The sampling period is 20msec and can be initiated remotely. A step signal indicates when sampling is completed and a result is held available; the input current is then below 0·15μA.

The digitizer is primarily designed to receive a d.c. voltage, as an input, sample its level, and provide the result rapidly and accurately in digital form for application to a decoder and line printer or electric typewriter, thereby enabling a record to be obtained for subsequent reference. Used in this way, it finds particular application in repeatedly monitoring the level of radiation at a number of points in a nuclear power station. Voltages derived from points to be monitored are applied in turn to the digitizer, and need only be steady over each sampling period.

The high speed capabilities of the digitizer also make it suitable as a convertor translating an analogue voltage into digital form for application to data processing equipment.

EE 22 769 *for further details*

MULLARD LTD

Mullard House, Torrington Place, London W.C.1

ALLOY DIFFUSION TRANSISTORS

The alloy diffusion technique has already been employed successfully in the large scale production of the OC170 and OC171 transistors which are widely used for i.f. and r.f. amplification in h.f. and v.h.f. receivers. A number of devices has now been developed for use in telecommunications, computing and industrial equipment, including the following:-

Transistors for V.H.F. Communications

The v.h.f. transistor type AFZ11 has been developed specifically as a low-noise transistor for 100Mc/s amplification in professional communications equipments. The particular manufacturing technique used for this device enables the extrinsic base resistance and the feedback capacitance to be kept at very low values, and at 100Mc/s the transistor will give a power gain greater than 10dB. The noise figure, at the same frequency, does not exceed 6dB. The average current gain is 50.

Maximum collector-base voltage is −20V and maximum collector-emitter voltage is −12V at the full current rating of 10mA. Collector leakage current is 4μA (average) with collector-base voltage of −6V and at a temperature of 25°C. Maximum junction temperature is 75°C and the thermal resistance is 0·6°C/mW, permitting a dissipation of 50mW in an ambient temperature of 45°C. The transistor is housed in a metal envelope.

The AFZ11 will be followed soon by a further transistor designed for the same professional applications but at twice the frequency. In the 200Mc/s of v.h.f. communications equipment this transistor will provide a power gain greater than 10dB, with a noise figure of less than 9dB.

Very Fast Computer Logic

An alloy-diffused transistor with good large-signal characteristics will shortly be available for the logic circuits of very high-speed computers.

This transistor is an extremely sophisticated device which will meet the requirements of future computers with clock pulse rates up to 10Mc/s. It will have a frequency cut-off of several hundreds of megacycles per second, excellent bottoming characteristics at 50mA collector current, and an adequate dissipation rating for its applications.

This transistor was demonstrated in low-level voltage-switching logic circuits and in a 10Mc/s decade counter.

Fast, High-Current Core Driving

This transistor is under development for driving high-speed ferrite memory systems in computers, in conjunction with the new high-speed logic transistor mentioned above.

It is capable of delivering output pulses of several hundreds of milliamps with extremely fast rise times. Its maximum collector current rating is ½A and the value of its f_1 parameter is greater than 60Mc/s.

A demonstration featured the device in a high-speed, high-current pulse generator providing ½A output pulses with average rise-times shorter than 40mμsec.

Its high-frequency, high-power characteristics also suit it for carrier telephony and transmitting equipment. In these applications it provides 3W output at 8Mc/s, or up to 1W at higher frequencies. A demonstration showed it used in the driver and output stages of a 2·5W transmitter working at 8Mc/s.

PNPN Switching Transistor

In the ATZ10 the alloy-diffusion technique has been developed to provide a 4-layer pnpn switching transistor.

Its properties enable 'flip-flop,' astable and monostable circuits to be designed using only a single pnpn transistor in place of the two conventional 3-layer devices normally required, with consequent savings in space and associated components. Further, it is also suitable for use as a switched speech path in automatic telephone systems.

The ATZ10 is a three-terminal device, and its action is similar to that of a combination of a pnp and an npn transistor with internal regeneration. This construction results in a pronounced negative impedance characteristic over a usefully wide current range and enables the device to be pulse-operated as a high-speed electronic switch: once switched 'on' it automatically remains 'bottomed' until switched 'off' by a following pulse.

The ATZ10 has a typical off-on impedance ratio higher than 3 million to one. Its maximum voltage rating and peak current rating are in the region of −35V and 25mA respectively. The maximum junction temperature is 55°C, which allows a maximum total dissipation of 17mW in an ambient temperature of 45°C.

At the exhibition a decade ring counter using ten ATZ10's and capable of counting speeds of 30kc/s was demonstrated.

Avalanche Transistor with 1mμsec Rise-time

This device does not function by normal transistor action, but by the 'avalanche' phenomenon. It represents a specialized development of the alloy-diffusion technique that has valuable applications where unusually fast rise-times are required—for example, in high-speed sampling oscilloscopes.

A peak current of 50mA can be reached with a rise-time of only 1mμsec, which represents a rate of current rise of 50A/μsec.

At the exhibition it was demonstrated in an avalanche circuit producing 50mA pulses with a rise-time of about 1mμsec, at a p.r.f. of 4Mc/s. These pulses were displayed on a sampling oscilloscope which itself uses an avalanche transistor in the sampling circuit.

Switching Transistor

The ASZ20, already in production, is intended for switching applications, including computer logic circuits.

It has a peak current rating of 25mA, and a maximum collector-base voltage of −40V. The f_1 value is greater than 40Mc/s, and the minimum large signal current gain is 33 at 25mA collector current. Maximum junction temperature is 75°C and the thermal resistance is 0·6°C/mW, allowing a dissipation of 50mW in an ambient temperature of 45°C. Collector leakage current at −6V is less than 5μA for a temperature of 25°C.

EE 22 770 *for further details*

LETTERS TO THE EDITOR

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

Diffraction Grating Measurement

DEAR SIR,—Mr. Wilde in his letter on diffraction grating measuring systems (ELECTRONIC ENGINEERING, September 1959) raises some points on which I should like to comment.

Mr. Wilde mentions the conventional d.c. type reading head using four photocells and producing a balanced pair of quadrature output signals. Apparatus of this type has been used successfully for a number of years on Ferranti machine tool control and digital measuring systems and is covered by British Patent 810,478 and is also covered by applications in several overseas countries.

The grating defects of fan distortion and progressive error occur on some prismatic replica gratings to the extent mentioned by Mr. Wilde, but these would be regarded as rejects and not as good commercial gratings. In the case of 2 500 line/in gratings giving a 0·0001in digit, the normal limit for progressive error is $\pm$0·0001in per ten inch length and for fan distortion $\pm$40sec. For 250 line/mm gratings giving a 1μ digit the limits are $\pm 1\mu$ per 20cm for progressive error and $\pm$15sec for fan distortion. A certain amount of selection has been necessary up till now to hold these limits, but with the improved range of master rulings now being made at N.P.L., it is expected that the yield will approach 100 per cent.

The modified arrangement of index grating shown, which presumably has ¼ pitch step in its line structure between quadrants, would appear to preserve quadrature at the expense of amplitude and in fact the amplitude would be zero with a tilt of approximately two minutes with a 2 500 line/in grating if Mr. Wilde's illustration is life size, since the anti-phase quadrants would be brought into phase by the existence of Moiré fringes with a pitch of 0·8in and would then annul each other. This amount of tilt would also cause collapse of the signal with the other arrangement, but a tilt of one minute would give a phase relationship of about 45° which is a usable signal.

If only four digits per pitch of the grating are being used, quite severe phase variations or waveform distortion cannot introduce errors greater than one digit. A much larger subdivision of the cycle using coarser gratings with a carrier system has its attractions, but it loses the merit of simplicity and the close relationship between grating structure and output pulses. No doubt there is a place for both systems, but in fields which are basically digital such as co-ordinate positioning, and in applications such as the evaluation of photographs of nuclear tracks where high precision slides can utilize fine pitch gratings to give 1μ accuracy, the d.c. system of reading has distinct advantages.

Yours faithfully,
A. T. SHEPHERD,
Ferranti Ltd,
Edinburgh.

The Correspondent Replies:

DEAR SIR,—I should like to thank you for giving me the opportunity to reply to Mr. Shepherd's letter which you print above. I will try to discuss his points in the order in which they occur in his letter.

The success of the Ferranti digital measuring system is undoubtedly established; however, I am sure Mr. Shepherd would be the last man to claim its ultimate perfection.

On the question of the accuracy one may expect from 2 500 line/in commercial gratings I think Mr. Shepherd has been fortunate. To get gratings of the quality he specifies, we have had to apply to the National Physical Laboratory directly although they are naturally reluctant to interrupt their very intensive programme of grating improvement to do this work at present. I understand that this work in fact will shortly enable them to license to grating manufacturers improved techniques which will enable the tolerances Mr. Shepherd quotes to be reliably maintained.

Incidentally, the biggest strides have been made in rather coarser gratings, particularly 1 000 line/in, which can now be reproduced by the N.P.L. to accuracies of about 10μin. To make full use of the potential accuracy of these scales it is, of course, essential to use an a.c. interpolation system. I will mention such systems again after I deal with the quadrant cursor.

I must apologize first for the brevity of my previous letter. I will try to be more explicit in the following analysis, which is the most simple method of explaining, though not the best method of using the system.

Consider two equal sized square cursors as shown in Fig. 1, (a) being quasi-quadrant arranged, (b) arranged as Mr. Shepherd advocates. Each has four zones of equal area, intended to read four quadrature spatial phases as shown.

If the squares are for convenience, represented as having sides 4 units long, and each cursor is now moved to an area of grating which has a pure fan error giving $a°$ phase shift per unit length, one may represent the spatial vector diagrams of each system as in Fig. 2(a) and 2(b) respectively.

Differencing the anti-phase pairs of Fig. 2(a), the quasi-quadrant system, gives two vectors in perfect quadrature, of amplitude $2 \cos a$. Differencing the anti-phase pair of Fig. 2(b), Mr. Shepherd's system, gives two vectors $a°$ out of quadrature, of amplitude $2 \cos a$.

System (a) will therefore work even when a is over 45°, system (b) would be working marginally, if at all, in this condition. What is more, the 'square' arrangement is by no means an optimum one, a true quadrant system (four quarters of a round cursor) can be easily shown to be even more advantageous. One does not, of course, simply remain content with the better quality signals. By considerably enlarging the cursor to, say, ¾in diameter a high signal level is possible with a more underrun lamp. This last item is most important, since local heating of the scale due to the lamp's radiant heat can cause appreciable error. In practice this system is quite fascinating to watch, the Lissajous figure of the two quadrature signals remaining perfectly circular on a grating, the imperfections of which are well outside the limits I quoted in my original letter. Most important also is the fact that it will be similarly effective if, with perfect gratings, a slideway inaccuracy causes the reading head to tilt.

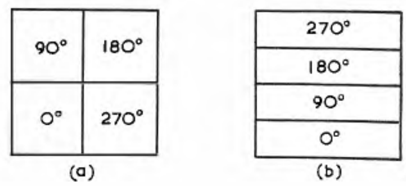


Fig. 1. Two equal sized square cursors

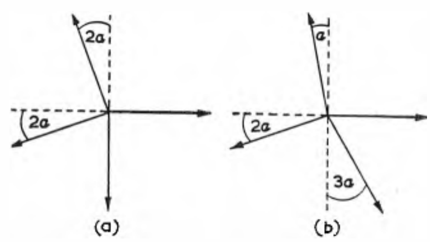


Fig. 2. The spatial vector diagrams

Concerning a.c. and d.c. reading systems I personally feel that the most recent developments of the a.c. system will shortly render the earlier digital equipments obsolescent on machine tools. The early a.c. systems, with their reading heads involving rotating gratings perhaps had as many drawbacks as the purely digital systems. Using detector modulation techniques, however, a system is now in existence which combines the electronic simplicity of an a.c. system, with the optical simplicity of a d.c. system. The reading head itself is extremely simple, the interpolation accuracy 1/50th of a ruling pitch or better, and the electronic output of such a form that the two fundamental operations for machine tools, positioning and contouring, become greatly simplified. Such an equipment has been demonstrated at this year's Physical Society Exhibition, on the N.P.L. stand.

Yours faithfully,

R. W. WILDE,

The Staveley Research Department,
Bedford.

A Variable Time Delay Recording/Reproducing System Using Magnetic Tape

DEAR SIR,—Normal magnetic tape time delay systems rely on a measured length as an analogue of the delay required, the length being variable if required. This is convenient for short delays, but is difficult to implement in very long delay systems, in which the delay is required to vary over the whole range. A possible system (Fig. 1) uses the alternative of counting the number of pulses between the record and replay heads and comparing this number with an input number which is proportional to the delay required.

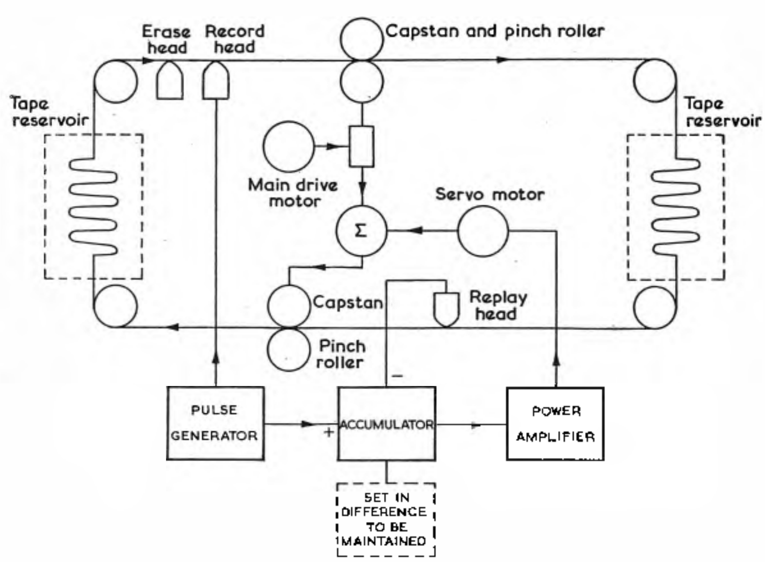
One track of the record head is supplied from the pulse generator with a train of pulses, which are also stored in an accumulator. The replay head of the same track subtracts the pulses from the accumulator; a balance is therefore set up which is a measure of the number of pulses between the two heads, or the delay. Comparison of this number with the delay demand will give sense and speed to the servo system and cause the replay capstan to adjust to the correct delay time by rotation added or subtracted through the differential.

Yours faithfully,

C. DAIN,

E.M.I. Electronics Ltd.,
Hayes, Middlesex.

Fig. 1. The variable time delay system



The Magnetic Drum Store of the 'Mercury' Computer

DEAR SIR,—I read with interest the article by K. I. Turner and J. E. Thompson which appeared in the January 1960 issue. The description of the discriminator circuit in Fig. 8 was somewhat confusing however, since no explanation was given of the semiconductor diodes connected across one half of both transformer secondaries. The reference given for the circuit (Millman, J., Taub, H. Pulse and Digital Circuits) makes no mention of these diodes and their use is not apparent from the schematic. Furthermore it would seem more advantageous to employ semiconductor diodes for the sampling diodes as well because of the lower voltage drops encountered. Could the authors clarify this point.

Yours faithfully,

E. L. SOOHOO,

Beverley, New Jersey, U.S.A.

The Authors Reply:

DEAR SIR,—We apologize for any confusion that we may have caused by omitting to mention the diode referred to. This in fact does not form part of the discriminator circuit as described by Millman and Taub but is part of the mechanism whereby the required strobe is produced. As stated in the article, this strobe is made by means of a transformer with low primary inductance and the diode acts as an asymmetric damping resistor for this.

Without this diode the circuit values are such that on turning off current in the primary the transformer would ring with a period of approximately 2μsec. The diode is connected in such a way that the initial pulse will bias it in a non-conducting direction. During the second half cycle, however, the overswing will cause the diode to conduct thus reflecting the low impedance into the primary and rapidly damping out any further cycles of oscillation. The diode also operates in the same way to prevent the formation of strobe pulses of the wrong polarity when turning on current in the primary.

With regard to the second point raised it was found necessary to employ thermionic diodes in the discriminator because of the high impedance of the associated circuits. In fact, the voltage drop is no great difficulty and the actual circuit used employs a thermonic diode to give the necessary high back impedance in series with a point contact diode which reduces the coupling due to the capacitance of the thermionic type.

Yours faithfully,

K. I. TURNER AND
J. E. THOMPSON,

Ferranti Ltd, Manchester

Reducing Ripple in Stabilized Supplies

DEAR SIR,—I read with great interest Mr. Potok's letter in the August 1959 issue. I myself studied and elaborated the problem of ripple reducing transistor circuits, but my results, calculated and measured, seem in some respects better than those of Mr. Potok.

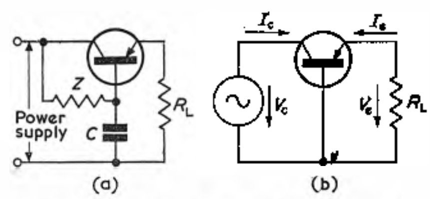


Fig. 1. (a) Basic circuit approximation (b) Equivalent circuit of (a)

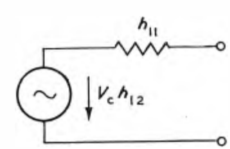


Fig. 2. Thevenin equivalent of the circuit in Fig. 1

The basic circuit and its equivalent is shown in Fig. 1.

To a first approximation suppose that capacitance C is so large, that the a.c. voltage in the base of the transistor is negligible. In this way one obtains a common-base circuit, which is driven by voltage V_c in the collector. Now examine voltage in the emitter loaded by R_L.

The simplest way of calculation is to use the common-base h parameters:

$$V_e = h_{11}I_e + h_{12}V_c \quad \ldots\ldots\ldots \quad (1)$$

$$I_c = h_{21}I_e + h_{22}V_c \quad \ldots\ldots\ldots \quad (2)$$

and

$$V_e = -I_eR_L \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

Substituting equation (3) into equation (1).

$$V_c/V_e = 1/h_{12}\,(1 + (h_{11}/R_L)) \geqslant 1/h_{12} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

and because the value of h_{12} lies between 10^{-3} and 10^{-4} depending on the type of the transistor, the ripple will be reduced by 60 to 80dB.

The Thevenin equivalent of the circuit is illustrated in Fig. 2.

Effecting now the exact calculations and regarding the ripple in the base due to voltage source λV_c, which has an output admittance Y, this circuit and its Thevenin equivalent is shown in Fig. 3. where

$$Y_1 = h_{22}' + \frac{Y(\lambda - h_{12}')(1+h_{21}')}{1 + Yh_{11}'} \qquad \ldots\ldots\ldots\ldots (5)$$

$$Y_2 = h_{22}' + \frac{Y(1 - h_{12}')(1+h_{21}')}{1 + Yh_{11}'} \qquad \ldots\ldots\ldots\ldots (6)$$

and $h'-s$ represents the common-emitter h parameters.

The circuit tried out was built with a Telefunken transistor, type OD603. The d.c. supplied was 800mA, the a.c. voltage in the collector was about 250mV and in the emitter below 0·3mV.

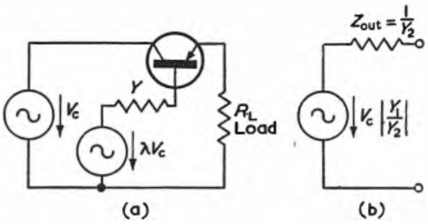


Fig. 3. (a) Exact ripple reducing circuit (b) The Thevenin equivalent

The circuit used is shown in Fig. 4. When calculating λ and Y, the following values are obtained:

$-\lambda \simeq 10^{-3}$

$Y = j0{\cdot}314$ mho ($f = 50$c/s)

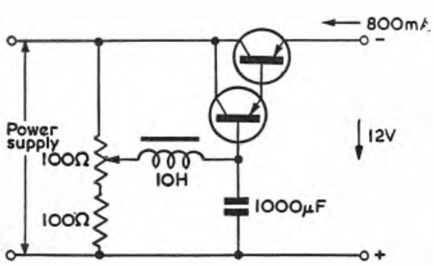


Fig. 4. The final circuit

Substituting these into equations (5) and (6) gives about 10^3—i.e. 60dB—reducing of ripple and hum in accordance with the measured values.

In my opinion this result differs from that of Mr. Potok, since he obtained a ripple reduction of $\beta(h_{21}')$ and this is only 20 to 30 with usual intermediate power audio transistors.

Yours faithfully,

P. May,
Institute for Electrical Power Research,
Budapest, Hungary.

The Correspondent Replies:

Dear Sir,—Mr. May appears to have misunderstood my letter. What I was concerned with was reduction in ripple of an existing supply. If one has a supply with too much ripple one proceeds to filter it out. A simple form of such a filter is an *RC* network with large *C* across the load. The ripple across the load will then be equal to the current through multiplied by the impedance of the capacitor. If the *RC* network were replaced by a transistor with the same capacitor in the base, then the current through *C* would be reduced β times and so because of the low output impedance of an emitter-follower the output ripple will be reduced β times as compared with the simple filter network. The trouble with Mr. May's circuit—which is not new—(vide the last paragraph of my letter and one of the references I mention there) is that it requires such a large time-constant. It is satisfactory to use a 10H choke and 1 000μF capacitor for a 12V supply, but this is not the case for a 2 000V or even a 300V supply. Finally, I did not make it clear in my letter because it seemed self-evident that when using two transistors cascaded the improvement will tend towards β^2, i.e. for $\beta = 30$ total improvement = 900, not far short of Mr. May's 1 000.

Yours faithfully,

M. H. N. Potok,
Royal Military College of Science,
Shrivenham.

A Transistor-Thermistor Feedback Quadrature Suppressor

Dear Sir,—The equipment described in the article by Hutcheon and Harrison in the February issue runs on parallel lines to a device which has been in use for some years in this Company's Laboratories[1].

Here transistor thermistor feedback loops are used to obtain automatically the dynamic balance of transducer bridges when plotting the response to vibration in aircraft structures. In tests involving a multiplicity of transducers, the device enables switching to any transducer without the need for the manual adjustment of 'in-phase' and 'quadrative' balance.

Basically, in the dynamic condition the transducer output is a balanced carrier-suppressed signal with an error of the 'in-phase' component due to misbalance of the transducer, and a 'quadrative' component due to pick-up, phase shift in the transducer etc.

This signal is amplified for eventual display, is further amplified and squared in a diode limited amplifier and is then fed to two gates, one operated at $\angle 0°$ and the other $\angle 90°$ to the carrier signal.

From the $\angle 0°$ gate comes blocks of square waves of one sign during positive excursions of the transducer and of opposite sign during negative excursions. These are integrated in the thermistor heater, and the magnitude of the resultant signal is a function of the misbalance of the transducer.

From the $\angle 90°$ gate each cycle of the carrier gives a signal which is a function of the phase difference between the transducer signal and the carrier. When integrated a signal of the mean phase difference over a cycle of the excitation frequency results.

The thermistors are arranged in arms across the two carrier phases and the resultant signal added to the transducer output in the primary of the amplifier input transformer.

Yours faithfully,

W. A. Winter,
Bristol Aircraft Limited,
Bristol.

REFERENCE

1. Bristol Aircraft Limited, Instrumentation Laboratories Memo. 51.

The Authors Reply:

Dear Sir,—We have been very interested to hear of the system described by Mr. Winter in which two thermistor devices suppress the in-phase and quadrature components of an a.c. carrier signal (but do not follow rapid variations in the in-phase component).

In our own application, only the quadrature component is suppressed by a thermistor device, while a self-balancing recorder displays variations of the in-phase component, which represents a physical variable such as fluid flow.

We do not, of course, lay any claim to initiating the use of thermistors for this sort of purpose and, in fact, one such application is described in Ref. 1 of our I.E.E. paper (i.e. Control Engineers Handbook Section 6.1.8. p. 77). We are, however, quite pleased with our circuit arrangement in which a third thermistor compensates for the effect of temperature changes on the two main thermistors and on all the transistors in the amplifier.

Yours faithfully,

I. C. Hutcheon and
D. N. Harrison,
George Kent Limited,
Luton, Beds.

Correction

In the Letter to the Editor by A. Goral on page 309 of the May issue the equation immediately below equation (2) should read:

$$\frac{B_A + B_B}{N/l\,(I_B - I_A)} = \Delta B/\Delta H$$

In equation (3) S_w is in Oe sec not Oe sec^{-1} as printed. In equation (4) ω^{-8} and in equation (5) ω^8 should read 10^{-8} and 10^8 respectively.

Correction

The left-hand side of equation (7) was printed incorrectly in both the article 'Fast Counting Circuits Using E1T Tubes' by V. Radeka, in the February issue, and the correction in the April issue. This should read:

$$\psi_9 = \ldots\ldots\ldots\ldots\ldots\ldots \quad (7)$$

BOOK REVIEWS

Laplace Transforms for Electronic Engineers

By J. G. Holbrook. 208 pp. 170 Figs. Demy 8vo. Permagon Press Ltd, New York. London. 1959. Price 50s.

A BOOK on a branch of mathematics by an engineer for other engineers is usually very useful and sometimes very good. This one is both and will appeal to a large number of engineers and physicists.

The reviewer found it exceedingly interesting and enlightening. The field of applications of Laplace transforms is quite new to him and by the end of the book he felt that he knew why electrical engineers find them so useful. The examples given are chosen carefully for their simplicity and general utility. It was enlightening that Dr. Holbrook presents these transforms in the traditional manner of mathematicians, using complex variable, and takes great pains to ensure that his readers were keeping up with him. The reviewer does not know whether engineering students approve of this manner of presentation but it does link together many strands in mathematical analysis, Laplace transforms, Fourier series and complex variable, and he is sure that the author is right.

It used to be difficult to convince teachers that the differential and integral calculus were necessary yet now the calculus is taught in the fourth forms at school. Now it is not easy to convince them that Laplace transforms are necessary for engineers, but there is no doubt that, given a clear introduction, they can be explained very easily in an electrical engineering course.

The author writes in a friendly and leisurely manner, as in older text books before the demand for maximum information in minimum space became an obsession with readers and publishers. He tells us not to read more than one section (of about three pages) each day, a suggestion which opens up all sorts of speculation. If each of us reads three pages a day of such books for each day of our lives, and remembered what we read, clearly we should by now be very well read in a limited field; but our book-shelves would be nearly empty. Would Pergamon Press approve?

G. J. KYNCH

Transistoren

By J. Kammerloher. 212 pp. 97 figs. Demy 8vo. C. F. Winter'sche Verlagsbandlung, Füssen. 1959. Price DM.15,60.

THIS small book, or rather monograph, gives a concise and to-the-point elementary introduction to transistor circuits.

The work is obviously intended to be used by junior engineers and students who have, perhaps, no higher qualifications than the equivalent of an ordinary national certificate. It cannot be claimed that any explanations are given, which will enable the reader to understand the action of transistors, nor can the book be described as a textbook, but rather as a guide to the application of transistors.

The initial chapters deal with the transistor characteristic curves and various aspects of transistor applications in amplifiers using the characteristic curves as qualitative design guides. This method is extended to the derivation of transistor equations for small signals connecting current and voltage.

Current, voltage and power amplifiers as well as input and output matching are then treated in an elementary quantitative way.

The h-parameters are then introduced and there follows a simple but fairly comprehensive description of a great variety of transistor applications in grounded base, emitter and collector circuits, stabilization, feedback, etc.

The final third of the book deals with many aspects of complete amplifiers using several stages of transistors, calculation of coupling transformers, power amplifiers and push-pull circuits, etc. Finally, noise in transistors and its significance in small signal amplifiers is described.

In all these matters the text is simple, concise and illustrated by many numerical examples.

The book will certainly enable many junior engineers to use a better-than-empirical approach to the application of transistors.

Unfortunately, apart from the language difficulty, the German habit of using Gothic characters for vector quantities is also used in this book, and this makes the printed equations look unfamiliar and more complicated than is necessary.

An equivalent work in English language and print would be highly desirable.

K. HOSELITZ

Handbook of Automation, Computation and Control, Volume 2

Edited by E. M. Grabbe, S. Ramo and D. E. Wooldridge. 1033 pp. 416 Figs. Medium 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd, London. 1959. Price 140s.

VOLUME 1 of this three-volume work was well received on its appearance in this country last year, and dealt, under the general title of "Control Fundamentals" with the mathematical background to the design and use of computing and control equipment. The present volume is sub-titled "Computers and Data Processing" and, although reference is naturally made to the topics treated in Volume 1, it may be regarded as a completely independent handbook. The forty-one contributors include, apart from the three editors, some other distinguished names in American electronic engineering.

Any handbook dealing with a rapidly developing subject must effect several compromises. Speed of production, allowing coverage of recent developments, conflicts with careful editing for accuracy and avoidance of duplication or omissions. Detailed information on specific equipments can become useless padding in a short time as the equipment is superseded, and there is always the problem of relative emphasis of the treatment of a wide variety of topics in a limited total space.

The present book is not immune from criticism on any of these grounds. Many errors have been noted which could have been eliminated by more checking (for example, an incorrect decimal-binary conversion in Chapter 2, and a figure in Chapter 15 where caption and associated text disagree). All these, however, are errors of detail rather than of substance and will not mislead the reader.

Chapter 2, by far the longest in the book, includes a lengthy and advanced treatment of programming in terms of "forming a correspondence between the original problem functions and the machine functions, and between the problem operator and the machine operator." While such a treatment provides a formal basis for consideration of automatic programming processes, few readers will agree that such a lengthy section should appear in a general work of this nature. For other reasons, the 50 or 60 pages devoted to the detailed order codes of nine machines (eight American and one Russian, nearly all machines now obsolescent) could also perhaps have been better used. For example, the electronic engineer will feel that more than the two pages allowed could have been justified for a fuller discussion of capacitor-diode and ferro-electric storage, super-conductive devices, the twistor and the parametron. Similarly the sections on transistor circuits give undue attention to surface barrier and even point contact types rather than to the much more widely used junction types.

These however, are detailed criticisms of a book which gathers together a wealth of information on most aspects of digital and analogue computing devices (including useful sections on mixed systems and digital differential analysers). Very adequate references are given in most chapters to supplement the text itself and there will be few workers in any way associated with computing systems who do not find the book a valuable addition to their reference shelves.

A. C. D. HALEY

Physics and Electronics in Physical Medicine

By A. Nightingale. 283 pp. 325 Figs. Demy 8vo. G. Bell & Sons Ltd. 1959. Price 30s.

WITH the increasing application of Physics and Electronics in medical practice, there is a need for an elementary book to explain the principles of the subject and to show the relationship between the two disciplines. When one considers the number of text books necessary for studying any one of these subjects, it is obvious that such a book cannot give the reader a full understanding of each. The primary object of the author has been to produce a text book suitable for Physiotherapists studying for the examination of the chartered society and for postgraduate students reading for the Diploma in Physical Medicine of the Royal College of Physicians and Surgeons.

The basic principles of physics are covered in some detail and are explained in a very clear and concise manner. The reader will particularly appreciate the principles being given in conjunction with their medical applications. The principles of electronics are given briefly and although probably adequate for the student, would need to be given in more detail for the worker in medical physics. The multivibrator is described in some detail but no mention is made of the monostable or bistable conditions which are as widely used. Basic techniques commonly used in Physical Medicine such as the cathode follower and the rejection of interfering signals could have been explained in more detail. Had the standard definition of frequency bands been used rather than the medical interpretation the anomalous term 'High frequency long wave diathermy' could have been avoided.

The bibliography has adequate references to Physical Medicine, but could have been supplemented with recommended text books for further study of the basic subjects described.

The main object of the author has certainly been achieved and in addition the book should be of value to those interested in medical physics who do not wish to make an intensive study of the subject.

W. J. PERKINS

Millimicrosecond Pulse Techniques

By I. A. Lewis and F. H. Wells. 417 pp. 195 Figs. Demy 8vo. 2nd Edition. Pergamon Press Ltd, New York. London. 1959. Price 50s.

THE reviewer had the pleasure of reviewing the first edition of this book in 1954. During the intervening years the work has become widely appreciated, so that now it is considered required reading in many laboratories engaged on fast pulse work. The second edition, considerably expanded to include new developments, is therefore most welcome.

It is inevitable that a book which leans heavily towards the review approach should treat some aspects of the subject in a superficial manner. While this criticism would scarcely be levelled against the chapters on transmission lines and transformers, the reader may often puzzle as to the detailed operation of certain of the circuits which appear in the later sections. Often the circuit treatment is superficial and non-critical. However, the authors are to be particularly commended on the exhaustive reference section. Not only does this provide a comprehensive literature source for the serious worker, but it enables the reader to make reference to the original paper should he require a more complete treatment.

As the authors correctly predict, semiconductor devices are destined to make important contributions to fast pulse techniques. Indeed, the transistor trigger circuit is now faster, in general, than its vacuum tube counterpart. The avalanche transistor switch may well be surpassed in speed by the diffused-base mesa transistor which, in turn, may give way to the tunnel diode. Both the tunnel diode and the parametric amplifier offer interesting wide-band amplifier possibilities. It is scarcely fair to expect the authors to have included all the advances mentioned above, for some of them remain active research projects. Yet, in view of such significant activities, the treatment of semiconductors must be regarded as slight.

All in all there can be no doubt as to the value of this new edition of "Millimicrosecond Pulse Techniques." It carries on the worthy tradition of its predecessor and should continue to be the standard reference on the subject.

N. F. MOODY

Masers

By G. Troup. 146 pp. 20 Figs. Foolscap 8vo. John Wiley & Sons Inc., New York. Methuen & Co. Ltd, London. 1959. Price 13s. 6d.

THE title of this book is the abbreviated form used for the devices employed in obtaining Microwave Amplification and Oscillation by Stimulated Emission of Radiation.

A discussion of the stimulated emission process is followed by an outline of methods used to obtain the necessary conditions for amplification. The effects of various physical processes upon amplifier efficiency and upon the effective frequency response of a molecular transition used to amplify are described. Derivations of the gain, bandwidth, and noise factor of the travelling-wave line type of stimulated emission amplifier and of the resonant cavity type of amplifier are also given, together with a review of the experimental work on 'masers' and a comprehensive bibliography.

Crystal Growth

365 pp. 195 figs. Medium 8vo. 2nd Edition. Butterworths Scientific Publications. 1959. Price 60s.

THIS general discussion of the Faraday Society was first published in 1949. It has been out of print for a long time and has been reprinted in response to the many requests received by the Faraday Society.

Stereo Handbook

By G. A. Briggs. 146 pp. 88 figs. Demy 8vo. Wharfedale Wireless Works Limited, Bradford. 1959. Price 10s. 6d.

THIS book, which has been augmented by contributions from R. E. Cooke, N. H. Crowhurst, S. Kelly, C. E. Watts and R. L. West, is non-technical throughout and is intended to help the amateur to understand stereo and its applications. Maximum space has been allocated to pick-ups, loudspeakers and recording techniques in that order of importance.

Elsevier's Telecommunication Dictionary (in six languages)

By A. Visser. 1011 pp. Medium 8vo. D. Van Nostrand Company Ltd. 1960. Price 147s.

IN the technical dictionary published in 1955 by the Netherlands General Post Office for its own staff, Dutch was used as the basic language. In the present edition, however, English has been chosen as the basic language. All terms used in the telecommunication world can be found in the basic table arranged on an English alphabetical basis. The equivalents in French, Spanish, Italian, Dutch and German are then given. Following this basic table are lists for each language, each word being keyed to the basic table so that the user can find the equivalent term in English and all the other languages. The system of thumb-indexing adopted makes it easy to find any language at once.

Grundlagen und Anwendungen der Informationstheorie (Principles and Applications of the Information Theory)

By W. Meyer-Eppler. 404 pp. 178 Figs. Medium 8vo. Springer-Verlag, Berlin. 1960. Price DM98.

INFORMATION Theory is a very young but lusty child: just how lusty is indicated by the appearance of this book, because it is the first volume of a new series on Cybernetics modelled on the famous "yellow series" Grundlehren der mathematischen Wissenschaften. This new series is bound in a distinguished looking grey cloth and, like the yellow series, published by the Springer-Verlag; the book is, accordingly, superbly printed and handsomely produced, and the price is just over 8 gns.

The editor of the series is also the author of this first volume. He collects and presents material from very many widely dispersed sources, the majority of which are in English and more easily accessible to the readers of ELECTRONIC ENGINEERING than to those at whom the book is aimed.

The presentation makes much use of mathematical formulation, but the mathematics is descriptive rather than deductive, and the reader is not expected to follow any deep mathematical reasoning. Ample references help him to find a more thorough treatment if he has the taste for it.

About half the book is concerned with those ramifications and applications of the theory that deal with the sense organs and their rôle in communication. It is in this field that the author's own original contributions have been made, and this is perhaps the most rewarding part of the book. An auspicious beginning to a brave new venture.

B. H. NEUMANN

Principles of Frequency Modulation

By B. S. Camies. 147 pp. 87 Figs. Demy 8vo. Iliffe & Sons Ltd. 1959. Price 21s.

THE author of this book states in the preface that he is writing 'primarily for students, radio engineers and radio amateurs'. The term 'engineer' is unfortunately used as an umbrella to shelter a wide variety of people and here it appears to mean radio servicemen; certainly the professional radio engineer will find nothing of value to him in this 'popular' account of the principles and applications of frequency modulation.

There are, in effect, six chapters covering basic principles, noise and interference, generation and detection of f.m. signals, the v.h.f. receiver and non-broadcasting applications. The last chapter deals very superficially with f.m. applied to radar, telegraphy and facsimile. Chapter 2 on the principles of f.m. shows some of the difficulties in trying to appeal to the student and amateur at the same time. The difference between the amplitude-modulated carrier ($A(1+m \cos pt) \cos \omega t$) system and suppressed-carrier ($Am\cos pt.\cos\omega t$) system is heavily underlined and yet a few pages later Bessel functions are introduced. A modulation factor m is used to multiply maximum deviation frequency in order to indicate any intermediate value of modulation but the similarity to amplitude modulation factor is not pressed home so that its point is lost. It also leads to the use of the term deviation ratio for the ratio of maximum deviation to modulating frequency with no statement as to whether the modulating frequency is maximum, minimum or indeterminate. The abscissa scale of Fig. 2.8 is named deviation ratio but the caption refers to modulation index.

The student is again presented with more difficulties in Chapter 3 where interference from an unwanted carrier is defined in terms of peak phase ratio but the numerical example (immediately following) is quoted in terms of peak-frequency deviation ratio. Fig. 4.3b will also raise queries in the student's mind for L_1 is paralleled by L_3 when in fact it should be in series, and he will not know what to make of Fig. 5.20 where the diode current can apparently be reversed by reversing the connexions of an electrolytic capacitor across the load resistance.

This book is likely to appeal mainly to the technician who seeks a nodding acquaintance with the subject and desires a descriptive approach; the serious student will find that it raises as many problems for him as it solves.

K. R. STURLEY

Principles of Optics

By M. Born and E. Wolf. 803 pp. 365 Figs. Crown 4to. Pergamon Press Ltd. 1959. Price 120s.

THIS is a new book incorporating the results of recent advances in classical theory and is not a translation of the earlier book in German by M. Born entitled OPTIK, although the aim has been retained of presenting optics deductively as a system based on Maxwell's equations. It deals with those optical phenomena which can be described with the aid of a continuous distribution of matter and, inside this frame, gives a complete picture of the present knowledge of optics.

Progress in Dielectrics

Edited by J. B. Birks. 225 pp. 73 Figs. Royal 8vo. 2nd Edition. Heywood & Co. Ltd. 1960. Price 55s.

THE stated object of this series is to co-ordinate current knowledge of dielectric phenomena, materials and techniques and to review recent progress. It is intended to serve the needs of anyone with an interest in dielectrics, whether it be from the electrical engineering angle or from the point of view of the chemist concerned with manufacture. Review papers of the type presented in a volume of this kind are invaluable in that they provide a convenient guide to the ever increasing literature. It is obvious that the authors of such papers must remember that their readers will not in general be specialists in the subject being discussed and that adequate background material must be provided. Judged from this standpoint, the present volume is admirable and all the papers should be intelligible to the average electronic engineer. Much of the material is at a level appropriate to a text book and this will undoubtedly make the book useful to the wide range of readers at which it is aimed.

The first paper gives a concise summary of the theory of polarization and absorption in dielectrics and provides an excellent introduction to this subject. Particular attention is given to the thermodynamical aspects. Three papers deal with specific materials, polymers, glass and high permittivity ceramics. All of these materials have applications in the electronic field, the last group being of rapidly increasing importance in providing compact capacitors. Each of these papers deals with dependence of the dielectric properties on the chemical structure and makes excellent reading for the electronic engineer who wishes to know the reasons for the properties on which he relies. A further paper deals with the effects of irradiation on polymers, again a subject of practical interest in that irradiated polythene is becoming increasingly used as a dielectric when the possibility of temperatures above 100°C has to be considered. The final contribution surveys the properties of artificial dielectrics as used in microwave lenses.

The first volume of this series has been well received and the present one maintains the same standard. The papers are sufficiently authoritative and supported by comprehensive bibliographies to be of value to the specialist and also of sufficient clarity to satisfy the general reader.

J. BROWN

Microwave Data Tables

By A. E. Booth. 61 pp. Crown 4to. Iliffe & Sons Limited. 1959. Price 27s. 6d.

THIS book comprises a collection of accurately computed tables which are primarily intended for use by engineers and scientists engaged in research and development work involving waveguides and similar transmission lines. Twenty-six tables are included, with over 23 000 calculations which were carried out on an electronic digital computer to at least one more figure than in the tables themselves, thus ensuring their accuracy. All the independent variables are measurable quantities and are given to the limit of measuring accuracy, while the dependent variables are given to one more figure than would be required for most practical purposes. Guide wavelength tables are given for nine sizes of rectangular waveguide and two sizes of circular waveguide—the choice of sizes being directly influenced by current usage. The velocity of light used in the calculations is the most recently accepted value. The mathematical basis of all formulae used in the computations is stated and comprehensive notes on the use of the tables are included.

Short News Items

Gresham Transformers Ltd, Hanworth, Middlesex, have just built what is believed to be the world's largest electro-magnetic control system or transductor. This equipment, which weighs over twenty tons, will be used in conjunction with germanium rectifiers manufactured by Westinghouse Brake and Signal Co. Ltd, to control an output current of 12 000A. It is to be used for research work in Holland. The value of the order is £17 500.

The United Kingdom and the United States have begun co-ordination of their time and frequency transmissions. The purpose of this synchronization is to provide a uniform system of time and frequency transmissions, which is needed in the solution of many scientific and technical problems in such fields as radio communications, geodesy, and the tracking of artificial satellites. Participating in the project are the Royal Greenwich Observatory, the National Physical Laboratory, and the Post Office Engineering Department in the United Kingdom, and, in the United States, the U.S. Naval Observatory, the Naval Research Laboratory, and the National Bureau of Standards. This programme follows previous co-operative efforts of these agencies to achieve uniformity and simplification in procedures. The following time signal services are involved:

U.K.

Rugby GBR 16kc/s
Rugby MSF 2·5, 5 and 10Mc/s 60kc/s

U.S.A.

Beltsville WWV 2·5, 5, 10, 15, 20, 25Mc/s
Canal Zone NBA 18kc/s
Hawaii WWVH 5, 10, 15Mc/s

Co-Ordination began in January. It is expected that by the end of 1960 the time signals from all the participating stations will be emitted in synchronism to the thousandth of a second. Such accuracy has been needed for some time in tracking artificial satellites on a world- wide basis.

The Department of Nuclear Science and Technology of the Royal Naval College at Greenwich, which was set up last year to meet the Navy's requirements for nuclear trained officers has now equipped its laboratories with a sub-critical reactor and a pulsed neutron generator marketed by Miles Hivolt and based on a design produced by Birmingham University Physics Department with the assistance of the Department of Scientific and Industrial Research. The nuclear courses at the College are of post-graduate standard and the equipment will enable a wide range of nuclear physics experiments and research to be carried out by the advanced course students and staff.

Ultra Electronics Ltd is to sponsor two University scholarships. One, a two-year research scholarship for an honours graduate in Engineering, is available for the term commencing next Autumn. The second scheme is an undergraduate sandwich course. This is open to students who pass advanced level G.C.E. this summer and who are accepted by a University for enrolment in the autumn of 1961 to read electrical engineering. The student will complete one year in the Engineering Department of Ultra Electronics Ltd before taking his University course and will return for a similar period afterwards.

The Signals Research and Development Establishment are arranging to hold a Battery Symposium at The Pavilion, Bournemouth on October 18-20 this year. About 25 papers will be presented covering a wide field of subjects concerned with Research, Development and Application of Battery Systems. Further details can be obtained from Mr. F. C. Wells S.R.D.E. Somerford, Christchurch, Hants.

The British Computer Society is holding its Second Annual Conference at the Sun Pavilion, Harrogate on July 4-7, 1960. Abstracts of the main papers will be available at the Conference. Requests for application forms should be made to Miss D. E. Pilling, The Electronic Computing Section, The University, Leeds 2, Yorkshire.

The National Physical Laboratory announces that the name of the Control Mechanisms and Electronics Division is to be changed to the Autonomics Division. The word 'Autonomics', meaning 'Self governing', is particularly apt as a description of the new types of device that the Division aims to develop. The Division already has an extensive record in the field of automatic control and in pioneering digital computers, but now that industry has adopted these techniques, research has moved forward into new fields. Existing machines are deterministic in that their behaviour can be predicted from their design and their environment. Conversely they can be designed only after a detailed mathematical analysis of the factors involved. By working closely with biologists in studying the methods which living physical systems use to achieve control, the newly named Autonomics Division hopes to discover underlying principles of machines which can improve their own performance can be built. The autonomic devices of the future will be self-governing machines which discover for themselves how to recognize, how to translate and how to control.

Hatfield Technical College, Herts has now received official confirmation from the National Council for Technological Awards that it has recognized Hatfield for the conduct of courses leading to the Diploma in Technology. The courses so far approved are in Aeronautical Engineering and Electrical Engineering, and students studying these subjects at the college in sandwich courses will eventually qualify for the award of Dip. Tech. (Eng).

Ferranti Ltd, Edinburgh and Bendix Aviation Corporation of America have concluded an agreement for the sale in the U.S.A. of the Ferranti systems of machine tool control. These systems are the transistor-hydraulic continuous-path machine tool control equipment, numerical position control equipment and inspection machines. Bendix will also market and service the British built systems throughout the U.S.A. and will set up a computer centre initially in Detroit to supply magnetic tapes for this equipment. The inspection machine which will be exhibited at the forthcoming British Exhibition in New York utilizes transistorized electronic counting circuits and optical measuring gratings to inspect machined parts to an accuracy of 0·001in. The operator has under his control an inspection probe with interchangeable tips suitably shaped to locate at any point on a machined part. This probe is supported from a rigid cantilever beam projecting over the work table and is free to move over the length and breadth of the table. Movement of optical gratings attached to both longitudinal and transverse axes of the machine create a Moiré fringe pattern that activates photocells. Changes in light intensity are converted into pulses that are stored by transistorized bi-directional decade counters.

Royston Instruments Ltd, of Byfleet, Surrey and Lockheed Aircraft Services Ltd, of Ontario, California have signed a joint marketing agreement for the Midas range of Magnetic Tape Data Recording equipment. Lockheed Aircraft Services have agreed to handle the full range of Midas data recorders and data reduction equipment throughout the Western Hemisphere and Australasia. This agreement follows closely on the Federal Aviation Agency regulation requiring all turbine aircraft in the United States, flying at heights of over 25 000ft to carry flight recording equipment. The new range will supplement the Lockheed Aircraft Services 109C flight recorder, already delivered or on order for nearly 400 American jet aircraft. The LAS 109C was developed to meet the FAA mandatory requirement for a crash recorder only, and it is able to record only on a comparatively limited scale. Although the Lockheed recorder is the cheapest known method of meeting the legal requirements imposed by the FAA, the Midas system provides a much wider variety of data recording and handling equipment and is considered to be at least three years ahead of anything being developed in this field.

The Benelux Section of the Institute of Radio Engineers is to hold an international symposium on data transmission at Delft, Holland on September 19-20, 1960. The symposium will be concerned with the problems of transmitting and receiving information in digital form. Particular emphasis will be placed on the behaviour of practical communication networks—including existing telephone systems, existing and planned military systems, and schemes of the future, such as those that use satellites. The symposium will be conducted in English and papers already promised indicate good representation of the work being carried on in the U.S.A., as well as that conducted in Europe. All correspondence regarding the symposium should be addressed to The Secretary, The Benelux Section I.R.E. Postbus 174, The Hague, Holland.

Air Marshal Sir Raymund G. Hart, Director, Radio Industry Council, has from April 1 taken over the functions of secretary to the Council in succession to Mr. G. B. Campbell whose duties as secretary, Radio Industry Exhibitions Ltd, no longer allow time to act in his former capacity with the Council. Air Marshal Sir Raymund Hart was appointed Director, R.I.C., in March 1959 on his retirement from the Royal Air Force.

The British Investment Casters' Technical Association of Sheffield has collected from its member companies some typical examples of investment castings having dimensionally accurate and good mechanical properties and of particular application to the electronic industry. Components produced by the investment (lost wax) casting technique are being used in a variety of electrical and sound reproduction mechanisms. For the latter applications permanent magnet alloys and materials of highly magnetic permeability have been widely used. Substantial economics have been effected by the adoption of investment castings for these applications since the accuracy which can be achieved has obviated or considerably reduced machining operations. The ability to cast to a high standard of accuracy is particularly useful in the case of alloys which are difficult to forge and machine.

The BBC's Orkney V.H.F. sound broadcasting station at Netherbutton, near Kirkwall was brought into full power service on May 2. A directional aerial is used giving a maximum effective radiated power of 25kW and the transmissions are horizontally polarized. Orkney receives the three V.H.F. sound programmes direct from the BBC V.H.F. station at Thrumster, near Wick which in turn receives them from the V.H.F. station at Meldrum, Aberdeenshire. The new station is expected to serve almost the whole of the Orkney Islands and the northern part of Caithness where its coverage will link up with that of the BBC's V.H.F. station at Thrumster near Wick, which was brought into full power service on 1 March 1960. The combined population coverage of the two stations will be approximately 43 000.

The Transitron Electronic Corporation of Wakefield, Massachusetts, announces the formation of a European Sales Subsidiary, Transitron Electronics S.A. The headquarters of this company are in Zug, Switzerland and it has sales subsidiary offices in London, Paris and Munich with warehouse facilities in Amsterdam. The London Office, Transitron Electronic Ltd, is located at Bilton House, Ealing, London W.5 and this company will from now on be responsible for marketing the complete range of the parent company's semiconductor and other products in the United Kingdom.

The Institution of Electronics (Northern Division) Fifteenth Annual Electronics and Instruments Exhibition and Convention is to be held at the Manchester College of Science and Technology on 7 to 13 July, 1960. The Manufacturers' Section will include an extensive display of British and Overseas Electronic Devices and Instruments and the Scientific and Industrial Research Section will include displays by the Admiralty, the Radio Research Station of D.S.I.R. (work done on 'Earth Satellites'). The Convention will include a programme of lectures and film shows. Complementary admission tickets and copies of the 'Preview of the Exhibition and Convention' (post free 6d and available from 10 June) and of the 'Handbook of the Exhibition and Convention' (post free 3s and available from 1 July) may be obtained from W. Birtwistle, The General Secretary, The Institution of Electronics, 78 Shaw Road, Rochdale, Lancs.

Correction

It is regretted that in the article 'Effect of Minority-Carriers on the Dynamic Characteristic of Parametric Diodes' by I. Hefni on page 226 of the April issue, Figs. 5 and 6 are upside down and should read Figs. 8 and 7 respectively, while Fig. 7 should read Fig. 5 and Fig. 8 should read Fig. 6.

(Pages 389 to 400 French and German Sections)

PUBLICATIONS RECEIVED

GUIDE FOR BUYERS has been issued by The National Federation of Engineers' Tool Manufacturers, Light Trades House, Melbourne Avenue, Sheffield 10, and gives details of the suppliers of British made engineers' cutting tools. A feature of the publication is the alphabetical list of tools in four languages—English, French, German and Spanish. Overseas buyers will have no difficulty in finding in the Guide the names and addresses of British suppliers of all types of engineers' cutting tools. Part 4 of the Guide contains an alphabetical list of trade marks, trade names and brand names. Price 5s.

EUROPEAN FREE TRADE ASSOCIATION is a new F.B.I. booklet which gives a brief historical background leading up to the signing of the Outer Seven Convention and provides a commentary of the terms of the Convention itself. The booklet sets out objects of the Convention then deals with the sections covering import duties; import quotas; definition of origin and deflexion of trade, and the escape clauses included to safeguard members of the Outer Seven in the event of any deterioration in their balance of payments of difficulties experienced in any particular sector of industry. Appendices give the basic materials list, details of the special provision in the Convention for Portugal and a series of charts giving vital economic data on each of the member countries at a glance. Federation of British Industries, 21 Tothill Street, London S.W.1.

BTH ESCORT MARINE RADAR is a fully illustrated brochure describing the main features of 'Chart-Plan' true-course displays, and including a technical specification and list of world service agents. An important feature of Escort Marine Radar is that the operation of its true-course generator remains automatic under all conditions. Copies of this brochure (publication No. 4231-71) may be obtained from A.E.I. Electronic Apparatus Division, Radar Sales, Blackbird Road, Leicester, or any A.E.I. Regional or District Office.

THERMAL STRESSES IN LONG PRISMS BY RELAXATION METHODS by J. D. Cummins is an unclassified article issued by the Winfrith Secretariat, Atomic Energy Establishment, Winfrith, Dorchester, Dorset, which is available from H.M.S.O. price 7s. Also available from the same source at a cost of 6s. is an article entitled AUTO-CONTROL EXPERIMENTS ON DIDO USING DISCONTINUOUS FEEDBACK by L. A. J. Lawrence, in which experiments on auto-controlling the reactor DIDO are described and the equipment design discussed in some detail. The experiments are carried out to show the suitability of an on/off type of control for the maintenance of (a) a constant flux level in the presence of noise and (b) constant period during power change.

'PAXOLIN' TUBES AND CYLINDERS is a revised edition of the booklet published by The Micanite and Insulators Co. Ltd, Empire Works, Blackhorse Lane, Walthamstow, London E.17. This pamphlet deals with tubes and cylinders of bonded paper grouped according to their characteristics into two classes—Paxolin and Elephantide. Paxolin is the registered trade name for those manufactured by the above Company which are classified by the British Standards Institution as 'Synthetic resin-bonded sheets and tubes'. While the information in this pamphlet relates chiefly to insulation for electrical purposes, the light weight, combined with good mechanical properties of Paxolin tubes, opens up applications in other fields.

G.E.C. VALVES AND CATHODE RAY TUBES, 1960 EDITION is a catalogue produced by the M.O. Valve Co. Ltd, a subsidiary of The General Electric Co. Ltd. Several hundred devices are listed under some twenty main groups and there are many photographic illustrations. Additions include a comprehensive index, a list of principal overseas distributors, and selection charts for some valve ranges. The General Electric Co. Ltd, Magnet House, Kingsway, London W.C.2.

VOLTAGE STABILIZATION is a booklet dealing with the constant voltage transformers manufactured by Advance Components Ltd, Roebuck Road, Hainault, Ilford, Essex. The main function of this static electro-magnetic voltage regulator is to provide a constant voltage output from its secondary winding which shall be unaffected by changes in the input to the primary.

Le Salon de L'AUTOMATION DE L'ELECTRONIQUE ET DES INSTRUMENTS

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

Traduction des pages 376 à 380

ALMA COMPONENTS LTD

55 Holloway Road, London N.19

RÉSISTANCES BOBINÉES

Un certain nombre de nouveaux modèles de résistances bobinées, dont les suivantes, furent exposées:

Type Y

Une nouvelle série de résistances sur bobines de format américain pour assurer leur interchangeabilité. Evaluées à $\frac{1}{4}$, $\frac{1}{3}$ et $\frac{1}{2}$ W et réglées 0,1%, ces résistances sont d'un haut degré d'excellence de fabrication et leurs coefficients de température sont garantis supérieurs à 0,002%/°C.

Type A4

Première de la série Type A, cette résistance est hermétiquement scellée et tropicalisée et elle est en cours d'homologation par DEF.5113. Elle fonctionne dans la gamme −55°C à +120°C et ses caractéristiques sont de 100 ohms à 400 k. ohms, réglées à 0,1%.

Type C

Une résistance bobinée composée, avec coefficient de température garanti de zéro ±4 parties par million par °C, dans la gamme de 0°C à 100°C. Bobinées sur bobines normales ALMA Type S, ces résistances sont évaluées à 1W (Type C1), $\frac{1}{2}$ W (Type C2) et $\frac{1}{4}$ W (Type C4). Les tolérances sont de ±0,1%.

EE 22 751 *pour plus amples renseignements*

ASSOCIATED ELECTRICAL INDUSTRIES (Woolwich) LTD

155 Charing Cross Road, London W.C.2

TRANSISTORS

Les dispositifs récemment réalisés pour des applications industrielles comprennent quatre transistors de puissance. Les caractéristiques de courant de pointe de deux de ces dispositifs sont de 10A et celles de tensions de base collecteur sont de 80 V et 100 V (types XC155 et XC156). Quant aux deux autres transistors, leurs caractéristiques de courant de pointe sont de 3A et celles de leurs tensions de base collectrice sont de 40 V et 60 V (types XC141 et XC142).

Une gamme de trois transistors de dérive a été également introduite en vue d'applications exigeant la commutation à grande vitesse. La fréquence minima à laquelle le gain de courant est unité (ou f1) pour ces transistors est de 20 MHz, 40 MHz et 60 MHz respectivement. Les dispositifs présentés comportaient, d'autre part, deux transistors de commutation du modèle "Mésa", à fréquence minima de 25 MHz et 35 MHz respectivement (types XA 161 et XA 162), offrant une résistance de saturation réduite, une dissipation au collecteur maxima absolue de 75 mW a 55°C et une tension de base collecteur maxima absolue de 13 V.

Pour la commutation de basse tension à vitesse réduite, le transistor type XB 121, à tension de base collecteur de 105 V, est d'un intérêt considérable.

Une gamme de trois transistors en alliage de germanium npn, pour circuits de commutation supplémentaires, fut également présentée. Ce sont les types XA 701, XA 702 et XA 703, qui ont des fréquences de coupure minima de 3 MHz et 10 MHz et une dissipation collecteur de 70 mW a 55°C.

Le domaine des transistors au silicium fut représenté par une série de transistors à jonction diffuse, à dissipations de l'ordre de 30 W à 100°C.

On a pu voir en outre des transistors de dérive à amplification haute fréquence allant jusqu'à 100 MHz, ainsi que la gamme, plus classique et plus connue, de transistors par alliage pnp pour hautes et basses fréquences, dont certains sous moulage d'un intérêt particulier pour les utilisateurs de circuits imprimés.

La plupart des nouveaux transistors industriels sont d'ailleurs sous moulage JEDEC, pour l'emploi dans des circuits imprimés.

EE 22 752 *pour plus amples renseignements*

AVO LTD

Avocet House, 92-96 Vauxhall Bridge Road, London S.W.1

APPAREIL DE MESURE DE CARACTÉRISTIQUES DE LAMPES

(Illustration à la page 376)

Un appareil perfectionné de mesure de caractéristiques de lampes vient d'être réalisé.

Il s'agit du Mark 4 qui, tout en gardant tous les avantages des modèles précédents, comporte les perfectionnements suivants:

(a) La possibilité considérablement étendue de réglage correct des potentiels de grille négatifs, particulièrement aux basses valeurs.

(b) Des moyens plus précis et plus concis pour déterminer le courant de grille, en particulier lorsque la lampe soumise à l'essai accuse une émission primaire ou inverse.

(c) Des tensions anodiques supplémentaires pour vérifier les lampes à basse tension anodiques récemment créées.

(d) Réglage amélioré de l'alimentation en tension anodique et d'écran, afin de reduire les problèmes d'impédance réfléchie.

(e) Disjoncteur à maxima pour les mouvements.

L'appareil est prévu, en outre, pour des tensions anodiques de 12.6 V à 400 V par plots, des tensions d'écran de 12,6 V à 300 V par plots, une tension de grille a variation continue de 0 à 100 V, un courant anodique de 2,5 mA à 100 mA de déviation sur la totalité de l'échelle, des courants d'écran dans la même gamme, un courant de grille de 0 à 100 μA, des tensions de chauffage de 0,625 à 117 V, une conductance mutuelle de 0,1 à 60 mA/V et il comprend, enfin, un dispositif spécial de marche/arrêt.

Les diodes et les redresseurs peuvent être vérifiés dans des conditions de charge de filtrage de 1 mA à 180 mA.

L'appareil est fourni complet avec manuel d'utilisation détaillé.

EE 22 753 *pour plus amples renseignements*

B. & K. LABORATORIES LTD

4 Tilney Street, Park Lane, London W.1

CHARGEUR DE CADRE À BANDE MAGNÉTIQUE

Ce chargeur serait le plus petit qu'on ait conçu à ce jour pour le traitement de cadres à bande ayant jusqu'à 120 pieds de longueur. Il assure une conversion simple, depuis l'enregistreur/reproducteur bobine-a-bobine de laboratoire, jusqu'au dispositif à boucle continue. Il contient des rouleaux pouvant recevoir facilement des bandes de 6,35 mm, 12,7 mm ou 25,4 mm, atteignant jusqu'à 36 m de longueur.

Le parcours de la bande est extrêmement simple et se prête à une mise en place rapide. Les longues boucles peuvent être interchangées rapidement et toutes les vitesses de bande normales, de 47,6 mm/sec à 524 mm/sec, peuvent être choisies instantanément.

Le chargeur peut être retiré de l'Enregistreur P.S.200 et remplacé par un chargeur bobine-a-bobine, courant de 266, 7 mm.

Ce nouveau chargeur de bande permet d'utiliser l'enregistreur P.S. 200 pour des applications de retardement et des analyses de spectres de grandes quantités de données, et il peut loger jusqu'à 14 pistes d'informations. De telles applications ont nécessité jusqu'à présent de volumineux appareils, montés sur supports et comportant des éléments suspendus de guidage fort encombrants ou des pistes de bande entrelacées, au réseau compliqué. Grâce au nouveau chargeur, les utilisateurs pourront commencer par l'emploi d'un instrument à cadre ou à bobine-contre-bobine et étendre plus tard la série des applications.

EE 22 754 *pour plus amples renseignements*

ENREGISTREUR À ENCRE

L'Oscillomink 230B est un appareil enregistreur à encre, pouvant enregistrer jusqu'à 1 kHz, avec une sensibilité variant de 0,2 mm/mV à 4 mm/100 V. Le système d'écriture directe est basé sur un nouveau principe, selon lequel on fait gicler à grande vitesse un jet d'encre extrêmement fin hors d'un ajutage de diamètre capillaire, placé au centre moteur du galvanomètre. Le jet libre ne faisant pas corps avec le moment d'inertie du système, une haute fréquence de résonance, d'environ 650 Hz est donc réalisable. La ligne de base peut être déplacée vers une position appropriée sur le papier sans affecter la linéarité. Lorsqu'on enregistre avec plusieurs voies, les traces peuvent se croiser ou être enregistrées avec une ligne de base commune sans brouillage. L'enregistreur est également muni d'un réglage individuel de la largeur de la ligne. Les vitesses d'avance du papier sont de 5, 10, 20, 50, 100 et 200 mm/sec et elles peuvent être réglées en cours de fonctionnement. La précision est supérieure à 1%.

Des accessoires sont fournis pour des vitesses d'avance du papier de 0,5, 1 ou 2 m/sec, commandées à distance ou sur l'appareil même.

L'Oscillomink 230B est doté de deux amplificateurs push-pull jumelés de courant direct, pour le réglage approximatif ou précis de la sensibilité, la correction de phase et d'amplitude, une bonne discrimination entre les signaux symétriques et assymétriques et un étage final à modulation par fréquence porteuse avec couplage à détente.

Il existe d'autres modèles comprenant des enregistreurs à 8 et 12 voies et des amplificateurs jumelés.

EE 22 755 *pour plus amples renseignements*

R. H. COLE (Overseas) LTD

2 Caxton Street, London S.W.1

MESURE DU NIVEAU DE FRÉQUENCE PORTEUSE

(*Illustration à la page 377*)

Cet equipement, fabriqué par Siemens & Halske, se compose d'un oscillateur de niveau et d'un enregistreur de niveau sélectif. Il permet d'effectuer des mesures sélectives de niveau, d'atténuation et de gain à des fréquences entre 30 kHz et 15 MHz. Ce spectre comprend, par exemple, des systèmes téléphoniques à fréquence porteuse multivoies, utilisant des cables coaxiaux ou hertziens, le groupe de base du CCITT (60 à 108 kHz, 12 voies vocales), le supergroupe de base du CCITT (312 à 552 kHz, 60 voies vocales), le groupe principal de base du CCITT (812 à 2044 kHz, 300 voies vocales) et le super-groupe principal (8,516 à 12,388 MHz, 900 voies vocales).

Les qualités particulières de cet équipement sont la grande précision de la fréquence du générateur et le dispositif d'accord du détecteur, obtenus grâce à un système de verrouillage de fréquence électronique piloté par quartz. A n'importe quel point de la gamme totale de fréquences, parcourue en un balayage de l'écran, une lampe indicatrice s'éclaire au moment où la fréquence s'est fixée sur la fréquence correspondant à la division d'échelle devant être établie (tous les 100 kHz). Ceci accélère considérablement le réglage des fréquences. Des valeurs de fréquence entre les divisions d'échelle peuvent être réglées jusqu'à environ 200 Hz, à n'importe quel point de la gamme totale, au moyen d'une commande d'accord différentiel supplémentaire.

Grace au dispositif d'accord automatique des commandes de fréquences différentielles et principales de l'enregistreur de niveau, toutes les mesures, lorsqu'il n'y a pas nécessité de séparer physiquement le générateur et le détecteur, peuvent être effectuées à partir du générateur aussi simplement et aussi rapidement que des mesures à large bande.

La gamme de mesure étendue et la grande sélectivité, alliées au rejet élevé de la fréquence image et à la faible distorsion non linéaire résiduelle, permettent des mesures de diaphonie et d'atténuation de distorsion; le dispositif de verrouillage de fréquence électronique ajoute à la facilité de ces mesures. La stabilité avec le temps de ces réglages est de grande utilité pour le contrôle continu.

Pour les besoins moins importants de précision de fréquence, il existe des modèles de cet equipement sans le dispositif de verrouillage de fréquence électronique, avec étalonnage néper ou décibel.

L'oscillateur de niveau peut être modulé en amplitude à des fréquences de 0 à 100 kHz par une source extérieure.

Vu que l'appareil de mesure de niveau répond aux valeurs efficaces, il peut être utilisé pour des mesures de bruit quelconque, par exemple, de concert avec le générateur de parasites intermittents type Rel. 3W432.

EE 22 756 *pour plus amples renseignements*

EKCO ELECTRONICS LTD

Southend-on-Sea, Essex

ÉCHELLE AUTOMATIQUE

(*Illustration à la page 377*)

Le nouvel appareil Ekco Type N60 est une échelle automatique et un compte-temps combinés, incorporant une alimentation haute tension stabilisée, un amplificateur d'entrée à grand gain et un analyseur de hauteur d'impulsion. Il comporte également un compteur avec choix, par bouton-poussoir, de gammes de 1000, 10 000 et 100 000 coups/seconde.

L'echelle peut être arrêtée automatiquement à des comptages prédéterminés allant jusqu'à 1000 000 et à des moments prédétermines allant jusqu'a 100 000 sec. Le compte-temps peut, d'autre part, être utilisé aussi comme échelle auxiliaire avec son propre discriminateur de hauteur d'impulsion, le système de minuterie étant rapporté à la fréquence du secteur, quoiqu'il soit possible d'utiliser un étalon de fréquence extérieur. La sensibilité d'entrée est de 5 mV en sens négatif et 5 V en sens positif. Le gain d'amplification choisi par interrupteur est de 25, 50, 100, 250, 500 ou 1 000 et il y a une alimentation haute tension stabilisée de 250 à 2 000 V en deux gammes.

EE 22 757 *pour plus amples renseignements*

ENGLISH ELECTRIC CO LTD

Marconi House, Strand, London W.C.2

SYSTÈME D'INTERDISJONCTION TRANSISTORISÉ

(*Illustration à la page 377*)

Ce système d'interdisjonction entièrement transistorisé coûte environ la moitié du prix d'un matériel classique. Il fonctionne d'après le principe du glissement de fréquence et il a été prévu pour servir de lien de protection dans les réseaux de puissance.

L'appareil comporte deux circuits de protection et ne mesure que 50,8 cm × 30,5 cm × 30,5 cm. Il est donc d'un encombrement plus réduit que n'importe quel autre matériel du genre.

Les circuits logiques transistorisés garantissent un service sûr, une faible consommation de courant et un temps de fonctionnement rapide. Dans un secteur électrique normal, le temps mis pour relayer un signal afin de disjoindre un circuit n'est que de 55 m/sec et, même dans les conditions les moins favorables, il ne dépasse pas 85 m/sec.

Une application caractéristique est celle où un transformateur abaisseur est relié à l'extrémité éloignée d'une ligne de 33 kV.

En pareil cas, une panne sur le côté à tension basse du transformateur ne sera pas nécessairement decelée par le relais protecteur à l'extrémité d'alimentation de l'artère et, dans ces conditions, des difficultés se produisent dans la disjonction de l'interrupteur a l'extrémité d'alimentation. Ceci se vérifie en particulier lorsque le niveau de dérangement sur le côté à basse tension du transformateur est limité par une résistance de mise à la terre. On obvie à cet obstacle par l'emploi d'un appareil d'interdisjonction mis en œuvre sur le côté à tension basse du transformateur.

Chacun des organes terminaux de ce nouvel équipement, type SL54, a son propre émetteur et récepteur. Dans les conditions normales, un signal continu sur une fréquence donnée est transmis, à travers les lignes d'interconnexion des P.T.T., au récepteur à l'autre extrémité de la ligne. Lorsqu'un défaut se produit, cette fréquence se déplace et ce changement, détecté par le récepteur approprié, actionne le fonctionement d'un relais qui, à son tour, déclenchera l'action nécessaire.

En cas de dérangement dans le système d'interdisjonction, tout fonctionnement erroné du relais de déclenchement est empêché et un signal d'alarme est donné.

EE 22 758 *pour plus amples renseignements*

ERICSSON TELEPHONES LTD

Beeston, Nottingham

TACHYMÈTRE

(Illustration à la page 378)

Construit avec des sous-éléments de la susdite société, cet instrument emploie une fréquence de référence de 100 Hz pour des mesures à partir d'une limite inférieure, jusqu'à un maximum de 1200 000 tours/minute, qui ne dépendent que de l'efficacité du transducteur employé. La fréquence de 100 Hz peut être produite intérieurement ou extérieurement, selon le choix.

L'instrument comprend deux voies d'entrée pour des amplitudes d'impulsions de 0,1 V a 1 V ou de 1 V à 100 V. Les deux voies peuvent recevoir soit des impulsions sinusoïdales soit des impulsions à direction positive d'une durée de plus de 4 μsec.

La période de comptage et la durée d'affichage peurent être réglés de 1 sec à 10 sec, le comptage et les durées d'affichage étant complémentaires à un total de 11.

EE 22 759 *pour plus amples renseignements*

TRANSDUCTEUR ENTRAÎNÉ PAR ARBRE

(Illustration à la page 378)

Le Rotapulse est un transducteur photoélectrique entraîné par arbre, qui convertit la vitesse angulaire de l'arbre en un nombre connu d'impulsions électriques par tour.

La rotation de l'arbre produit un mouvement cyclique de la courbe d'interférence causée par la lumière brillant à travers des grilles d'émission, ce mouvement étant détecté par des phototransistors montés en paires diamétralement opposées. Chaque paire étant donc soumise à des signaux antiphase, les deux paires forment conjointement un dispositif différentiel produisant un nombre maximum d'impulsions. L'instrument courant produit 1000 impulsions par tour, mais des instruments avec d'autres maxima peuvent être fournis.

Ainsi qu'on a pu le voir au Salon, le Rotapulse est monté de telle façon que son arbre sort du centre d'un disque étalonné en 1000 unités. Une indication du nombre d'impulsions déclenchées par le Rotapulse, résultant de la rotation de l'aiguille fixée à l'arbre, est donnée par un compteur bidirectionnel construit avec des éléments de comptage transistorisés, Type 116A.

EE 22 760 *pour plus amples renseignements*

CELLULES DE CHARGE

(Illustration à la page 378)

Produites pour répondre à la demande accrue de dispositifs de pesage rapides, précis et robustes, les cellules de charge NOBEL sont prévues pour assurer une stabilité extrême contre les effets de température, une dérive à long terme de la sortie, et le fluage sans charge soutenue. Ces caractéristiques ne sont pas affectées par l'emploi dans des atmosphères corrosives et le boîtier de la cellule de charge est construit pour porter des charges latérales comparables avec la charge nominale maxima de la cellule.

Lorsque plusieurs cellules Nobel sont utilisées dans une installation de multipesage, un seul instrument peut être employé pour indiquer la sortie d'une ou de toutes les cellules. Un précision supérieure à $\pm 25\%$ est normale, et ce degré de précision est maintenu pendant de longues périodes. Cinq formats étalon de cellule, couvrant tous les poids jusqu'à 2268 kg, ont été réalisés et furent exposés comme présentation statique.

EE 22 761 *pour plus amples renseignements*

FERRANTI LTD

Hollinwood, Lancashire

MACHINE DE CONTRÔLE COORDONNÉ

(Illustration à la page 378)

Cette machine de contrôle coordonné facilite grandement le travail des services de contrôle car elle peut vérifier les proportions d'un objet jusqu'à un degré de précision de 0,001″ dans la moitié du temps requis par des moyens de contrôle classiques tels que les calibres à cadran des ingénieurs, etc.

Un palpeur de contrôle, avec embouts interchangeables et de forme apropriée, peut être placé à n'importe quel point d'une pièce usinée. Ce palpeur est tenu par une poutrelle en porte à faux rigide, surplombant la table de travail sur laquelle est serré l'article devant être contrôlé.

Des unités sensibles électro-optiques sont fixées aux axes x et y de la machine. Tout mouvement le long d'un des deux axes provoque la traversée de cellules photoélectriques par le diagramme de franges moirées produit par le réseau de diffraction connexe. Les changements d'intensité lumineuse sont alors convertis par les cellules en contre-parties électriques. Des compteurs de décades bidirectionnels à transistorsun pour chaque axe—enregistrent la variation cyclique.

Les données de mesures sont affichées de façon continue sur les deux tableaux électroniques à cinq chiffres, la position du palpeur, à n'importe quel point de sa course x et y, pouvant être lue directement jusqu'à une précision de 0,001″.

EE 22 762 *pour plus amples renseignements*

THE GENERAL ELECTRIC CO LTD

Magnet House, Kingsway, London W.C.2

(Illustration à la page 379)

MICROMÈTRE À LECTURE DÉCIMALE

Au cours d'une démonstration, on a pu voir comment des indicateurs peuvent être engrenés l'un dans l'autre pour obtenir une haute résolution. Quatre indicateurs décimaux à code binaire du format 11 furent employés, donnant un affichage numérique de la lecture micrométrique en dix-millièmes de pouces. Tous les circuits sont transistorisés. L'engrènement des indicateurs n'est pas critique car ils sont munis de bobinages supplémentaires disposés, à l'aide d'un circuit connexe, de façon à empêcher toute erreur. Ceci s'effectue par l'utilisation du chiffre le plus significatif de chaque paire d'indicateurs précis pour commuter l'excitation de l'indicateur approximatif d'un bobinage à l'autre.

Les éléments sont excités au moyen d'ondes carrées produites par un multivibrateur à transistors. Les sorties qui résultent des indicateurs sont amplifiées, leurs phases sont détectées, puis elles sont emmagasinées dans des circuits bistables et, enfin, décodées au moyen de relais actionnant l'affichage numérique.

Ainsi qu'il fut montré, chaque indicateur a un bloc à part de circuits de sortie qui lui est rattaché. Dans de nombreuses applications, on peut exciter les éléments en séquence et employer un seul bloc de circuits de sortie sur une base de partage de temps. Ceci peut conduire à une économie considérable d'appareillage, particulièrement dans des systèmes comprenant un nombre élevé d'indicateurs.

EE 22 763 *pour plus amples renseignements*

APPAREIL D'ENTRAÎNEMENT DE VIBREUR

Cet appareil a été étudié pour remplir toutes les fonctions nécessaires pour l'entraînement d'une vibreur VEM.5.2. ou tout vibreur similaire, d'une manière propre à l'emploi par le personnel d'une usine.

Le vibreur est actionné dans la gamme de fréquences de 30 Hz à 10 kHz par un amplificateur à faible distorsion, à sortie nominale de 2,5 kW. Le niveau d'accélération de la table du vibreur peut être contrôlé très étroitement par un dispositif de commande à réponse rapides, fonctionnant à partir de n'importe lequel des trois accéléromètres, avec indication indépendante de niveau sur un appareil de mesure à échelle logarithmique de 2 a 100 g.

On peut employer une source de signaux extérieure ou bien une des deux sources intérieures. L'oscillateur intérieur en dents de scie entraîné par moteur, est commandé entièrement à partir du panneau avant et il peut être balayé entre toutes les limites actuelles, dans la gamme de 30 Hz à 10 kHz, à des vitesse de balayage de 4 à 140 sec/octave. Une commande manuelle peut être employée pour régler n'importe quelle fréquence requise. Une fois réglée, cette fréquence peut être maintenue indéfiniment. Le magnétophone intérieur, à double piste et inversion automatique, fournit un réenregistrement continu de 30 minutes, répété indéfiniment avec une durée d'inversion de moins de cinq secondes.

EE 22 764 *pour plus amples renseignements*

GENERAL PRECISION SYSTEMS LTD

Aylesbury, Buckinghamshire

ACCESSOIRES DE CALCULATEURS ANALOGIQUES

La série d'accessoires de calculateurs analogiques pour courant continu comprend notamment:

Générateur de Fonctions

Ce générateur électronique de fonctions, du type à interrupteur à diode, est étudié pour recevoir une entrée maxima de ± 100 V c.c. Sa sortie atteint aussi ± 10 V c.c. et elle peut être réglée pour constituer n'importe quelle fonction voulue de la tension d'entrée. La fonction est approchée par une série de lignes droites. Ceci s'effectue en divisant la gamme en 10 sections par 9 "points de rupture" qui peuvent être fixés à n'importe quelle valeur de la tension d'entrée. Chacune des sections suit une loi linéaire dont la pente peut varier de zéro à 10 dans 1. Toutes les commandes de réglage sont presque indépendantes les unes des autres, à tel point qu'en ne réglant chacune qu'une seule fois, on obtient une précision globale de $\pm 2\%$. La largeur de bande est de 2 kHz. L'appareil est complet avec ses propres amplificateurs intérieurs.

Des fonctions peuvent également être produites par l'emploi des cartes de résistances a prises dont munies les éléments servo-multiplicateurs.

Servo-Multiplicateur

Ce servo-multiplicateur à vitesse réduite se compose d'un convertisseur vélodyne actionnant trois potentiomètres. L'un de ceux-ci est employé pour la commande du moteur qu'un interrupteur permet de débrancher lorsqu'on veut l'intégration. Il y a en outre une boîte de vitesses standard à deux vitesses, à laquelle on peut ajouter d'autres boîtes de vitesses, donnant une variation possible de la gamme d'intégration de ½ cycle par seconde à 1 cycle par huit heures.

Les deux autres potentiomètres sont à 9 prises, de sorte que diverses fonctions peuvent être approchées par charge résistive appropriée. Deux potentiomètres supplémentaires peuvent être ajoutés.

Il existe aussi un servo-multiplicateur à action rapide, utilisant des potentiomètres rotatifs actionnés par servomoteur MES. Une vitesse de 10 Hz sur la totalité de l'échelle peut être atteinte et soit trois soit six potentiomètres peuvent être fixés. Dans chaque cas, l'un des potentiomètres sert à la commande du moteur, un autre est à 9 prises et le reste est à prise centrale.

EE 22 765 *pour plus amples renseignements*

LIVINGSTON LABORATORIES LTD

Retcar Street, London N.19

OSCILLOSCOPE D'ÉCHANTILLONNAGE

(Illustration à la page 379)

L'oscilloscope Hewlett-Packard Type 185A, allié à l'amplificateur à double trace Type 187A, forment un excellent système pour l'observation de phénomènes à impulsions rapides, exigeant des largeurs de bande de 500 MHz. Des échantillons successifs de l'impulsion soumise à l'essai sont prélevés à intervalles toujours plus étendus. Les echantillons sont ensuite reconstitués sur une longue base de temps, donnant un système avec un temps de montée effectif de 0,7 mμsec.

Le nombre d'échantillons à prélever par trace peut être choisi par une commande sur le panneau avant. On peut aussi examiner en détail les durées d'étalonnage et d'affaiblissement des impulsions grâce aux dispositif d'étalement de balayage et de déclenchement multiple dont est pourvu l'oscilloscope.

En outre, on peut effectuer des mesures précises de temps de montée, de durée et d'espacement, par l'emploi du second canal pour un affichage simultané de la sortie de l'étalonneur de temps. Un sortie est également prévue pour actionner un appareil de traçage des courbes X-Y, ce qui permet d'obtenir un enregistrement permanent de phénomènes à grande vitesse.

Chaque canal a une sensibilité indépendante avec des gammes étalonnées de 10 mV/cm à 200 mV/cm, d'une précision de $\pm 5\%$, et une impédance d'entrée d'environ 100 k. ohms shuntée par 2 pF.

EE 22 766 *pour plus amples renseignements*

CAMÉRA D'OSCILLOSCOPE

La caméra d'oscilloscope Hewlett-Packard Type 196A avec dos Polaroid Land, permet d'obtenir un enregistrement photographique positif d'une trace d'oscilloscope, dans l'espace d'une minute. Grâce au capot de vision, on peut observer la trace durant la pose, et la lentille spéciale f/1.9 donne une image de la largeur totale d'un micromètre, soit 10 cm, sans dégradation appréciable de la résolution cathodique. Le rapport de l'objet à la grandeur de l'image est de 1 à 0,9.

Les problèmes de fuite de lumière ont été éliminés par un soufflet de caméra normal et l'on peut déplacer la lentille sur 11 positions de détente, afin de faciliter le travail des poses multiples.

Quoiqu'essentiellement étudié pour être fixé à des oscilloscopes de grande puissance, un adapteur peut être fourni avec la caméra pour pouvoir la fixer à des oscilloscopes TEKTRONIX.

EE 22 767 *pour plus amples renseignements*

MARCONI INSTRUMENTS LTD

St. Albans, Hertfordshire

COMPTEUR

(Illustration à la page 379)

Cet instrument (Type TF 1292) indique le taux de comptage ou le comptage intégré de détecteurs de particules et de radiations nucléaires. Des gammes de taux de comptage de 30, 100, 300 et 1000 par seconde, sur la totalité de l'échelle, sont prévues. Avec ces dernières, des constantes de temps de 1,5 ou 20 secondes peuvent être choisies. Les gammes de taux de comptage intégré sont de 1000, 3000, 10 000 et 30 000 coups sur la totalité de l'échelle.

L'instrument est muni de deux sorties réglables très haute tension, l'une pour un compteur de particules ou de radiations et l'autre pour une chambre de précipitation. Il existe aussi une alimentation haute et basse tension pour un préamplificateur extérieur.

Le compteur Type TF 1292 convient à de nombreuses applications de mesure de radiation dans les processus industriels et les laboratoires. On peut l'employer, entre autres, dans un système de contrôle de gaz réfrigérant dans une centrale nucléaire pour la détection d'éléments de combustible défectueux. Utilisé à cette fin, il mesure la sortie d'un compteur à scintillations réagissant aux particules radioactives dans le gaz. Les particules, étant ionisées, sont précipitées sur un câble à très haute tension d'attraction fournie par le compteur.

Le compteur Type TF 1292 peut être télécommandé et lu à distance. En outre, un circuit avertisseur peut être mis en œuvre selon n'importe quelle indication prédéterminée ou en cas de panne de courant. En plus de l'affichage sur panneau du taux de comptage, il y a deux sorties proportionnelles pour usage extérieur donnant 50 V et 100 mV sur la totalité de l'échelle. La sortie inférieure est particulièrement prévue pour un enregistreur.

EE 22 768 *pour plus amples renseignements*

CONVERTISSEUR ANALOGIQUE-NUMÉRIQUE

(Illustration à la page 380)

Le convertisseur analogique-numérique TF 1325 est un ordinateur entièrement transistorisé, calculé pour fournir à un appareil indicateur, enregistreur ou de dépouillement de données, la valeur d'un potentiel de courant continu appliqué de 0 à 0,999 V. La sortie décimale à code binaire de trois chiffres donne une précision de ± 1 mV. La période d'échantillonnage est de 20 msec. et elle peut être mise en train à distance. Un signal à plots averrit lorsque l'échantillonnage a été complété et qu'un résultat peut être livré; le courant d'entrée est alors au-dessous de 0,15 μA.

L'indicateur a été conçu principalement pour recevoir une tension en direct, en tant qu'entrée, dont il échantillonne le niveau, puis fournit le résultat, rapidement et avec précision, sous forme numérique. Ce dernier est appliqué à un décodeur et imprimeur linéaire ou à une machine a écrire électrique, permettant ainsi d'obtenir un enregistrement auquel on peut se référer par la suite. Utilisé de cette façon, l'indicateur trouve une application particulière dans le contrôle répété du niveau de radiation à un certain nombre de points d'une centrale nucléaire. Les tensions parvenant de points devant être contrôlés sont appliquées à leur tour à l'indicateur et elles n'ont besoin d'être stables que pendant chaque période d'échantillonnage.

La grande vitesse dont l'indicateur est capable le rend apte à servir de convertisseur, traduisant une tension analogique sous une forme numérique pour être appliquée à un équipement de dépouillement de données.

L'indicateur est normalement livré dans un coffret pour fonctionnement sur banc d'essai, mais il peut aussi être fourni pour montage sur châssis.

EE 22 769 *pour plus amples renseignements*

MULLARD LTD

Mullard House, Torrington Place, London W.C.1

TRANSISTORS À JONCTION PAR DIFFUSION EN ALLIAGE

La technique de jonction par diffusion a déjà été appliquée avec succès pour la production sur une grande échelle des transistors OC170 et OC171, qui sont largement utilisés pour l'amplification m.f. et h.f. dans les récepteurs pour fréquence élevée et hyperfréquences. Un nombre de transistors a maintenant été mis au point pour l'emploi dans les télécommunications et dans le matériel industriel et de calcul électronique. Ces transistors comprennent en particulier:

Transistors pour communications hyperfréquences

Le transistor pour hyperfréquences type AZ11 a été réalisé spécifiquement comme transistor à faible bruit pour l'amplification à 100 MHz dans le matériel de communications professionnelles. La technique particulière de construction utilisée pour ce transistor permet de garder la résistance de base extrinsèque et la capacité réactive à de très basses valeurs, de sorte qu'à 100 MHz, le transistor donnera un gain de puissance supérieur à 10 dB. Le niveau de bruit à la même fréquence, ne dépasse pas 6dB. Le gain de courant moyen est de 50.

La tension maxima de base collecteur est de—20 V et la tension au collecteur de l'émetteur est de—12 V, à un régime de plein courant de 10 mA. Le courant de fuite au collecteur est de 4 μA (moyenne) avec une tension de base collectrice de—6 V et à une température de 25°C. La température de jonction maxima est de 75°C et la résistance à chaud est de 0,6°C/mW, permettant une dissipation de 50mW dans une température ambiante de 45°C. Le transistor est logé dans un boîtier métallique.

Le transistor Type AFZ11 sera bientôt suivi d'un autre transistor conçu pour les mêmes applications professionnelles, mais à une fréquence double. Aux étages de 200 mHz du matériel de communication hyper-fréquences, le nouveau transistor fournira un gain de puissance supérieur à 10 dB, avec un niveau de bruit inférieur à 9 dB.

Calcul logique ultra-rapide

Un transistor à jonction par diffusion en alliage, avec de très bonnes caractéristiques de grand signal sera bientôt offert pour les circuits logiques de calculateurs à action ultra-rapide.

Ce transistor constitue un élément des plus perfectionnés, conçu pour répondre aux besoins des calculateurs futurs ayant des taux d'impulsions d'horloge atteignant 10 MHz. Il aura une fréquence de coupure de plusieurs centaines de mégacycles par seconde, d'excellentes caractéristiques de saturation pour un courant au collecteur de 50mA, et un taux de dissipation suffisant pour ses applications.

Ce nouveau transistor a pu être vu en action dans des circuits logiques de commutation de tension à niveau bas et dans un compteur de décades de 10 MHz.

Commande de systèmes à mémoire

Ce transistor, qui est en cours de réalisation, pourra commander des systèmes à mémoire en ferrite à action rapide, de concert avec le nouveau transistor logique à action rapide dont mention a été faite plus haut.

Il pourra fournir des impulsions de sortie de plusieurs centaines de milliampères avec des temps de montée extrêmement rapides. Son régime de courant au collecteur maximum est de $\frac{1}{4}$ A et la valeur de son paramètre de fréquence minima est supérieur à 60 MHz.

On a pu le voir utilisé, au cours d'une démonstration, dans un générateur d'impulsions à courant élevé, fournissant des impulsions de sortie d'une demi ampère avec des temps de montée moyens inférieurs à 40 mμsecs.

Ses caractéristiques de grande puissance à haute fréquence le rendent également indiqué pour la téléphonie par courants porteurs et le matériel des transmissions. Pour ces applications, il fournit une sortie de 3 W à 8 MHz ou de 1 W aux fréquences supérieures. Une démonstration a permis de le voir en usage aux étages de commande et de sortie d'un émetteur de 2,5 W fonctionnant à 8 MHz.

Transistor de commutation PNPN

Dans le semi-conducteur Type ATZ10, la technique à jonction par diffusion et alliage a été appliquée à la réalisation d'un transistor de commutation pnpn à 4 couches.

Ses propriétés permettent la construction de circuits monostables et astables univibrateurs, en n'utilisant qu'un seul transistor pnpn au lieu de deux transistors conventionnels à 3 couches normalement requis, ce qui a pour résultat une économie d'espace et de pièces connexes. De plus, il peut aussi être employé comme voie de parole commutée dans les systèmes téléphoniques automatiques.

Le ATZ10 est un transistor à trois bornes et son effet est similaire a celui d'une combinaison de transistor pnp et npn avec rétablissement intérieur. Cette construction résulte en une caractéristique d'impédance négative prononcée, dans une gamme d'une largeur fort utile, et permet le fonctionnement par impulsions de ce transistor sous forme de commutateur électronique à action rapide. Une fois branché il demeure automatiquement 'à fond' jusqu'à ce qu'il soit 'débranché' par l'impulsion suivante.

Le transistor ATZ10 a un rapport d'impédance arrêt-marche supérieur à 3 millions: 1. Ses caractéristiques nominales de tension maxima et de courant de pointe sont d'environ 35 V et 25 mA respectivement. La température de jonction maxima est de 55°C, ce qui autorise une dissipation globale maxima de 17 mW dans une température ambiante de 45°C. Un compteur annulaire de décades, utilisant dix transistors ATZ10 et pouvant compter à des vitesses de 30 kHz, a été montré au Salon.

Transistor à avalanche avec temps de montée de 1 mμsec

Cet élément n'est pas mis en œuvre par l'action normale du transistor mais par le phénomène dit "d'avalanche." Il représente une réalisation spéciale de la technique de diffusion, des plus utiles pour les applications nécessitant des temps de montée exceptionnellement accélérés comme, par exemple, pour les oscilloscopes d'échantillonnage à action rapide.

Un courant de crête de 50 kA peut être atteint avec un temps de montée de 1 mμsec seulement, ce qui représente un taux de montée de courant de 50 A/μsec.

Il a été montré au Salon dans un circuit à avalanche produisant des impulsions de 50 mA avec des temps de montée d'environ 1 mμsec. à une fréquence d'impulsions de 4 MHz. Ces impulsions furent affichées sur un oscilloscope d'échantillonnage, utilisant lui même un transistor à avalanche dans le circuit d'échantillonnage.

Transistor de commutation

Le transistor ASZ20, qui est déjà en cours de production, a été prévu pour les applications de commutation, y compris les circuits logiques de calculateurs.

Il a une valeur nominale de courant de pointe de 25 mA et une tension de base collectrice maxima de −40 V. La valeur de fréquence minima est supérieure à 40 MHz et le gain de courant à grand signal minimum est de 33 à 25 mA de courant au collecteur. La temperature de jonction maxima est de 75°C et la résistance à chaud est de 0,6°C/mW, autorisant une dissipation de 50 mW dans une température ambiante de 45°C. Le courant de fuite au collecteur à −6 V est inférieur à 5 mA pour une température de 25°C.

EE 22 770 *pour plus amples renseignements*

Résumés des Principaux Articles

Un sondeur à ultrasons transistorisé miniature par R. N. B. Gatehouse

Résumé de l'article aux pages 336 à 340

Cet article décrit la mise au point et les principales caractéristiques de construction d'un très petit et fort robuste sondeur ultrasonore, prévu pour des profondeurs de 0,60m à 58,5m. Cet appareil convient à des applications telles que les levés hydrographiques et les sondages pour la pose de tuyaux sous-marins, ainsi que pour le pilotage des yachts et des petits bâteaux de pêche. Le poids de l'appareil complet n'est que de 5,5 kg, sa consommation électrique est de 0,1W et sa précision est de ±3% ou 15cm, selon le degré de précision le plus élevé.

Un circuit façonneur de formes d'ondes passives par M. J. Wright

Résumé de l'article aux pages 341 à 343

L'auteur décrit un circuit simple pouvant exécuter une variété de fonctions de façonnage de formes d'ondes. Quoique n'utilisant aucune pièce active, ce circuit peut fournir de vives transitions de tension de sortie, retardées dans le temps par rapport à la tension d'entrée. Ses applications sont, entre autres, l'allongement de la durée d'impulsions, le changement de phase de 90° d'ondes carrées, le filtrage et l'écrêtage combinés, la modulation marque/espace d'une entrée d'onde carrées.

Affichage multidimensionnel simple par D. M. Mackay

Résumé de l'article aux pages 344 à 347

Des fonctions de deux variables ou davantage peuvent être affichées sur un écran cathodique sous forme de projection bi-dimensionnelle de l'hypersurface qu'elles définissent. L'angle de projection peut être modifié par des potentiomètres tournants qui déterminent les proportions dans lesquelles les différentes tensions d'entrée contribuent aux déviations X et Y.

Quoique le meilleur processus d'affichage soit de se servir de potentiomètres à quatre curseurs et d'amplificateurs d'accord, un résultat remarquablement bon peut être obtenu à l'aide de circuits beaucoup plus simples, ne nécessitant que des potentiomètres linéaires à double jumelage et quelques résistances choisies avec soin. Les sorties pour un affichage entièrement stéréoscopique peuvent également être obtenues par un circuit à peine un peu plus compliqué.

La simulation des fonctions de transfert dans un calculateur analogique par F. C. Harbert

Résumé de l'article aux pages 354 à 355

Cet article décrit un circuit à amplificateur unique qui exécute l'opération de double intégration ainsi qu'une synthétisation des fonctions de transfert polynômes dans un calculateur analogique.

La construction d'instruments numériques optiques par I. R. Young

Résumé de l'article aux pages 359 à 365

L'article examine d'abord diverses considérations touchant à la fabrication d'instruments optiques, en soulignant l'importance d'une lecture immédiate en cas de besoin, ce qui est particulièrement souhaitable pour les systèmes diviseurs de temps. Quelques instruments à faible résolution sont ensuite décrits, avec des indications sur l'utilisation des données obtenues. Une description succincte des problèmes se rapportant à la construction d'instruments à très haute résolution conclut cet article.

Les effets de l'enrobement sur les pièces électroniques par F. P. Campbell

Résumé de l'article aux pages 366 à 371

On sait depuis longtemps que l'enrobement des circuits électroniques affecte leur rendement. L'auteur étudie dans cet article les effets de l'enrobement sur des pièces détachées d'usage courant.

Les tubes peuvent être enrobés en toute sécurité, à condition de suivre la technique correcte pour supprimer la chaleur autoproduite et pour protéger l'enveloppe de verre. On rencontre peu de difficultés dans le cas des transistors et des capacités. Par contre, les résistances bobinées se prêtent mal aux revêtements de résines liquides lorsqu'elles sont à forte dissipation. L'auteur décrit aussi les effets de l'enrobement sur les transformateurs et les bobines d'arrêt.

L'enrobement des résistances au carbone et au carbone crevassé affecte de manière différente certains de ces éléments. Ceci est illustré par la description d'une expérience. L'auteur résume, enfin, les causes des changements dans les composantes.

Une alimentation stabilisée et commandée par programme à réaction par M. Pacák

Résumé de l'article aux pages 372 à 373

Une alimentation stabilisée pour aimant de spectographe de masse est décrite dans cet article. La sortie est soit à variation continue de 3 à 800V, soit commandée par programme au moyen d'un dispositif résistance-capacité. L'amplificateur du stabilisateur est à réaction.

Die Ausstellung für INSTRUMENTE ELEKTRONIK UND AUTOMATION

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 376 bis 380

ALMA COMPONENTS LTD

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DRAHTWIDERSTÄNDE

Eine Anzahl neuer Drahtwiderstände wurde ausgestellt, unter denen sich folgende Typen befanden:

Type Y

ist eine neue Widerstandsreihe auf amerikanischen Normspulen, also austauschbar. Bei $\frac{1}{4}$, $\frac{1}{3}$ und $\frac{1}{2}$ W Nennlast und 0,1% Toleranz haben die mit hoher Präzision hergestellten Widerstände einen garantierten Temperaturbeiwert unter 0,002%/°C.

Type A4

Diese Widerstände sind die ersten der A-Serie, hermetisch verschlossen und tropenfest, deren Musterzulassung nach Vorschrift DEF. 5113 erwartet wird. Sie sind von 100Ω...400 kΩ mit 0,1% für einen Temperaturbereich von −55°C ... +120°C lieferbar.

Type C

ist ein Widerstand mit Verbundwicklung auf Alma-Standardspulen Type S mit einem garantierten Temperaturbeiwert von ±4 Teile pro Million pro °C über den Bereich von 0...100°C. Diese Widerstände sind für Nennlasten von 1 W (Type C1), $\frac{1}{2}$ W (Type C2) und $\frac{1}{4}$ W (Type C4); Toleranzen ±0,1%.

EE 22751 *für weitere Einzelheiten*

ASSOCIATED ELECTRICAL INDUSTRIES (Woolwich) LTD

155 Charing Cross Road, London W.C.2

TRANSISTOREN

Unter den kürzlich für industrielle Zwecke eingeführten Elementen befinden sich vier Leistungstransistoren. Typen XC 155 und XC 156 haben einen Spitzenstromwert von 10 A bei 80 bzw. 100 V Kollektorbasisspitzenspannung. Typen XC 141 und XC 142 sind für 3 A Spitzenstrom bei 40 bzw. 60 V Kollektorbasisspannung.

Für Schnellschaltvorgänge wurde eine Serie von drei Drift-Transistoren herausgebracht. Für diese Transistoren (Typen XA 141, XA 142 und XA 143) ist die Mindestfrequenz f_1, bei der die Stromverstärkung eins wird, 20 MHz, 40 MHz bzw. 60 MHz. Ausgestellt wurden auch zwei Mesa-Schalttransistoren mit Minimumfrequenzen f_1 = 25 MHz bzw. 35 MHz (Typen XA 161 und XA 162), die einen niedrigen Sättigungswiderstand, eine absolute Kollektor-Höchstverlustleistung von 75 mW bei 55°C und eine absolute Kollektorbasis-Höchstspannung von 13 V bieten.

Für Hochspannungsschalten bei geringeren Geschwindigkeiten ist Type XB 121 mit einer Kollektorbasisspannung von 105 V sehr interessant.

Auch drei Ge-npn-Legierungselemente für zusätzliche Schaltkreise wurden gezeigt und zwar Typen XA 701, 702 und 703 mit Mindestfrequenzen f_1 = 3 MHz, 5 MHz bzw. 10 MHz und einer Kollektorverlustleistung von 70 mV bei 55°C.

Zum ersten Mal zeigt die Firma Silizium-Transistoren in Form einer Entwicklungsserie von npn Diffusions-Flächentransistoren für Verlustleistungen in der Grössenordnung von 30 V bei 100°C.

Drift-Transistoren für HF-Verstärkung bis zu 100 MHz, sowie die konventionellen und gut eingeführten pnp Legierungs-Transistoren für NF- und HF-Verwendung wurden z.T. in wahlweisen Verkapselungen ausgestellt, die sich besonders für gedruckte Schaltungen eignen.

Die meisten der neuen industriellen Elemente für Verwendung in gedruckten Schaltungen werden in JEDEC Gehäusen geliefert.

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AVO LTD

Avocet House, 92-96 Vauxhall Bridge Road, London S.W.1

RÖHREN-KENNDATENMESSER

(Abbildung Seite 376)

Der verbesserte Röhren-Kenndatenmesser wurde vor kurzem eingeführt.

Alle Grundzüge früherer Typen wurden beibehalten und die folgenden zusätzlich eingebaut:

(a) Der negative Gitterpegel kann, besonders für niedrige Werte, viel genauer eingestellt werden.

(b) Eine genauere und kürzere Methode zur Bestimmung des Gitterstroms, besonders wenn die zu prüfenden Röhren Umkehr- oder Primäremission haben.

(c) Zusätzliche Anodenspannungen zur Prüfung der neuen Röhren mit niedrigen Anodenspannungen.

(d) Verbesserte Regulierung der Anoden- und Schirmgitterspannungen, um den Einfluss der Rückwirkungsimpedanz herabzusetzen.

(e) Überlastungsschutz für die Anzeigegeräte.

Anodenspannungen sind zwischen 12,6 V und 300 V in Stufen einstellbar. Die Gitterspannung kann von 0-100 V stufenlos eingeregelt werden, Anoden- sowie Schirmgitterstrom kann von 2,5 mA Vollausschlag bis 100 mA Vollausschlag gemessen werden; Gitterstrom 0...100 µA; Heizfadenspannungen von 0,625...117 V; Steilheit von 0,1...60 mA/V; Anzeigevorrichtungen für "Gut/Ausschuss" sind auch im Gerät vorgesehen.

Dioden und Gleichrichter können von 1...180 mA unter Ladebelastung geprüft werden.

Jedem Gerät liegt ein umfassendes Daten-Handbuch bei.

EE 22 753 *für weitere Einzelheiten*

B. & K. LABORATORIES LTD

4 Tilney Street, Park Lane, London W.1

ENDLOS-TONBANDKASSETTE

Das Modell soll die bisher kleinste Kassettenkonstruktion sein, die Endlos-Bänder bis zu 36,5 m Länge aufnehmen kann. Durch die Kassette können Labor-Aufnahme-/Wiedergabegeräte mit normalen Spulen auf einfache Weise in Endlos-Tonbandgeräte umgewandelt werden. Bis zu 36,5 m lange und 6,35 mm, 12,7 mm oder 25,4 mm breite Tonbänder können bequem über die in die Kassette eingebauten Rollen eingelegt werden.

Der Bandweg ist äusserst einfach und ermöglicht schnelle Bandeinlage. Lange Schleifen können schnell gegen einander ausgetauscht und die Standard-Bandgeschwindigkeiten von 47,5 cm/s...52,5 cm/s augenblicklich eingestellt werden.

Die Kassette kann abgenommen und beim P.S.200 Tonbandgerät durch eine andere ersetzt werden, in der das Band von einer normalen 227 mm Spule zur anderen läuft.

Mit der neuen Endlos-Kassette kann das P.S.200 Aufnahme-/Wiedergabegerät für Verzögerungssysteme und Spektralanalyse grosser Datenproben eingesetzt werden, wobei bis zu 14 Informationskanäle Verwendung finden können. Für solche Zwecke waren früher grosse Gestellanlagen mit schwerfällig gefederten Leitrollen und komplex durchflochtenen Bandwegen erforderlich. Bei der neuen Kassette kann der Benutzer entweder mit einem normalen Gerät oder einem Endlos-Bandgerät beginnen und dann den Anwendungsbereich ausweiten.

EE 22 754 *für weitere Einzelheiten*

TINTENSCHREIBER

Der Oscillomink 230B ist ein Tintenschreiber, der Vorgänge bis zu 1 kHz Empfindlichkeiten von 0,2 mm/mV ... 4 mm/100 V registrieren kann. Dieses Direktschreibsystem arbeitet nach einem neuen Prinzip, bei dem ein haardünner Strahl Schreibflüssigkeit mit hoher Geschwindigkeit aus einer am beweglichen Organ des Messwerkes befestigten Düse austritt. Da der freie Strahl nicht in das Systemträgheitsmoment eingeht, kann eine hohe Resonanzfrequenz von ungefähr 650 Hz erreicht werden. Die Nullinie kann auf dem Papier in jede gewünschte Lage verschoben werden. Bei Mehrkanalaufzeichnungen können sich die Spuren ohne Beeinflussung kreuzen oder in Bezug auf eine gemeinsame Nullinie aufgezeichnet werden. Die Stärke der einzelnen Spuren kann unabhängig geregelt werden. Die Papiervorschübe von 5, 10, 20, 50, 100 und 200 mm/s können während des Betriebs umgeschaltet werden. Die Genauigkeit ist besser als 1%.

Durch Zusatzgeräte können eigen- oder ferngesteuerte Papiervorschübe von 0,5, 1 oder 2 m/s erreicht werden.

Das Gerät ist mit Zwillings-Gleichspannungs-Gegentaktverstärkern mit Grob- und Feineinstellung der Empfindlichkeit, regelbarer Phasen- und Amplitudenkorrektur, guter Unterscheidung symmetrischer und unsymmetrischer Signale und einer trägerfrequenzmodulierten Endstufe mit Ausweitungskopplung ausgerüstet. Ausserdem sind 8- und 12-kanälige Schreiber und Zwillingsverstärker lieferbar.

EE 22 755 *für weitere Einzelheiten*

R. H. COLE (OVERSEAS) LTD

2 Caxton Street, London S.W.1

TRÄGERFREQUENZ-PEGELMESSUNG

(*Abbildung Seite 377*)

Das von Siemens und Halske hergestellte Gerät besteht aus einem Pegeloszillator und einem selektiven Pegelmesser. Pegel-, Dämpfungs- und Verstärkungsmessungen können für Frequenzen von 30 kHz ... 15 MHz ausgeführt werden. In diesem Spektrum liegen z.B. die Mehrkanal-Trägerfrequenztelefoniesysteme mit koaxialem Kabel oder Funkbrücken der CCITT Grundgruppen (60 ... 108 kHz, 12 Sprechkanäle), CCITT Übergruppen (312 ... 552 kHz, 60 Sprechkanäle), CCITT Hauptgruppen (812 ... 2044 kHz, 300 Sprechkanäle) und CCITT Überhauptgruppen (8516 ... 12 388 MHz, 900 Sprechkanäle).

Ein besonderer Vorteil dieses Gerätes ist die hohe Genauigkeit der Geberfrequenz und der Detektorabstimmung, die durch ein quarzgesteuertes elektronisches Frequenzmitnahmesystem erreicht wird. Der gesamte Frequenzbereich wird mit einem Durchstimmen der Skala bestrichen, wobei eine Anzeigelampe aufleuchtet, wenn die Frequenz mit einer der 100 kHz Skalenteilungen entsprechenden Frequenz übereinstimmt. Dadurch erfolgt die Einstellung der Frequenz wesentlich schneller. Frequenzwerte zwischen den Teilstrichen können überall im Gesamtbereich mittels einer weiteren Zusatzabstimmung mit ungefähr 200 Hz Genauigkeit eingestellt werden.

Durch die automatische Abstimmvorrichtung für Haupt- und Zusatzabstimmung des Pegelmessers können alle Messungen genau so einfach und schnell vom Geber aus ausgeführt werden wie Breitbandmessungen, soweit Oszillator und Detektor dabei nicht voneinander getrennt benutzt werden sollen.

Grosser Messbereich und hohe Selektivität ermöglichen in Zusammenwirkung mit hoher Spiegelfrequenzsicherheit und geringem Restklirrfaktor Messung des Nebensprechens und der Klirrdämpfung; die elektronische Frequenzanzeige erleichtert die Messungen noch mehr. Die Zeitstabilität der Einstellung ist für Überwachungszwecke wertvoll.

Wo geringere Anforderungen an Frequenzgenauigkeit gestellt werden, können Geräte ohne elektronische Frequenzmitnahme mit Neper- oder Dezibeleichung geliefert werden.

Der Pegel-Oszillator kann von einer Fremdquelle mit Frequenzen von 0 .... 100 kHz amplitudenmoduliert werden.

Da der Pegelmesser auf Effektivwerte anspricht, kann er auch für Rauschmessungen, z.B. zusammen mit Rauschgenerator Type Rel 3 W 432 Verwendung finden.

EE 22 756 *für weitere Einzelheiten*

EKCO ELECTRONICS LTD

Southend-on-Sea, Essex

AUTOMATISCHER UNTERSETZER

(*Abbildung Seite 377*)

Das neue Ekco Gerät Type N610 ist ein zusammengebauter automatischer Untersetzer und Zeitmesser mit stabilisierter Hochspannung, einem Leistungsverstärker und Impulshöhenanalysator. Ein Ratemeter mit Drucktastenwahl der 1000, 10 000 und 100 000 Impuls/Sek.-Bereiche ist auch eingebaut.

Der Untersetzer kann automatisch bei vorbestimmten Zählstössen bis zu 1 000 000 und vorbestimmten Zeiten bis zu 100 000 Sek. angehalten werden. Der Zeitmesser kann auch als zweiter Untersetzer mit eigenem Impulshöhendiskriminator benutzt werden, dessen Zeitmesser auf die Netzfrequenz bezogen ist; jedoch kann auch ein Fremdfrequenznormal eingesetzt werden. Die Eingangsempfindlichkeit ist 5 mV negativgehend und 5 V positivgehend. Durch Schalter können Verstärkungen von 25, 50, 100, 250, oder 1 000 eingestellt werden. Ein stabilisiertes Netzgerät gibt 250...2 000 V in zwei Bereichen.

EE 22 757 *für weitere Einzelheiten*

ENGLISH ELECTRIC CO. LTD

Marconi House, Strand, London W.C.2

TRANSISTORISIERTES ZWISCHENAUSLÖSUNGSGERÄT

(*Abbildung Seite 377*)

Das volltransistorisierte, mit Frequenzänderung arbeitende Zwischenauslösungsgerät kostet ungefähr die Hälfte der üblichen Geräte und wurde als Schutzverbindung für Kraftnetze entwickelt.

Das zwei Schutzkreise enthaltende Gerät ist mit 508 × 305 × 305 mm kleiner als jedes bisherige Gerät.

Die transistorisierten logischen Schaltungen sind betriebssicher, haben geringen Stromverbrauch und kurze Ansprechzeit. In einer normalen Kraftleitung ist die Übertragungszeit für das Auslösesignal unter 55 ms und unter den schlechtesten Bedingungen 85 ms.

Als typisches Verwendungsbeispiel kann ein Abwärtstransformator gegeben werden, der mit dem Ende einer 33 kV-Leitung verbunden ist.

Ein auf der Unterspannungsseite des Transformators auftretender Fehler wird nicht unbedingt von der Schütze am Anfang des Speisekabels entdeckt. Unter diesen Umständen treten Probleme im Zusammenhang mit der Auslösung des Trennschalters am Speisepunkt auf. Das trifft besonders dort zu, wo der unterspannungsseitige Fehlerpegel durch einen Erdungswiderstand begrenzt ist. Diese Schwierigkeit wird durch das Zwischenauslösungsgerät überwunden, das von der Unterspannungsseite des Transformators betätigt wird.

Jeder Endsatz des neuen Gerätes Type SL54 hat seinen eigenen Sender und Empfänger. Wenn alles in Ordnung ist, wird über verbindende Fernsprechleitungen ein Dauerton bestimmter Frequenz zum Empfänger am anderen Ende der Leitung übermittelt. Wenn ein Fehler auftritt, wird diese Frequenz verschoben, und die vom entsprechenden Empfänger entdeckte Änderung betätigt ein Relais, das die erforderliche Tätigkeit einleitet.

Störungen im Zwischenauslösesystem werden angezeigt und falsche Betätigung der Schütze verhindert.

EE 22 758 *für weitere Einzelheiten*

ERICSSON TELEPHONES LTD

Beeston, Nottingham

Drehzahlmesser

(Abbildung Seite 378)

Das aus ETL-Bausteinen zusammengestellte Gerät benutzt eine 100 Hz-Bezugsfrequenz zur Messung von einer unteren Grenze, die nur von der Wirksamkeit des Gebers abhängt, bis zu 1 200 000 UPM. Die 100 Hz können von einer eingebauten oder Fremdquelle kommen.

Zwei Eingangskanäle sind für Impulsamplituden von 0,1 V ... 1 V oder 1 V ... 100 V bestimmt. Beide Kanäle akzeptieren sinusförmige sowie positivgehende Impulse von mindestens 4 μs Dauer.

Sowohl Zählintervall als auch Anzeigezeit können zwischen 1 und 10 s eingestellt werden, jedoch müssen sie zusammen eine Gesamtzeit von 11 s ergeben.

EE 22 759 *für weitere Einzelheiten*

Geber Mit Schaftantrieb

(Abbildung Seite 378)

Der „Rotapulse" ist ein photoelektrischer Geber mit Schaftantrieb, der die Winkelgeschwindigkeit seines Schaftes in eine bekannte Anzahl elektrischer Impulse pro Umdrehung umwandelt.

Durch die Rotation des Schaftes wiederholen sich die durch Durchleuchtung des Übertragungsgitters hervorgerufenen Interferenzbilder periodisch; diese Wiederholung wird von einander gegenüberliegenden, gegensinnigen Phototransistorpaaren wahrgenommen. Da also jedes Paar Signalen in Gegenphase ausgesetzt ist, bilden beide zusammen ein Differenzsystem, das die höchste Impulszahl hervorruft. Das Gerät in Standardausführung gibt 1000 Impulse pro Umdrehung ab; Ausführungen mit anderen Höchstwerten sind jedoch auch lieferbar.

In der gezeigten Form durchdringt der Schaft des Rotapulse eine in 1000 Einheiten geeichte Scheibe. Drehen des auf dem Schaft befestigten Zeigers ruft eine Anzahl vom Rotapulse abgegebener Impulse hervor, die auf einem aus transistorisierten Zählerbausteinen Type 116A gebildeten, in zwei Richtungen registrierenden Zähler angezeigt werden.

EE 22 760 *für weitere Einzelheiten*

Lastgeber

(Abbildung Seite 378)

Die Nobel Lastzelle wird zur Befriedigung des wachsenden Bedarfs für eine schnelle, genaue und robuste Waage hergestellt und wurde für höchste Stabilität gegen Temperatureinflüsse, Drift der Ausgangsspannung innerhalb langer Zeitspannen und Änderungen unter Dauerbelastung entwickelt. Diese Kennzeichen bleiben auch in korrodierender Atmosphäre erhalten. Das Gehäuse des Lastgebers ist so konstruiert, dass es Seitendrücke aushalten kann, die mit der Nennlast der Zelle vergleichbar sind.

Wenn verschiedene Nobel-Zellen in e i n e r Mehrfach - Waageneinrichtung benutzt werden, kann ein einziges Messgerät die Ausgangsspannung jedes einzelnen oder aller Geber anzeigen. Genauigkeit innerhalb $\pm 0,25\%$ ist durchaus normal und kann über lange Zeitspannen eingehalten werden. Alle Lasten bis zu 22 680 kg können mit den hergestellten fünf Typen gewogen werden, die ausgestellt, aber nicht vorgeführt wurden.

EE 22 761 *für weitere Einzelheiten*

FERRANTI LTD

Hollinwood, Lancashire

Koordinaten-Prüfmaschine

(Abbildung Seite 378)

Mit Hilfe dieser Koordinaten-Prüfmaschine, die Dimensionen von Gegenständen mit einer Genauigkeit von 0,0254 mm in der Hälfte der mit Messuhren erforderlichen Zeit messen kann, werden Messungen in Prüfabteilungen erheblich beschleunigt.

Ein Messfühler mit auswechselbaren Spitzen geeigneter Form kann jede Stelle eines bearbeiteten Teiles berühren. Er wird auf einem über den Arbeitstisch herausragenden starren Träger geführt. Das zu prüfende Werkstück ist auf dem Tisch aufgespannt.

Die x- und y-Achsen der Maschine sind mit elektro-optischen Abtastelementen ausgerüstet. Durch zugeordnete Beugungsgitter ruft Bewegung entlang jeder Achse Moiré-Fransenmuster hervor, die Fotozellen überstreichen, in denen Lichtstärkewechsel in elektrische Grössen umgewandelt werden. Die periodischen Schwankungen werden in Transistor-Dekadenzählern in beiden Richtungen registriert (ein Zähler pro Achse).

Die Messwerte werden kontinuierlich in zwei fünfstelligen Zählerreihen angezeigt. Die Position des Fühlers kann für jeden Punkt des x- und y-Durchgangs mit einer Genauigkeit von 0,001 Zoll (0,0254 mm) direkt abgelesen werden.

EE 22 762 *für weitere Einzelheiten*

THE GENERAL ELECTRIC CO. LTD

Magnet House, Kingsway, London W.C.2

Mikrometer Mit Dezimalablesung

(Abbildung Seite 379)

Die Vorführung zeigte, wie Analog-Digitalumsetzer zur Erzielung höherer Auflösung verzahnt werden können. Je einer der vier binärverschlüsselten Dezimal-Digitalumsetzer Grösse 11 stellt eine Dezimalziffer in einer numerischen Darstellung der Mikrometereinstellung in zehntausendstel Zoll (0,00254 mm) dar. Alle Schaltungen sind transistorisiert.

Die Verzahnung zwischen den Umsetzern ist nicht kritisch, da diese zusätzliche Wicklungen haben, die mittels einer Hilfsschaltung Mehrdeutigkeit verhindern. Dazu wird die geltende Ziffer des feineren jedes Umsetzerpaares benutzt, um die Erregung des gröberen Umsetzers von einer Wicklung zur anderen zu schalten.

Die Umsetzer werden durch Rechteckwellenformen erregt, die in einem Transistor-Multivibrator erzeugt werden. Die hervorgerufenen Ausgangsspannungen der Umsetzer werden verstärkt, die Phase festgestellt, in bistabilen Schaltungen gespeichert und dann mittels Relais entschlüsselt, die die numerische Anzeige treiben.

EE 22 763 *für weitere Einzelheiten*

Vibrator-Antriebsgerät

Die Konstruktion dieses Gerätes, das alle für den Betrieb eines VEM 5,2 oder ähnlichen Vibrators erforderlichen Funktionen erfüllt, ermöglicht Bedienung durch Fabrikarbeiter.

Der Vibrator wird von einem verzerrungsfreien Verstärker mit einer Nennleistung von 2,5 kW im Frequenzbereich

30 Hz ... 10 kHz betrieben. Die Beschleunigung kann mittels eines schnellansprechenden g-Reglers, der von jedem der drei Beschleunigungsmesser gesteuert werden kann, in sehr engen Grenzen kontrolliert werden. Sie wird unabhängig auf einem Messgerät mit logarithmischer Skala von 2 ... 100 g angezeigt.

Sowohl eine Fremdsignalquelle als auch eine der beiden eingebauten Signalquellen kann benutzt werden. Der eingebaute, motorgetriebene Kipposzillator wird vollkommen von der Frontplatte aus geregelt und überstreicht alle vorgewählten Grenzen im 30 Hz ... 10 kHz Bereich mit Ablenkgeschwindigkeiten von 4 ... 140 s/Oktave. Mittels eines handbedienten Reglers kann jede gewünschte Einzelfrequenz eingestellt und danach unbegrenzt festgehalten werden.

Das eingebaute doppelspurige Magnettonband hat eine automatische Umkehrvorrichtung und eine Wiedergabedauer von 30 Minuten, die mit einer Umkehrzeit von unter fünf Sekunden unbegrenzt wiederholt werden kann.

EE 22 764 *für weitere Einzelheiten*

GENERAL PRECISION SYSTEMS LTD

Aylesbury, Buckinghamshire

ANALOGRECHNER-BAUSTEINE

Aus dem Fertigungsprogramm der Gleichstrom-Bausteine für Analogrechner wurden gezeigt:

Funktionsgeber

Dieser diodengeschaltete elektronische Funktionsgeber ist für Eingangsgleichspannungen bis zu ±100 V konstruiert. Die Ausgangsgleichspannung kann auch bis zu ±100 V gehen und für jede gewünschte Funktion der Eingangsspannung eingestellt werden. Die Funktion ist näherungsweise eine Reihe gerader Linien. Das wird durch Unterteilung des Bereiches in 10 Abschnitte erreicht, die duch 9, für jede Grösse der Eingangsspannung einstellbare „Knickpunkte" getrennt sind. Jeder Abschnitt folgt einer geradlinigen Funktion, deren Neigung von 0 bis 10 : 1 variabel ist. Die Einstellelemente sind nahezu unabhängig voneinander, und die einmalige Einstellung jedes Elementes ohne Nachregelung gibt eine Gesamtgenauigkeit von ±2%. Die Bandbreite ist 2 kHz. Das Gerät ist komplett mit eingebautem Verstärker.

Funktionen können auch mittels angezapfter Widerstandskarten durch Servo-Multiplikatoren erzeugt werden.

Servo-Multiplikatoren

Dieser langsame Servo-Multiplikator besteht aus einem Velodyne-Motorgenerator, der drei Potentiometer treibt. Eins der Potentiometer wird für Motorenregelung benutzt und kann für Integration abgeschaltet werden. Ein zweigängiges Getriebe ist Standardausrüstung, andere Getriebe können jedoch angebaut werden, so dass ein Integrationsbereich von ½ Periode pro Sekunde bis 1 Periode pro 8 Stunden möglich ist.

Die anderen zwei Potentiometer haben 9 Abgriffe, so dass mit geeigneter ohmischer Belastung die verschiedensten Funktionen näherungsweise eingestellt werden können. Zwei weitere Potentiometer können angebaut werden.

Ein schneller Servo-Multiplikator, dessen Drehpotentiometer durch einen MES-Servomotor getrieben werden, ist auch lieferbar. Eine max. Geschwindigkeit von 10 Hz wird erreicht, und drei oder sechs Potentiometer können anmontiert werden. In jedem Fall dient ein Potentiometer der Motorregelung, eins hat 9 Abgriffe, und der Rest hat Mittelabgriffe.

EE 22 765 *für weitere Einzelheiten*

LIVINGSTON LABORATORIES LTD

Retcar Street, London N.19

SAMPLING-OSZILLOGRAPH

(Abbildung Seite 379)

Der Hewlett-Packard Oszillograph 185A bildet zusammen mit dem Doppelspurenverstärker 187 ein System zur Sichtbarmachung schneller Impulsvorgänge, die eine Bandbreite bis zu 500 MHz voraussetzen. Aufeinanderfolgende „Teilbilder" (samples) des untersuchten Impulses werden bei jeder Wiederkehr etwas später abgetastet. Die Teilbilder werden dann auf einer langen Kippschwingung rekonstruiert, wobei das System eine effektive Anstiegszeit von 0,7 μs hat.

Die Anzahl der Teilbilder pro Spur kann auf der Frontplatte eingestellt werden. Durch umfangreiche Trigger- und Ablenkdehnungsmöglichkeiten können Anstiegs- und Rückflanken der Impulse eingehend untersucht werden.

Durch gleichzeitige Darstellung des Zeiteicherausgangs unter Benutzung des zweiten Kanals können genaue Messungen von Anstiegszeit, Dauer und Intervall ausgeführt werden. Auch für Betrieb eines X-Y-Schreibers ist eine Ausgangsspannung vorgesehen, so dass Daueraufzeichnungen schneller Vorgänge ausgeführt werden können.

Die Empfindlichkeit jedes Kanals ist unabhängig mit geeichten Bereichen von 10 mV/cm...200 mV/cm und ±5% Genauigkeit; die Eingangsimpedanz ist ungefähr 100 kΩ parallel zu 2 pF.

EE 22 766 *für weitere Einzelheiten*

OSZILLOGRAPHKAMERA

Mit der Hewlett-Packard Oszillographkamera 196A mit „Polaroid-Land"-Rückseite können positive fotografische Aufzeichnungen einer Oszillographenspur in wenigen Minuten hergestellt werden.

Die Spur kann während der Aufnahme durch einen Tubus beobachtet werden; die f/1,9 Speziallinse gibt die volle 10 cm-Rasterbreite ohne merkliche Herabsetzung der Schirmauflösung auf dem Bild wieder. Das Grössenverhältnis zwischen Schirmbild und Aufnahme ist 1 : 0,9.

Verwendung normaler Kamerabalgen vermeidet Lichteinfallprobleme; für Mehrfachbelichtung kann die Linse durch 11 gerastete Positionen gedreht werden.

Die Kamera wurde in erster Linie für HP-Oszillographen entwickelt. Für Einsatz mit Tektronix-Oszillographen ist jedoch ein Zwischenstück lieferbar.

EE 22 767 *für weitere Einzelheiten*

MARCONI INSTRUMENTS LTD

St. Albans, Hertfordshire

RATEMETER

(Abbildung Seite 379)

Das Gerät (Type TF1292) gibt eine Messgerätanzeige für Zählraten oder integrierte Zählung von Kernteilchen und Strahlendetektoren. Die Zählratenbereiche sind bei Vollausschlag 30, 100, 300 und 1000 pro Sekunde. Für diese Bereiche kann eine Zeitkonstante von 1,5 oder 20 Sekunden eingestellt werden. Bereiche für integrierte Zählung zeigen bei Vollausschlag 1000, 3000, 10 000 und 30 000 Impulse an.

Eine der beiden einstellbaren Hochspannungen ist für einen Kernteilchen- oder Strahlenzähler bestimmt, die andere für eine Niederschlagskammer. Ausserdem sind Anoden- und Heizspannung für zusätzliche Vorverstärker vorgesehen.

Das Gerät ist für viele Strahlenmesszwecke in industriellen Verfahren und Laboratorien geeignet. Es kommt besonders für Überwachung der Kühlmittelgase in Kernkraftwerken infrage und kann daher fehlerhafte Brennstoffelemente entdecken. In diesem Fall werden die Ausgangsimpulse eines Szintillationszählers gemessen, der auf die radioaktiven Teilchen im Gas reagiert; die ionisierten Teilchen schlagen sich auf einem Draht nieder, der vom Ratemeter mit einer anziehenden Hochspannung gespeist wird.

Fernsteuerung und -anzeige sowie Alarmschaltungen für eine vorgewählte Messgerätanzeige oder Stromversorgungsausfall sind vorgesehen. Ausser der Instrumentanzeige der Zählrate oder des Zählstosses auf der Frontplatte sind zwei proportionale Ausgangsspannungen von 50 V und 100 mV (Vollausschlag) für andere Geräte vorgesehen. Die niedrigere Spannung ist besonders für Schreiber gedacht.

EE 22 768 *für weitere Einzelheiten*

ANALOG-DIGITALUMSETZER

(Abbildung Seite 380)

Der volltransistorisierte Analog-Digitalumsetzer TF 1325 wird benutzt, um Anzeige-, Registrier- und Verarbeitungsgeräten die Werte angelegter Gleichspannungen von 0...0,999 V zuzuleiten. Der binärverschlüsselte (+10/0 V) dreistellige Dezimalausgang hat eine Genauigkeit von ±1 mV. Die Abtastperiode ist 20 ms und kann ferngesteuert werden. Ein Stufensignal zeigt an, wann das Abtasten beendet ist und ein Resultat zur Verfügung steht. Der Eingangsstrom ist dann unter 0,15 µA.

Der Umsetzer ist hauptsächlich dazu bestimmt, eine Eingangsgleichspannung aufzunehmen, den Pegel abzutasten und das Resultat schnell und genau in Digitalform zur Verarbeitung in Entschlüssler, Zeilendrucker oder elektrischer Schreibmaschine weiterzugeben, was eine Aufzeichnung für späteren Nachweis sicherstellt. Auf diese Weise findet er besonders für wiederholte Überwachung des Strahlungsniveaus an verschiedenen Stellen eines Kernkraftwerkes Verwendung. An den einzelnen Stellen auftretende Spannungen werden der Reihe nach in den Umsetzer gespeist und brauchen nur während der Abtastperiode gleichmässig zu sein.

Wegen seiner grossen Schnelligkeit eignet sich der Umsetzer auch als Translationswandler einer Analogspannung in die für Datenverarbeitungsanlagen erforderliche Digitalform.

Normalerweise wird der Umsetzer in einem Gehäuse für Messplatzarbeiten geliefert; er kann aber auch für Einbau in ein Gestell bestellt werden.

EE 22 769 *für weitere Einzelheiten*

MULLARD LTD

Mullard House, Torrington Place, London W.C.1

LEGIERUNGS-DIFFUSIONS-TRANSISTOREN

Die Legierungs - Diffusions - Technik wurde schon bei der Massenproduktion der Transistoren OC170 und OC171 benutzt, die weitgehend für ZF- und HF-Verstärkung in KW- und UKW-Empfängern Verwendung finden. Unter den kürzlich für Nachrichtenverkehr, Rechner und Industriezwecke entwickelten Elementen befinden sich:

Transistoren für den UKW-Nachrichtenverkehr

Der UKW-Transistor AFZ11 wurde besonders als geräuscharmer Transistor für 100 MHz-Verstärkung in Nachrichtenanlagen entwickelt. Durch die besondere Fertigungstechnik wurden Basisstörwiderstand und Rückkopplungskapazität niedrig gehalten; bei 100 MHz gibt der Transistor eine höhere Leistungsverstärkung als 10 dB. Für dieselbe Frequenz ist der Rauschfaktor nicht höher als 10 dB. Die mittlere Stromverstärkung ist 50.

Für den vollen Nennstrom von 10 mA ist die Kollektorspitzenspannung −20 V und die Emitterspitzen −12 V. Bei 25°C und einer Kollektorspannung von −6 V ist der durchschnittliche Kollektorreststrom 4 µA. Die max. Kristalltemperatur ist 75°C und der Wärmewiderstand 0,6°C/mW. Das erlaubt eine Verlustleistung von 50 mW bei 45°C Umgebungstemperatur. Der Transistor wird in einem Metallgehäuse geliefert.

Dem AFZ11 wird bald ein weiterer Transistor für dieselben Anlagen folgen, der bis zur doppelten Frequenz eingesetzt werden kann. In den 200 MHz-Stufen der Nachrichtenverkehrsanlagen wird der Transistor bei einem Rauschfaktor unter 9 dB eine Leistungsverstärkung über 10 dB haben.

Sehr schnelle Rechenlogik

Ein Legierungs-Diffusions-Transistor mit guten Kenndaten für Grossignale wird in Kürze für logische Schaltungen sehr schneller Rechner zur Verfügung stehen.

Der Transistor ist ein äusserst hochentwickeltes Element, das den Ansprüchen zukünftiger Rechner mit Taktfrequenzen bis zu 10 MHz gerecht werden wird. Ausserdem wird er eine Grenzfrequenz von mehreren hundert MHz, ausgezeichnete Kniespannungskenndaten für 50 mA Kollektorstrom und eine für seine Verwendungszwecke ausreichende Verlustleistung haben.

Der Transistor wurde in einer logischen Schaltung mit niedrigem Spannungspegel in einem 10 MHz Dekadenzähler vorgeführt.

Schneller Hochstrom-Kerntreiber

Dieser Transistor wird als Treiber für schnelle Rechner-Magnet-Kernspeicher im Zusammenhang mit dem oben erwähnten schnellen Logik-Transistor entwickelt.

Er kann Ausgangsimpulse von mehreren hundert mA mit äusserst kurzen Anstiegszeiten abgeben. Der Kollektorspitzenstrom ist ½ A und der f_1 Parameterwert höher als 60 MHz.

Das Element wurde in einem schnellen Hochstrom-Impulsgenerator für 0,5 A Ausgangsimpulse mit mittlerer Anstiegszeit unter 40 µs vorgeführt.

Auf Grund seiner HF-Leistungskenndaten ist der Transistor auch für Trägerfrequenzfernsprecher und -sender geeignet, in denen er 3 W bei 8 MHz und bis zu 1 W bei höheren Frequenzen abgibt. Er wurde in Treiber- und Endstufen eines 2,5 kW-Senders bei 8 MHz vorgeführt.

PNPN-Schalttransistor

Für den ATZ10 wurde die Legierungs-Diffusions-Technik zur Herstellung eines 4-schichtigen pnpn-Schalttransistors weiterentwickelt.

Seine Eigenschaften ermöglichen die Entwicklung von Flip-flop, nichtstabilen und monostabilen Schaltungen, in denen statt der normalerweise benötigten zwei herkömmlichen 3-schichtigen Elemente ein einziger pnpn-Transistor verwandt wird und dadurch Platz und zugehörige Bauelemente gespart werden. Ausserdem ist er für geschaltete Gesprächswege in automatischen Fernsprechanlagen geeignet.

Der ATZ10 ist ein Transistor mit drei Anschlüssen, dessen Wirkungsweise einer Kombination eines pnp- und eines npn-Transistors mit interner Regeneration ähnlich ist. Diese Konstruktion gibt ausgesprochen negative Impedanzkennlinien über einen breiten Netzstrombereich und macht den Transistor für Verwendung als schnellen elektronischen Schalter im Impulsbetrieb geeignet: „ein"-geschaltet bleibt er automatisch unter dem Knie, bis durch einen Impuls „aus"-geschaltet wird.

Für den ATZ10 ist ein höheres „aus-ein" Impedanzverhältnis als 3 Millionen zu 1 kennzeichnend. Spitzenspannung und -strom sind ungefähr −35 V bzw. 25 mA. Die max. Kristalltemperatur ist 55°C und erlaubt eine max. Gesamtverlustleistung von 17 mW in einer Umgebungstemperatur von 45°C.

Auf der Ausstellung wurde ein mit zehn ATZ10 bestückter Dekaden-Ringzähler mit einer Impulsfrequenz von 30 kHz gezeigt.

Lawinen-Transistor mit 1 ns Anstiegzeit

Dieser Transistor arbeitet nicht wie die üblichen Transistoren und benutzt den „Lawinen"-Vorgang. Er stellt eine Sonderentwicklung der Legierungs-Diffusions-Technik dar und ist überall dort nützlich, wo ungewöhnlich kurze Anstiegszeiten erforderlich sind, wie z.B. in Sampling-Oszillographen.

Ein Spitzenstrom von 50 mA kann nach einer Anstiegszeit von nur 1 ns erreicht werden. Das entspricht einer Stromanstiegsrate von 50 A/µs.

Auf der Ausstellung wurden in einer Lawinen-Schaltung 50 mA-Impulse mit einer Anstiegszeit von ungefähr 1 ns bei einer Impulsfrequenz von 4 MHz erzeugt und mittels eines Sampling-Oszillographen vorgeführt, der selbst mit einem Lawinen-Transistor in Abtastschaltung bestückt ist.

Schalt-Transistor

Der bereits lieferbare ASZ20 Transistor ist für Schaltzwecke einschliesslich logischer Rechnerschaltungen bestimmt.

Der Spitzenstrom ist 25 mA und die Kollektorspitzenspannung −40 V; f_1 ist höher als 40 MHz, und bei 25 mA Kollektorstrom ist die min. Grossignal-Verstärkung 33. Die max. Kristalltemperatur ist 75°C und der Wärmewiderstand 0,6°C/mW. Das ergibt eine Verlustleistung von 50 mW in 45°C Umgebungstemperatur. Für −6 V ist der Kollektorstrom bei 25°C unter 5 µA.

Die Demonstration zeigte den ASZ20 in einer stromschaltenden Logik, in der die Verzögerung pro Stufe nur 25 ns ist.

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Zusammenfassung der wichtigsten Beiträge

Ein transistorisiertes Miniaturecholot von R. N. B. Gatehouse

Zusammenfassung des Beitrages auf Seite 336-340

Der Beitrag beschreibt Entwicklung und Hauptkonstruktionsmerkmale eines sehr kleinen und robusten Echolots, das in 3,5 . . . 58,5 m Tiefe benutzt werden kann. Das Gerät ist für Hydrografie, Vermessungen zum Verlegen von Unterwasserrohrleitungen, Navigation von Jachten und kleinen Fischereischiffen und ähnliche Zwecke geeignet. Es wiegt komplett 2,5 kg, verbraucht 0,1 W und hat eine dem jeweils grösseren Wert entsprechende Genauigkeit von ±3% oder 15 24 cm.

Eine passive wellenformende Schaltung von M. J. Wright

Zusammenfassung des Beitrages auf Seite 341-343

Eine einfache Schaltung, die verschiedene wellenformende Funktionen ausführen kann, wird beschrieben. Trotzdem keine aktiven Bauelemente Verwendung finden, können scharfe Ubergangssprünge der Ausgangsspannung erzielt werden, die in Bezug auf die Eingangsspannung verzögert sind. Zu den Verwendungs möglichkeiten gehören: 1) Impulsverlängerung, 2) 90° Phasenänderung einer Rechteckwelle, 3) kombiniertes Ausfiltern und Begrenzen, 4) Impulslängenmodulation.

Eine einfache mehrdimensionale Sichtanzeige von D. M. MacKay

Zusammenfassung des Beitrages auf Seite 344-347

Funktionen von zwei oder mehr Variablen können auf dem Schirm einer Elektronenstrahlröhre als zweidimensionale Projektionen der Oberfläche, die sie bestimmen, dargestellt werden. Der Projektionswinkel kann durch Rotieren von Potentiometern geändert werden, die bestimmen, in welchem Verhältnis die verschiedenen Eingangsspannungen zur X- und Y-Ablenkung beitragen.

Am besten wird das mittels Sinus-Kosinus Potentiometern mit vier Schleifkontakten und Anpassungsverstärkern getan, jedoch werden überraschend zufriedenstellende Resultate mit einfacheren Schaltungen erreicht, die nur lineare Tandempotentiometer und einige sorgfältig ausgesuchte Widerstände benötigen. Die Ausgangssoannungen für eine vollstereoskopische Darstellung können mit geringem zusätzlichen Schaltungsaufwand erreicht werden.

Die Nachbildung von Ubergangsfunktionen auf einem Analogrechner von F. C. Harbert

Zusammenfassung des Beitrages auf Seite 354-355

Der Beitrag beschreibt eine Schaltung mit einem Verstärker, in der Doppelintegration vorgenommen wird, und eine Methode zur künstlichen Darstellung vielgliedriger Ubergangsfunktionen mittels eines Analogrechners.

Die Durchbildung optischer Digitalmessgeräte von I. R. Young

Zusammenfassung des Beitrages auf Seite 359-365

Zuerst werden Konstruktionserwägungen für optische Messgeräte besprochen. Es wird betont, wie wichtig es ist, dass die Ablesung sofort vorhanden ist, wenn sie verlangt wird, da dies für Abtastsysteme (time sharing) wünschenswert ist. Einige Instrumente niedriger Auflösung werden beschrieben und die Verwendung der von ihnen abgelesenen Daten angegeben. Schliesslich werden die mit Durchkonstruktion von Instrumenten hoher Auflösung zusammenhängenden Probleme kurz erwähnt.

Der Einfluss der Giessharzumüllung auf elektronische Bauelemente von F. P. Campbell

Zusammenfassung des Beitrages auf Seite 366-371

Es ist schon lange bekannt, dass das Vergiessen elektronischer Schaltungen die Leistung beeinflusst. Die Wirkung des Vergiessens auf üblicherweise benutzte Bauelemente wird beschrieben.

Röhren können ohne Bedenken vergessen werden, solange zur Ableitung eigenentwickelter Wärme und zum Schutz des Glaskolbens die richtigen Techniken angewandt werden. Transistoren und Kondensatoren bieten kaum Schwierigkeiten. Der Einfluss des Vergiessens auf Transformatoren und Drosseln wird beschrieben. Drahtwiderstände eignen sich nicht zum Vergiessen, wenn sie eine hohe Verlustleistung haben.

Der Einfluss auf Masse- und Schichtwiderstände ist für die verschiedenen Fabrikate unterschiedlich und ein Versuch wird beschrieben. Die Ursachen für Anderungen in den Bauelementen werden zusammengefasst.

Ein stabilisiertes und programmgesteuertes Netzgerät mit positiver Gegenkopplung von M. Pacáck

Zusammenfassung des Beitrages auf Seite 372-373

Ein stabilisiertes Netzgerät für den Magneten eines Massenspektrometers wird beschrieben. Die Ausgangsspannung ist von 3 . . . 800 V stufenlos regelbar oder kann durch eine RC-Schaltung programmgesteuert werden. Im Verstärker des Stabilisators findet positive Gegenkopplung Anwendung.